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STUDY OF ADVANCED COMMUNICATIONS SATELLITE SYSTEMS BASED ON SS-FDMA

Prepared for:

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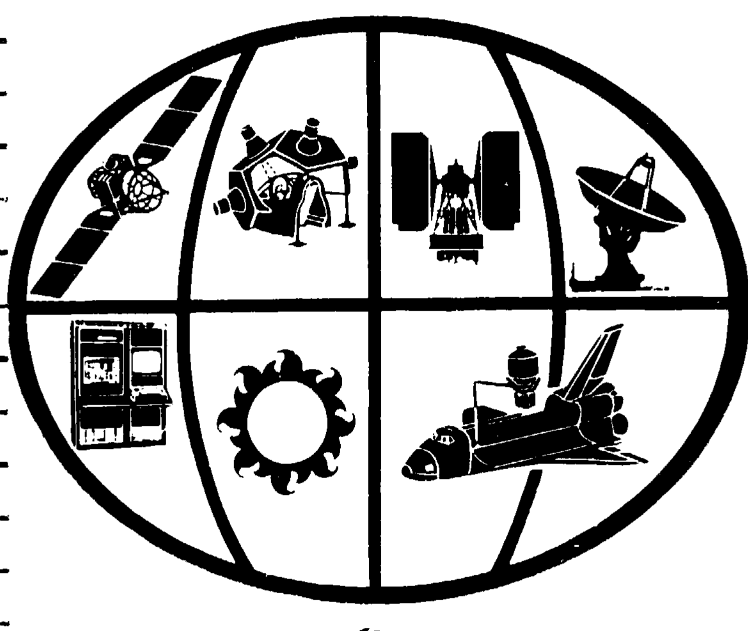

space division

GENERAL ELECTRIC

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STUDY OF ADVANCED COMMUNICATIONS SATELLITE SYSTEMS BASED ON SS-FDMA

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FOREWORD

This contract was performed under the direction of Mr. Dale Pope, Lewis Research Center, Cleveland, Ohio. The author wishes to acknowledge the substantial contributions made by Mr. Peter Foldes (GE) in the spacecraft antenna analyses, Mr. Wilbur Popowsky (GE) in the analysis of satellite switching technology, and Mr. James Rautio (GE) in the analysis of fading effects.

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SECTION 1

INTRODUCTION AND OVERVIEW

SECTION 1

INTRODUCTION AND OVERVIEW

1.1 BACKGROUND

1.1.1 PURPOSE

The purpose of this study is to explore a satellite communication system based on the use of a multiple, contiguous beam satellite antenna and Frequency Division Multiple Access (FDMA). Emphasis is on the evaluation of the feasibility of SS-FDMA technology, particularly the multiple, contiguous beam antenna, the on-board switch and channelization and on methods to overcome the effects of severe Ka-band fading caused by precipitation. This technology is evaluated and plans for technology development and evaluation are given in Section 6. In addition, the application of SS-FDMA to domestic satellite communications (fixed services) also was evaluated. Due to the potentially low cost earth stations SS-FDMA is particularly attractive for thin route applications - up to several hundred kilobits per second - and offers the potential for competing with terrestrial facilities at low data rates and over short routes. The on-board switch also provides added route flexibility for heavy route systems; however, this application was not studied thoroughly because these attributes of FDMA are well known. The key beneficial SS-FDMA strategy is to simplify and thus reduce the cost of the direct access earth station at the expense of increased satellite complexity.

The multiple beam antenna provides complete coverage of the service area such as the continential U.S. and off-shore locations while affording the possibility of frequency reuse through isolation of co-frequency beams by antenna sidelobe control and possibly through the use of orthogonal polarization. Consequently, a satellite system based on such an antenna system can have large capacity because of the frequency reuse but also will have interference because of imperfect isolation. For two way telecommunications the higher antenna resolution or gain also reduces earth station EIRP and G/T requirements which suggests that low cost earth stations may be achievable. A further consequence of the use of a multiple contiguous beam antenna is the need to route the various communication signals received by the satellite to the proper destinations via the associated downlink antenna beam. The use of FDMA, an important subject in this study, requires that these access paths within the satellite be dynamically selected on the basis of the individual signal's position within the available spectrum, e.g., different destinations are provided by different frequency bands. This routing function is provided by filtering and switching on the satellite.

While this concept affords the possibility of both heavy route and thin route communication systems, emphasis in this study is on thin route applications especially those affording direct access to the satellite system by a small, low cost earth station somewhere on the user's premises (e.g., roof top, parking lot, field, etc.). Properly implemented, such a system can be low cost, reliable, flexible with regard to service use, bandwidth, modulation, etc., and can be implemented at any microwave telecommunication frequency band. It can contain its own integral communication switch independent of terrestrial tails, interconnects, long haul facilities or switches. This study then is an inquiry into the characteristics and technology of such a direct access system in order to:

- Define basic system characteristics and utility,
- Explore performance bounds with regard to transmission impairments, effects of precipitation attenuation, cross talk on user channel isolation, and the relationship of these to the system concept,
- Identify technology required for demonstration or operation in terms of feasibility,
- Define potential user costs, and
- Define a program that could evolve a system demonstration.

While sometimes referred to as a "switchboard in the sky" and other terms, the system described herein is really a new adaptation of the basic FDMA system which is the work horse of present day communication satellites. Consequently, the term "satellite switched" frequency division multiple access or SS-FDMA is adopted to describe the technique.

1.1.2 HISTORY

Some of the fundamental limitations in the utilization of space communications, particularly for the U.S., are the available electromagnetic spectrum, the usable segment of the geostationary arc, and the flexibility, reliability, and cost of service to the user. Satellite telephony and data services particularly have experienced constricted growth because of the high cost relative to alternative terrestrial means, caused principally by the congested C-Band which limits satellite power flux density and earth station antenna diameter.

NASA has had extensive prior involvement in the initial development and early evolution of U.S. space communications beginning with Syncom and Project Relay which demonstrated satellite use of active repeaters (and use of synchronous orbit, e.g., Syncom), to the ATS-6/CTS satellite series demonstrating the utility of satellite spot beams and high power. The immediate technical problem left undone was the method of routing signals through the satellite multiple antenna beams, in a way compatible with low user costs. Consequently, the real payoff in low cost satellite communications was never realized. So-called public service experiments on ATS-6 and CTS, reaping the technical benefits of spot beams and high satellite power were frustrated in the transition to operational (e.g., carrier-operated) satellite systems because these satellites were not designed to provide the requisite antennas gain and power, but rather were designed with wide area coverage antenna patterns and low transponder power to favor use of large expensive earth stations designed for trunking or heavy route traffic.

1.1.3 DEFICIENCIES OF EXISTING COMSAT SERVICES

Concentration of traffic and any desired processing or switching is performed by the terrestrial system. The satellite system with only minor exception* is relegated to

* All the satellite carriers also have direct access services via relatively expensive terminals which show no sign of great proliferation.

providing long haul trunk service, emulating radio relays and thus providing only a small portion of the total communications service normally desired by users. Switching, interconnection and interface still are provided only by the terrestrial system.

SBS is making a step forward with a system based on TDMA with a limited random access (e.g., switch) capability; however, access to this system will still be through terrestrial interconnects and tails, and the SBS terminals are relatively expensive.

In addition, Intelsat has made use of FDMA with microwave filtering and limited switching which is only practical for use with heavy route systems and satellites with relatively few beams. Also satellite switched TDMA, (SS-TDMA) also has come into prominence (Advanced Westar), however, low user costs with dedicated terminals are not favorable because of the cost of (1) storage (for bursting), (2) high speed MODEMS (60 mbps to 600 mbps), (3) high speed switching, and precise timing, (4) framing, and (5) earth station high power amplifiers to support the high burst rate. Use of TDMA results in "propagating complexity," power, backup power, safety, monitoring, control, alarms, upkeep are all very much complicated by the presence of the TDMA/ high power technology.

The system studied herein overcomes these problems and results in flexible routing and low user costs. In fact, there appears to be no alternative but to relieve the earth station of the switching and signalling burden, to concentrate these in either the satellite or in a central ground facility. This requires a new approach to earth station design, emphasizing simplicity, modularity, standardization, mass production, standard interfaces, and low EIRP and G/T, and a new approach to satellite design in which complexity is acceptable if flexibility and reliability are preserved and earth station design remains simple.

1.2 SS-FDMA SYSTEM CONCEPT

1.2.1 OPERATIONAL DESCRIPTION

In its simplest form the user earth station has a single communications MODEM, and single-thread uplink and downlink microwave hardware. If the user signal is 32 kbps the MODEM (except for slight changes in rate to accommodate coding), operates at 32 kbps and the earth station EIRP and G/T are sized for this rate - requiring a nominal 2 meter antenna. Since the earth station is working in conjunction with a multiple (contiguous) beam satellite the spectrum for the user antenna beam is divided into RF paths. These paths define particular routes through the satellite to the various downlink beams. In the "n" beam satellite each uplink beam is provided with at lease "n" RF paths to the "n" downlink beams resulting in at least n^2 possible routes through the satellite. These paths are not necessarily equal to bandwidth at any instant and total path bandwidth must be changed to follow the traffic demand.

In its simplest form the user signals the satellite system over a common signalling channel indicating destination, originator, and desired signal bandwidth. The common signalling channel responds by assigning frequencies to the originating and destination earth terminals which are in specific frequency bands corresponding to correct RF paths through the satellite. After call completion the frequencies are available for similar services by other earth stations. The earth station is thus a simple SCPC-DAMA type FDMA terminal.

Signalling emulates that of the terrestrial system and, in fact, can be designed to interface the satellite system into terrestrial facilities of comparable performance. The signalling system is thus centralized, either at a convenient earth terminal or even in the satellite. Called a Network Control Center (NCC), the NCC actually performs the functions of a telecommunication switch, routing calls, billing, rendering operator assistance, busying out, tracking traffic intensity, etc. These functions are accomplished automatically by a computer.

It may be recognized that "path" switching is a great simplification over the conventional telephone switch. The latter must connect each telephone to any other telephone. This is accomplished (for such a vast network) by switching, and tandem switching of unbelievable complexity. This technology is not suitable for satellites in the foreseeable future. Fortunately, it is not necessary to switch on a subscriber by subscriber basis since the satellite access destination problem can be solved by properly routing signals among the "n" beams. The final subscriber connection, e.g., switching in the communications sense, is accomplished by the SCPC/DAMA system by the assignment of compatible transmit and receive frequencies to communication earth stations, using a common signalling channel. In the satellite, it is only necessary to provide RF paths from each of the uplink beams to each of the downlink beams. A variety of communications signals may be contained within a specific path - one or more analog [illegible] gital carriers, in combination. Thus, the satellite is still a "bent pipe." This arrangement allows a maximum flexibility to the user since demodulation, decoding, [illegible] in the satellite is neither necessary or desirable.

The on-board channelization and switching is not simple. First it must be expected that a large disparity in traffic demand per beam will be experienced in any real system. A beam that illuminates the New York City area will certainly carry a large traffic load, and will likely be the first beam to saturate at the end of satellite life. On the other hand, a beam illuminating the northwest, say the Montana and Wyoming area, will have relatively little traffic. If there are many beams, say 20 or more, it is possible, particularly for a specific carrier's system, that some beams will carry no traffic at all. This disparity of traffic amongst the beams can be efficiently accommodated by SS-FDMA. Second, traffic predictions are only approximate and acquisition of new users or accommodating to changing user needs may require reconfiguration of the routes. For example, a route normally carrying no traffic may be required to provide occasional full motion interactive TV for teleconferencing. Such contingencies also can be provided for in the SS-FDMA system. In essence, the routes are made up of combinations of fixed and switched RF paths, of standard bandwidths, say 36 MHz, 10 MHz, 5 MHz, etc. Depending on the traffic load predictions these RF paths are assigned by switches or fixed (preassigned). The former can be fully switched

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The bandwidth of the routes (a strong function of the number of beams "n") dictates the onboard switch technology. For those RF paths which are 10 MHz or less, which occurs for $n > 10$, the switching is done at low RF frequencies, say 10 MHz where complementary metal oxide semiconductor, silicon on sapphire (CMOS-SOS) LSI technology is attractive - many switches can be provided on a chip. LSI technology is critical to the implementation of the SS-FDMA system because the number of switch points can be very large. For example, suppose we wish to provide ten 10 MHz paths from each of 20 uplink beams to each of 20 beam downlink beams. Converting all these incoming paths, 20 x 10 = 200 in all, to a common frequency, say 10 MHz, filtering them to individually identify each route using ceramic or SAW* (Surface Acoustic Wave) filters results in 200 input connections to a "cross-point" switch and 200 output connections. Connecting any one input to any one output requires 40,000 switches. It is apparent that recent developments in LSI are crucial to the implementation of such switch matrices. For 36 MHz paths different technology, perhaps field effect transistors in LSI configurations can be used; in general the 36 MHz paths are substantially fewer in number so that less dense packaging can suffice. All this is certainly new satellite technology and the arrangement of these switches to satisfy the communication system needs yet be compatible with satellite requirements in terms of weight, power, dissipation, reliability, etc., is an important aspect of this study and of future development.

* The channelization need not occur at 15 MHz.

Finally, it should be emphasized that the on-board switch is generally not involved with setting up individual calls. If, for example, the RF paths in the routes are prearranged to satisfy the requirements of the "busy hour" then no on-board switching is involved in setting up individual calls! The call setup, as discussed previously, is accomplished solely by the ground SCPC/DAMA system. The on-board switch is exercised only to change the routing bandwidth to adapt the satellite to a new situation. This indirectly involves the on-board switch in the SCPC/DAMA system since the SCPC/DAMA or signalling/switching computer also will likely control the on-board switch too, via a common signalling channel.

1.2.2 SS-FDMA ADVANTAGES

The advantages of the SS-FDMA system are:

1. Low cost earth station and low user costs
2. Simple signalling and access equipment
3. Flexible routing, on demand
4. Transparent medium - e.g., no demodulation or decoding is required
5. Flexible assignment of data rate or bandwidth, this is particularly advantageous for direct access systems because high performance channels with wide bandwidth can be assigned almost instantaneously - on demand; this is a service which terrestrial systems cannot provide. The bandwidth need not be the same in both directions.
6. Complete coverage of the service area through the use of a multiple contiguous beam antenna
7. Large aggregate traffic can be achieved even though individual user demands are small
8. SS-FDMA is compatible with any of the frequency bands now allocated to communications satellites
9. The SS-FDMA is diffuse. While not obvious at this point, the SS-FDMA system is not as subject to single points of catastrophic failure. Shorting switch failures merely result in pre-assigned routes, open switch failures reduce flexibility but parallel paths are available.
10. High availability despite severe fading is feasible. Again the diffuse nature of SS-FDMA can be used to advantage in overcoming the effects of precipitation attenuation at Ku-Band or Ka-Band. Since each carrier, in general, has a discrete origination and destination at any instant the vast majority of these

carriers are involved with earth stations not being subjected to intensive precipitation such as that found in thunderstorms (which causes this attenuation). Consequently, on a dynamic basis, some of the margin of each of these carriers can be applied to those carriers experiencing attenuation. This is tantamount to giving these carriers more power, at the expense of the others, provided by increased output from the earth station HPA which generates this carrier (controlled by the SCPC/DAMA signalling system). Increased HPA power also can be used to overcome the effects of downlink precipitation attenuation. What results is a high continuity of service, $\sim$0.9999 even at Ka-Band, with only a modest increase in satellite power and earth station complexity. No antenna diversity (e.g., two earth station antennas) is envisioned.

11. Independent private networks can be formed with restricted interconnectively having dedicated modulation and signalling characteristics.

1.2.3 SS-FDMA DISADVANTAGES

The disadvantages of SS-FDMA are:

1. Weight and power of on-board channelization and switching can contribute significantly to the overall satellite weight and power.
2. Weight of the multiple beam antenna can be significant, as can be the impact of the antenna geometry on the spacecraft configuration (of course this is not endemic to SS-FDMA).
3. FDMA operation of the satellite output amplifier requires back off and/or linearization to control intermodulation power. All in all this is not a significant cost item to the user. In fact, variable power amplifiers in one form or another which consume power more or less in proportion to the actual traffic signals passing through them can help to minimize prime power requirements and minimize battery energy for eclipse. Taken in total this can result in a lighter overall transponder amplifier implementation than a transponder amplifier with a single FDM or TDM carrier (and especially in a fading environment).
4. The antennas, switch and channelization require development and demonstration.
5. Spectrum Utilization. The satellite spectrum will never fully emulate the instantaneous traffic demand and, hence, there will be wasted spectrum (a trunking FDM-FM carrier also "wastes" spectrum if it is not fully loaded). Also, path channelization will reduce spectrum efficiency further (this corresponds to wasted time in TDMA systems). The achievable spectrum utilization in SS-FDMA is an important subject for future study.

1.2.4 TECHNOLOGY DEVELOPMENT REQUIREMENTS

Technology development requirements are strongly related to the desired satellite capacity (e.g., number of contiguous antenna beams to cover the service area) and to the operating frequency band. What may be envisioned as a future operational satellite is a 10-30 beam, Ka-Band satellite, providing perhaps as much as 3 to 10 frequency reuses (idealized), and working in conjunction with low cost earth stations having 2 meter dishes. The following enabling/enhancing developments are foreseen.

1.2.4.1 Switch Development

Develop and test wideband (36 MHz) and narrowband ($\approx$ 5 MHz) switch matrices, packaged for space use, with controls. In essence this involves development of LSI chips.

1.2.4.2 Channel Filter

Develop and test the last stages of frequency conversion and channelization with sufficient adjacent channel elements to evaluate adjacent channel effects. It should be noted that items 1.2.4.1 and 1.2.4.2 require a substantial systems effort to develop specifications, the circuit arrangement, and frequency plan so that weight and power are minimized and spurious signals are avoided.

1.2.4.3 Multiple Contiguous Beam Antenna

The study results indicate this function can be achieved with an offset fed parabolic antenna. While high confidence can be placed in the analysis, it seems prudent to build and test a model to confirm performance in the presence of inter feed coupling and expected mechanical misalignment and distortions. Only a portion of the feed network need be active. Worst case beam interference ratios of 25 dB to 30 dB are needed.

1.2.4.4 Satellite High Power Amplifier

Amplifiers required for a fully operational Ka-Band system with a usable bandwidth per beam of 833 MHz, and operating with 2 meter earth station antennas and providing a high availability in a fading environment require a substantial amplifier saturated power. This power cannot be precisely identified because of the myriad system choices available and because of uncertainties in subsystem performance such as earth station receiver noise temperatures. Estimated saturated power of the order of 100 to 400 watts are needed.* Lightweight, efficient, linearized amplifiers at Ka-Band of this power level are not available and need to be developed. In addition, there is a need to demonstrate multi level power operation either with a single tube or several tubes.

1.2.4.5 Satellite Low Noise Receiver

A low noise receiver at 30 GHz is needed to minimize the cost of the earth station HPA. A peltier cooled FET or paramp is needed and these are not presently available.

1.2.4.6 High Power Satellite Technology

The study indicates that several kilowatts of prime power is needed for operational system. However, the diurnal variations in dissipation can be substantial if variable power satellite amplifiers are employed. Consequently, an active thermal system using louvers and heat pipes may be necessary. This may involve new technology or at least test and demonstration.

1.2.4.7 Low Cost Earth Stations

The projected costs can only be achieved with standardized configurations with standardized interfaces making maximum use of LSI and MIC (microwave integrated

* Transponder powers substantially less than that would be needed for a demonstration system involving thousands of earth stations.

circuits) and by production - say thousands of terminals - in a lot. The problem is complicated by the present state-of-the-art at Ka-Band where devices of adequate performances may not yet exist. The centralized signalling and switching concept using SCPC/DAMA and a common signalling channel also must be standardized. These problems suggest that in operational systems the carriers are likely to provide these facilities since individual users could not be expected to use sufficient numbers of terminals to achieve the low cost or to understand the complex system interfaces. These suggest a radically different approach to earth station production than is now the case where these are built one or several at a time by either a carrier or a user. It should be noted that standardization and productization does not necessarily mean lack of flexibility. Several standard transmission rates and single and multiple services are envisioned and can be accommodated with the standardized concept. Since direct user systems use narrow band transmission the satellite and earth station oscillators must have low phase noise and good frequency stability. This requirement can involve oven-stabilized oscillators and active tracking loops, which can be significant cost items and is a subject meriting considerable study.

1.2.4.8 Signalling and Switching (SCPC/DAMA)

A direct access system must provide its own signalling and switching independent of terrestrial facilities. It is believed this independence is a particularly attractive feature of direct access systems. The SS-FDMA requirements for signalling and switching are increased over previous systems because of the need to detect and report fading, control the earth station HPA and control the satellite switching. The provision of signals of varying bandwidths, on demand also imposes additional requirements. Demonstrations of these capabilities are essential to proving the flexibility and utility of SS-FDMA for a great variety of different services.

1.2.4.9 Services

There appears to be a minimum constraint on services provided by SS-FDMA. Since SS-FDMA is basically a direct user system, the traffic is composed mostly of narrow band signals, say 32 kbps to 56 kbps, with some smaller amounts of high data rate signals. With DAMA all of the known telephone, data and video services can be accommodated (subject to the effects of satellite system delay), if they are attractive economically to the user. These are:

- Private line voice, data
- Computer interaction
- Point of sale (POS) transactions
- Electronic fund transfer
- Teleconferencing, wideband (video)
- Teleconferencing, narrow band (facsimile, etc.)
- Telemail, etc.

Users are government, business, and public service, the latter consisting of medical and educational services of various types and possibly some forms of emergency services involving portable (not mobile) terminals. There may also be a need to transmit bulk (heavy route) traffic already concentrated because of economy or history. The SS-FDMA system is capable of efficiently providing these services. In fact, it is quite possible to conceive of a satellite system consisting of SS-FDMA and some form of trunking such as SS-TDMA to provide a mix of direct access and trunking services. There appears to be no technical restrictions in implementing a variety of services and systems.

These services or new types of services capable of implementation by SS-FDMA stem from elimination of terrestrial facilities and their restrictions. Impulsive noise, caused by power line switching, normally present on terrestrial lines is avoided, as are the bandwidth and performance restrictions of the terrestrial lines, switches, etc., designed to optimize telephone communication. Literally dial-a-bandwidth services appear feasible - any bandwidth, anytime, anywhere - and this is truly a new dimension in communications.

1.3 SUMMARY OF RESULTS

This summary of the study results are based on evaluations of the feasibility of the multiple contiguous beam antenna, on-board LSI switching, narrow band channelization (such technology evaluations formed the largest part of the study effort) in the context of a high capacity operational Ka-Band satellite having either 10 or 100 beams covering the 48 states. However, each major problem area has been investigated sufficiently so that the technical problems are well understood. While much technology development, test, and demonstration remains to be done, there are no discernible breakthroughs needed to realize the performances predicted herein for SS-FDMA.

1.3.1 ANTENNA

Computer performance predictions for multiple contiguous beam antennas indicate cochannel sidelobe interference levels in the range of 25 to 30 dB. While this interference must be taken into account in system operation, these levels are generally acceptable. A number of attractive beam topologies (arrangements and combinations of beams) are identified. All antenna configurations are based on offset feed parabolic reflectors with multi-horn feeds which are recommended for this application.

1.3.2 SWITCH

Requirements for the switch vary according to the number of antenna beams and the system capacity. There appears to be a need for several different path bandwidths, namely 36 MHz and something of the order of 5 - 10 MHz, the latter enables "trimming" the route bandwidth in order to conserve spectrum. For low frequency switching CMOS-SOS technology enables fabrication of crosspoint switch arrays of minimum weight and power. Smaller numbers of wideband (36 MHz) switches can be fabricated using more conventional PIN diodes or FETS because weight and power are less critical. Both switch types need to be demonstrated. Frequency conversion and frequency synthesis are also challenging engineering problems and should not be neglected.

1.3.3 CHANNELIZATION

Surface Acoustic Wave (SAW) filters are the preferred solution for both wideband and narrow band paths. These filters are lightweight, and can provide the performance needed to channelize the RF spectrum with high spectrum use efficiency. Other filters such as quartz or ceramic bulk filters also may be used but are heavier.

1.3.4 SIGNALLING

A conventional SCPC-DAMA signalling system is envisioned, with added capability to control the earth station HPA power level and to control the satellite switch matrices. The latter requirements do not add significantly to the processing or signalling equipments. A dedicated MODEM is likely to be required for signalling so that signalling functions (such as fade compensation) can be provided even when the communications MODEM is in use. Signalling can use the same HPA and LNR as the communications channel. Signalling paths are only between each earth station and Network Control (e.g., the System Routing Center) so many such paths (as needed) can be provided in parallel.

1.3.5 RAIN COMPENSATION

In SS-FDMA statistical advantage is taken of the many individual carriers to alleviate margins on a dynamic basis. These carriers undergoing either uplink or downlink fading (due to precipitation attenuation), are given more power. This is achieved by dynamically increasing earth station HPA power. Downlink fading at 20 GHz can be compensated such that availabilities of 0.9999 can be achieved at most U.S. locations except those in the vicinity of the Gulf of Mexico. Uplink fading at 30 GHz can also be compensated to achieve similar availabilities; however, a tube (TWT) type amplifier may be needed depending on the earth station antenna diameter and the number of satellite antenna beams. With a 2 meter earth station antenna and a 10 beam satellite, a 30 GHz HPA power in the range of 5 - 10 watts is needed for 32 kbps.

1.3.6 USER COSTS

Using previous studies to predict earth station and satellite costs, user end-to-end annual costs were computed for various situations. Figure 1-1 depicts such costs vs earth station antenna diameter for a postulated range of single MODEM earth station costs from $10 K to $30 K, with and without DAMA. The scales at the right of Figure 1-1 represent existing 2.4 kbps, 9.6 kbps and 32 kbps tariffs. The end-to-end user costs using SS-FDMA are very competitive with terrestrial wideband facilities even at short route miles. SS-FDMA appears competitive even at 9.6 kbps rates provided earth station costs are between $10 - 20 K. Figure 1-2 shows similar results for an SS-FDMA earth station with ten 32 kbps MODEMS and two 64 kbps MODEMS, (448 kbps total). In this case the SS-FDMA system is competitive at rates as low as 2.4 kbps at very short route miles. Actually, earth station costs could be significantly higher without affecting overall competitiveness.

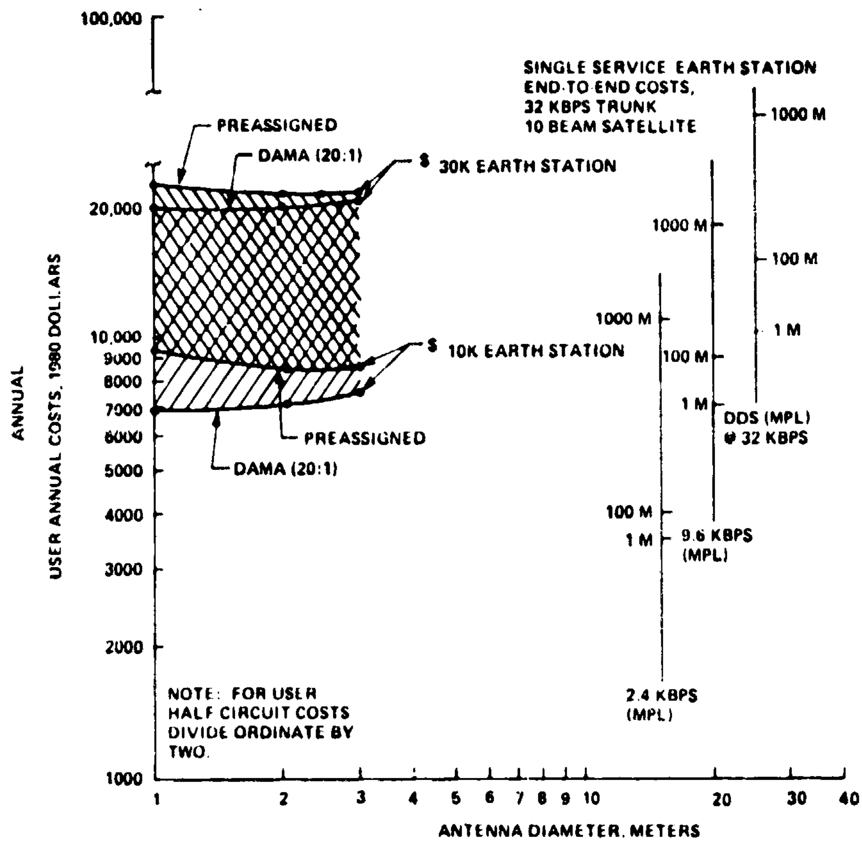

Figure 1-1. Single MODEM Earth Station

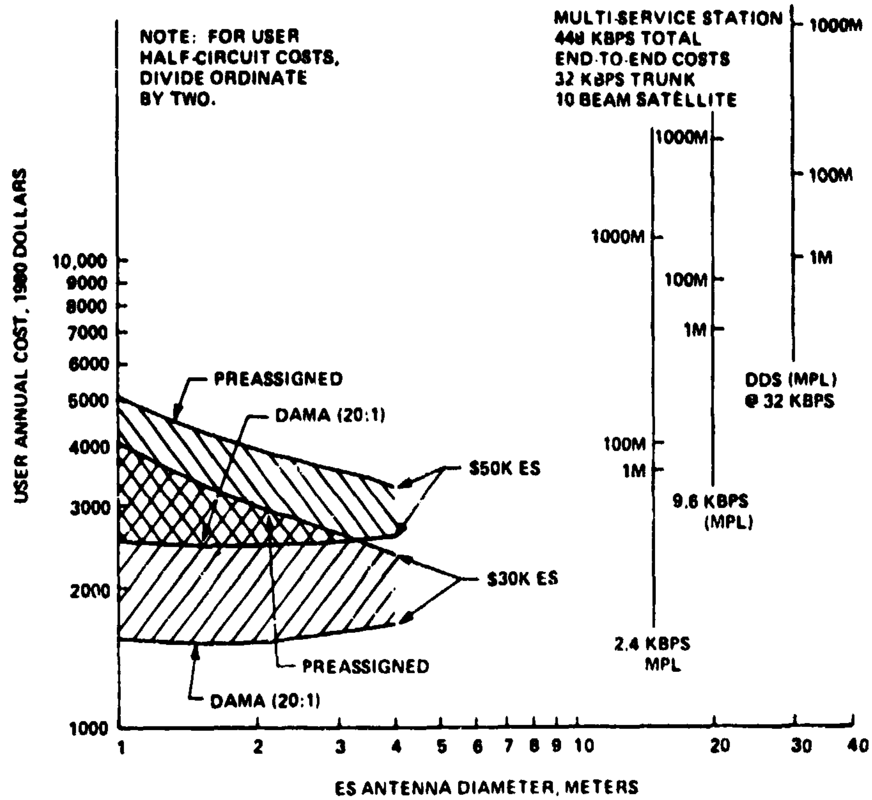

Figure 1-2. Multi-MODEM Station

1.3.7 TRUNKING SYSTEMS

While SS-FDMA can also be used for wideband trunking this case was not analyzed specifically in order to emphasize the direct to user application. SS-FDMA can improve the connectivity and flexibility of trunking FDMA systems; however, an examination of present tariff structures for interconnects or "tails" revealed the latter are dominant costs. Consequently, it may not be important in terms of user costs whether FDMA or TDMA is used (the former generally increases satellite costs while the latter generally increases earth station costs).

1.4 ORGANIZATION OF THE REPORT

Section 2 contains an initial tutorial description of the SS-FDMA system concept based on multiple contiguous antenna beams. Both satellite and earth station overall characteristics are described as well as an initial operating scenario. This is intended as an introduction to SS-FDMA systems and does not represent the results of a detailed system study. In fact, the report emphasizes the onboard switch and multiple contiguous beam antenna technology. However, some system optimizations were performed to illustrate the potential advantages and benefits of the SS-FDMA. A major section compares terrestrial systems, trunking satellite systems and direct access satellite systems under a variety of circumstances and draws important conclusions regarding potential benefits derived from the application of advanced technology. Section 2 concludes with a description of a method to overcome the effects of Ka-Band precipitation attenuation fading so that high availability, approaching 0.9999 can be achieved with single antenna earth stations with minimal acceptable system and satellite complexity and minimum cost.

Section 3 describes the technology and performance of the contiguous multiple beam antenna starting with beam topologies which offer different values of frequency reuse and antenna self-interference. Practical antenna technology is described and examples of mechanical configurations given. Included in this section is an earth station antenna design capable of reducing adjacent satellite interference such that 1^{o} satellite spacing can be envisioned. An example of a mechanical configuration also is described.

Section 4 describes the on-board filtering and switching technology including the assumed system performance requirements. Limitations of LSI switch implementation are described and physical features given. An example packaging design is presented. Channel filtering based on Surface Acoustic Wave (SAW) filters and ceramic piezoelectric filters also is described. Methods for controlling the switch, performing frequency conversion and frequency synthesis also are described.

Section 5 describes the operational characteristics of a future, hypothetical Ka-Band SS-FDMA system. Satellite concepts for 10 and 100 beam configurations are summarized and costed. Typical user terminals also are identified and costed, and total user costs in terms of 1980 dollars are computed based on various operational scenarios.

These are compared with existing terrestrial tariffs in order to identify future attractive satellite services and applications and to further identify salient system features which can provide guidance to the technology development. Section 5 concludes with a summary of SS-FDMA system characteristics regarding service flexibility, user cost, technical feasibility and operational advantages.

Section 6 describes recommendations for further development of the SS-FDMA system with emphasis on enabling technology and suggestions for flight experiments.

An appendix deals with the ramifications of SS-FDMA technology to potential applications of satellite aided land mobile systems.

SECTION 2

SYSTEM CONCEPTS, SS-FDMA

SECTION 2

SYSTEM CONCEPTS, SS-FDMA

2.1 OVERALL CONCEPT

Future U.S. communications satellites will make use of multiple beam antennas in order to provide more frequency reuse and therefore higher capacity satellite systems. A concommitant benefit of frequency reuse through multiple beam antennas is increased satellite antenna gain enabling a corresponding reduction in earth station antenna aperture. However, an access problem is then created in the satellite relating to the routing from the corresponding uplink beams to the corresponding downlink beams. Each uplink signal arriving at the satellite through one of several antenna uplink beams must find its way to the desired downlink antenna beam. These signals, are those associated with two way communication, either telephony or data, between two, (or just a few) earth stations. Distribution of television signals over wide coverage areas, another DOMSAT application is generally not benefitted by spot beam antenna systems.

While multiple antenna beam systems are not new, the number of antenna beams needed to carry the traffic can have a profound influence on satellite and system design. In a multiple beam FDMA satellite such as Intelsat V (to be launched in 1980) the routing is accomplished by assigning different transponders (e.g., different frequency bands) to different beams. Sophisticated switching is included to change the FDMA routing to accommodate the traffic demand. The switching (millisecond speed) is accomplished by a microwave switch network, switching 36 MHz and 72 MHz transponders. In a multiple beam TDMA satellite such as Advanced Westar (to be launched in 1981-1982) the routing is accomplished by assigning different time slots to different routes. Sophisticated on board switching connects each uplink beam to each downlink beam on a sequential non-interferring basis such that sufficient time is allocated to each route. This switching, (at nanosecond speeds) also is accomplished by microwave switch networks, switching RF paths containing 40 MBPS to 600 MBPS data streams.

Because of high earth status costs, conventional SS-TDMA systems are associated with trunking systems, e.g., systems having heavy routes or concentrated bundles of traffic. Consequently, the numbers of antenna beams of approximately 10-15 result in realizable practical satellite switch networks by present technology standards. Attempts to broaden the application of this technology to thin route or direct access terminals must involve techniques for reducing earth station costs. GE has performed a conceptual study for Western Union[(1)] using a combined TDMA/FDMA concept having the potential application to low cost earth terminals using TDMA.

FDMA on the other hand is widely used for both heavy route and thin route applications. In the former, single and multiple carrier transponders and digital or analog carriers form the basis for these systems. Thin route systems using single channel per carrier, (SCPC) with digital or analog modulation, and with and without DAMA are available. The Intelsat SPADE System is a digital SCPC/DAMA system used for thin route application. Algeria also has installed such a system for domestic applications but with a centralized processor. The General Electric Company has pioneered in the development of SCPC - DAMA using a centralized processor. The U.S. domestic carriers also use SCPC for thin route and direct access applications although, except in unusual circumstances such as in the Alaskan system, they have

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(1) Low Cost Earch Station Study Final Report, J. Kiesling and J. Swana, by General Electric Space Division for Western Union Telegraph Co., Contract WU 4748A (R-4-76).

not found wide application because of cost. All these SCPC systems are based on single antenna beam systems.

FDMA has considerable appeal for thin route and direct access systems because of the low potential cost of such earth terminals. Particularly in direct access systems where the user may have to bear the entire earth station cost (perhaps even the cost of two or more earth stations), earth terminal cost is critical. Use of FDMA earth terminals have the following advantages:

1. Narrow band: The incoming data or voice stream is transmitted to and received from the satellite at the same rate except for minor changes in rate due to error correction coding, signalling, etc. This translates to low MODEM cost and low power amplifier cost.
2. Equipment Simplicity: There is no complicated TDMA framing, satellite timing and no burst MODEMS as in TDMA. Signalling to accomplish DAMA is also simple and low cost at least with regard to user terminals when central routing processors are used.
3. SCPC earth terminals avoid the "propagating complexity" of TDMA terminals - these are additional costs incurred for power (back-up power for high-powered amplifiers), monitoring, alarm and control.

The inherent simplicity and low cost of an SCPC terminal suggests that it is an appropriate selection for thin route or direct access systems where terminal costs dominate overall transmission costs. In considering the adaptation of SCPC to multiple beam satellites, a satellite solution is sought which preserves the simplicity and low cost of the SCPC terminal even at the expense of higher satellite complexity, weight and cost. This is exactly the approach taken in the study, and the acronym satellite switched frequency division multiple access or SS-FDMA is coined to describe the general method.

2.2 SYSTEM ROUTING CONCEPT

While thin route or heavy route services can be provided by SS-FDMA through the use of isolated multiple satellite beams as evidenced by Intelsat V a more general application consists of direct access service to an area covered by a mosaic of contiguous satellite antenna beams as illustrated in Figure 2-1. Similar mosaics can also provide off-shore coverage, however such considerations are obvious and will not be considered further in this report. Frequency reuse is achieved in the arrangement of Figure 2-1 if the co-frequency beams are sufficiently far apart so that adequate isolation is achieved, either through sidelobe level isolation, or through the use of orthogonal polarization or both. The closer the co-frequency beams, the more frequency reuse is achieved, however, in general the interference levels also are higher. The insert in Figure 2-1 is an example of frequency reuse achieved only through sidelobe isolation. The available bandwidth is divided into three parts, labeled "1", "2 . d "3" and only one of these three bandwidths is used in any one beam.* Total bandwidth available is $\frac{NB}{3}$ where: N = number of beams, B = available bandwidth. A 100 beam satellite with a 2500 MHz allocation can have a total available bandwidth of 83250 MHz! Co-frequency beams in

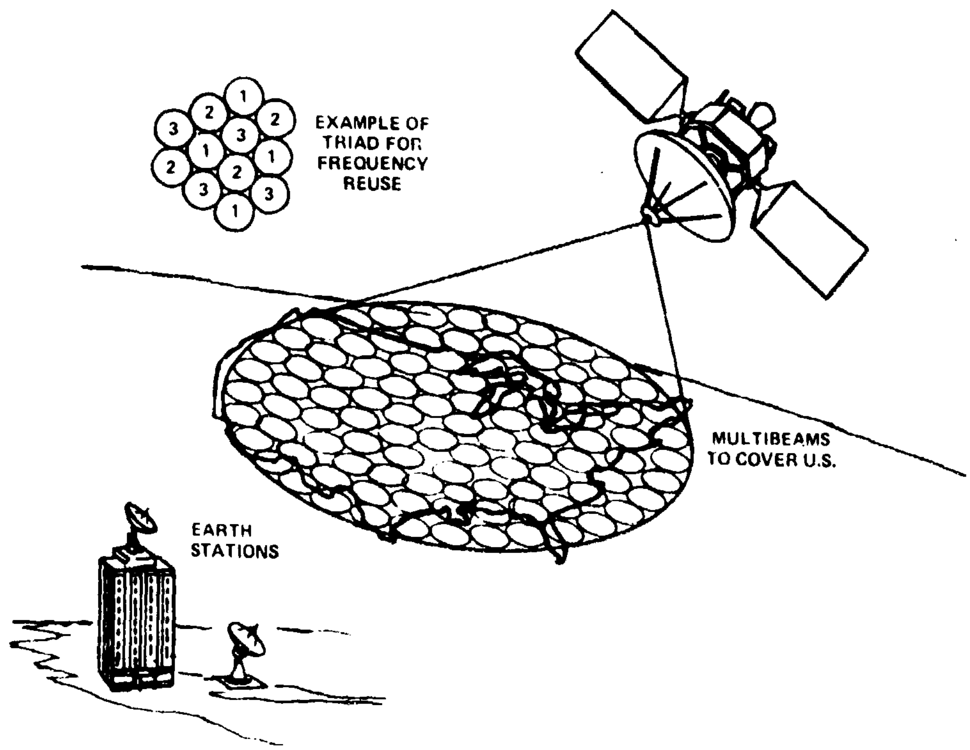

Figure 2-1. Concept of Two-Way Communications Service to a Coverage Area Through The Use of a Mosaic of Multiple Contiguous Antenna Beams

* The total bandwidth available to an earth station is B/3.

Figure 2-1 are less than a beam width apart and the co-frequency interference level can be high. Alternatively, orthogonal polarization can be used to gain additional isolation. Unfortunately, at Ka Band, the orthogonality is severely degraded in rain so that equal level, co-frequency, co-beam but orthogonally polarized signals can have nearly the same magnitude during heavy rain. On the other hand, orthogonal polarization can be used to achieve additional isolation between two different beams where isolation also is provided by sidelobe level control. Such arrangements, both singly and dually polarized, called beam topology, are an important subject of this report and are described in Section III with regard to topologies that maximize frequency reuse, or minimize interference or combinations thereof.

Figure 2-2 depicts the SS-FDMA satellite concept. Each uplink beam of a contiguous multiple beam antenna contains a 30GHz receiver which essentially establishes the input noise temperature and which converts the received band of FDMA

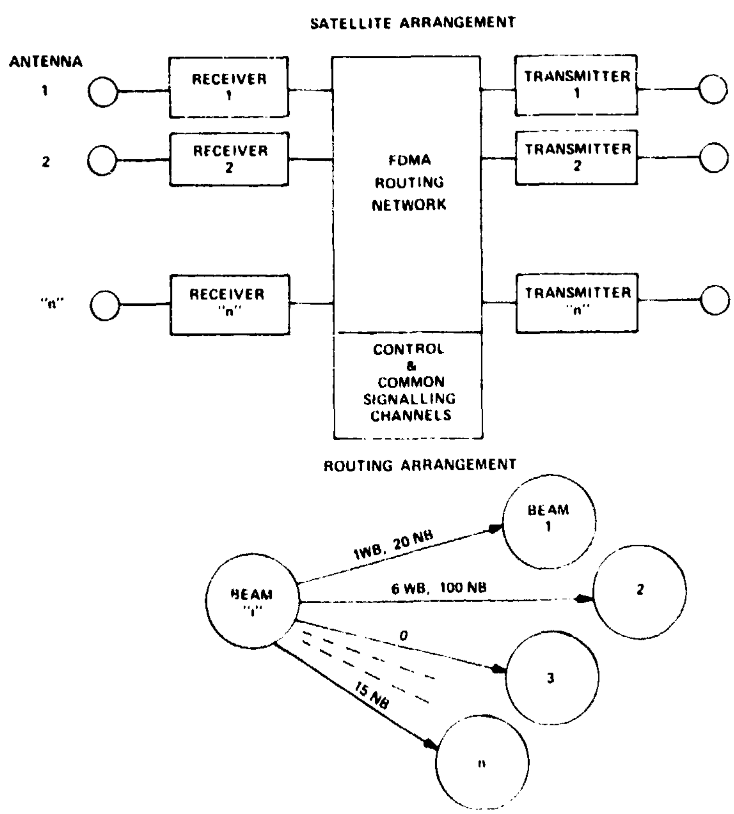

Figure 2-2. Satellite Arrangement for SS-FDMA

signals to a convenient intermediate frequency. Correspondingly, each 20GHz downlink beam is driven by an amplifier/frequency converter which amplifies the FDMA carriers for that beam to the proper power level while providing control of intermodulation noise. The routing of the FDMA signals through the satellite, e.g., each uplink signal from each uplink beam, to its destination downlink beam is controlled by the FDMA network. The arrangement resembles that for SS-TDMA hence the acronym SS-FDMA. However, the FDMA routing network actually contains networks of frequency changers, RF filters and switches to accomplish the routing function, all controlled by the Satellite Control System which also contains elements of the system Common Signalling Channel, and telemetry and command functions.

At first glance, the SS-FDMA system appears to be a "switchboard in the sky", an old idea advanced to emulate terrestrial systems. However, this approach, which involves a switch to connect any subscriber to any other subscriber is a very complex technology, even in terrestrial systems and is a hopeless proposition for space implementation in the foreseeable future. And it is not necessary. In SS-FDMA, it is only necessary to route the signals properly amongst "n" beams, the remaining subscriber connectivity can be provided by action of the earth station network. To accomplish this, RF paths are set up by the FDMA network through the satellite to the proper downlink beams, FDMA signals for the "ij" route are merely assigned frequencies within the bandwidth of this route, call a "path". If the path bandwidth is not adequate to satisfy the instantaneous requirement, additional "paths" can be switched in--perhaps at the expense of other paths where the demand is lighter. In concept, the available spectrum within each antenna beam can be divided (on an FDMA basis) into distinct bandwidths or paths as illustrated in Figure 2-3. A single carrier or multiple carriers may be assigned within each path depending on the subscriber or user demand. Most of the traffic in a Direct Access System is at moderate speeds say in the range of 2.4KBPS to 56KBPS, however, there may exist heavy routes of "bundled" or multiplexed traffic or a

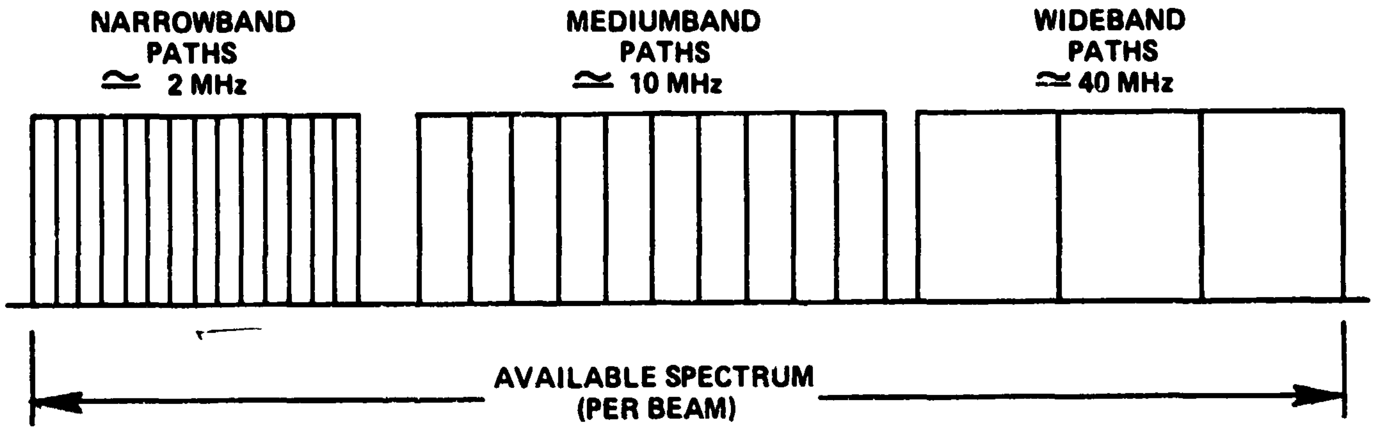

Figure 2-3. Concept of FDMA Path Allocation in Representative Antenna Beam

demand for occasional high speed service such as interactive television (40MBPS). Therefore, there is a need to provide combinations of path bandwidths. The existence of 36 MHz paths does not necessarily mean that these allow only single carrier services because it also is convenient, in very active paths, to assign multiples of slower data rate carriers to these wider band paths, e.g., if the total required bandwidth is large--even though this is composed of many slower speed carriers--it is more convenient to provide a single wider bandwidth path then to connect several narrow band paths.

Within the satellite, the FDMA spectrum illustrated by Figure 2-3 is routed to the other beams on a path by path basis. In Figure 2-2b, beam "1" received paths are distributed by a switch network (not shown) to "n" downlink beams, some downlink beams will be provided with a lot of paths, perhaps of differing bandwidths while others may be provided with substantially less, or with none, depending on the instantaneous traffic demand. In essence, the satellite switch behaves as a "bent pipe", e.g., no demodulation or decoding takes place, and within any given RF path, any reasonable user format (analogue or digital modulation and/or coding) can be accommodated, at any speed, provided the satellite path has adequate bandwidth.

The satellite design problem then is to define these paths and switch them to the proper (destination) downlink beam. In essence this is accomplished by down converting the RF path to frequencies which are low enough so that convenient lightweight filters can be used to separate each individual path from all the others, and that advantage can be taken of LSI (large scale integration) to construct large switch matrices. An example arrangement is depicted in Figure 2-4 which illustrates a contiguous FDMA stack of RF paths applied to a bank of mixers supplied with oscillator power from a frequency synthesizer. Each RF Path is converted to the same frequency where it is filtered or channelized and then applied to a crosspoint switch (fully connected n by n switch) where each individual RF Path is connected to a new FDMA "stack" in its destination beam. In general, each incoming RF Path can be connected to any downlink beam, although in practice this is not always necessary. Figure 2-4 is illustrative only, a detailed design for a specific application may indicate the need for several down conversion steps, differing bandwidths of RF paths, several different switch networks, etc. However important technology development is indicated by the Figure 2-4 concept. The switching network can consist of thousands of switch elements with tens of thousands of filters and of course, the frequency conversion step itself, while not requiring new technology is in itself a challenging engineering problem. However, the RF path switching has significant advantages for a satellite implementation which are worth summarizing here.

1. RF path switching substantially reduces channelization and switching requirements so that implementation in the foreseeable future is feasible.

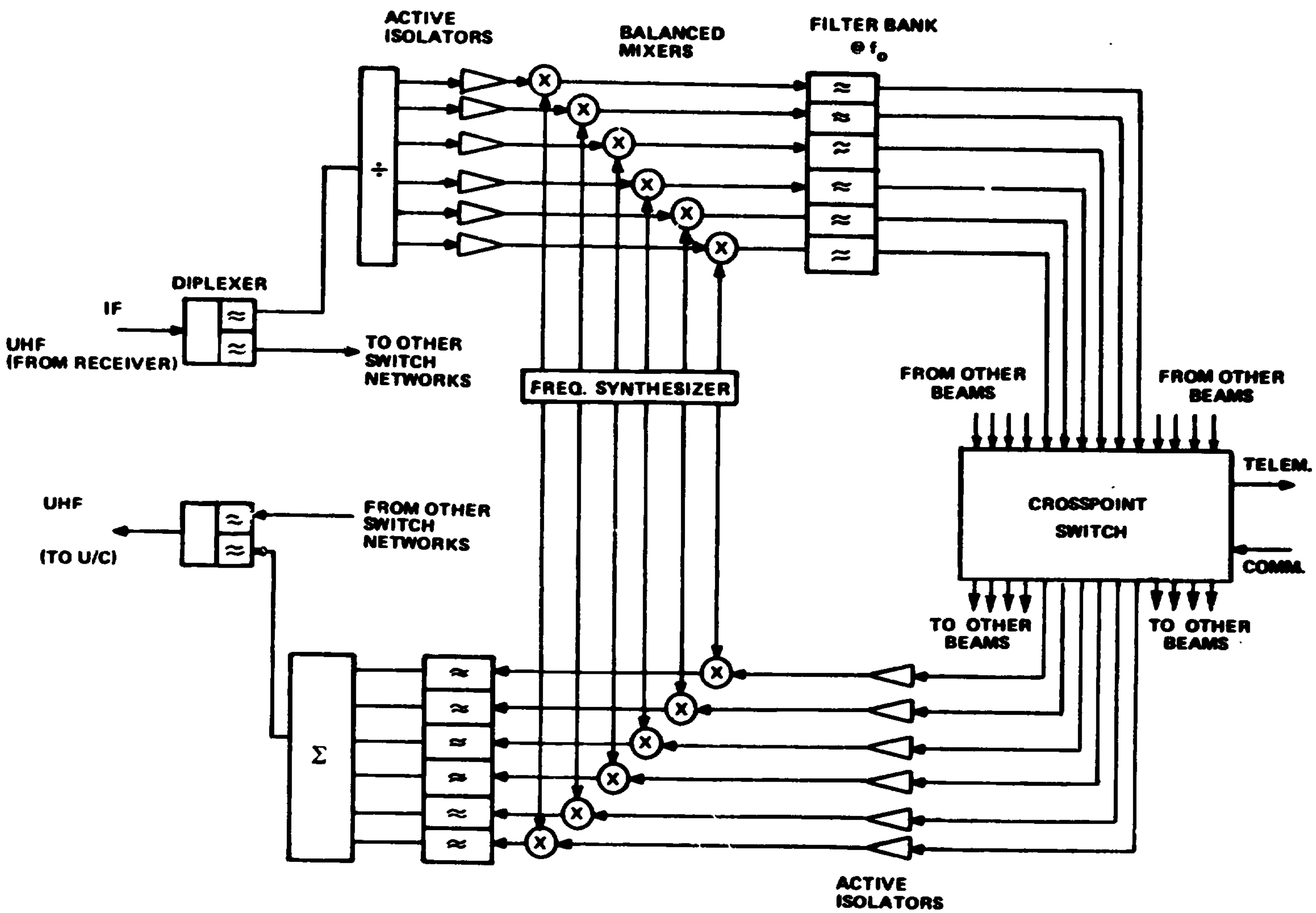

Figure 2-4. Example of Typical Switch Network

2. RF path switching, by avoiding demodulation and decoding (and remodulation and recoding) in the satellite imposes minimum restrictions on user formats; within a broad range of possibilities the user may use analogue or digital modulation at any bandwidth or rate.

3. RF path switching can enhance reliability since multiple paths are provided for many routes and additional paths can be switched in on demand. Switch failures, or in general path failures reduce flexibility and eventually capacity, but do not cause catastrophic failure of the network.

A key to the achievement of these advantages is operation at sufficiently low frequencies such that new solid state technology (SAWS,* ceramic-piezoelectric filters and LSI) may be applied.

In summary, the SS-FDMA routing concept, applicable to a multiple beam satellite is a new system concept offering potential benefits to user systems. Some of these benefits are:

1. It avoids the expense of complex TDMA terminals.

2. It avoids the need of high-powered amplifiers in earth stations because no bursting is required. 1 and 2 suggest that SS-FDMA favors the use of direct access terminals.

3. It has the potential for high flexibility and reliability.

4. It has the potential for use either as a direct access or trunking system and can support heavy route or thin route traffic (or high or low data rates).

5. The RF paths (bent pipe) do not limit user modulation and coding formats.

6. The application of SS-FDMA to a direct access satellite with SCPC-DAMA provides the system with a complete switching capability which is completely independent of terrestrial facilities (interconnects, entrance lines, switches, etc).

7. While not fully discussed yet, SS-FDMA also offers advantages to Ka Band or Ku Band systems because of carrier power sharing so that availabilities in the range of 0.999 to 0.9999 can be achieved with single antenna earth stations in all but the most extreme U.S. climates, with little cost impact to the user.

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*Surface acoustic wave or SAW filters.

8. Satellite amplifier linearization to control intermodulation noise (through back-off, feed forward or feed back techniques) is a FDMA penalty, however, if the VOX factor is considered along with the power sharing for fade compensation, and the ability to adjust the amplifier RF and DC input levels to satisfy the traffic demand also is considered the penalty for linearization is not significant. A FDMA transponder even though "backed off" will weigh less and consume less power than a TDMA amplifier. Batten weight also can be reduced if the eclipse time traffic is small and the transponder power can be reduced during these periods.
9. If desired the carrier level at the earth station can be adjusted so that the satellite carrier EIRP is the same no matter where in the satellite beam the destination earth station is located.

Disadvantages of SS-FDMA are that it significantly increases the complexity of the satellite--in switching, channelization, and frequency conversion. While these require the further development of new but existing technology there does not appear to be a substantial increase in satellite cost above that of an SS-TDMA direct access satellite (both with multiple beam antennas). Thus, there appears to be an opportunity to demonstrate improved satellite communications system through the application of R&D.

2.3 DIRECT ACCESS SYSTEM USING SS-FDMA

2.3.1 INTRODUCTION

Since SS-FDMA permits the lowest cost earth station a truly cost effective direct access system might be attainable using this access method. By direct access we mean an earth station located on the user premises which communicates directly with its destination earth station via the satellite with no intervening terrestrial facilities. Most of the present satellite carriers operate some direct access terminals, mostly high-speed pre-assigned data terminals. However, the bulk of the traffic is carried by so-called trunking systems in which the traffic is first concentrated at the earth station via terrestrial lines. At C-Band, additional entrance links are required because the terminals are located outside cities, away from RF conjestion. Trunking stations contain a lot of multiplex equipment either FDM or TDM for the purpose of "bundling" the traffic into heavy routes. The question naturally arises, given a user having a certain data rate, use factor and route distance, concerning the best communications media; an all terrestrial system, a trunking system or a direct access system. It is difficult to analyze the question in a general sense because users have complex service requirements, e.g., different traffic requirements to different cities and users often use combinations of different facilities to satisfy their needs. However, some insight can be gained by a general comparison of the three generic configurations. Figure 2-5 illustrates the three configurations. The terrestrial systems, consisting of complex tandem connections of wires, switches, radio relays and multiplexing can be represented for traffic purposes by a simple line. The trunking system illustrates that access is provided by terrestrial facilities. In the following, it is assumed that the three services have comparable performance characteristics, and satellite delay is not a factor.

2.3.2 SATELLITE COSTS

Satellite costs used in this analysis are conjecture. Present use of 56 Kbps direct-access systems results in about 100 circuits per $1.2M transponder or costs of approximately $12K/year. However, the values used in these analyses presume a satellite system designed specifically for SCPC/DAMA using small earth terminals. For bandwidth limited operation ($1.2M, 36 MHz) the following results for both trunking and direct access satellites using representative present day charges.

for 56 Kbps	$2700/circuit/year
9.6 Kbps -	$462/circuit/year
2.4 Kbps	$116/circuit/year

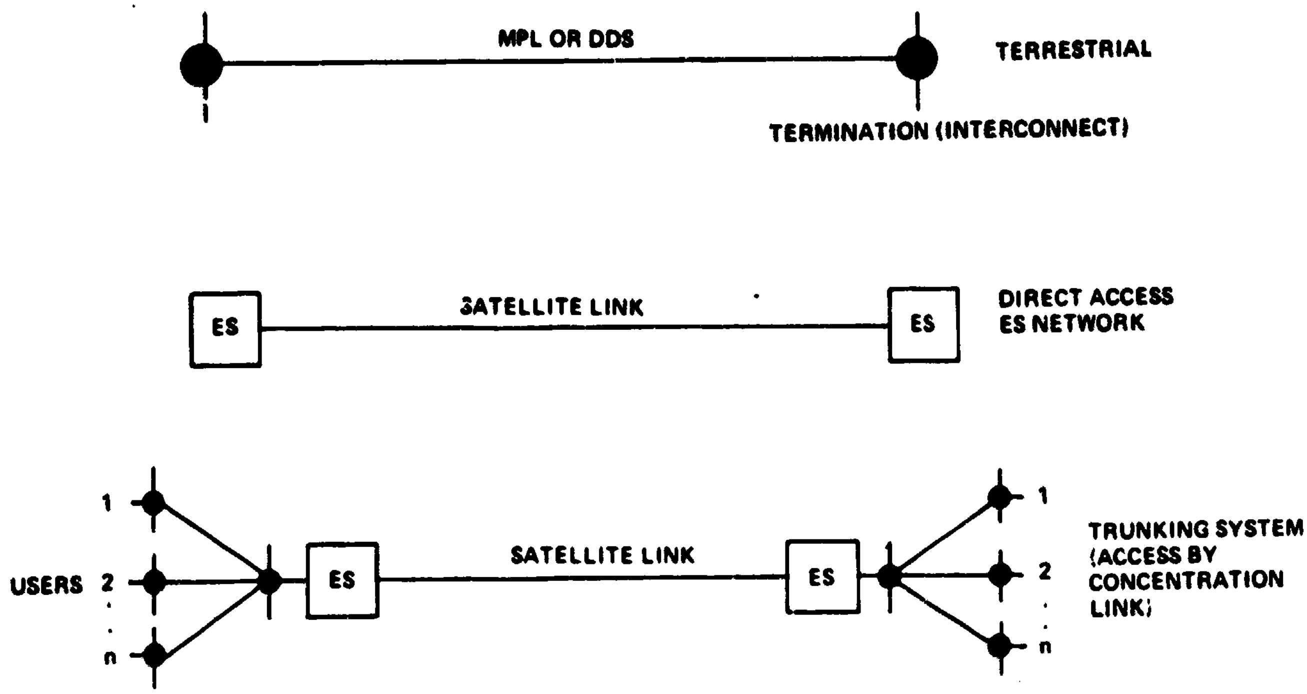


Figure 2-5. Communications Facilities

These assume preassigned channels. Operation with VOX/DAMA can substantially reduct these charges. Satellite charges used for the analyses of direct access systems are assumed to be \$11.7K or \$ 1.17K per year for 56 Kbps and \$2000 or \$200 per year for either 2.4 Kbps, 4.8 Kbps or 9.6 Kbps (the latter assumes these data rates will use 9.6 Kbps as a standard rate, with buffering if necessary). These charges straddle the range of possibilities and enable an evaluation of the sensitivity of user end-to-end costs to these charges. At this point in the study the range of satellite charges is quite arbitrary and will need to be reviewed later.

With regard to full video services the charges in Table 2-4 are too high for interactive services such as teleconferencing. Consequently, a direct-access or concentrated satellite system will obviously be more cost effective provided DAMA is used.

2.3.3 TERRESTRIAL SYSTEM COSTS (TARIFFS)

Table 2-1 lists our understanding of the present standard AT&T tariffs for DDS/MPL (Digital Dataphone Service/Multiple Private Line) for 2.4 Kbps, 4.8 Kbps, 9.6 Kbps, and 56 Kbps. It should be noted at the onset that tariffs are analogous to prices, and are not necessarily costs. There are two important charges, one relating to distance (so-called mileage charge) and a second relating to terminating the lines at a user's premise, (a third charge relating to installation is small and is neglected).

Table 2-2 lists similar characteristics for the Series 8000, a medium speed data service, and Table 2-3 lists the characteristics of Telpak "C" and "D", high speed data services. Telpak may be discontinued subject to a present court dispute between the Government and AT&T. In any event, it is not available to new customers (since 1977).

Table 2-4 lists the Series 7000 tariff for continuous and occasional TV. It is apparent that these are too expensive for direct-to-user services such a teleconferencing.

Using a hypothetical 500 mile, 2-way teleconferencing link the Series 7000 tariff for preassigned service results in a monghly charge (mileage, termination and local charges) of $60,000. If occasional service, say 8 hours per week, is assumed then the monthly charge (mileage/hourly, termination and local charge) is $37,240.

The various data service tariffs are plotted in Figure 2-6 versus mileage. These are characterized by a fixed termination charge (per pair of ends), called herein T_o, and a mileage rate T over a mileage distance L. For a direct-access system to break even the comsat charge must equal $T_o + TL$, e.g., there is a break-even distance at which the two services are equivalent in projected charges. The situation is different for trunking systems which concentrate the traffic at the earth stations. In this case, a pair of terminations is required at each earth station to connect each circuit. For Break-even, the trunking systems must reduce the long haul charges sufficiently to overcome the effect of the extra termination charge. Adding more circuits only adds more termination charges. In fact, it is shown that the

TABLE 2-1. DDS/MPL (2 WAY) TARIFFS
(monthly charges)

Mileage	(A-A City) 370 Cities Schedule 1	(A-B City) Not Available Schedule 2	(B-B City) Not Available Schedule 3	56 Kbps DDS
0-1	51	52	53	225
2-14	51 + 1.80/mi	52 + 3.30/mi	53 + 4.40	255 + 9/mi
15-24	76 + 1.50/mi	98.20 + 3.10/mi	114.60 + 3.80	372 + 7.50
25-39	91.20 + 1.12/mi	129.20 + 2.00/mi	152.60 + 2.80	450 + 5.60
40-59	108 + 1.12/mi	159.20 + 1.35/mi	194.60 + 2.10	534 + 5.60
60-79	130.40 + 1.12/mi	186.20 + 1.35	236.60 + 1.60	646 + 5.60
80-99	152.80 + 1.12/mi	213.20 + 1.35	268.60 + 1.35	758 + 5.60
100-999	175 + 0.67/mi	240.20 + 0.67	295.60 + 0.68	870 + 3.35
≥1000	769.20 + 0.40/mi	834.20 + 0.40	907.60 + 0.40	3885 + 2.0
MODEM Termination	2400 $84/mo/end (2016/yr) 4800 160 (3840) 9600 281.33 (6752)	$25/end/mo (no MODEM)	Same as 2	$650/mo/end (15600) Install $180.76/end
	+ Install $128.75/end	$54.15/end	Same as 2	

TABLE 2-2. SERIES 8000 (2 WAY) - 40.8 - 50 KBPS

	12 Ckts (11 Data + 1 Voice)	
	$16.20/mi/no/ to 250 miles	
	11.40 251-500	
	8.15 7500	
Termination + MODEM		$460/end/mo
Installation		$216/end

TABLE 2-3. SERIES 5000* (TELPAK)

	Telpak "C"	Telpak "D"
	60 Ckts or 240 kHz (743 Kbps)	240 Ckts or 960 kHz (3 Mbps)
Mileage	$32.50/mi/mo	92.05/mi/mo
Termination	$43.30/end/mo/ckt	43.05/end/mo/ckt
Installation	$5415/end/mo	54.15/end/mo

*Not available with independent companies for example in Rhode Island or Vermont.

TABLE 2-4. SERIES 7000

TV (1 Way Only)

Continuous - 525 Line, + Audio

Mileage	$55/mi/mo
Termination	$1500/mo/end
Local Charge	$1000/mo/end

Occasional Service - 1 Way

$0.75/hour/mile, 1 hour minimum

Termination Charge	$80/hour/end
Local Charge	$500/ch/day/end - min. charge
	$500/day/end
	max $1000/end/mo

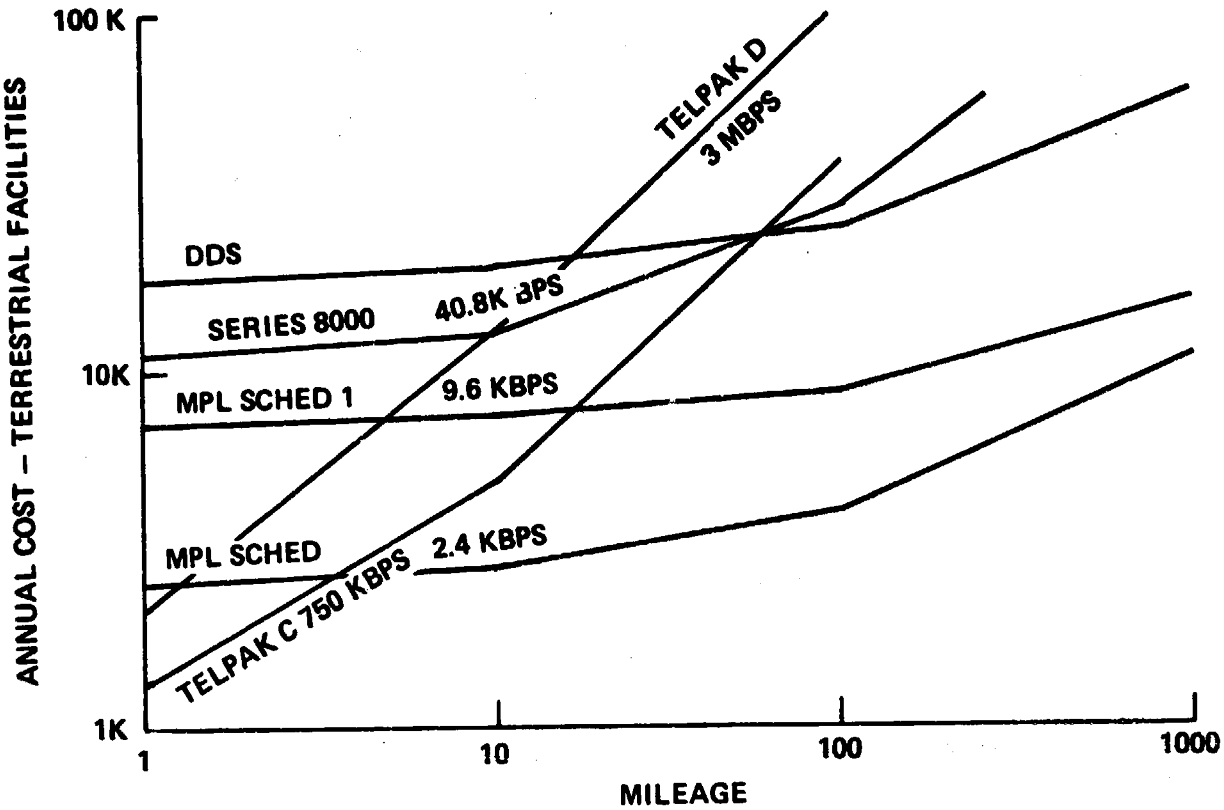


Figure 2-6. Various Tariff Schedules vs. Mileage

termination charge (or interconnect charge) is a significant cost parameter for concentrated systems. This break-even relationship can be approximately expressed as:

$$2\left(nT_o + \frac{P_e}{3}\right) + nS' = nLT + nT_o$$

where

P_e = earth station installed cost

n = number of circuits (e.g., two-way)

T_o = termination charge per circuit (both ends)

S' = satellite charge per circuit

L = break-even mileage

T = average mileage rate for an equivalent terrestrial link

For the direct-access system the break-even relationships can be expressed as:

$$2\frac{P_e}{3} + nS'' = nLT + nT_o$$

Note that in both cases, earth station installed costs are multiplied by 0.33 to obtain annual costs (including depreciation, O&M and return on investment). These equations can be manipulated to show important relationships concerning economic viability.

2.3.4 RESULTS OF COMPARISON

Figure 2-7 shows such relationships for a data rate of 2.4 Kbps, plotting earth station installed cost versus break-even distance. Both trunking networks and direct-access networks are shown. A postulated value of $20,000 is shown for a

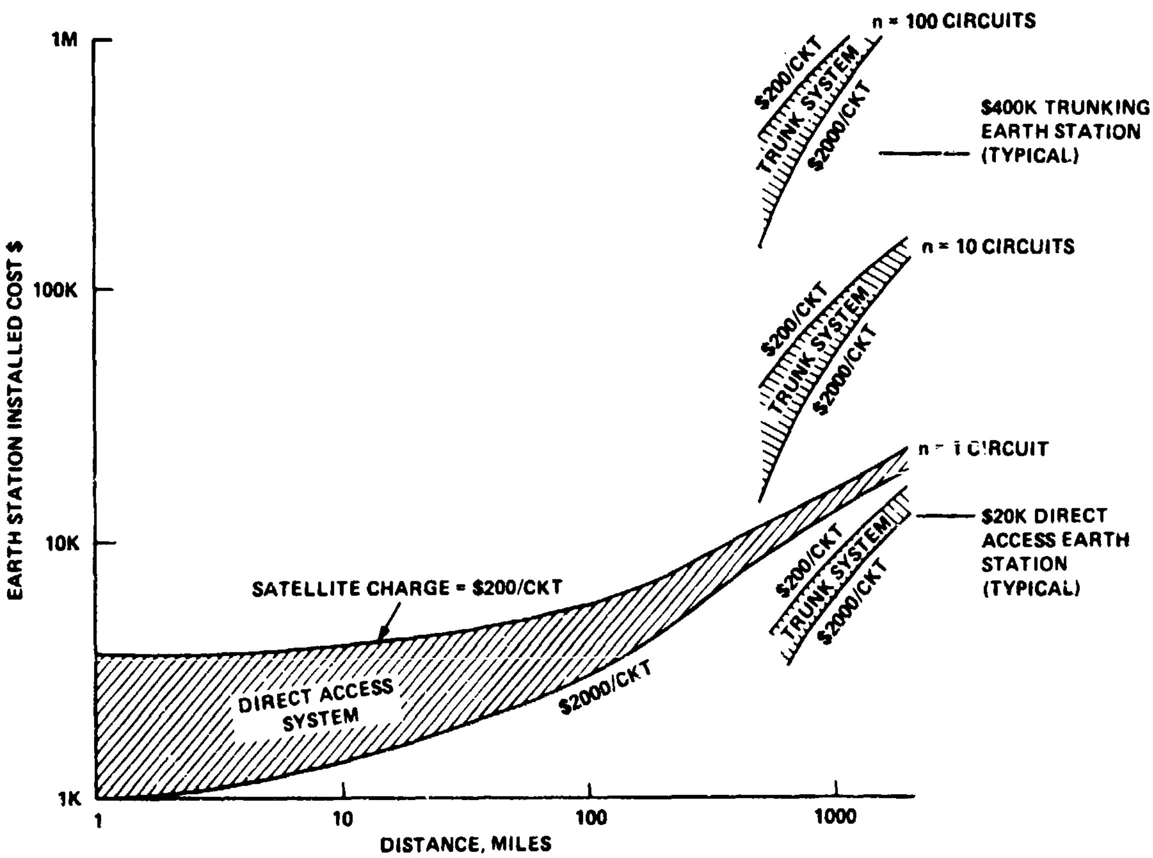


Figure 2-7. Breakeven Cost Comparison 2.4 Kbps MPL

typical direct-access terminal and $400,000 for a typical terminal with concentration, located within a city in its service area. An out-of-city trunking station costing $5M to $10M including radio relay and multiplex is believed to be typical of that situation. The trunking terminals have a mileage asymptote setting a minimum break-even distance, typically around 500 to 1000 miles. Even if the pro-rata cost of the satellite portion of the link is zero there is still a break-even distance, when $T_o = LT$, e.g., break-even is determined essentially by the terrestrial system. This may be viewed as the terrestrial mileage necessary to overcome the cost of the extra pair of terminations in the trunking system. The break-even distance cannot be less than this. There is no asymptotic break-even for direct-access systems. However, there is a cost break-even; for an earth station cost of $20,000 (beyond present state of art and experience), the break-even distance is approximately 1000 miles. For trunking systems, break-even is 500 to 800 miles for about 100 circuits depending on the satellite circuit charge. For 2.4 Kbps, concentration is the only practical satellite solution since direct-access earth station costs substantially below $20,000 are not believed to be feasible.*

However, the slope of the curves for the direct access system are small; a reduction of earth station cost, from $20,000 to $10,000 changes the breakeven distance substantially--to approximately 100 miles! This might be achieved in a situation requiring two 2.4 Kbps lines, e.g., $20,000 earth station with two channels, or a charge of $10,000 per earth station. Note that the satellite charge also is critical because of the small slope. Reduction of satellite charges to several hundred dollars per circuit, by advanced satellity technology, or DAMA, or both can be significant. The trunking system characteristics on the other hand have large slopes around 1,000 miles. Changes in earth station costs or satellite charges have almost no effect on the breakeven distance. This does not mean that the

*Even an "occasional" service, in which the space segment charge approaches zero does not change the break-even distance significantly.

trunking system is not cost effective. At an earth station cost of $400K, and with 100 2.4 Kbps trunks, the trunking system is competitive provided the average mileage is greater than 1,000 miles. In essence, the abscissa of Figure 2-7, the mileage, also is proportional to user cost. Thus, an important implication of Figure 2-7 is that the direct access system has high potential for reducing user costs providing the required earth station costs and satellite charges can be achieved.

Figure 2-8 shows similar results for 9.6 Kbps. In this case for a $400K trunking station at least 100 circuits are needed for the trunking system to achieve a break-even distance of 1000 miles with an asymptote of 580 miles*. The direct-access system with a $20,000 earth station, on the other hand, breaks even between 500 and 900 miles.

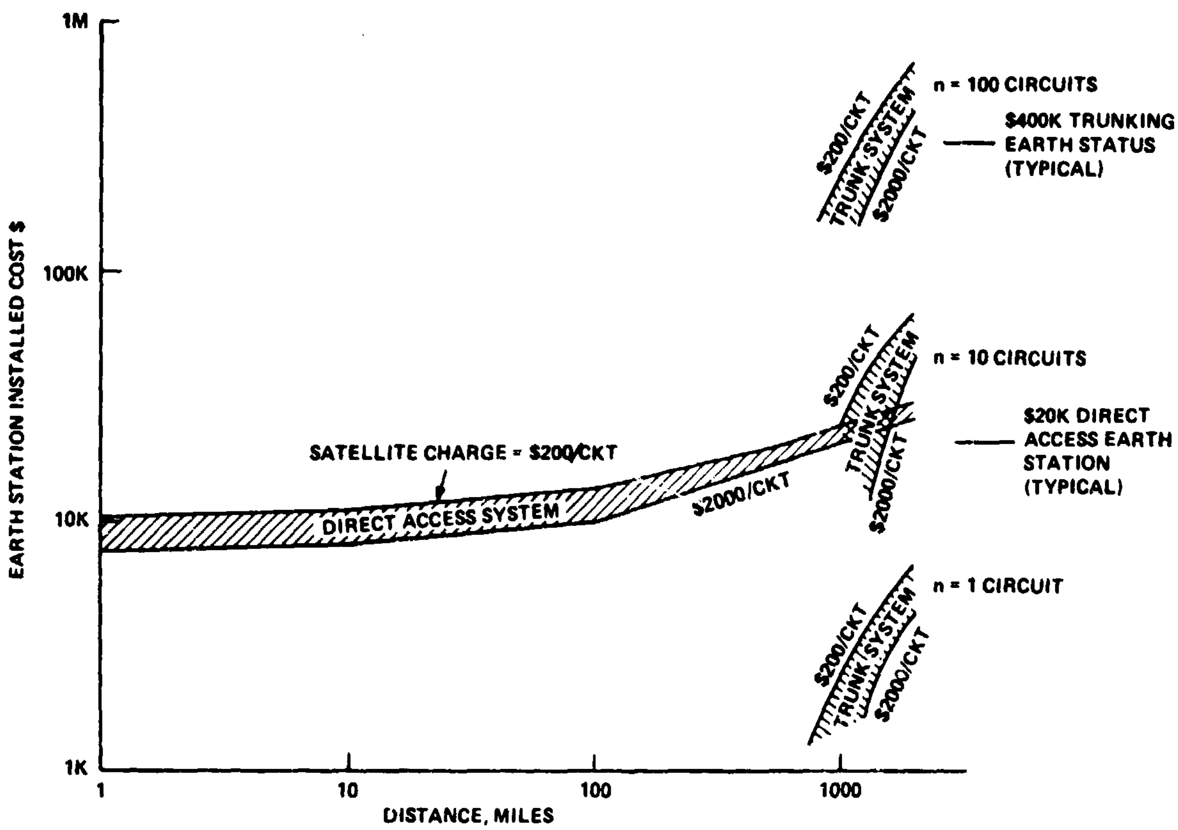


Figure 2-8. Breakeven Cost Comparison 9.6 Kbps MPL

* The 10 circuit case for trunking requires an earth station cost under $100K which is not believed to be achievable.

Again, the direct access system achieves a low breakeven distance if its earth station cost can be reduced to $10,000 (either by a $10,000 earth station cost or by the use of two 9.6 Kbps channels in a $20,000 earth station). Again there is a large payoff to reducing the satellite charge. In fact, breakeven distances less than 10 miles can be achieved. The trunking system can never achieve these low costs because of the terrestrial interconnects.

Figure 2-9 shows similar results for 56 Kbps considering the use of both DDS and Series 8000 as interconnect and DDS for long haul. For concentrated traffic, a trunking system breaks even at 10 circuits at about 1100 miles. With lower cost interconnections (Series 8000) this is reduced to 600 miles with an asymptote of 310 miles. Note that the lower priced interconnects are just as effective as lower satellite costs in reducing overall costs. Figure 2-9 also shows that a $10M trunking

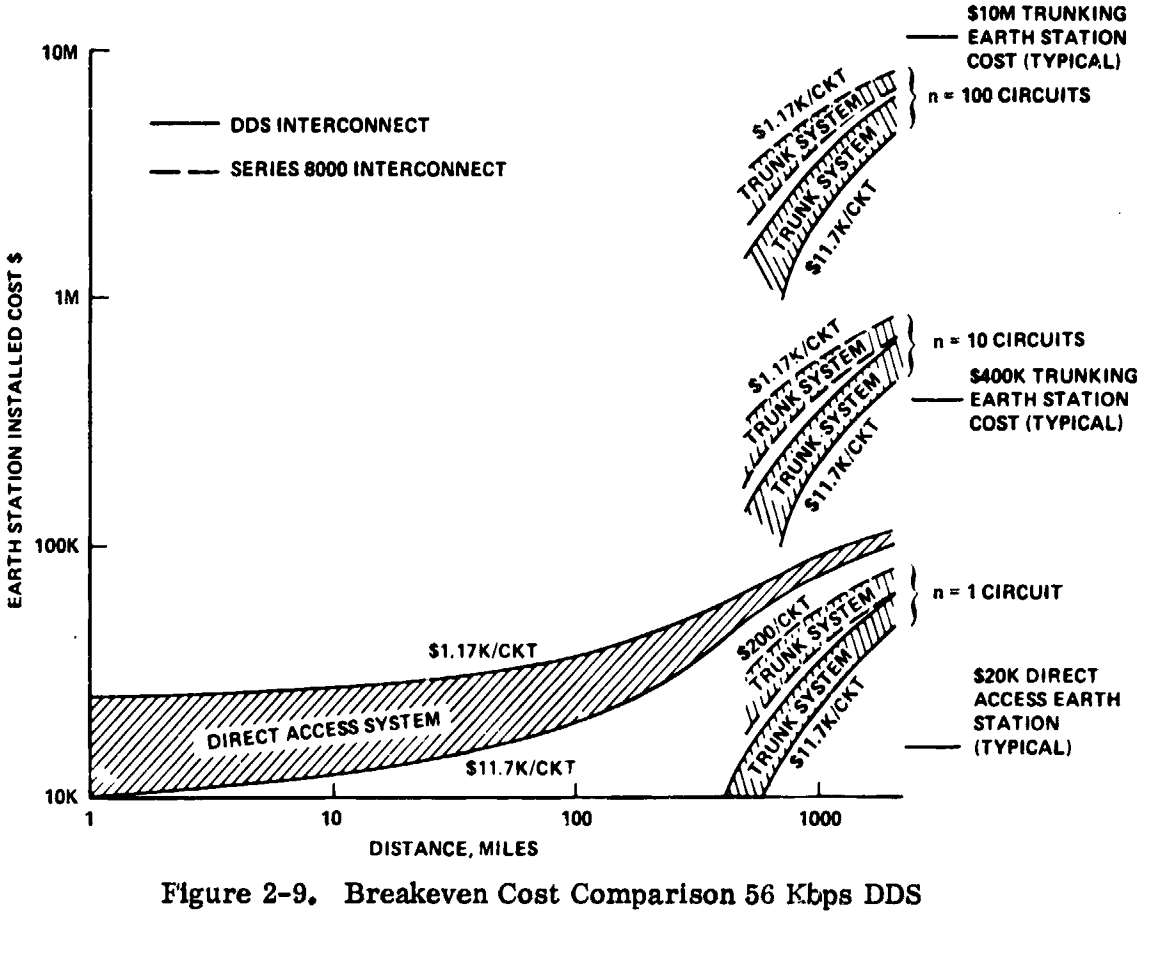


Figure 2-9. Breakeven Cost Comparison 56 Kbps DDS

terminal also can break even but at a level greater than 100 circuits. The direct-access system break-even is attractive at 100 miles or less. With sufficiently low satellite charges the direct-access system has the apparent potential of providing intra-city traffic (with Series 8000 or Telpak this would not be true). In fact, with break-even distance less than 10 miles, the direct-access system is by far the lowest in cost.

Again, the direct access system has the best potential for providing low user costs and in providing a real alternative to the use of terrestrial facilities. Multi service (more than one channel per earth station and/or use of DAMA) can significantly improve this situation. In all three situations (2.4 Kbps, 9.6 Kbps and 56 Kbps), the direct user system has the potential for lower user costs than a trunking system--irrespective of network size. Consequently, a serious review is warranted of the significant economic factors--the potential satellite charge of an SS-FDMA satellite and the potential cost of the direct access earth station.

Present day direct access C-Band terminals cost of the order of $100,000, (less for telephony, more for data). Figures 2-7, 2-8 and 2-9 show that even with lower satellite charges than are attainable today (satellite charges are high for this service because the satellite transponders do not have sufficient power to operate with smaller earth station antennas, even if these were permissable at C-Band), these earth station costs result in breakeven distances considerably greater than 1,000 miles and costs that are higher than those of trunking systems. Consequently they are attractive for use only in situations where the average circuit length is high, the data rate high (approximately 56 Kbps) and where trunking facilities are not available. Lower earth station costs can only be achieved by higher performance satellites which reduce the earth station antenna and amplifier costs (in bands where small earth station antennas are permitted) and by maximum application of LSI (large scale integration) and MIC (microwave integration circuits) to large volume production of standarized terminals.

The direct access and trunking systems also can be compared directly; this can provide additional insight into the salient characteristics of the two systems.

This can be evaluated by finding the concentration factor (n circuits) and other key parameters at which the two approaches break even. This is given by:

$$n = \frac{1}{(P_e{}'' - 3T_o) - 3/2\ (S' - S'')}$$ and results are shown in Figure 2-10.

where

$P_e{}'$ = earth station acquisition cost for concentrated traffic

$P_e{}''$ = earth station acquisition cost for direct-access

S = corresponding satellite circuit annual charges

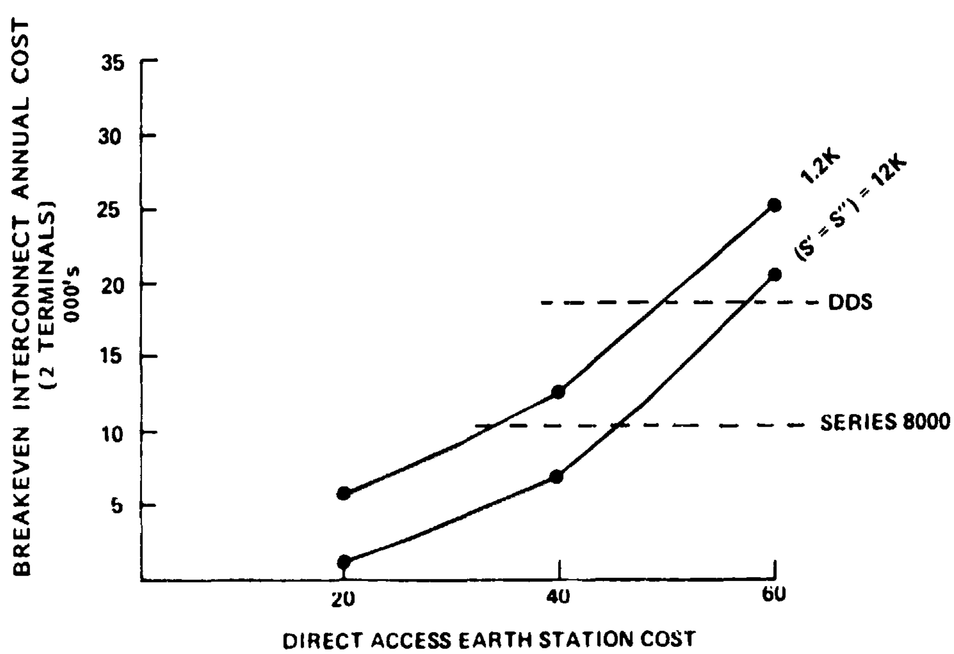


Figure 2-10. Interconnect Cost vs. Direct Access Terminal Cost for n → ∞ for 56 Kbps Circuit

2.4 USER EARTH STATION CHARACTERISTICS

2.4.1 INTRODUCTION

Several different earth stations are required to provide service. The most critical with regard to cost is the single MODEM (or single service earth station). When additional services are added, additional MODEMS, CODECS, etc., also are added, and the HPA power also increases, however, there is still much equipment not requiring alterations such as the antenna, low noise receiver, converters, signalling system, etc., so that the earth station installed cost per MODEM (and its corresponding annual cost) is reduced as services are added.

Figure 2-11 is a simplified block diagram of a single service earth station consisting of an antenna, "single thread" microwave and "single thread" MODEM, CODEC and DAMA. It may be used for a variety of services such as those indicated, but only one at a time, at the MODEM rate and to a single destination. Additional flexibility can be provided by a smart multiplexer, a processor-controlled time division multiplexer that can accept a variety of simultaneous inputs provided the MODEM rate is not exceeded. Several channels of data or voice also can be provided with such a configuration (again provided the MODEM rate is not exceeded) but if the destinations are in different beams of the satellite multiple beam antenna then additional switch capability is needed in the satellite in order to switch the single carrier (or multi-destination carrier) into two separate downlink antenna beams. This discussion indicates that such a terminal is very flexible and the flexibility is provided by the multiplex which is not a costly item since it is amenable to production via LSI. While more sophisticated earth stations with many MODEMS also will be needed, say to handle communications for large complexes of people and computers etc., the cost per MODEM channel will be even less since the common equipment items will

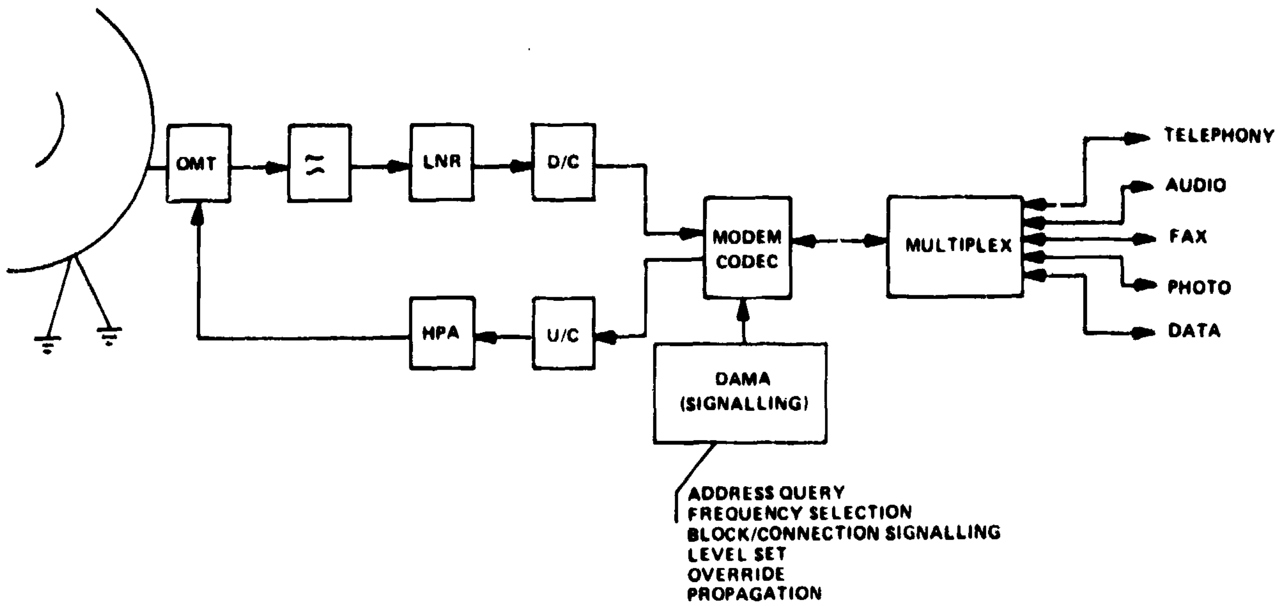


Figure 2-11. Low Cost User Typical Earth Station Configuration

be shared among many channels. Consequently, it is particularly pertinent to determine the cost of the single service (e.g., single MODEM) earth station type. This problem will be addressed in subsequent paragraphs.

2.4.2 SYSTEM OPTIMIZATION

The achievement of low cost earth stations through system optimization is beyond the scope of this report, although it will be apparent that certain optimizations have been attempted, for example in earth station antenna/satellite sizing and in overcoming the effects of precipitation attenuation.

The satellite system parameters can be arranged to result in a low earth station cost and (hopefully simultaneously) achieve a low user cost. For example, the satellite must have adequate power to permit a cost effective earth station antenna. For a bogey $20K goal (1980 dollars) this antenna should be in the range of 1 meter to 3 meters, otherwise antenna cost, including its shipping and installation cost will be too high. Actually we are interested in the G/T of the station (antenna gain, divided by the system noise temperature) however previous studies indicate that a transistor low noise receiver using GaAsFets will likely be the most cost effective configuration.

Such devices are barely in the laboratory test phase now, but presumably will be available in quantities for operational Ka-Band systems. The satellite also must be designed for use with small earth stations. This normally means high power, or to be more precise, the system should be designed so that it is not power limited and with sufficient capacity so that the satellite charge is small. Assignment of channels on demand, or DAMA also can help since this enables the users to efficiently share a limited facility. Finally a balance between earth station antenna diameter (and hence cost) and satellite charge should achieve a minimum user cost. If the earth station antenna is too small the satellite charge can be too great, if the earth station antenna is too large then the earth station cost will be too great.

A small earth station antenna can also increase the cost of the earth station high power amplifier or HPA. It has been shown to be a secondary consideration to optimizing G/T. However, it is desirable that the HPA be solid state to improve reliability and life and reduce maintenance. Solid state amplifiers, either GaAsFets or IMPATTS can be considered for use in future operational Ka-Band systems but again, the system must limit the power levels to several watts. Fortunately, the use of satellite multiple beam antennas reduces earth station HPA requirements so that it appears that simple configurations of solid state amplifiers can be considered. For multi-service earth stations where the earth station must generate several simultaneous RF carriers, it may be necessary to parallel amplifier stages consisting of solid state devices, or multiplex single carrier amplifiers at the microwave band, (which places limitations on frequency planning) or resort to tube types such as traveling wave tube amplifiers with sufficient power to allow backoff for linearization purposes. Fortunately, the multi-service earth station can have a high cost and still be economical in terms of user costs.

Operation at Ka-Band also is characterized by severe precipitation attenuation fading due principally to local, intense rain activity characterized by thunderstorms. Since most telecommunication users require real time services of high availability, say in the range of 0.999 to 0.9999 methods for overcoming precipitation attenuation effects must be provided at the system level so as not to require two earth station antennas and connecting links e.g. space diversity. Such a solution imposes too high a cost burden on the user earth station. Similarly, fixed margins imposed on the satellite design would likewise substantially increase its cost. At Ka-Band, this also could mean difficulty in developing enough transponder power or prime power or both, in the satellite. Fortunately there is a solution for the direct access system that allows use of single antenna earth stations by dynamically balancing the available margin for both fading and non-fading carriers. This is discussed subsequently in Section 6.

Earth station availability also is impacted by equipment reliability. The postulated cost is based on a single thread design so that availability is dependent both on equipment failure rate and time to repair it. This requires careful analysis. Present experience with single purpose terminals indicates that availabilities better than 0.999 are attainable with solid state designs. Perhaps these can be improved. Probably the least reliable item is the MODEM/CODEC itself because of the large piece-part count and the complex functions performed. Possibly, MODEMS can be simplified by using noncoherent PSK; this might be acceptable if satellite charges are low. This is another example of a system method for reducing hardware complexity. Finally, it should be noted that users of terrestrial facilities normally do not duplicate their computers or processors or MODEMS or lines for the sake of higher availability and therefore may be willing to accept equipment availabilities in the postulated range. It should be noted that the situation is quite different in trunking systems or in terrestrial common facilities such as radio relays or switches. Where bundles of traffic exist, reliabilities and availabilities must be very high. In the direct access system concentration exists only in the satellite which must be very reliable from cost considerations alone.

2.4.3 MICROWAVE INTEGRATED CIRCUITS (MIC)

Microwave integrated circuits are a technology for reducing production costs. Transmission lines and reactances, device connection and matching, and certain special functions such as DC blocking, bias circuit filtering, etc. are capable of implementation in MIC format. Production costs are lowered to the point where piecepart costs dominate and circuit tolerances are capable of being precisely controlled. Low noise transistor receivers, solid state transistor or IMPATT amplifiers, frequency converters and interconnecting transmission lines, impedance transformers and isolation filters can be combined into "super components" to minimize hand assembly and test operations. Another important consideration is the frequency stability or drift and phase noise contributed by various earth station and satellite subsystems which are particularly critical in (relatively) narrow band systems such as direct access systems. In present SCPC DAMA systems satellite and down converter oscillator errors are tracked out using active pilot carrier loops. Once locked up, simple offsets are used to tune to the desired channel. While the design of the frequency system is beyond the present scope it is clear that operation at Ka Band will impose more stringent requirements on both satellites and earth station oscillators. It may be necessary to dynamically control the satellite oscillators to eliminate long term drifts, and derive earth station uplink and downlink carriers indirectly from the satellite, possibly using the common signalling channel whose pilot carrier is always present in each beam.
It is clear that the entire system design, both satellite and earth station must achieve high stability but also must be amenable to low cost earth stations. Improvements in phase or short term stability of approximately five times is believed to be within the state of the art.

2.4.4 LARGE SCALE INTEGRATION LSI

There exist today examples of CODECS, multiplexers, and frequency synthesizers which have been reduced to LSI form. Once this is accomplished, circuit production costs and testing costs are small. The MODEM itself is the most difficult subsystem to accomplish in LSI because so much of it is composed of analogue circuits--frequency changers, filters, mixers, etc. Also present narrow band MODEMS are non-standard because the market for their use is small. However, it is believed that, given the proper production incentive, and corresponding standardization, much could be done to reduce the manufacturing and test costs of these devices. Perhaps in conjunction with a standard MODEM rate such as 32Kbps for either voice or data, a multiplexer--a relatively simple low cost device-also could provide for 2.4, 4.8, 9.6 Kbps inputs. This is especially attractive if the satellite charge for transmission at buffered rates is not significant. The cost characteristics of Section 5 indicate this is the case.

For the present, using Reference (2) a compendium of earth station costs can be computed to at least indicate the range of possibilities. This report, based on work performed in 1975 and 1976 is based on vendor equipment surveys, literature research and experience factors to predict the cost of various earth station subsystems and components at different performance levels, for different production quantities and where applicable, different frequency bands. Assuming similar component and subsystem costs at Ku-Band and Ka-Band the cost of several representative earth stations can be computed. These are given in Table 2-5 for:

- A single modem 32Kbps earth station
- A multi-modem earth station with 10 32Kbps MODEMS, 2 64 Kpbs MODEMS and a 60 watt TWT HPA.
- An interactive full motion high quality color TV teleconferencing station.

Table assumptions are based on:

- Mass production and test of single design or standardized earth stations in lots of 1000,000 (1000 for TV).
- Factory test of antenna and antenna mounted equipment (microwave circuits) and of user located equipment (MODEMS, CODECS, Interface etc).
- Low cost installation by carrier-trained technicians (of relatively low-grade skill) using standardized mounting techniques, followed by turn on and checkout with Network Central. This is assumed to be similar to installation tests performed by telephone installers.

TABLE 2-5. PRICE OF REPRESENTATIVE EARTH STATIONS

	EARTH STATION #1	EARTH STATION #2	EARTH STATION #3
	QTY 100,000 SINGLE SERVICE 32 KBPS DUPLEX (AVD) .9999 A 999 DAMA 20:1	QTY 100,000 MULTI-SERVICE (COST ALLOCATED BY BANDWIDTH) 10 32 KBPS DUPLEX (AVD) 264 KBPS DUPLEX (AVD) DAMA 20:1	QTY 1000 SINGLE SERVICE TV TELECONFERENCING 40 MBPS DUPLEX .9999 A 990
ANTENNA DIAM, M	2 METERS	2 METERS	6 METERS
HPA POWER	1 WATT	60 WATTS	400 WATTS
PRICES (1980 DOLLARS)*			
HPA	$1500	$4500	$40,000
ANTENNA	1100	1100	15,300
LNR & D/C (300°K)	2000	2000	2000
UP CONVERTER	1500	1500	1500
MODEM	1000	13000	5000
DAMA/SIGNALLING	1000	2000	1000
MISCELLANEOUS (ASSY, TEST & INSTALLATION)	1500	3000	25,000**
	$9600	$27100	$89,000

*QUANTITY 100,000
**INCLUDES CAMERA AND MONITOR

- Low-cost O&M provided by automatic periodic diagnositc test of earth stations and modular (card) replacement of defective subassemblies. The modules or cards are returned to a depot for repair.
- Automatic functional control of earth stations by Network Control. The only user function is to dial his destination, he exercises no control over his terminal and does not have access to the equipment.

The above outlined concept is not new with regard to the communications and broadcasting industry but there has been little experience or incentive in the U. S. DOMSAT industry for mass production and widespread maintenance. It does appear however that the standardization requirement, and the requirement to buy in lots, and the single signalling interface with the Network Central (switching and billing center) and the problem of O&M, that the earth stations will likely be owned by the carrier and not by the user.

It is not certain at this time whether the prices listed in Table 2-5 are achievable for Ka-Band earth stations. Much work remains to be done even to identify all the functional requirements and specifications that SS-FDMA systems will impose on them. In this report, the single service station is assumed to range in price from $10,000 to $30,000 and the multipurpose earth station from $30,000 to $50,000 because it is much too premature to predict this proce accurately. As background, GE was under contract with the Government of Iran to build prototypes of an earth station that would be built in large lots at a later time -- the total production being 10,000 terminals. Each Ku-Band terminal had a two-meter antenna, 4 watt solid state HPA and a low noise GaAs Fet receiver. Modular in construction each earth station could receive up to four television broadcast cahnnels and also provide up to four SCPC/DAMA subsystems at 32 Kbps for telephony. Examination of this design and related manufacturing cost data indicates that a single MODEM earth station (no TV) would cost approximately $22,000. More intensive application of MIC and LSI technology than was then contemplated plus benefits from the advancing state-of-the-art in MIC and LSI promise to reduce these costs further. On the other hand, the availability of Ka-Band devices at production costs, or even reasonably accurate predictions concerning these are not expected to be available in the near future and this creates a degree of uncertainty and concern.

2.4.5 COMPARISON WITH TDMA

Use of conventional TDMA or SS-TDMA (as discussed previously) for direct access systems is not favorable because of high earth station costs. TDMA common equipment, for burst rates of 40 Mbps to 600 Mbps is expensive, costing in the order of \$100,000 to \$300,000 per redundant terminal. Second, the high burst rate requires high power amplifiers (TWT's or Klystrons) which with the present state of the art are of the order of 1 Kilowatt in order to provide the peak burst power. These cost in the order to \$100,000 with redundancy. Monitoring, switching (for redundancy), fault detectors, safety, backup power, etc., are examples of "propagating complexity" that add further to the station complexity, operation and maintenance problems and cost. The analysis of direct user services in Section D identifies an allowable earth station cost. If a multi-service earth station consisting of ten 32 Kbps MODEMS, (AVD) and two 64 Kbps data MODEMS for a total data rate of 448 Kbps was ever implemented via conventional TDMA technology, the earth station cost--in small quantities--will be approximately \$400,000 (1980 dollars), or approximately \$29,000 per 32 Kbps channel. The tariff comparison indicates such a terminal can be cost effective, at least for the higher data rates. However, the service is limited to the few biggest system users. An FDMA implementation will cost substantially less and therefore will compete with less traffic per earth station.

Previous studies have indicated that lower cost TDMA terminals can be built by using combinations of TDMA and FDMA. In this case, the earth station burst rate is reduced and simultaneous bursts are "stacked" in an FDMA format to fill the desired spectrum. This reduces MODEM costs and HPA costs and still maintains the simplicity of a TDMA or SS-TDMA satellite, (albeit satellite transponder efficiency now corresponds to multi-carrier FDMA). Further cost reductions can be achieved by restricting the flexibility of the earth station. However, if it is recognized that system flexibility for either FDMA and TDMA systems is advantageous and it is desirable not to restrict the DAMA capability unnecessarily it is hard to believe that the TDMA terminal cost can be less than that for an FDMA terminal based on the following considerations.

- TDMA MODEM rate will be much higher, tens of megabits per second vis a vis 50 Kbps to 100 Kbps for FDMA.
- HPA costs will still be higher at the higher burst rates, tube type amplifiers are likely to be required.
- TDMA MODEMS must be designed for bursts, e. g., fast turn on and fast lockup. The FDMA MODEM is essentially continuous except for conversational modes--lock-up times can be tens of milliseconds.
- Burst timing must be precise--of the order of microseconds, and timing caused by satellite drift around its prescribed station must be compensated. For direct access terminals this timing and correction can be provided by a master station (timing error corrections are different for different geographically located terminals).
- Use of carrier power diversity, discussed in Section F will likely not work well because there may not be enough FDMA carriers.
- Route reconfiguration requires frequency and timing changes which may effect all the earth station simultaneously.

This supports the basic precept of this study, which is at the heart of the SS-FDMA concept which is identifying the importance of any technique or concept which lowers the cost of the direct access terminal. FDMA does result in the lowest cost terminal and the results of Section 5 show that corresponding increases in satellite complexity and cost are not significant. It should be noted that the satellite costs are high in absolute terms principally because the satellite is high powered. A direct access system based on TDMA/FDMA would evidence similar costs.

2.5 SIGNALLING/SWITCHING

2.5.1 INTRODUCTION

Signalling is important for a Direct Access System because it allows, through switching, for a flexible interconnection (on demand) of the system users. Better interconnectivity, and more efficient use of the available facilities results. In the latter case, users share common facilities on demand so that the satellite can handle many more users than it could if its trunks were pre-assigned. The user benefits because he pays for the satellite trunks only when he uses them. In the context of this study, the communications signalling and switching functions are carried out principally by the SCPS/DAMA System, augmented to include the satellite on-board switching, and to control the earth station HPA power. The signalling and switching concept outlined herein is a first cut at sucn a design concept, emphasizing overall systems design, which is by no means optimized for use by a particular carrier. The design is in sufficient detail so that basic system characteristics and hardware implications can be understood.

In the conventional SCPC/DAMA signalling and switching system using centralized network control, each user communicates to network control via a common signalling channel, normally a two-way dedicated channel. Netowrk control responds to call requests by assigning complementary frequencies to the caller and called stations. This sets up the call. After completion, the complementary frequencies are available for other calls. In the conventional SCPC/DAMA system, the process of signalling and frequency assignment involves the whole generic signalling and switching process in the communications sense. That is, SCPC/DAMA is a signalling and switching system, and as far as the user is concerned, he is connected to a fully connected signalling and switching system.

Additional signalling and switching functions must be performed in a SS/FDMA system. In general, each area covered by a satellite beam must have a common signalling channel; all of these are routed to network control. In addition, the network control must control the satellite on board switch so that the RF routes or paths are adequate to meet the instantaneous traffic demand. Since the system is not constrained as in SCPC/DAMA to a single bandwidth, but can assign bandwidth on demand, the network control must have a plan by which these assignments can be made while preserving FDMA efficiency. Finally, network control must receive earth station precipitation attenuation information so that the HPA levels may be controlled through the common signalling channel.

2.5.2 SIGNALLING FUNCTIONS

Two functions of signalling are:

- Supervisor (Initiate, terminate connections and show call status)
- Address Signalling (Indicate destination of call)

In telephone systems, this was done historically by DC, e.g., simple on-off or ± signalling, later by SF (Single Frequency) signalling, basically in the on-off mode, and finally by MF (Multiple Frequency) or a series of tones in the voice band, which provides 10 digits of address and up to 6 control signals. While all signalling is initially performed with the telephone "off hook," once a thorough connection is attained no further signalling is possible, and such a system is vulnerable to interference (and fraud). Recently, the Bell System has added CCIS (Common Channel Interoffice Signalling)(1) containing stored program control in a dedicated communications channel.

Some of the signalling functions presently required are:

- Identification of sender and receiver
- Priority calls

(1) CCIS, BSTJ, November, 1960. PGS 1381-1444.

- Change routing
- Billing information (standard plus automatic, collect calls, calls billed to a third party, or credit card calls)
- Maintenance data
- International dialing, via No. 6 Signalling System
- Path continuity
- Blocking signals
- Dual verification
- Path continuity, call set-up verification

Normally, the signalling functions are completed in about two seconds. However, in some terrestrial networks, signalling can take up to 20 seconds. Satellite facilities, despite earth-satellite delays, can be equally fast because there are no tandem networks. Reliability in signalling is very important. Error correction via retransmission, requiring error correction, or forward error correction, are needed to achieve acceptable common signalling channel performance. A raw BER of 10^{-5} with a coded error rate of 10^{-8} is typical. However, some errors (presumably billing errors) require undetected signal error rates of 10^{-10}. Reliability requirements also are high, particularly in the common portions of the system. For example, earth station malfunctions, including malfunctions in the earth station signalling system, are acceptable if the station availability is typically better than .995. On the other hand, failure of the signalling/switching computers or similar devices which cause the entire system to fail are of a catastrophic nature and must be minimized. In this case, yearly outages are measured in minutes, even seconds.

2.5.3 MODIFIED SCPC/DAMA SYSTEM

Signalling System Functions and Features

The operating signalling and switching system has the following functions and features:

1. Fully variable (automatic) DAMA, including bandwidth request
2. Central control via (mini) computer, including rapid update and network expansion
3. Certain control functions shared in earth stations using microprocessor (for example earth station alarms, HPA power control, etc.)
4. Common channel signalling
5. At least one duplex signalling RF channel per beam
6. 1-second polling cycle
7. Rapid call set-up, several seconds; operator assist, where necessary
8. Logging (calls, trouble reports, etc.)
9. Supervisory traffic servicing and network maintenance functions
10. Status read out, control of each earth station
11. Earth station HPA control
12. Ancillary control of satellite switch
13. Traffic monitors
14. Centralized automatic fault diagnosis
15. Network status and traffic intensity
16. Blockage of 0.01 (typical)

Centralized Control

The heart of the centralized signalling and switching system is a fully redundant and switchable computer system controlled by software (added channels, users are accommodated quickly by software changes in user date base, e.g., number plan, call constraints, etc). This computer is connected via a common signalling channel to the controller of each earth station in the network operating in the DAMA mode. A centralized system is envisioned for a situation where earth station cost is the paramount economic tradeoff parameter.

Centralized control also is compatible with large earth station networks, Centralized DAMA is recommended because of the following considerations. Low cost includes capitalization, reconfiguration, and maintenance. Simplicity is the key. The addition of a centralized DAMA capability to an SCPC earth station is estimated to be approximately $100 except for the added cost of the signalling MODEM. The complicated computations (logging, etc.) are thus performed by the centralized network control. Any processing which could be performed centrally, but is not, must be performed in each earth station. The centralized DAMA control system offers major cost advantages in network implementation flexibility, since these are accomplished via software in the central computers. No changes to the earth stations are required. Call set-up times are not substantially penalized by central control. Set-up times of 2.5 seconds or less are achievable and these compare favorably with terrestrial systems, using rotary dial instruments. The networks themselves will be composed of subnetworks of private systems which may require isolation, e.g., it may be required that one subnetwork not be able to communicate to others. This can be controlled by the look up tables in a centralized DAMA system:

2.5.4 DESIGN CONCEPTS

The baseline system is depicted in Figure 2-12. The NCC controls all circuit switching activity via dedicated satellite channels that are terminated at the NCC and at each (DAMA) earth station with Signalling Transmission Units (STU). Signalling and supervision relating to originating trunk/users are sent to the NCC on the Common Signalling Channel (CSC). Control messages, along with signalling and supervision messages, are transmitted from the NCC to the earth stations on the common signalling channel. The CSC functions as a polling channel, polling each earth station each second or so. Since the satellite has N beams, there are N CSC's: however, all terminate at one end at the NCC.

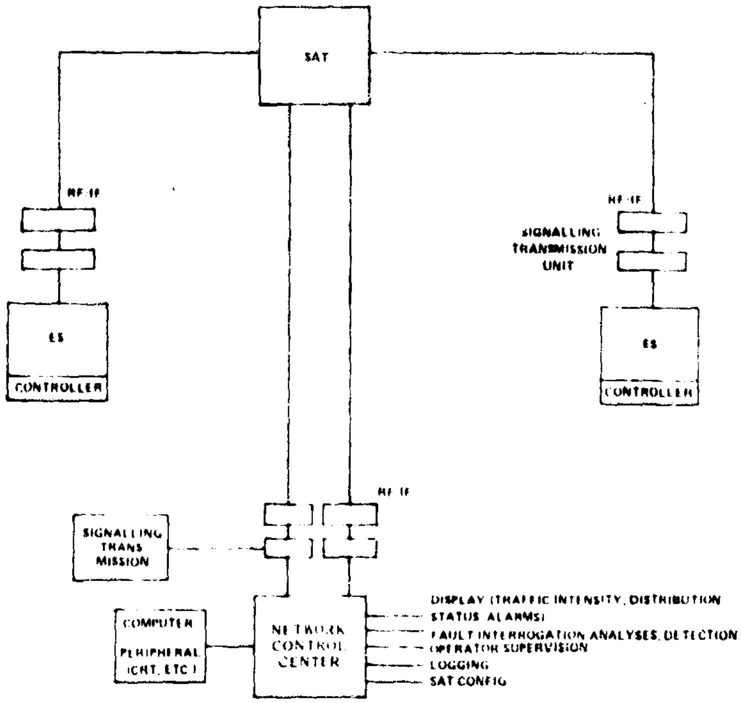


Figure 2-12. Block Diagram Concept for Centralizer Signalling and Switching System.

Consequently, all signalling, control logging, and status information is available to the NCC, which is located within one of the N satellite beams. The NCC processes information for display of important network functions. These capabilities are indicated in the block diagram of Figure 2-13, showing fully redundant, switched computers and peripherals, including a capability to interrogate particular earth stations for fault diagnoses and maintenance. The interreliability of these functions is depicted in Figure 2-14.

The earth station controller, depicted in Figure 2-15 and 2-16 controls the frequency synthesizer and the signalling display, and is controlled in turn via the common signalling channel by the NCC and by the dialer. Normally, in on hook condition, the MODEM guards the dedicated frequency of its common signalling channel in its subnetwork, with the network covered by one of "N" satellite antenna beams. Call

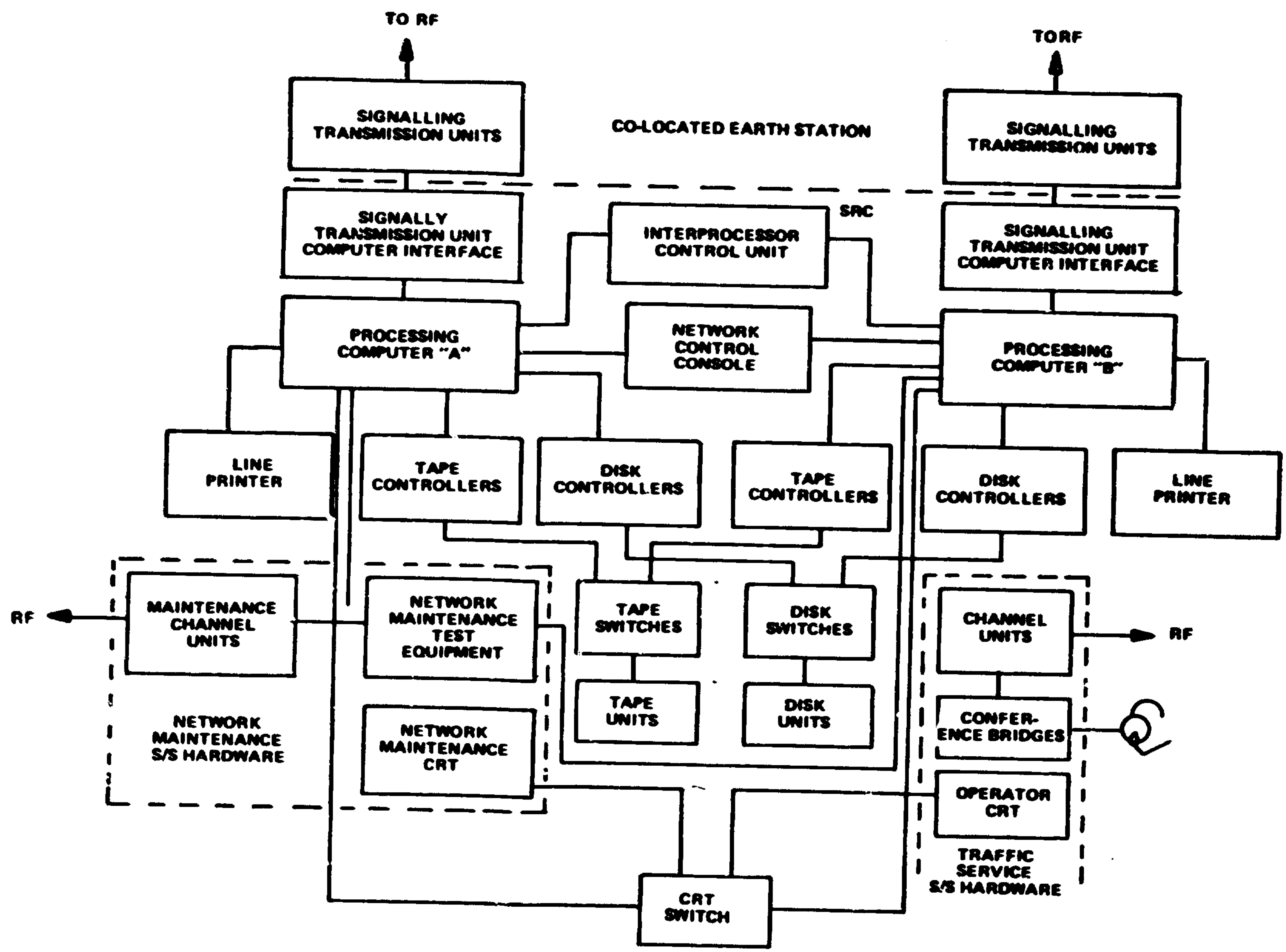


Figure 2-13. NCC Block Diagram

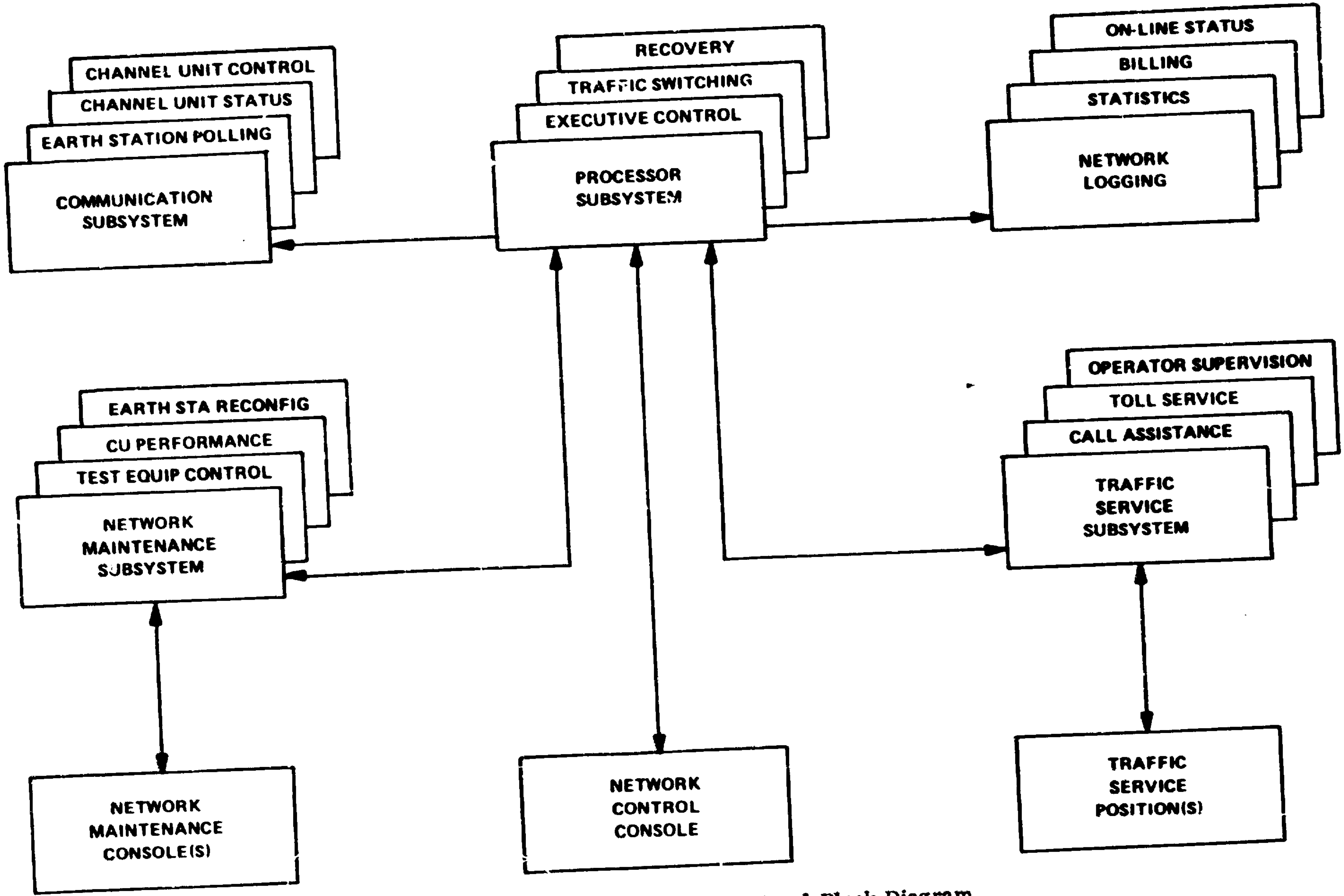


Figure 2-14. NCC Functional Block Diagram

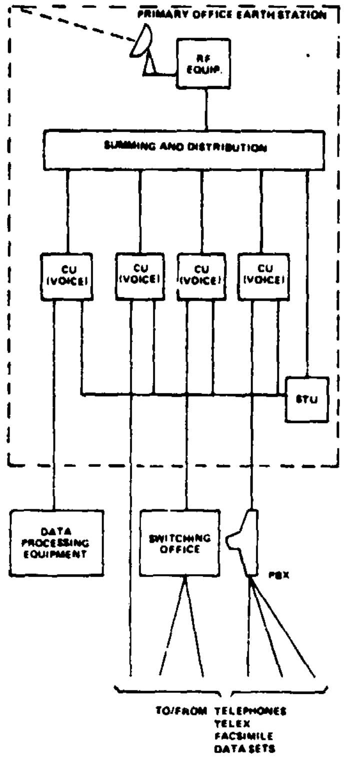


Figure 2-15. Earth Station SCPC/DAMA Log

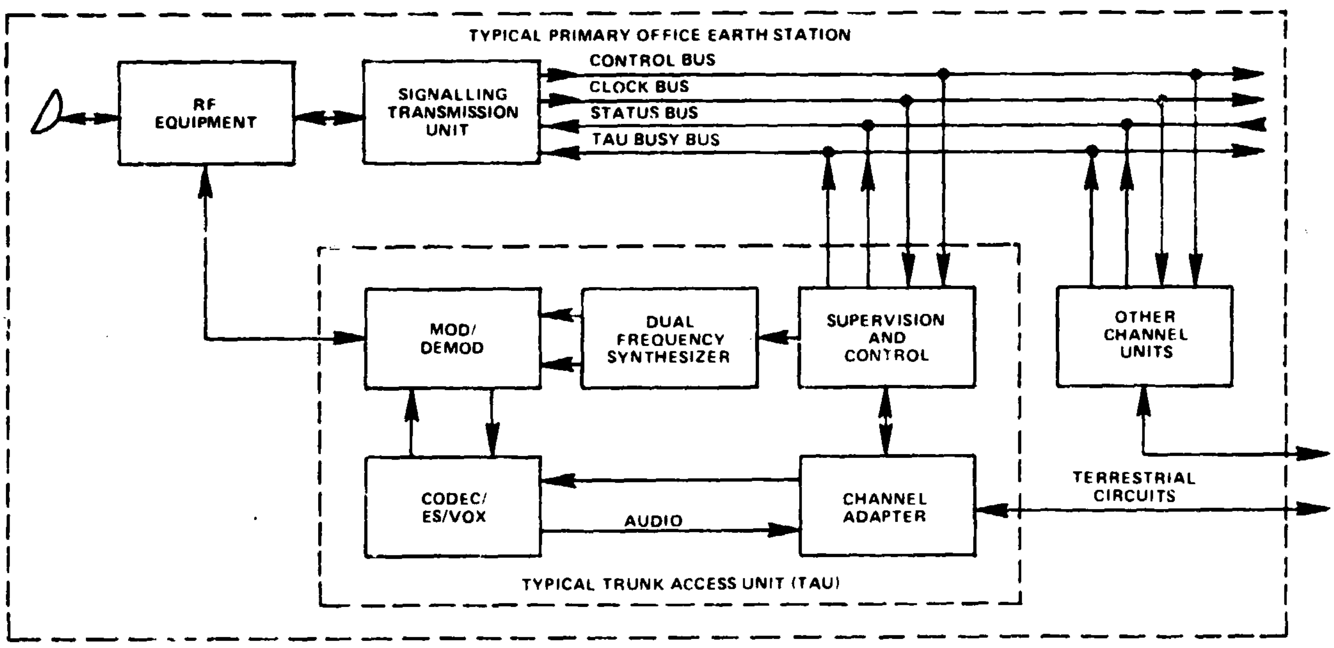


Figure 2-16. Earth Station Supervision and Control

request, set-up, and ringing are provided via the CSC, culminating in the actuation of the frequency synthesizer to tune in the assigned frequency pair (for receive and transmit) at which time the common signalling channel is disabled (until call completion). After call completion, the frequency synthesizer is reset so that the common signalling channel for that earth station is restored. An alternative, particularly for multi-MODEM earth stations, or earth stations with critical traffic, is to provide a dedicated MODEM for continuously available common signalling channel. Note that all earth stations communicate directly only with the NCC. A sequence of events might be as follows:

An off-hook signal to a channel unit results in a dial tone automatically being sent to the calling party. Off-hook signals are also sent to the NCC via the CSC. The NCC logs the number of the calling channel unit and its earth station, and then awaits the arrival of the dialing digits identifying the called earth station. The NCC checks its dynamic files to locate a compatible channel unit or MODEM (in bandwidth) at the called earth station; if available, the NCC seizes it via a control word. At the same time, a pair of frequencies are selected from a dynamic table of frequencies, and these assignements are transmitted to the channel unit controllers before the called party answers. This allows the voice path to be present while ringing and when the called party answers.

The NCC logs the off-hook at the called channel unit for traffic statistics and billing purposes. At the end of the call, the channel units receive on-hook signals from the calling or called party and transfers this condition via the CSC to the NCC. The NCC. acknowledges the on-hook status and transmits termination control words to the channel unit for release of the assigned frequencies. The channel units and operating frequencies are then removed from the "lists" of those being used.

Typically, the NCC may consist of a 16-bit word, medium size mini-computer. Core size varies with the size and traffic intensity of the network. However, core can be added in 8K word modules.

Network maintenance is provided by several facilities. Each earth station is monitored with regard to status (or health), and this information is telemetered back to the network control via the common signalling channel. Status of such items as temperature, voltage, security (tampered seals), AGC, various signal levels such as HPA power, receiver AGC, phase lock loop lock, etc. indicate the health and general dynamic state of the earth station. In addition, various telemetry points also should indicate the status of the channel units, e.g., on-hook, off-hook, etc.

2.5.5 TYPICAL CALL SEQUENCE

Table 2-6, showing a typical earth station call sequence, is divided into 5 columns: the Calling Party, Calling Channel Unit (CU), Network Control Center (NCC), Called CU, and the Called Party. The signals are assumed to originate from typical telephone dialers, however, the actual call may involve automatic dialing between computers.

TABLE 2-6. CALL SEQUENCE CHART

Calling Party/ Telephone	Calling CU/ES	NCC	Called CU/ES	Called Party/ Telephone
Off Hook →	Recognizes off-hook, start of a new call. Returns dial tone to user.			
Dial Tone ←				
Digits Dialed →	Accepts dialed digits and accumulates them until sufficient digits are received to route the call.			
	Sends status messages to NCC with ES/CU identification, digits, and off-hook indication. →	Receives status message.		
	Starts timer to send additional repeats of status messages unless an acknowledge message is received.	1. Sends acknowledge message.		
	Acknowledge message received. Stops sending status message bursts.	2. Determines call routing and number of digits required.		
Additional digits if required - - - →	Accepts additional digits if required and transfers to NCC. Same sequence as above.			
		3. Selects called ES/CU. If unavailable sends message to calling party with a recorder tone command.		
Recorder Tone ← - - -	Message received. Sends recorder tone to calling party. ← - - -			

TABLE 2-6. CALL SEQUENCE CHART (Continued)

Calling Party/ Telephone	Calling CU/ES	NCC	Called CU/ES	Called Party/ Telephone
		4. Sends seize command to called party with frequency assignments.	Message received. Seizes called party and generates ring signal. Sets frequency to assigned channel.	Seize and Ring Signal
E. Lead Seizure	Message received. Seizes E lead. Sets CU frequency synthesizer to assigned channel.	5. Sends frequencies to calling CU and seize to E lead	Starts timer for off-hook from called party. If called party does not answer before timer expires, reverts to control channel and sends message to NCC indicating "NO ANSWER".	
Receives ring-back tone from called end	VOICE PATH ESTABLISHED			
			Stops "off-hook" timer	Off-Hook
	CALL IN PROGRESS			

TABLE 2-6. CALL SEQUENCE CHART (Continued)

Calling Party/ Telephone	Calling CU/ES	NCC	Called CU/ES	Called Party/ Telephone
Hangup	Hangup received. Releases E lead. Reverts to control channel frequencies. Sends status message to NCC indicating hangup.			
		Hangup received.		
	Acknowledge message receives. Stops sending status message bursts	1. Sends acknowledge message to calling party.		
		2. Logs hangup message-ES/CU end time.		
	Ready for new call.	3. Updates files of CU usage.		
		4. Waits for called party hangup.	Hangup received. Release E lead. Revert to control channel frequencies. Sends status message to NCC indicating hangup.	Hangup
		Hangup received.		
		1. Sends acknowledge message to called party.	Acknowledge message received. Stops sending status burst message	
		2. Logs hangup message ES/CU and time.		
		3. Updates files of CU usage and frequency usage.	Ready for new call.	

2.5.6 SIGNALLING STRUCTURE

The design of a signalling system is very complex and depends heavily on the system characteristics. What will be indicated here are the salient features of such a system and some suggested solutions. Important considerations are:

- Number of satellite beams and resultant geographic area covered
- Number of earth stations per beam
- Number of channel units per earth station and mix of preassigned and DAMA units and intended use (data or voice and typical conversational durations)
- Traffic intensity of subnetwork or erlangs
- Sundry characteristics such as dialing or answering delays, number of premature hangups, number of busy's, number of no answers, etc.
- Dialing plans including routing
- Telco interface (compatible signalling)
- Operator service and assist
- Party lines, or party connections

In the KaBand systems envisioned herein, a single antenna beam can have a bandwidth of 833.3 MHz (2,500 MHz allocation with one-third frequency reuse factor), which can result in an availability of 33,000 preassigned half-trunks using 32 kBPS variable slope delta modulation and 4ϕ CPSK modulation. If these trunks were all used for telephones, the system capacity, based on 100 second calls should approach one million calls per beam during the busy hour, with blocking of approximately .01. Some of the capacity is undoubtedly preassigned, and some of this capacity is for wide band data at rates of 32 kBPS or more, and where typical call durations, such as in teleconferencing, telemail and computer interaction will be much longer than 100 seconds. The largest MODEM community is likely to be at 32 kBPS. In order to gain an understanding of the signalling requirements, suppose there are 100,000 earth stations with at least one 32 kBPS MODEM. Consequently it requires 16 bits to identify each earth station. In addition, there are 10,000 maximum possible frequency pairs or 12 bits if 833 MHz were accessible to the 32 kBPS MODEMS. The principal requirement for signalling capacity is the call set up itself. This involves 32 bits for earth station identification and 12 bits for frequency assignment plus a few bits for controls - ringing, dial tone, busy signals, etc. A corresponding amount is required for ending the call. In addition, the earth station status is polled at regular intervals to ascertain its health. This might consist of 3 bits of HPA level, 3 bits for receiver signal level (CSC carrier) and perhaps 40 bits of go-no status bits. If this is read out each hour in a routine, this is an insignificant load for the CSC (and can be prempted during the busy hours). On the other hand, a change of status due to preciptitation attenuation or system failure triggers an automatic dial up, however, this will not occur often compared to ordinary communications signalling and therefore, still represents an insignificant load to the CSC. The principal CSC load then is telephone signalling. The total dialing message setting up the call is of the order of 100 bits compared to 3,200,000 bits for a typical 100 second call. If all 100,000 earth stations were active in one hour the total signalling throughput is $\frac{100 \text{ bits} \times 100,000}{3,600}$ = 2,780 bits per second if idle time could be avoided, (which it cannot). Idle time arises because it is not useful to preassign each earth station to a CSC. With a TDMA time frame of one second, at 32 kBPS, a maximum of only 320, 100-bit messages can be transmitted each second, per CSC carrier, with, one the average 28 (100,000 per hour) earth stations per second contending for the 320 slots. The probability of being blocked from the CSC (which should be less than the probability of trunk blocking, e.g., inadequate satellite capacity), depends on the network characteristics described previously. One particular problem arises because groups of earth stations may be assigned to confirmation by the NCC, without confirmation an earth station will continue to transmit its signalling message and therefore can seize a time slot thereby blocking all other earth stations assigned to that slot. This is avoided by a strategy that allows a blocked earth station to contend for a different slot.

2.6 PRECIPTATION ATTENUATION COMPENSATION (ADAPTIVE POWER SHARING)

2.6.1 INTRODUCTION

Any system operating a Ka-band will occasionally experience deep fades due to rain attenuation. In order to maintain a high grade of service, any system design must have a large available margin, on the order of 20 to 30 dB. One approach to this problem is to design the link with a constant (fixed) 20 or 30 dB margin. This, unfortunately, requires the spacecraft transponder to handle 100 to 1000 times more power.

A more attractive approach is to design the link with a relatively small fixed margin to overcome any small, high probability fades. Then, when an occasional deep carrier fade is experienced, the affected carrier's power can be increased to overcome the fade. The increased power is accomplished through increasing the output power of the earth station providing the carrier. This remedy serves for either uplink or downlink fades.

In essence each earth station measures the level of the fixed power common signalling channel; reductions in this level are interpreted as earth station fades, and the quantized level changes are transmitted back to the centralized Network Control for action. The Network Control may respond by communicating with the appropriate earth stations (in general this is the earth station detecting the fade (uplink fade) and the earth station transmitting toward the earth station encountering a fade (downlink fade) as appropriate. The transponder, linearized to control intermodulation power, operates at constant gain.

The advantage of this scheme for FDMA lies in the large number of carriers amplified by each satellite transponder which gives a statistical advantage in that only a few of the total number of carriers can be affected by severe fading at any given time, (severe fading is caused by local thunderstorm activity). In addition, the cost impact is small for measuring the fade, signalling this information to Network Control, analyzing a proper response and signalling the appropriate earth station HPA's, because this signalling capability already exists. The principal cost impact results from providing the "excess" HPA power capability. The cost impact will vary according to the desired link availabilities, transmission bandwidth, and number of satellite beams (satellite antenna gain).

Since the satellite transponder is not ideal, any link which is in the "excited" state will increase the intermodation power experienced by any other station using the same transponder. In addition, due to finite antenna sidelobes (and cross polarization isolation) on both up and downlink, stations in nearby satellite beams reusing the same frequency as an excited station, will experience increased cochannel interference. Both of these effects will deplete the fixed margin of unexcited, unfaded stations. Note, however, that, while the fixed margin is decreased (and perhaps even entirely depleted), the signal strength seen by these unexcited, unfaded stations remains unchanged. Since only a signal fade will cause a station to boost its signals, an unexcited, unfaded station may lose all its margin to intermods and interference and remain unexcited. Improperly designed, the fixed margin of an unexcited station could be exceeded. The probability that this will happen must be made very small. We can do this by making the fixed margin sufficiently large. A properly designed margin will then be able to overcome not only the small atmospheric fades, but also maintain the link under even the worst intermodulation and interference conditions caused by other stations which are excited. It is apparent that a power balance is being achieved between unfaded and faded carriers, so that the performance of each is being maintained; large amounts of satellite transponder power is being diverted to a few faded carriers at the cost of depleting a small fixed margin applied to the majority of unfaded carriers.

2.6.2 THE WORST CASE

The worst case for this analysis does not occur when a carrier is excited due to a fade. In this case, intermodulation and cochannel interference is no longer a problem since these are now reduced 20 to 30 dB below the "boosted" signal. As long as there is sufficient signal strength in the dynamic margin to overcome the fade, the link performance will be realized. It is important, however, to assure that any unexcited, unfaded link does not become unusable due to intermodulation and interference from a number of excited links.

The peak number of excited links is determined by probability theory and fade characteristics. Given the peak number of excited stations, the fixed margin must be determined which permits any unexcited, unfaded link under the corresponding peak interference conditions. In addition, additional margin is needed above and beyond the interference margin to handle any atmospheric fades which leave the link unexcited. Interference margin is defined as the margin required to overcome intermodulation and interference 99.99% of the time and fade margin as the margin required to overcome the small atmospheric fades described above.

The total fixed margin is the sum of the interference margin plus the fade margin. It is desirable to minimize the total fixed margin in order to minimize the satellite transponder power. It is apparent that such a minimum exists. If a very small fade margin is chosen, the probability of excitation is large and the peak interference will be large (e.g., many excited carriers) requiring a large interference margin. If a large fade margin is chosen the interference margin can be small, however, the transponder power must increase in order to provide the large fade margin.

2.6.3 OPTIMUM FIXED MARGIN WITH INTERMODULATION DISTORTION ONLY

The overall quiescent (no excited station) CNR is

$$\gamma = \frac{1}{1/2.5\,\gamma_i + 1/\gamma_N} \tag{1}$$

where

γ = overall link CRN

γ_I = quiescent intermodulation ratio

γ_N = overall link signal to (thermal) noise

The weighting factor of 2.5 is the "VOX" factor, assuming the traffic actively is similar to two way voice.

As an example consider $\gamma_I = \gamma_N = 30$ dB.

Then equation 1 becomes

$$\gamma = \frac{1}{\frac{1}{2.5 \times 1000} + \frac{1}{1000}}.$$

$$\gamma = 28.54 \text{ dB}$$

Now, if several stations become excited the transponder power increases and the intermodulation level will increase correspondingly, with the cube of the increase in transponder power. Thus, if the required transponder power increases by 10%, that is $\frac{P}{P_o} = 1.1$, the new overall CNR becomes:

$$\gamma = \frac{1}{(1.1)^3 \frac{1}{2.5 \times 1000} + \frac{1}{1000}}$$

Thus, the intermodulation has increased by the factor $(1.1)^3$ due to a 10% increase in transponder power.

Next an expression is derived for the increased power due to excited stations.

$$\frac{P}{P_o} = \frac{N_T + N_A (D_o - 1)}{N_T}$$

where

N_T = total number of carriers in the transponder

N_A = number of carriers in the excited state

D_o = dynamic margin

The dynamic margin, D_o, is the amount of signal is boosted when a fade greater than the fade margin is experienced. For this simple analysis, a dynamic margin of 20 dB is assumed. (In an operational system several different values of D_o might be more optimum.)

For example, if there are 12 excited carriers (levels increased by 20 dB) out of 1000 carriers using a particular transponder, then

$$\frac{P}{P_o} = \frac{1000 + 12 \cdot (100-1)}{1000}$$

$$\frac{P}{P_o} = 2.19$$

The transponder will be called upon to deliver a little more than twice it's quiescent power.

The equation for overall CNR becomes:

$$\gamma = \frac{1}{\left(\frac{P}{P_o}\right)^3 \frac{1}{2.5\,\gamma_I} + \frac{1}{\gamma_N}}$$

or:

$$\gamma = \frac{1}{\left(\frac{N_T + N_A (D_o - 1)}{N_T}\right)^3 \frac{1}{2.5\,\gamma_I} + \frac{1}{\gamma_N}} \qquad (2)$$

and in the example:

$$\gamma = \frac{1}{(2.19)^3 \frac{1}{2.5 \times 1000} + \frac{1}{1000}} = 22.84 \text{ dB}$$

Thus, if no carriers are excited ($N_A = 0$) we have the quiescent CNR of 28.54 dB. If 12 carriers out of 1000 have increased levels by 20 dB, the intermodulation noise will have increased, reducing the overall CNR to 22.84 dB.

Equation 2 can be plotted, showing the relationships between overall CNR (γ), the number of excited stations (N_A) for a given total number of stations (N_T), a given quiescent thermodulation ratio (γ_I) and a given quiescent signal to thermal noise ratio (γ_N). Such a plot is shown in Figure 2-17. γ_N can be varied to generate a family of curves. Notice that as the number of excited stations (out of 1000) increases, the interference increases, and the overall CNR (γ) decreases.

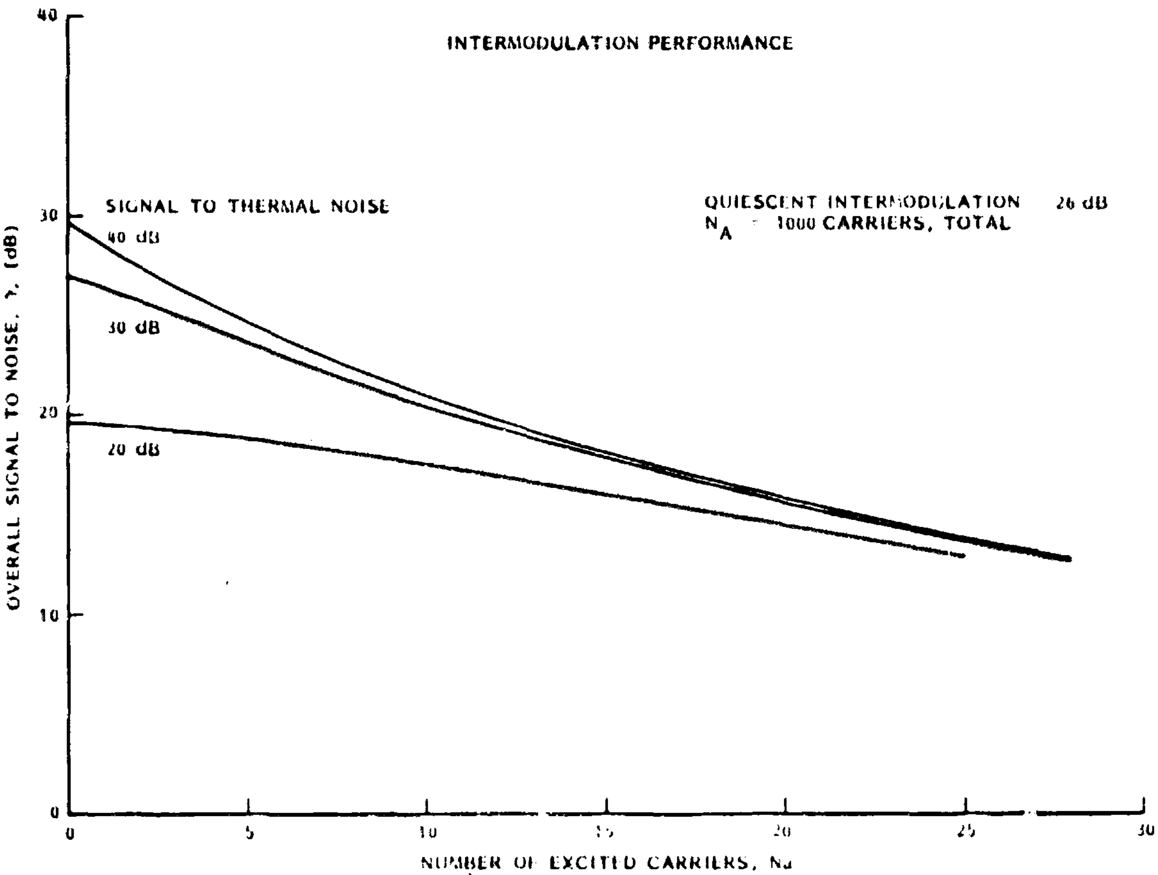


Figure 2-17. Overall Signal to Noise Ratio vs. the Number of Excited Carriers

Figure 2-17 is plotted for a quiescent intermodulation level of 26 dB. Including the VOX factor of 4 dB, this means that the overall CNR can not exceed 30 dB.

Using Figure 2-17, the minimum signal to thermal noise ratio can be chosen considering the interference and fade margins. To do this, the peak interference and hence the required interference margin must be determined. These are related directly to the maximum number of excited carriers which depends on the probability of any one station being excited, considering a given fade margin. Recall that a carrier will become excited if its signal strength fades by more than the fade margin. Thus, the probability that a carrier is excited equals the probability of a fade equal to or greater than the fade margin.

Rain fade statistics for our range of interest (0 - 5 dB) are not readily available, therefore, the probability distribution listed in Table 2-7 is used. For example, the probability that a rain fade will equal or exceed 2.0 dB is 0.361%. Thus, if we chose a fixed margin of 2.0 dB, the probability that any given carrier is excited is 0.361 %. If we assume that each carrier's excitation is independent of all other carriers, the total number of excited carriers is a sum of independent Bernoulli trials. A sum of independent Bernoulli trials generates a binomial distribution. Next the 99.99% level (or better) of the binomial distribution is evaluated. For example, from Table 2-7, a fade of at least 2.0 dB occurs with probability 0.361%. If there are 100 potentially excited carriers in the transponder, then, (from the third column of Table 2-7), the maximum number of excited carriers (with 99.99% confidence) is four (e.g., the probability of more than four excited carriers, which will exceed the interference margin, is less than 0.01%). Thus, given a 2.0 dB fade margin and 100 carriers, it is prudent to plan for four excited carriers, e.g., the link interference margin must be sufficient for $N_A = 4$. If there are 1000 carriers, then only 12 excited carriers should be anticipated (again from Table 2-7) and the link interference margin must be sufficient to handle 12 excited carriers. A base assumption is that carrier excitations are independent. This may not be strictly true for small fades. If the assumption is incorrect, then the peak number of excited carriers will be larger resulting in a larger fade margin. Also note that in the probabilities listed in Table 2-7, the probability of a fade 0.0 dB or more should actually be 100%, not 1.2%. Fortunately, this has no practical impact on the calculations.

TABLE 2-7. RAIN FADE STATISTICS USED IN THE ANALYSIS, THE PROBABILITY OF A FADE EQUALS THE PROBABILITY ANY SINGLE STATION IS EXCITED

Fade Magnitude (dB)	Prob. of Fade (%)	Num. of Excited Links (99.99% level) N_t=100	N_t=1000
0.0	1.200	7	27
0.2	1.064	7	25
0.4	0.944	6	23
0.6	0.837	6	21
0.8	0.743	6	19
1.0	0.659	5	18
1.2	0.589	5	17
1.4	0.518	5	15
1.6	0.459	5	14
1.8	0.408	4	13
2.0	0.361	4	12
2.2	0.321	4	12
2.4	0.284	4	11
2.6	0.252	4	10
2.8	0.224	3	10
3.0	0.198	3	9
3.2	0.176	3	8
3.4	0.156	3	8
3.6	0.138	3	7
3.8	0.123	3	7
4.2	0.097	3	6
4.6	0.076	2	6
5.0	0.060	2	5

Finally, when a fade exceeds the fade margin, two carriers actually become excited, the carrier of the station experiencing the fade (uplink fade), and (assuming a duplex link) the carrier of the station communicating with the faded station, (downlink fade). Since the full dynamic margin of the excited, faded station is not experienced by the spacecraft transponder, (assume that the fade is more or less compensated by the increased carrier level), only the excited, unfaded station situation is considered. For full accuracy, some allowance should be made for the additional intermodulation generated by the excited, faded station. For simplicity, only excited, unfaded stations are included in this analysis.

It is assumed that a minimum CNR = 13 dB is sufficient (4ϕCPSK). If the CNR falls below 13 dB an outage has occurred. In many cases the signal is still usable so this is an over simplification.

In the above example, if the fade margin is 2.0 dB, then the overall CNR (γ) must be at least 15 dB, even when there are up to 12 excited carriers (1000 carriers per transponder).

Referring back to Figure 2-17, for N_A = 12 excited carriers, and 20 dB signal to thermal noise ratio (γ_N), the overall CNR (γ) is 16.89 dB. This exceeds our requirement of 15 dB (13 dB + 2.0 dB fade margin). Alternatively, rewriting equation 2,

$$\gamma_N = \frac{1}{\frac{1}{\gamma} - \frac{N_T + N_A (D_o - 1)^3}{N_T} \frac{1}{2.5\,\gamma_I}} \tag{3}$$

γ_N may be found directly.

If the fade margin is 2.0 dB, then the overall CNR is 13 + 2 = 15 dB, under the worst case interference. With a fade margin of 2 dB and 1000 carriers per transponder, the worst case is with N_A = 12 (from Table 2-7) excited carriers. Letting the quiescent intermodulation ratio γ_I = 26 dB, the dynamic margin D_o = 20 dB (that is, an excited carrier will have 100 times more power than an unexcited carrier) then:

$$\gamma_N = \frac{1}{\frac{1}{31.62} - \left(\frac{1000 + 12.99}{1000}\right)^3 \frac{1}{2.5 \times 400}} \tag{4}$$

$$\gamma_N = 16.75 \text{ dB}$$

In Figure 2-17 for γ_N = 16.75 dB and N_A = 12 the overall link CNR (including intermodulation) is exactly 15 dB. Thus, given a quiescent intermodulation ratio of 26 dB an overall CNR of 15 dB can be met for 99.99% of the time.

The signal to thermal noise (as calculated above for a fade margin of 2 dB) versus fade margin is plotted in Figure 2-18 for quiescent intermodulation levels γ_I = 26 dB, a dynamic margin of 20 dB and 1000 total (N_T) carriers. The optimum fade margin can then be chosen.

From Figure 2-18 a fade margin of 2 dB is optimum although it is apparent that the choice is not critical. When a fade margin of 2 dB is specified, it means that any time a carrier fades by 2 dB or more its level will be increased by 20 dB. Further, from Figure 2-18 the unexcited carrier will have at least that 2 dB fade margin 99.99% of the time, even under peak interference conditions if we provide the link with a 16.75 dB signal to thermal noise ratio.

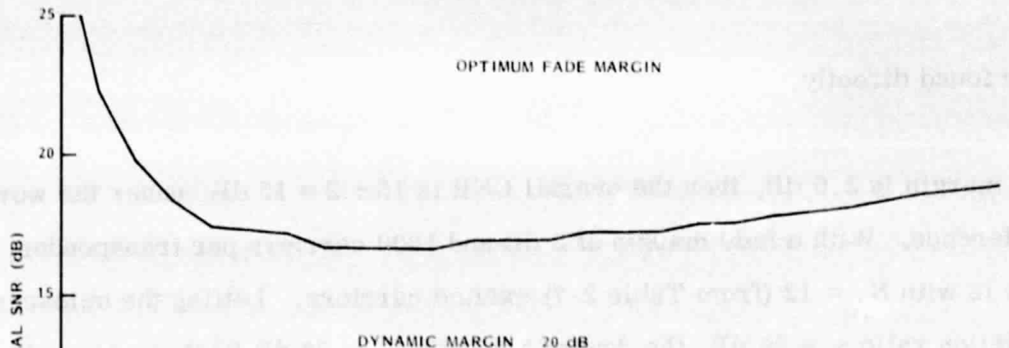


Figure 2-18. Thermal Carrier to Noise Ratio vs. Fade Margin

In the next section the effects of cochannel (antenna) interference will be added.

2.6.4 OPTIMUM FIXED MARGIN AND COCHANNEL INTERFERENCE

In the previous section the maximum interference generated by increased intermodulation distortion from excited carriers was considered. A fade margin was specified which determines when a carrier becomes excited. Also a thermal CNR was specified such that a fade margin exists even with maximum interference from intermodulation distortion.

The same algorithm used in the preceeding section for intermodulation distortion can also include the effects of cochannel (antenna) interference.

This interference is from two sources, unwanted reception from antenna sidelobes of the same polarization and unwanted reception due to loss of polarization isolation (perhaps induced by a rain fade). The first case is examined.

Cochannel antenna sidelobe interference occurs on both up and downlinks. With all stations in the quiescent state, there will be a quiescent level of cochannel interference, γ_C. This additional interference must be included in the previous expression (Equation 2) for overall link CNR.

For cochannel interference, there are a limited number of potentially disruptive interferers. It is unlikely that any of them will be excited. However, it is wise to design the link to handle at least one excited cochannel link. Using a 20 dB dynamic margin as before and assuming one of six cochannel stations is excited, the increase in cochannel interference over the quiescent state is, by inspection

$$\frac{105}{6} = 17.5 \text{ times more cochannel interference}$$

If it is assumed (worst case) that one cochannel interferer is always excited, then

$$\gamma = \frac{1}{\left(\frac{P}{P_o}\right)^3 \frac{1}{\gamma_I} + \frac{1}{\gamma_N} + 17.5 \left(\frac{1}{\gamma_{CU}} + \frac{1}{\gamma_{CD}}\right)} \tag{5}$$

where there is one excited cochannel interferer on both up and downlink. Combined with the worst case P/P_o (from the previous section), Equation 5 can be used to determine the optimum fade margin and the required thermal CNR to achieve that fade margin under peak interference conditions. Before doing that however, Equation 5 can be further modified to include the effects of degradation in polarization isolation. Since this effect can be severe at Ka-Band, it is assumed that each antenna beam is singularly polarized, then the degraded isolation is the sum of the degraded polarization isolation and the antenna sidelobe isolation.

A detailed investigation of this loss of isolation is beyond the scope of this analysis, consequently Equation 5 is modified to include γ_{PU} and γ_{PD}, the up and downlink minimum polarization isolation actually experienced, (not quiescent isolation). Then fade margin versus required thermal signal to noise for several values of polarization isolation can be examined to ascertain design objectives. Equation 5 now becomes:

$$\gamma = \frac{1}{\left(\frac{P}{P_o}\right)^3 \frac{1}{2.5\,\gamma_I} + \frac{1}{\gamma_n} + 17.5 \left(\frac{1}{\gamma_{CU}} + \frac{1}{\gamma_{CD}}\right) + \frac{1}{\gamma_{PU}} + \frac{1}{\gamma_{PD}}} \tag{6}$$

As an example of Equation 6 let

γ_I = 26 dB

γ_N = 20 dB

γ_{CU} = γ_{CD} = 30 dB

γ_{PU} = 26 dB

γ_{PD} = 30 dB

$\frac{P}{P_o}$ = 2, and

γ = 12.48 dB

With the above conditions a thermal CNR (γ_N) of 20 dB provides an overall link CNR (including intermodulation and interference) of only 12.48 dB. In this case interference has depleted the entire CNR leaving no fade margin.

It is more direct to solve Equation 6 for γ_N and plot it as a function of fade margin directly; therefore

$$\gamma_N = \frac{1}{\frac{1}{\gamma} - \left\{ \left(\frac{P}{P_o}\right)^3 \frac{1}{2.5\,\gamma_I} + 17.5\left(\frac{1}{\gamma_{CU}} + \frac{1}{\gamma_{CD}}\right) + \frac{1}{\gamma_{PD}} \right\}} \qquad (7)$$

Equation 7 is considered in the next section.

2.6.5 OPTIMUM FADE MARGIN FOR VARIOUS SYSTEM CONFIGURATIONS

Given various system parameters (dynamic margin, quiescent intermodulation ratio, quiescent cochannel interference and actual polarization isolation) an optimum fade margin and thermal CNR can be specified to satisfy availability requirements under worst case interference. The optimum fade margin is defined as the fade margin requiring the smallest thermal CNR, (by definition, whenever the signal fades by the fade margin, the carrier becomes excited).

First a set of standard link and transponder parameters will be defined. The optimum fade margin will then be determined for the standard link. Then the standard link parameters will be varied to determine parameter sensitivities.

The standard link is defined as:

1. Quiescent intermodulation distortion ratio (γ_I) = 30 dB.
2. Quiescent up and downlink cochannel interference ratio = 30 dB.
3. Dynamic margin = 20 dB.
4. Number of excited cochannel links, 1 each, up and downlink.
5. Actual polarization isolation ratio, up and downlink, = 30 dB, each.
6. Interference level chosen such that it will not exceed 99.99% of the time.

Item 4 above requires the standard link to handle 2 simultaneous excited cochannel carriers. Fortunately, two excited cochannel carriers and worst case intermodulation is an unlikely occurrance. For example, the model predicts that there is a probability of 0.459% that any given carrier is excited when we specify a fade margin of 1.6 dB. If there are twelve significant potential cochannel carriers (that is, six nearby antenna beam cells fall in, say, the first antenna sidelobe on uplink and another six on downlink), the binomial distribution predicts that no cochannel carriers will be excited with probability 94.63%. One cochannel carrier will be excited with probability 5.24% and two cochannel carriers will be excited with probability of 0.13%. Thus, two cochannel carriers excited and worst case (99.99%) intermodulation is extremely unlikely. Therefore, if the thermal CNR is sufficient for at least two excited cochannel carriers and worst case intermodulations simultaneously and still maintain a fade margin (with high probability), then the availability corresponding, to the 20 dB dynamic margin is achievable.

Figures 2-19 through 2-26 were generated using Equation 7 and the above parameters. These figures are identical to Figure 2-18 except cochannel interference and polarization isolation are included in addition to intermodulation distortion. In all figures, the solid line represents the standard link described above. The dashed lines are variations made in the standard link to determine parameter sensitivity.

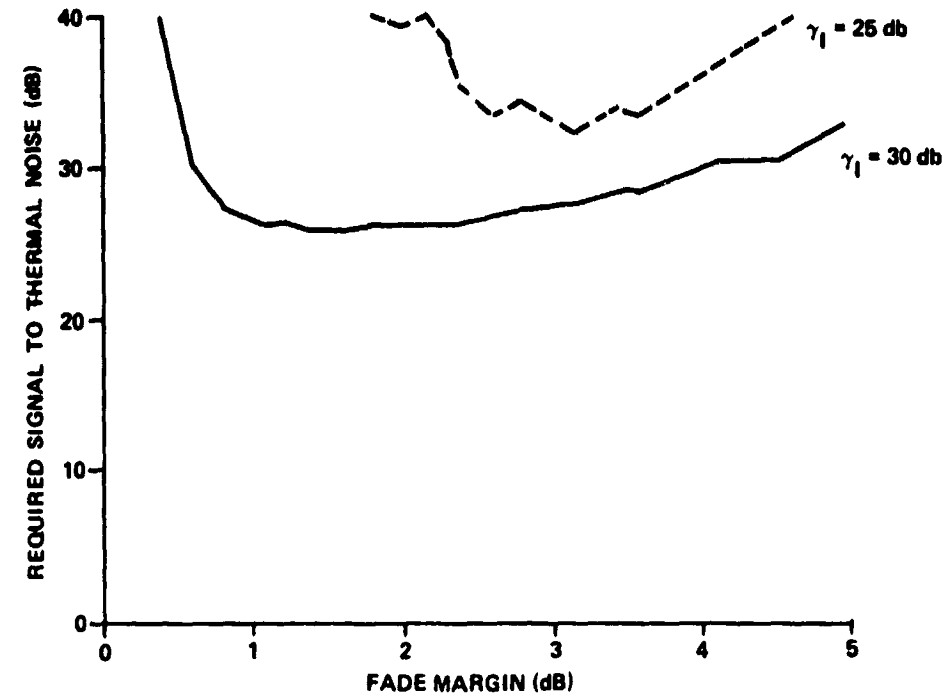


Figure 2-19. Thermal CNR vs Fade Margin

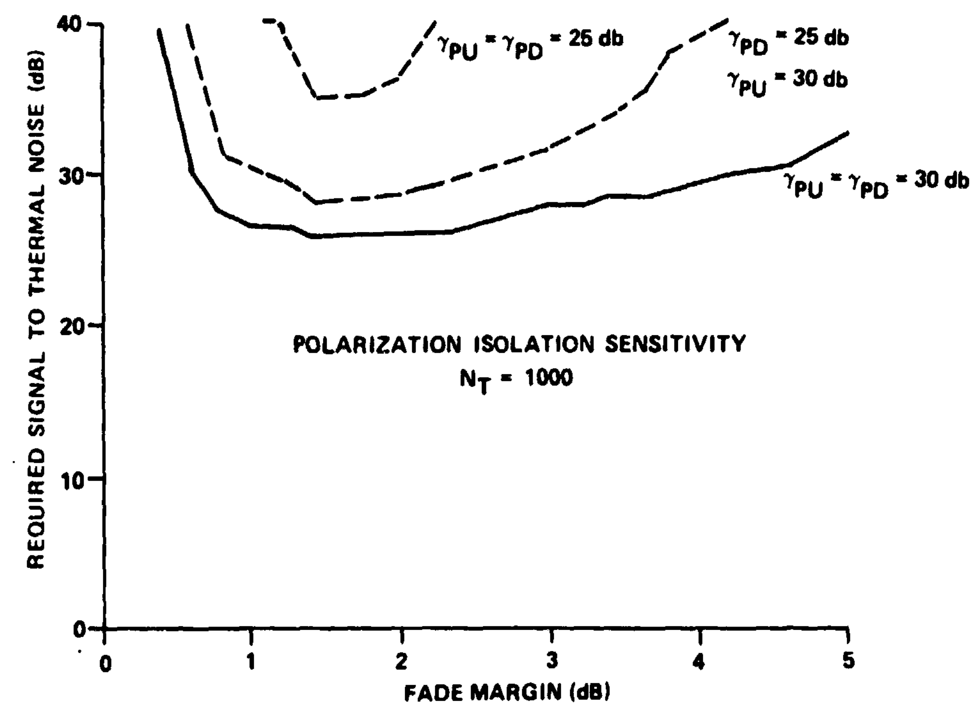


Figure 2-20. Thermal CNR vs Fade Margin

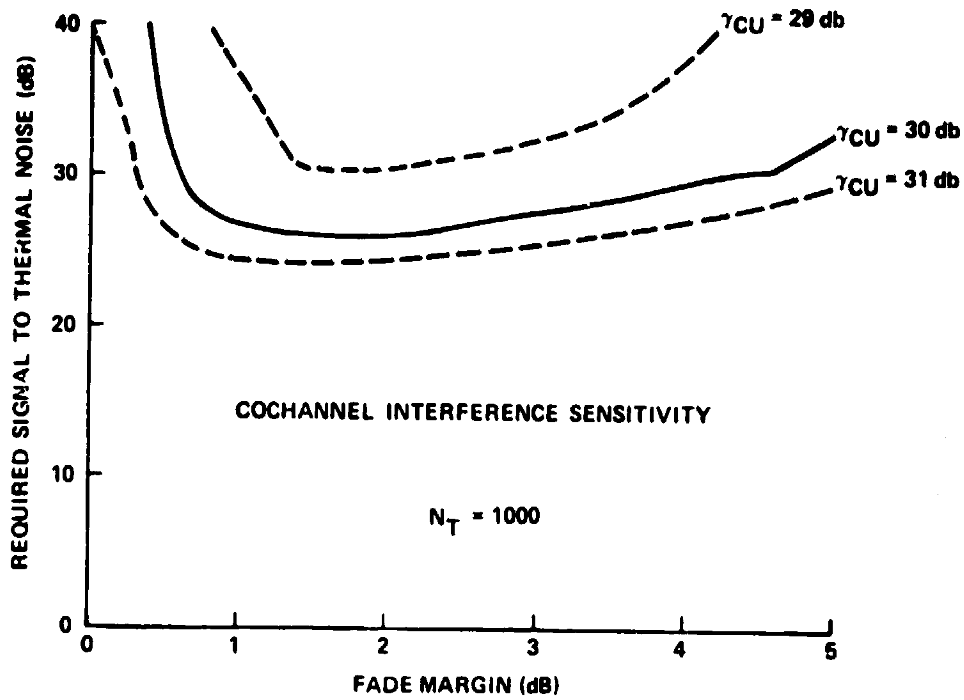


Figure 2-21. Thermal CNR vs Fade Margin

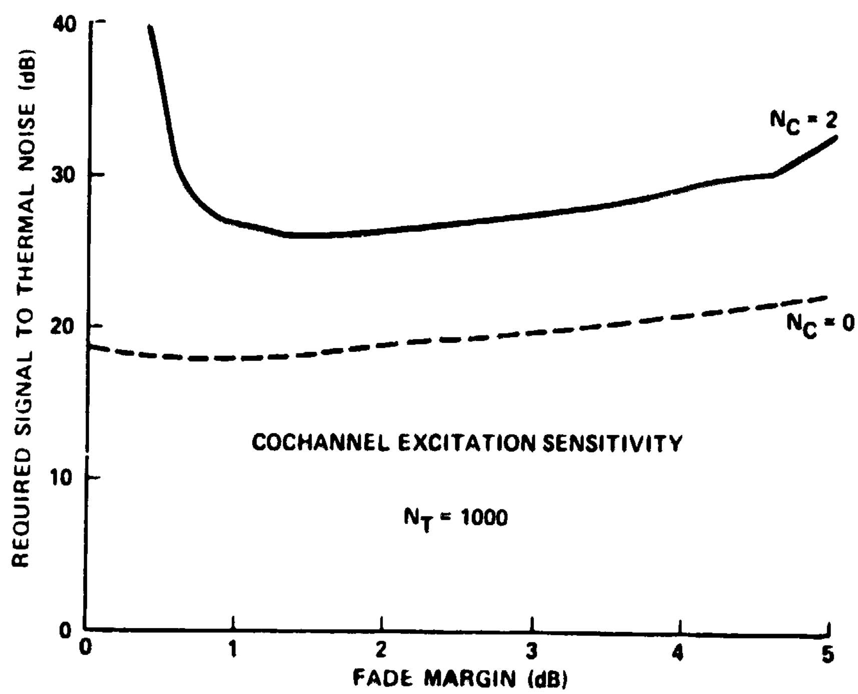


Figure 2-22. Thermal CNR vs Fade Margin

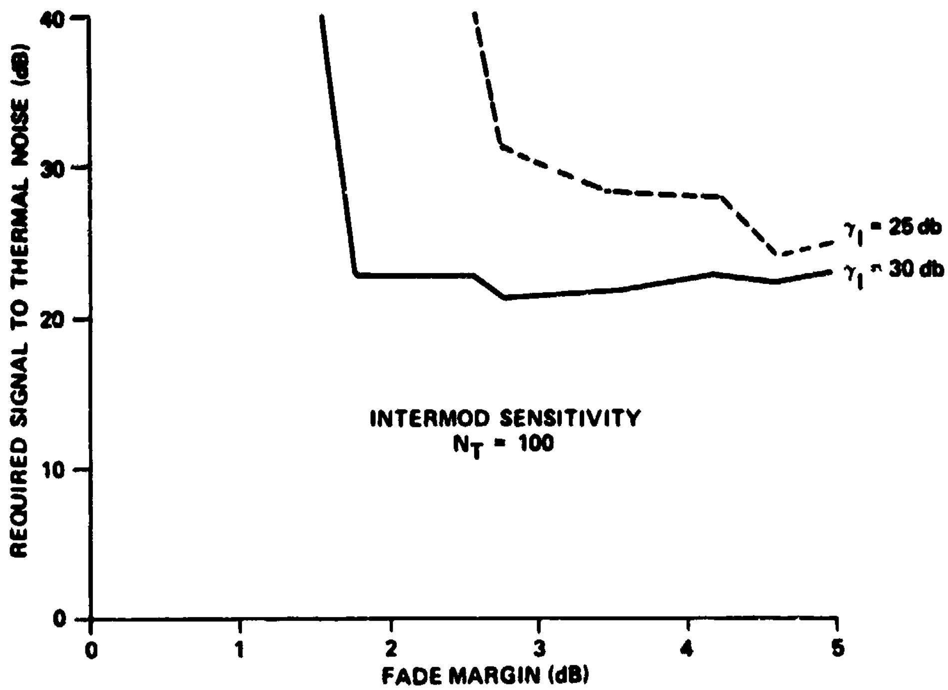


Figure 2-23. Thermal CNR vs Fade Margin

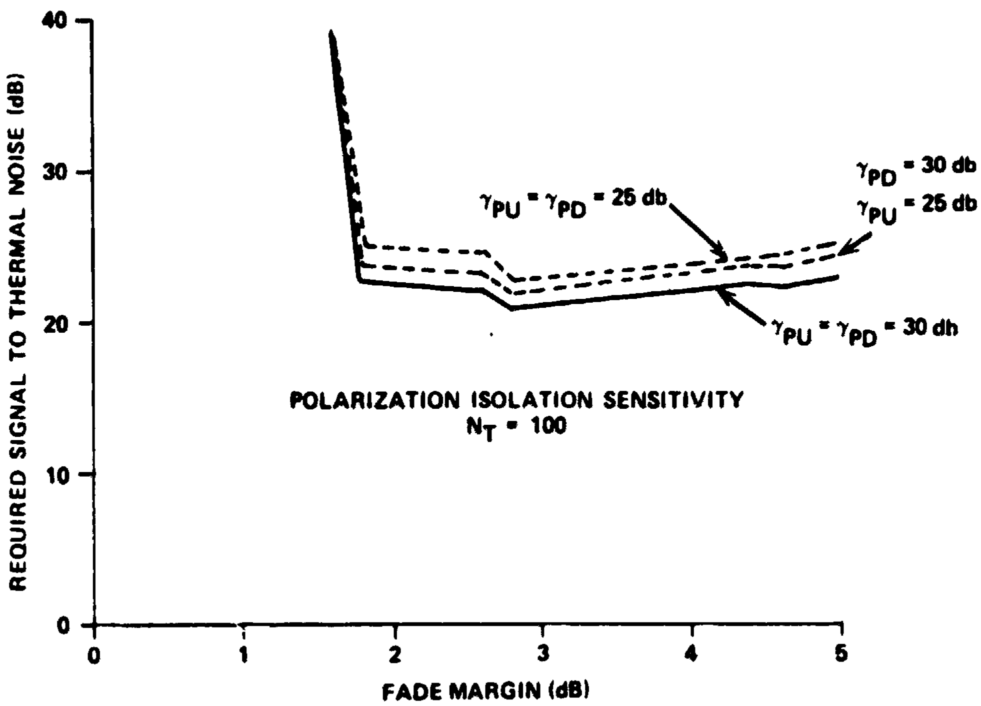


Figure 2-24. Thermal Intermodulation vs Fade Margin

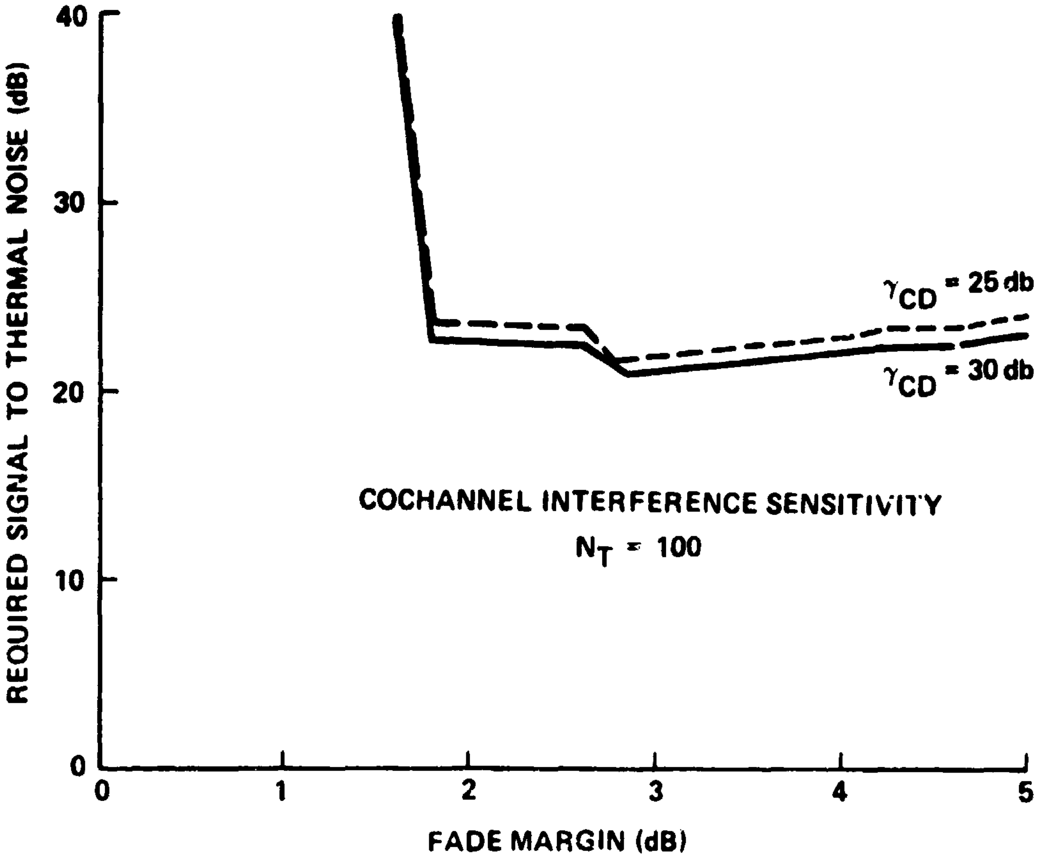


Figure 2-25. Thermal CNR vs Fade Margin

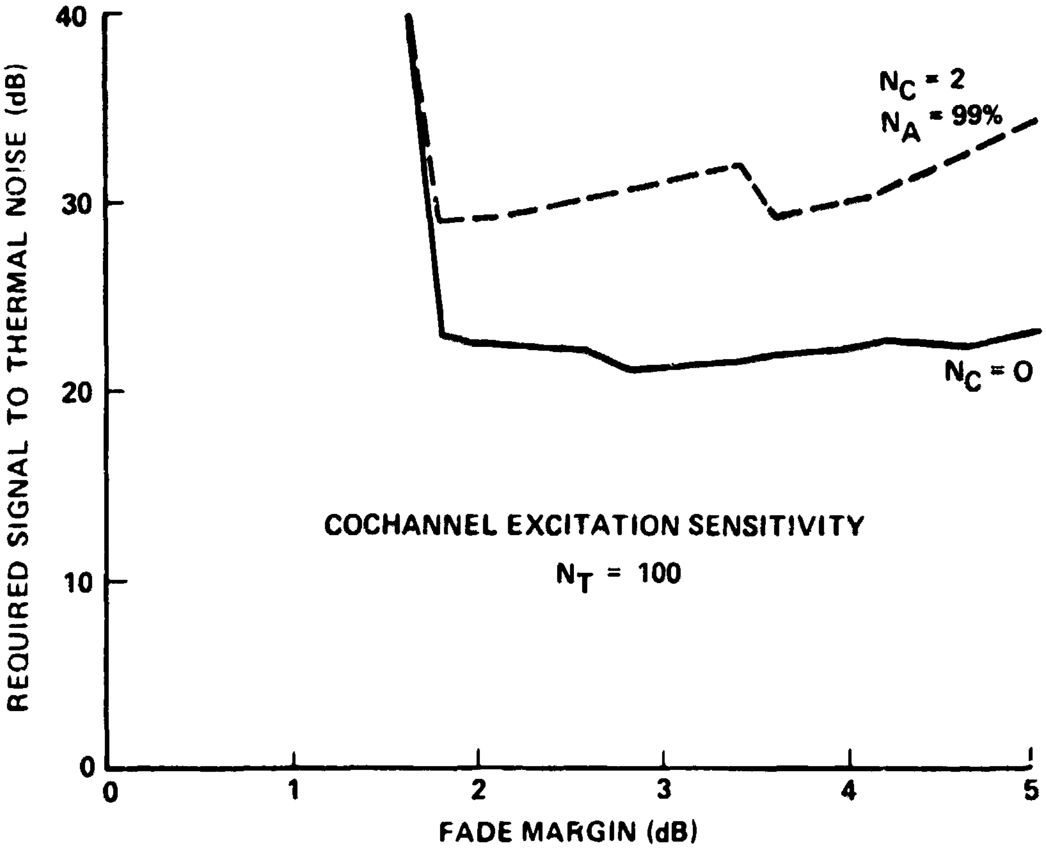


Figure 2-26. Thermal CNR vs Fade Margin

In Figure 2-20 the optimum fade margin for the standard link and 1000 carriers is 1.6 dB. This fade margin can be realized under maximum interference with a thermal CNR of 26.06 dB.

The dashed curve in Figure 2-19 illustrates the effect of reducing the intermodulation radio from 30 dB to 25 dB. The link plus fade margin is still realized under the maximum interference conditions. However, the thermal CNR must be increased in order to overcome the poorer transponder performance; specifically, $\gamma_N = 32.68$ dB with a fade margin of 3.2 dB. If smaller fade margins are chosen, the maximum number of excited carriers increases, thus increasing interference. If a sufficiently small fade margin is chosen, the maximum interference is so great, the link (plus fade margin) cannot be realized with maximum interference, regardless of thermal CNR.

Figure 2-20 also depicts the effect of changing polarization isolation. The standard link has 30 dB of polarization isolation (degraded polarization isolation plus sidelobe isolation) on both up and downlink. That is, the ratio of signal to the sum of all interference due to imperfect polarization isolation is 30 dB. The standard link, as before, is indicated with the solid line. If the polarization isolation is 25 dB on up (or down) link and 30 dB on down (or up) link, the thermal CNR must be increased to account for the poorer isolation. In this case the minimum required thermal CNR is 28.54 dB and the fade margin is 1.6 dB.

If the system has even less isolation, say 25 dB on both up and downlink, Figure 2-20 indicates the required thermal CNR is 35 dB. The optimum fade margin is still 1.6 dB. Clearly it is becoming more and more difficult to compensate for poorer isolation by simply increasing the thermal CNR.

Figure 2-21 indicates the sensitivity to intermodulation distortion levels which effect satellite transponder linearization considerations.

Figure 2-22 depicts cochannel excitation sensitivity. The standard link (solid line) assumes the unlikely situation that two cochannel carriers are excited causing greatly increased interference via the antenna sidelobes. Since there is no cochannel link excitation most of the time, a curve for no cochannel station excited (i.e., the factor of 17.5 in Equation 7 is set equal to 1) is plotted. Of course, the link interference margin should not be determined assuming that there are no excited cochannel stations.

The previous calculations are repeated for a 100 carrier transponder. The 100 carrier transponder with quiescent intermodulation of 30 dB cannot realize the link plus fade margin with worst case intermodulation and two simultaneous excited cochannel carriers. For this reason the standard link was changed to allow no excited cochannel carriers. The effect of excited cochannel carriers will be treated separately.

In figure 2-23 the solid line represents the standard link (with no excited cochannel links), with a quiescent intermodulation ratio of 30 dB. The standard link plus 2.8 dB of fade margin can be realized under worst case interference if γ_N = 21.05 dB.

If intermodulation is increased to 25 dB then γ_N = 24.29 dB in order to overcome increased worst case intermodulation. The fade margin must be increased to 4.6 dB to keep the peak number of excited carriers to a minimum.

Figure 2-24 compares the standard link (30 dB polarization isolation) with other degraded polarization isolations, as before. The system is nearly independent of polarization isolation, due to the lack of cochannel excitation stress.

Figure 2-25 depicts cochannel interference sensitivity. Only a small effect occurs if the total cochannel interference increases by 5 dB, since there are no excited cochannel links.

In Figure 2-26 the cochannel interference is increased by allowing two excited cochannel carriers and a level of intermodulation (determined by the number of excited stations using the same cell) which will not be exceeded 99% of the time (rather than

99.99%). Since the probability of two excited cochannel stations is about 0.13% the probability that this and the 99% level intermods are exceeded (probability = 1%) is on the order of 0.01%. Alternatively, under stressed conditions the potential cochannel interference can be blocked via the DAMA system (e.g., not assigning the cochannel interference slot). This will have a negligible effect on the system blockage even during the busy hour.

Figure 2-26 illustrates sensitivity of the link to excited cochannel stations. The solid line is again the standard link and the dashed line is the required thermal CNR under the worst case interference conditions described above. This curve has two minimums (due to the discrete nature of the binomial distribution), the minimum at a fade margin of 3.6 dB is chosen since the minimum at 1.8 dB is too close to an unstable region. From Figure 2-26 the fade margin of 3.6 dB is realized with γ_N = 29 dB under the maximum interference conditions described above.

From the preceeding discussion, it should be clear that a 100 carrier transponder will not work as well as a 1000 carrier transponder* (especially quiescent intermod ratio). The reason for this is that a greater percentage of stations will be excited in the 100 channel transponder than in the 1000 channel transponder under peak interference. Recall that the binomial has a mean = Np and a variance = Npq. Reducing the total number of stations by a factor of ten will cause the mean (Np) to decrease by ten, but the standard deviation ($\sqrt{Npg}$) will decrease by a factor of only $\sqrt{10}$. Thus, percentagewise, the number of excited carriers in a 1000 channel transponder will be a factor of $\sqrt{10}$ better (less) than the 100 channel transponder. Therefore, the adaptive link excells with high capacity transponders and becomes more difficult to reliably implement with lower capacity transponders.

* The average transponder capacity described herein for the 75,000 trunk case is 15,000 carriers.

2.6.6 ANCILLARY CONSIDERATIONS

Since the FDMA power sharing technique is complex and difficult to analyze certain simplifying assumptions were made. In particular, with regard to cochannel interference no attempt was made to take advantage of interstitial carrier spacing. In addition, the satellite beam bandwidth was assumed to be filled; this occurs only during the busy hour in saturated beams. Since the intermodulation maximum occurs infrequently (e.g., a maximum of excited carriers), these events do not necessarily occur during the busy hour so that potential cochannel interferors can simply be avoided by assigning different channel frequencies. Also, the 20 dB dynamic fade is exceeded at 20 GHz for only 0.01% of the time in the Middle Atlantic States. In a DAMA type FDMA system undergoing exponential capacity growth,reserve capacity is available except during the busy hour of the last system operating year. That is, the situation described herein is that of the busy hour during the last year of system operation. In all preceeding years the availability will be significantly better. Even if capacity is constant, the FDMA characteristics are such that maximum power demand, due to fading does not necessarily correspond to the busy hour. Thus, the 20 dB value at 20 GHz is not 0.01% but more like 0.003%, (the probability of a fade occurring during a busy hour period of, say, 6 hours is 1/3).

It is apparent from these considerations that significant further optimizations can be accomplished.

Since it is desired to minimize the fade margin it is important to identify the rainfall statistics in the 1 to 5 dB range to determine the influence of wide area rainfall which could "trigger" the network into a state having too many excited carriers. While this, in any case, will increase the fade margin, network overreaction can be avoided because the Network Control (by receiving the carrier levels measured at each of the earth stations) can ascertain the nature of the rain characteristics and can consequently enable a "measured response" to the situation.

2.6.7 STABILITY CONSIDERATIONS

The system as described possesses an inherent instability. One source is called cochannel interference instability and the other is called intermodulation distortion instability.

Cochannel Interference Instability

Cochannel interference instability is the first to appear and is the more benign of the two. This will occur when several links on any given channel are excited. If the resulting cochannel interference is sufficient to use all the fixed margin of any of the remaining unexcited cochannel stations, those stations will also become excited even though they are unfaded. Once these stations are excited, the cochannel interference will increase to the point where every station using that channel in every cell will become excited. However, the channel will still be usable as most (but not all) links will still realize acceptable signal to cochannel interference ratios. The dangerous aspects of cochannel interference instability is that it contributes to another kind of instability.

Intermodulation Distortion Instability

When all links on a given channel are excited by the cochannel interference instability, each transponder (assuming one cell per transponder) will have one additional excited link. The increased power required by the transponder for this link will result in increased intermodulation distortion for the entire cell. Once enough links become excited by means of the cochannel interference instability, the cell will begin to see links excited to do intermodulation distortion. Note that since the cochannel interference instability is on a per-channel basis, when one cell experiences intermodulation distortion instability it is likely that there are several other cells sharing the same channel which are also on the verge of intermodulation distortion instability. When a cell becomes unstable via intermodulation distortion, all links in that cell will become excited.

This situation, unlike simple cochannel interference instability will render the entire ceil useless. In addition, cochannel interference for all other neighboring cochannel cells will then increase making cochannel inteference instability more likely in cells which, up to now, were stable (a similar effect may be caused by reduced transponder power or poor satellite pointing).

Ultimately, it can be seen that all links in every cell will become excited leaving the entire system hopelessly mired in interference and distortion until the Network Control can observe the situation and override the excited state via the common signalling channel.

SECTION 3

SATELLITE AND EARTH STATION ANTENNA TECHNOLOGY

SECTION 3

SATELLITE AND EARTH STATION ANTENNA TECHNOLOGY

3.1 INTRODUCTION

Frequency spectrum conservation is a key element in the design of a contiguous, multi beam antennas.

The geographical area is taken as the lower 48 states. Extension to include Alaska, Hawaii and Puerto Rico increases the system complexity somewhat, however, this does not affect the overall conclusions, and hence will not be considered further.

110° West satellite longitude is assumed for the location of the synchronous orbit satellite. This is nearly optimum for the stated geometrical area.

For the selected examples the total average communication capability is identical. This total capacity can be distributed nearly uniformly within the angular gain of the coverage, however, such a uniform distribution is neither practical, nor desirable. The distribution of required communication capacity typically is proportional to the population density or density of industrial activity.

Propagation effects and spacecraft positioning errors, affect the availability of a communication channel and other parameters are not negligible, but their consideration is beyond the scope of this section.

The assumed frequencies are 18 and 28 GHz. The actual bandwidth available for the down and uplink is a function of the specific antenna configuration (beam topology plan).

In the following discussions an attempt will be made to survey the possible system implementation alternatives, calculate their potential characteristics, determine their implementation configuration and technology development requirements.

The discussion will start with a list of assumptions and definitions followed by a description of a number of beam topology plans. An exhaustive discussion of the various beam topology plans is not possible within the scope of the present study, but all the major canonic types are defined and analyzed. Characteristics of other beam topologies can be derived by interpolation from the treated examples. The discussion then will go on to present the detailed radiation characteristics of the various topologies. These characteristics are discussed in two groups. The first group of data includes the primary computed main and crosspolarized gain contour characteristics for the various antenna beams required to realize the different beam topology plans. The second group of data includes the derived characteristics, between beam and polarization isolation, antenna size, gain slope at the contour of coverage, and number of spectrum reuses. A series of antenna circuit block diagrams where necessary, give a better insight into associated circuit complexities and problems. On the basis of the presented information the various systems are compared and rated. From these processes one of the more attractive systems are selected and its operation described on the basis of a more detailed block diagram.

3.2 ASSUMPTIONS AND DEFINITIONS

The assumptions and definitions for the present study are summarized in Table 3-1.

A spot of shaped beam is formed by one or more "component" beams. A component beam in turn is associated with an individual antenna radiating element (horn). However, for the purpose of reducing sidelobes and the angular derivative of the pattern at the coverage contour ("slope") the beams are generally intentionally shaped and formed by more than one component beam. These component beams can be divided into two types, (1) A "main component beam" which is excited close to the maximum power level and (2) An "auxiliary

TABLE 3-1. ASSUMPTIONS AND DEFINITIONS

Satellite position: 110°W
Coverage: 48 States, USA
Nominal uplink band: 28 GHz
Nominal downlink band: 18 GHz

G_o	=	peak gain of satellite antenna shaped beam
ΔG_o	=	contour gain below peak gain
αVPD	=	loss in output VPD (Variable Power Divider) circuit of satellite antenna
αSD	=	loss in satellite downlink antenna circuit
D	=	satellite antenna diameter
n	=	number of feed horns or component beams in satellite antenna
ΔF	=	total up or downlink band occupied by the communication system (allocation bandwidth)
B	=	maximum contiguous frequency band available in one shaped beam (channel)

$n_s = \frac{\Delta F}{B}$ = total bandwidth to channel bandwidth ratio

$C_o = n_c B$ = composite system bandwidth

$N = \frac{C_o}{\Delta F} = \frac{n_c B}{\Delta F}$ = number of spectrum reuses

F_B	=	total maximum bandwidth available in component beam
α	=	cell diameter or component beam center to center separation
N_B	=	number of beams
S	=	slope of the pattern at its contour coverage
ϵ	=	angle between a point of observation and the location of peak of the beam (degrees)
C/I	=	carrier to interference noise ratio
α_{sat}	=	attenuation of feed circuit of the satellite beam
α_{total}	=	total signal degradation, including earth station receive losses
θ_{MA}	=	scan angle of the beam in azimuth
θ_{ME}	=	scan angle of the beam in elevation

component beam" which is excited typically 10 db to 25 db below a main component beam and its function is strictly sidelobe level (beam) isolation control.

In the following it is assumed that for best spectrum utilization and isolation both polarizations are used. However, only one polarization is used per spot beam in order to avoid the interference effect that otherwise would be caused by deterioration of polarization orthogonal during fading and isolation.

3.3 BEAM TOPOLOGY ALTERNATIVES

The basic problem in a multibeam antenna design is the selection of a beam topology plan which optimally meets the desired system objectives of maximum frequency reuse and maximum isolation. The beam topology provides a plan in which the number, cell size, relative location, frequency band and polarization assignment of individual component beams are arranged in a given manner. The beam topology itself is not concerned with the physical implementation, but the evaluation of a specific topology plan does require the knowledge of physically realizable radiation characteristics.

Possible beam topology plans can be divided into contiguous and non-contiguous plans. For the contiguous plans the sum of the shaped beam coverages cover the entire communication area. Such plans require the use of two orthogonal polarizations or/and the subdivision of the overall frequency band into subbands. Generally not all the subbands are available in all the component beams. The noncontiguous plans can be operated with one polarization and one frequency band but some of the component beams may not be available for communication, because they are needed for sidelobe control.

Eight basic topology plans are described for contiguous coverage. Additional derivatives and some noncontiguous plans also will be discussed.

The main characteristic in which these variants differ from each other is the normalized guard angle, (or adjacent beam separation) which is the angle from the maximum of the component beam to the point where the sidelobe level falls below a given value K

For two component beams of identical frequency and polarization, and given adjacent beam isolation I_{Badj}, beam separation must be such that at the edge of contour of beam 2 the radiation level from beam 1 is

$I_{Badj} = A_c + \kappa$, This is the worst case beam to beam interference.

A_c is the power difference between the contour value and peak for beam 2. If for instance $I_{Badj} = 27$ db and $A_c = -1$ db, then $\kappa = 28$ db and the corresponding ϵ angle is determined. When N beams make up the entire system, then the adjacent beam isolation between the first and k^{th} beam is

$$(I_{Badj})_{1k} = A_{c1} + \kappa_{1k}$$

and the resultant beam isolation is

$$I_B = \sum_{2}^{N} (I_{Badj})_{1k}$$

Generally the radiation of a given spot beam, say spot beam 1 will be formed from more than one component beam. Some of these component beams will cover the intended coverage region of shaped beam 1, and are called main component beams. The other component beams of shaped beam 1 may be used to control the resultant sidelobe level and are called auxiliary component beams. Since the auxiliary component beams are directed outside the coverage region of spot beam 1, they generally fall into the guard regions between coverage regions or into the coverage region of another beam. When a component beam is used by more than one spot beam, then it may be called a coupled component beam.

Table 3-2 defines 8 different beam topologies. These plans can be divided into different groups on the basis of

a. Use of polarization
b. Use of component beams in the shaped beams
c. Existence of beam overlap

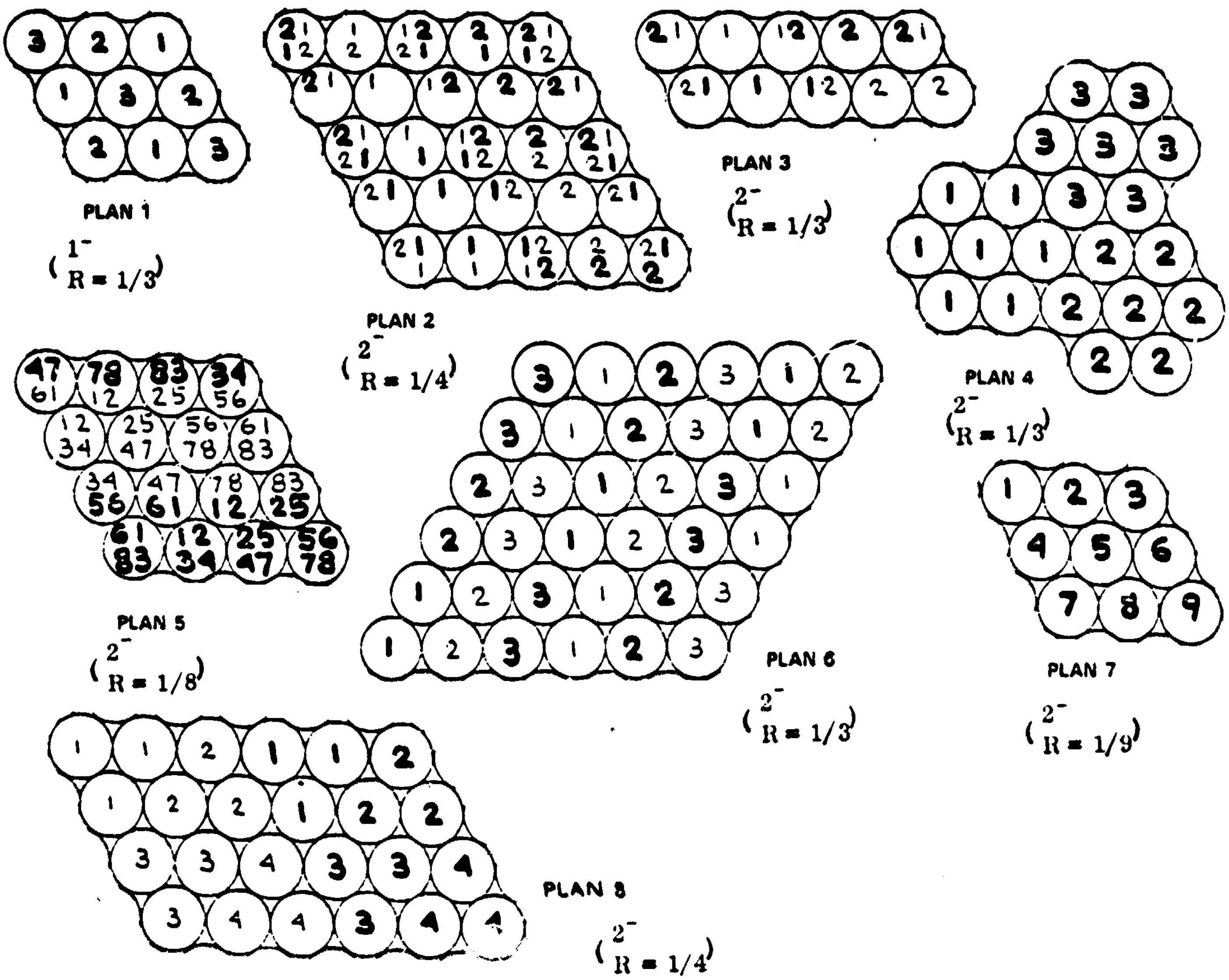
The systems can use single polarization (Plans 1, 4, 7) or dual polarization (Plans 2, 3, 5, 6, 8). Spot beams can be formed from one component beam (Plan 1, 7) or more than one component beam (Plans 2, 3, 4, 5, 6, 8). The shaped beams can be nonoverlapping types (Plans 1, 4, 6, 7, 8) or overlapping types (Plans 2, 3, 5).

TABLE 3-2. TOPOLOGY CHARACTERISTICS OF STUDIED BEAM PLANS

No.	Beam Plan	ϵ'/α Main Polarization	ϵ'/α Cross Polarization	Configuration	Number of Main Component Beams	Number of Auxiliary Component Beams in First Order Ring	Band Sub-division	Overlap
1	Triangular Layout Singlets Single Polarization	1.232	NA	Singlet	1	6	3	No
2	East-West Overlapping Novets Dual Polarization Polarization Reuse Strips	1.5	.5	Sextet or Novet	6 or 9	12 or 14	2	One Dimensional
3	East-West Overlapping Triplets, Dual Polarization Polarization Reuse Strips	1.7	.5	Line Triplet	3	10	2	One Dimensional
4	Triangular Layout Septets, Single Polarization	2.145	NA	Septet	7	12	3	No
5	Overlapping Quadruplets, Dual Polarization	2.345	1.232	Rhombic Quadruplet	4	10	8	Yes
6	Triangular Layout Doublets, Dual Polarization, Polarization Reuse Spots	2.145	.5	Doublet	2	3	3	No
7	Rhomboidal Layout Singlets, Single Polarization	2.5	NA	Singlet	1	6	9	No
8	Rhomboidal Layout Triplets, Dual Polarization	3.105	1.5	Triplet	3	9	4	No

TABLE 3-2. TOPOLOGY CHARACTERISTICS OF STUDIED BEAM PLANS (Cont'd)

Examples for Various Beam Topology Plans



Note: Parenthesis indicates approximate co-channel beam separation and R_1 the frequency re-use factor.

When multiple component beams are used to build a spot beam the number of main horns in practical systems of interest can be 2 (doublet), 3 (triplet), 4 (quadruplet), 7 (septet) or 9 (nontet). Larger systems can be also considered but the beam isolation characteristics for essentially symmetrical component beam configuration do not change drastically beyond the septet.

While the above listed categories are important the kev characteristic of all the plans listed in Table 3-2 is the ϵ/α ratio, which characterizes the smallest separation between identical frequency and identically polarized shaped beams. The beam isolation is increased with increasing ϵ/α, with increasing number of component beams in the spot beams with increasing "symmetry" of the spot beam. Also shown in Table 3-2 is the approximate frequency reuse factor; e.g., how much of the frequency allocation is available for each beam. Table 3-3 summarizes the main characteristics of the beam topology plans with regard to performance.

Figures 3-1 through 3-8 depict the eight typical topology plans defined in Table 3-2. The figures show how the layout of the plans vary with the cell diameter, in the

TABLE 3-3. FOUR CHARACTERISTICS OF BEAM TOPOLOGY PLAN PERFORMANCE

1. Angular Separation Between Identically Polarized Identical Frequency Cells.
2. Angular Separation Between Orthogonally Polarized Identical Frequency Cells.
3. Number and Configuration of Cells Forming Coverage Areas.
4. Crossover Level Between Nonidentical Frequency Cells.

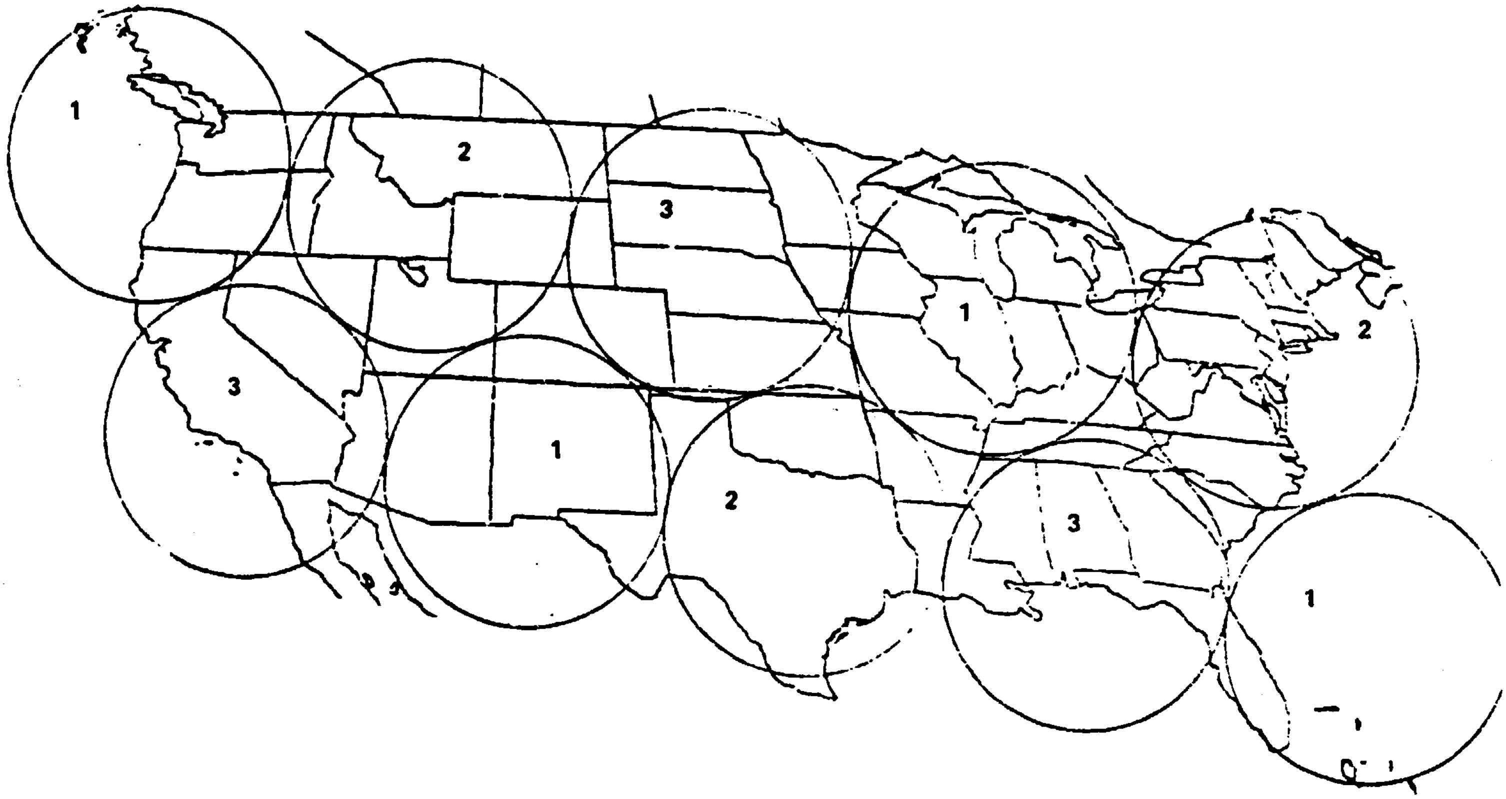


Figure 3-1a. **Beam Plan 1: triangular layout of singlets, single polarization.** $\alpha = 1.5°$, $n = 10$, $N_B = 10$, $N = 3.33$.

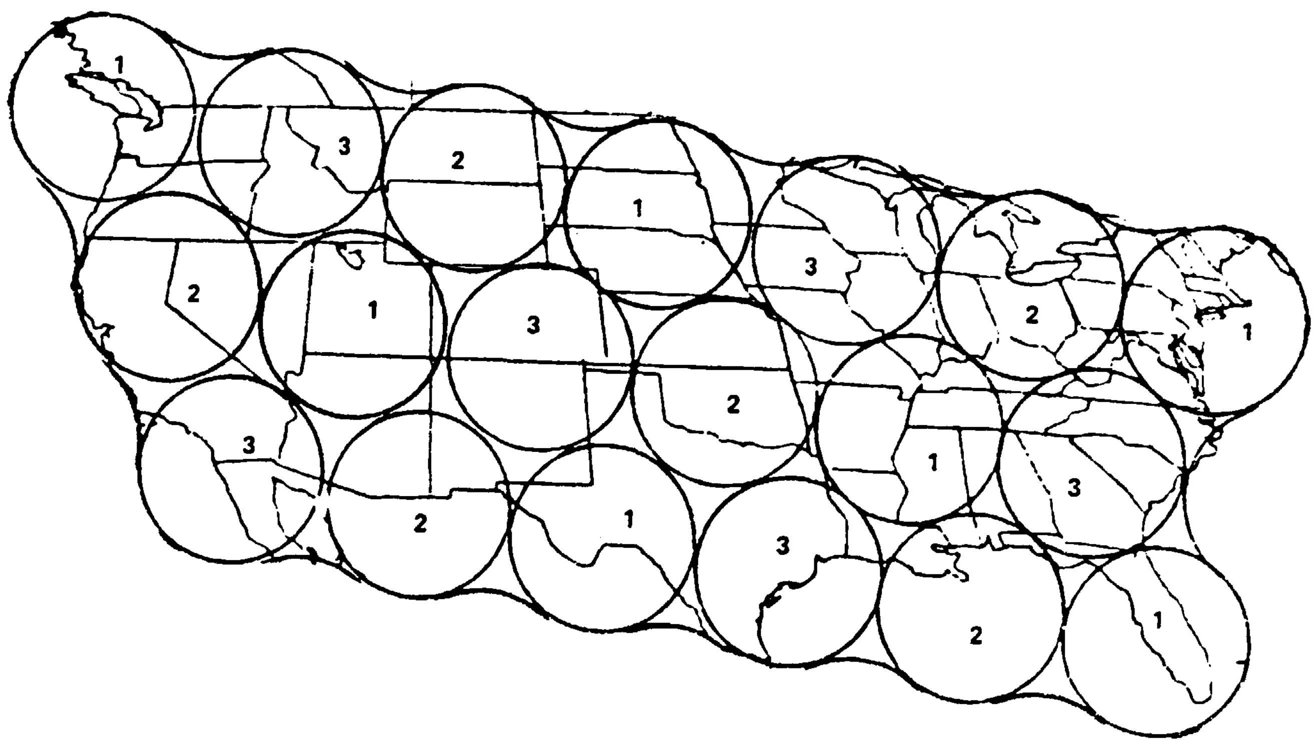


Figure 3-1b. **Beam Plan 1: triangular layout of singlets, single polarization.** $\alpha = 1°$, $n = 19$, $N_B = 19$, $N = 6.33$.

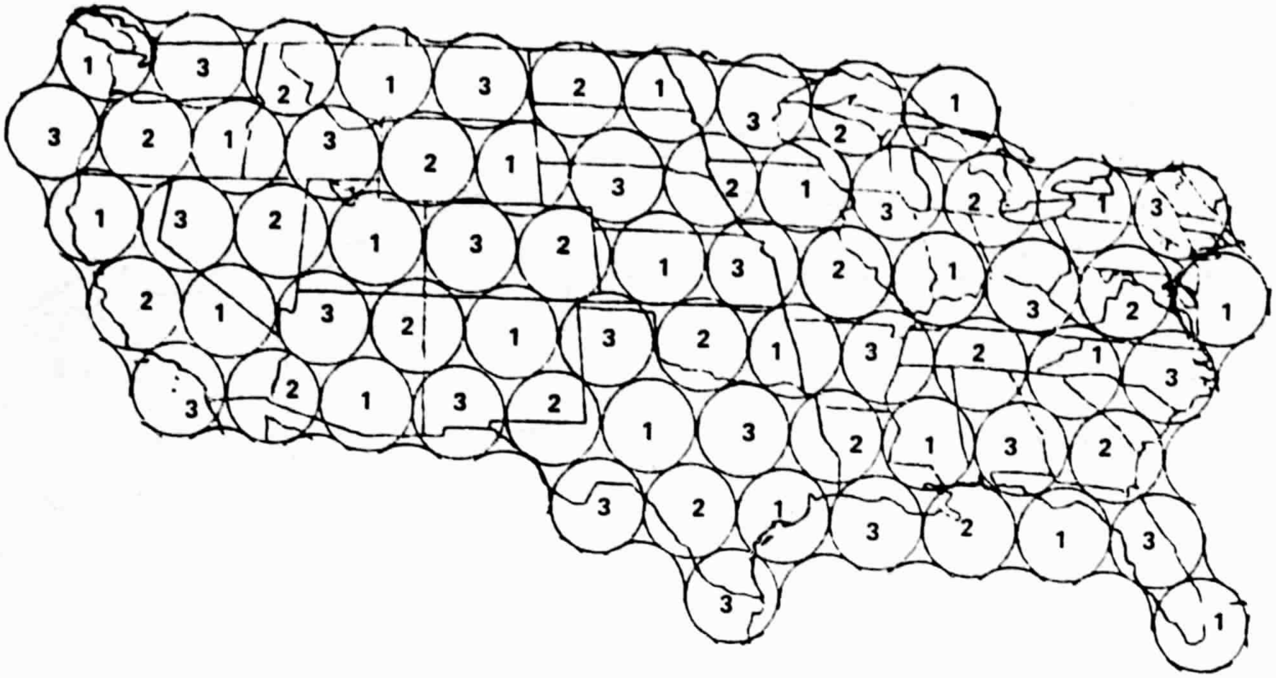


Figure 3-1c. Beam Plan 1: triangular layout of singlets, single polarization. $\alpha = .5°$, $n = 68$, $N_B = 68$, $N = 22.67$.

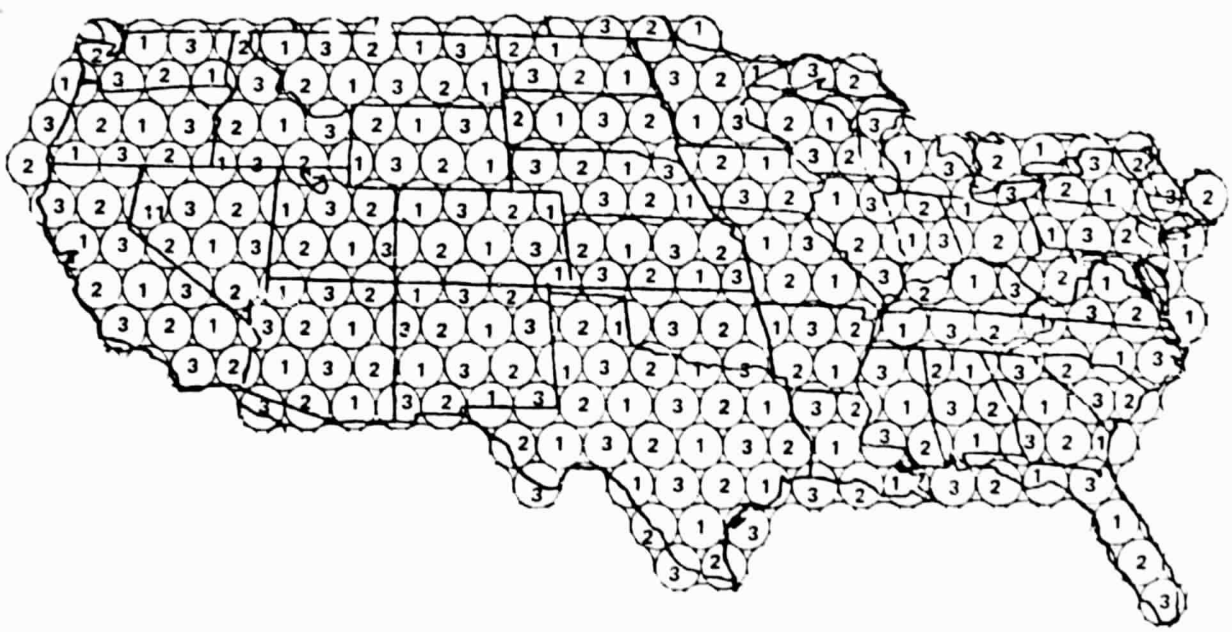


Figure 3-1d. Beam Plan 1: triangular layout of singlets, single polarization. $\alpha = .25°$, $n = 251$, $N_B = 251$, $N = 83.67$.

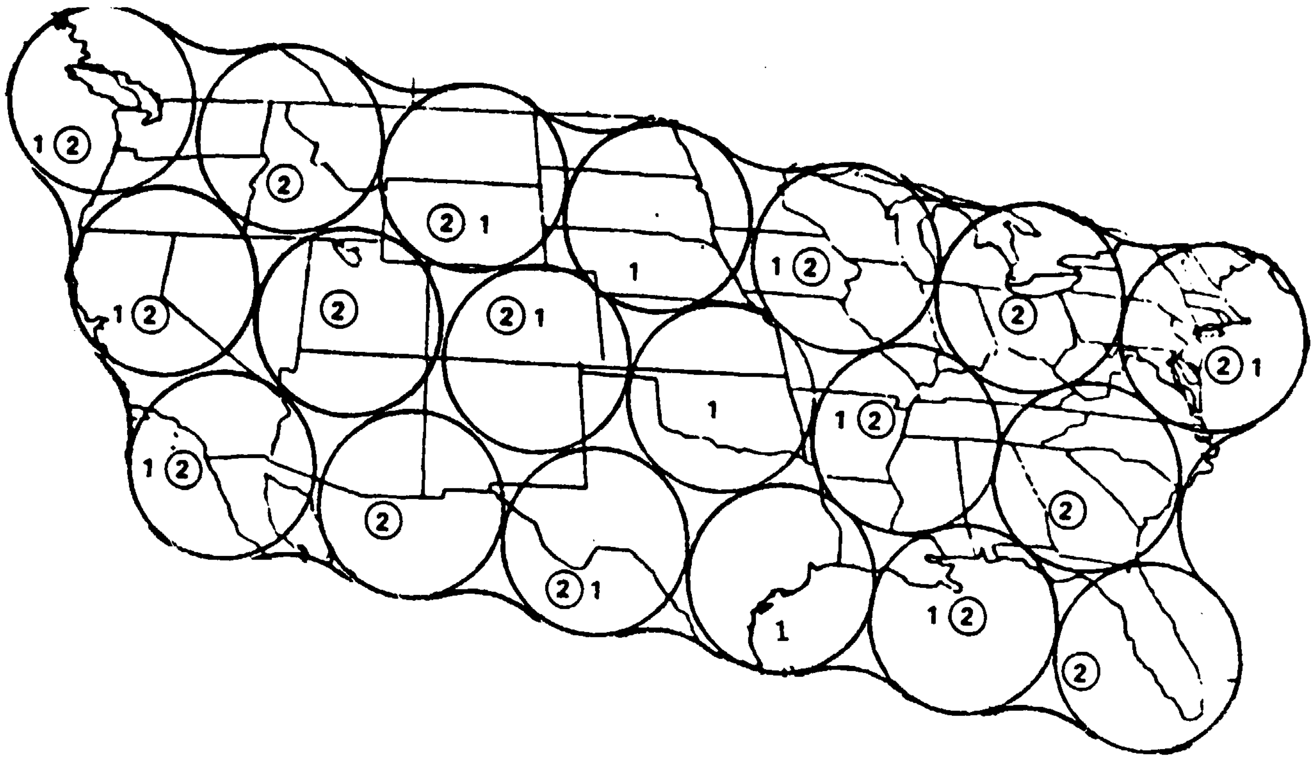


Figure 3-2a. **Beam Plan 2: East-West overlapping layout of novets, dual polarization, polarization reuse strips.**
$\alpha = 1°$, $n = 19$, $N_B = 5$, $N = 2.5$.

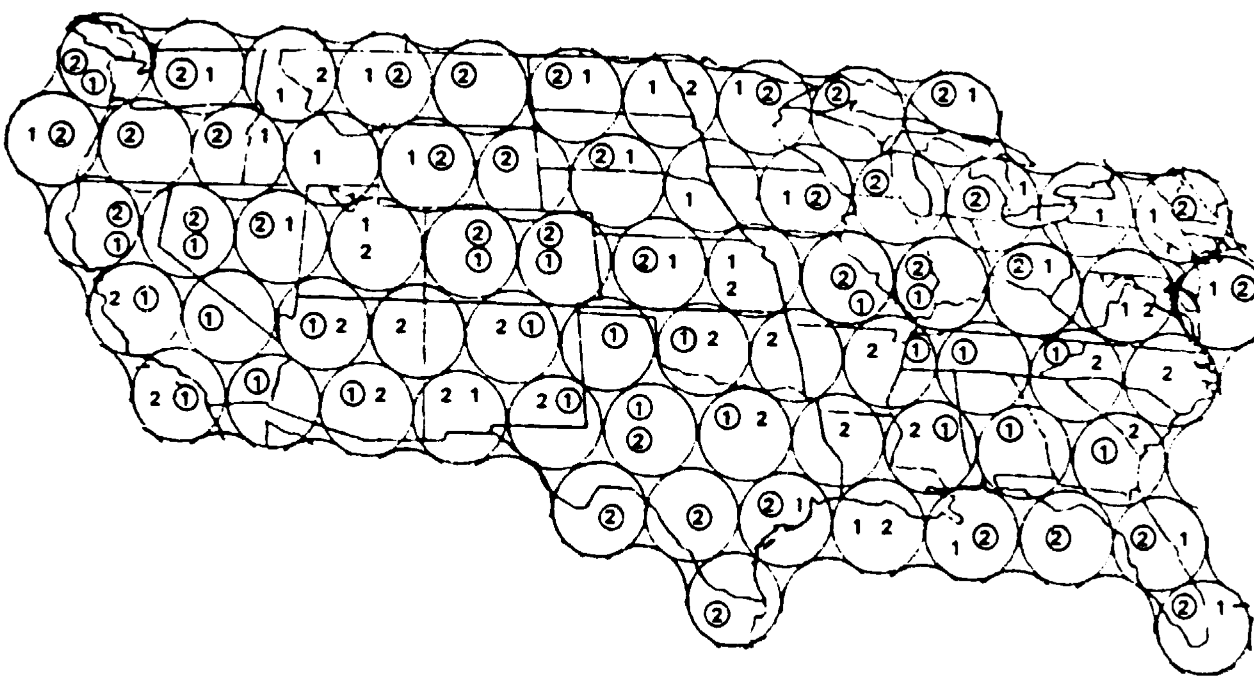


Figure 3-2b. Beam Plan 2: East-West overlapping layout of novets, dual polarization, polarization reuse strips.
$\alpha = .5°$, $n = 68$, $N_B = 22$, $N = 11$.

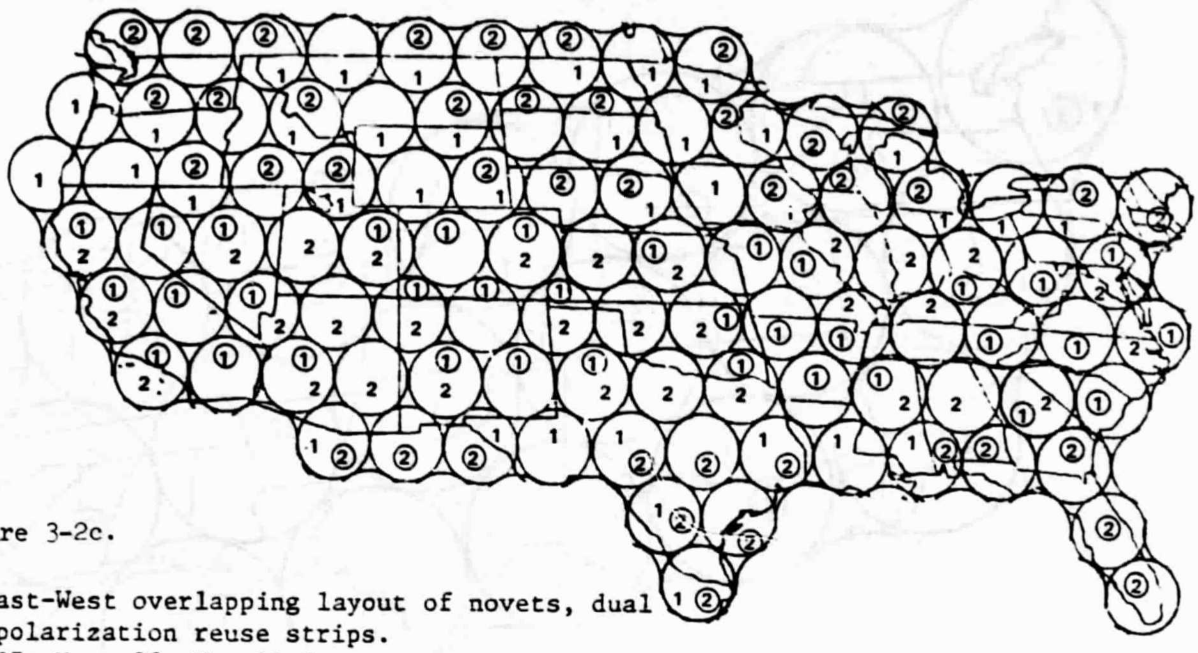


Figure 3-2c.

Beam Plan 2: East-West overlapping layout of novets, dual **polarization**, polarization reuse strips.
α = .41°, n = 97, N_B = 23, N = 11.5.

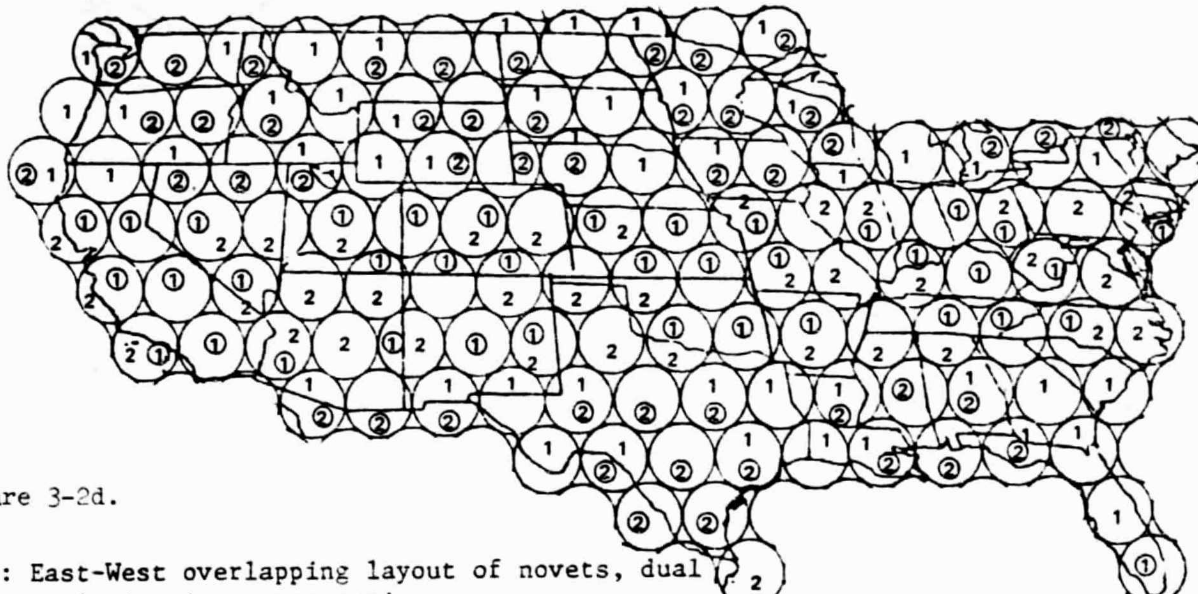


Figure 3-2d.

Beam Plan 2: East-West overlapping layout of novets, dual polarization, polarization reuse strips.
α = .37°, n = 117, N_B = 29, N = 14.5.

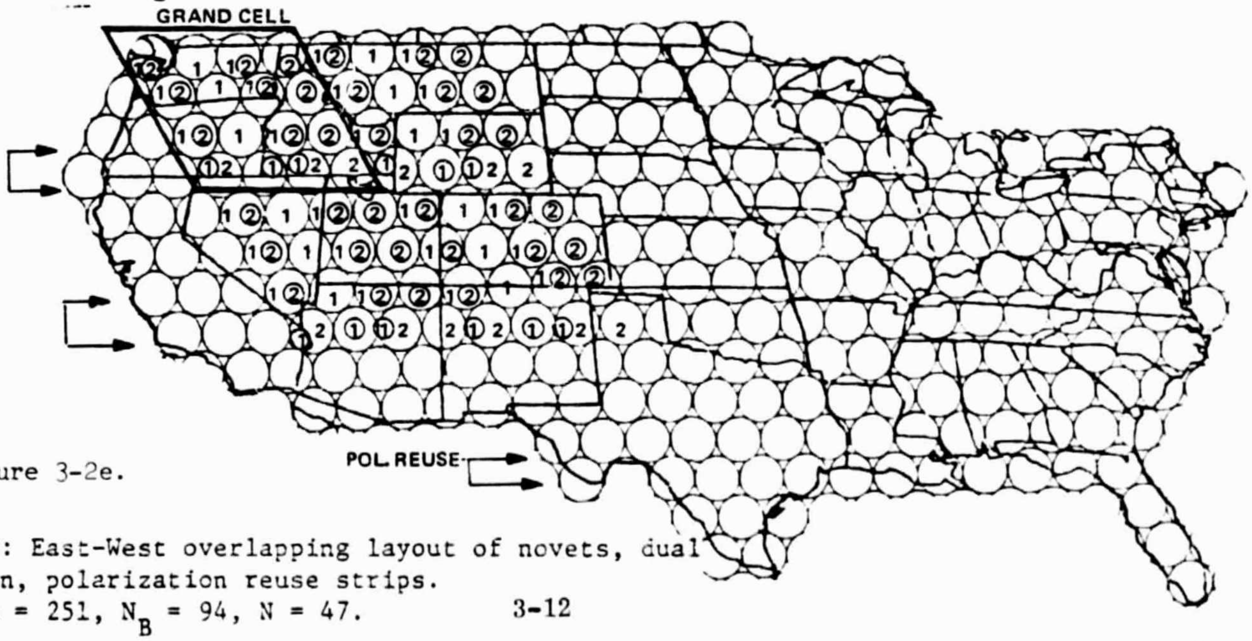


Figure 3-2e.

Beam Plan 2: East-West overlapping layout of novets, dual **polarization**, polarization reuse strips.
α = .25°, n = 251, N_B = 94, N = 47.

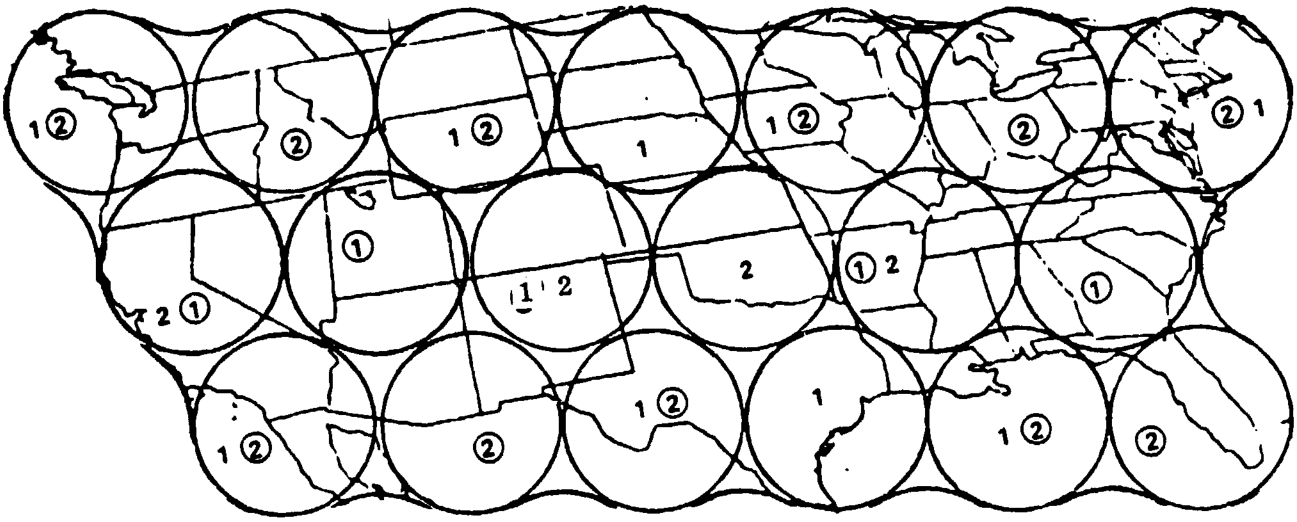


Figure 3-3a. Beam Plan 3: East-West layout of overlapping linear triplets, dual polarization, polarization reuse strips. $\alpha = 1°$, $n = 19$, $N_B = 12$, $N = 6$.

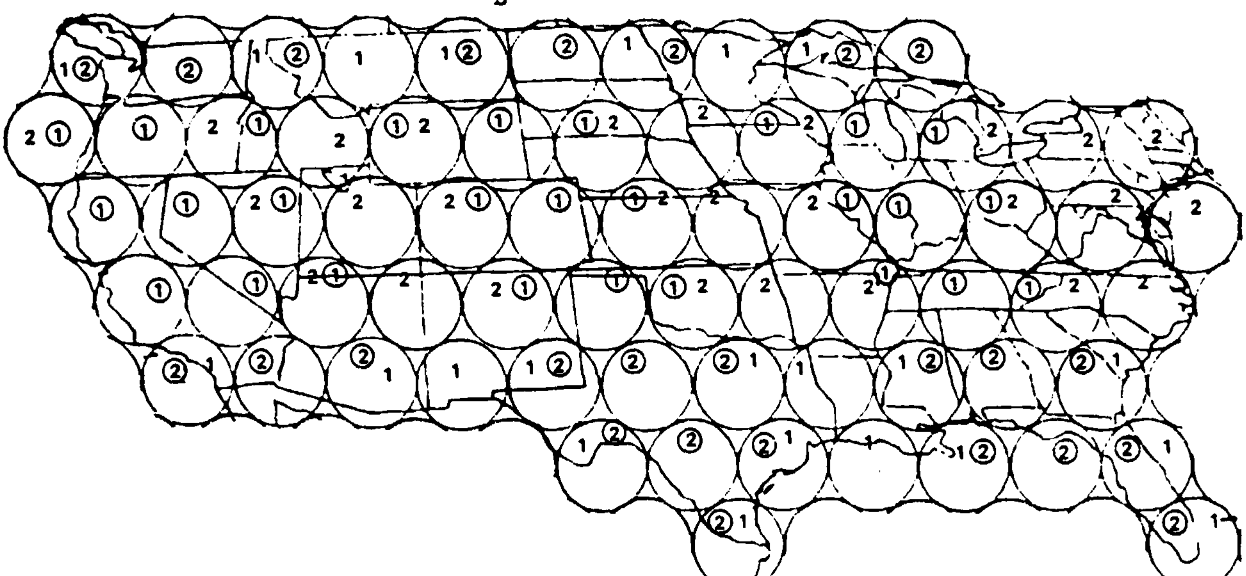


Figure 3-3b. Beam Plan 3: East-West layout of overlapping linear triplets, dual polarization, polarization reuse strips. $\alpha = .5°$, $n = 68$, $N_B = 20$, $N = 10$.

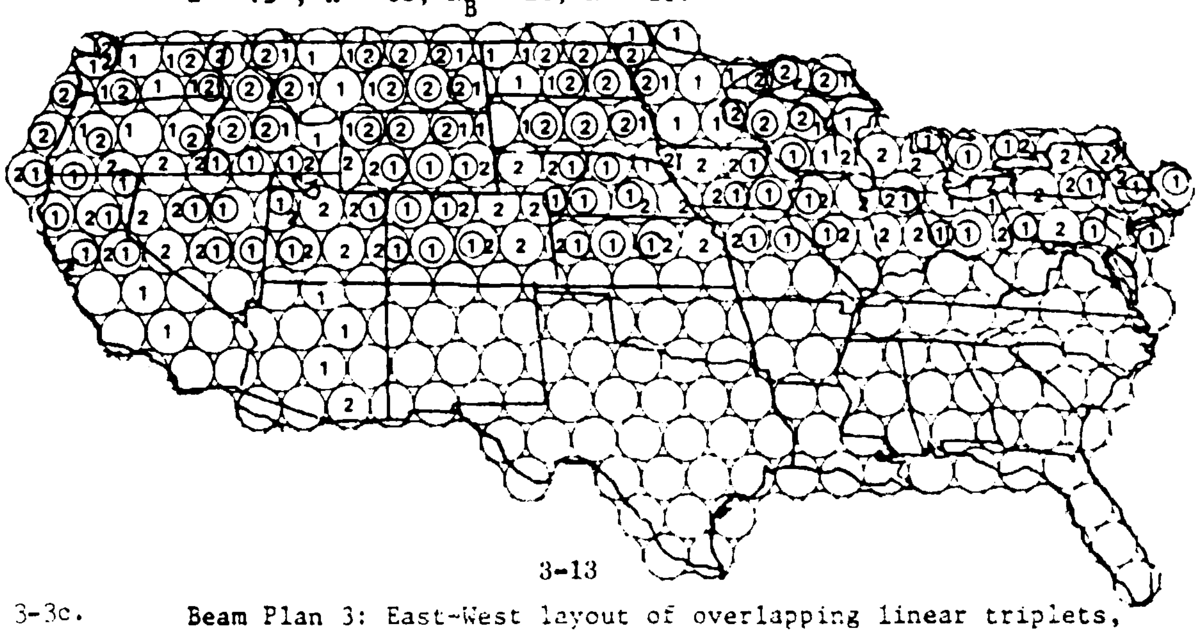


Figure 3-3c. Beam Plan 3: East-West layout of overlapping linear triplets, dual polarization, polarization reuse strips. $\alpha = .25°$, $n = 251$, $N_3 = 74$, $N = 37$.

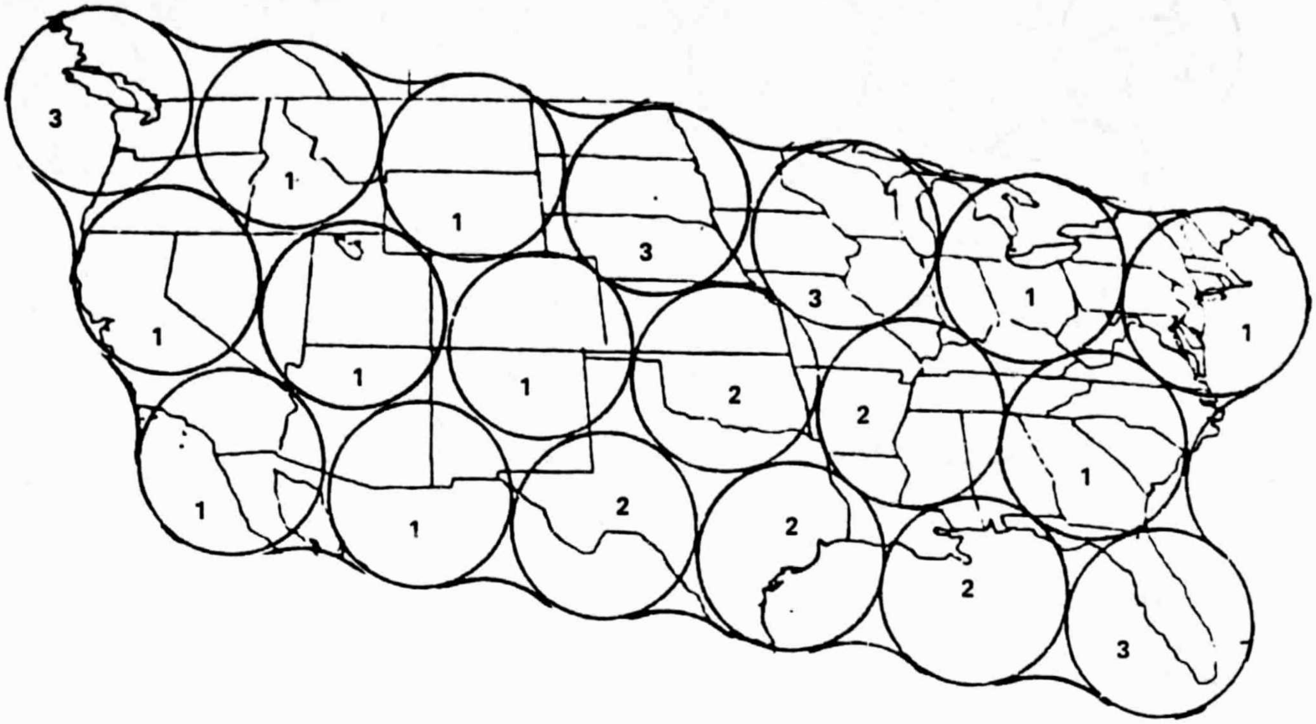


Figure 3-4a. Beam Plan 4: triangular layout of septets, single polarization. $\alpha = 1°$, $n = 19$, $N_B = 6$, $N = 2$.

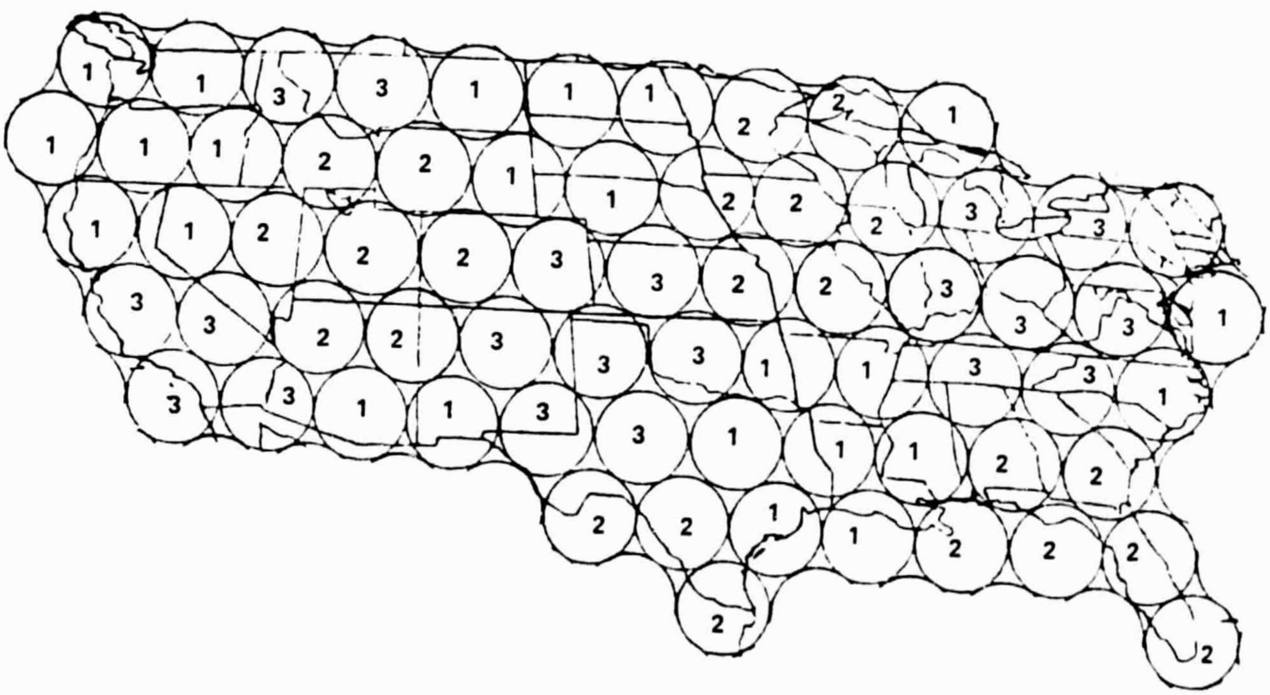


Figure 3-4b. Beam Plan 4: triangular layout of septets, single polarization. $\alpha = .5°$, $n = 68$, $N_B = 14$, $N = 4.67$.

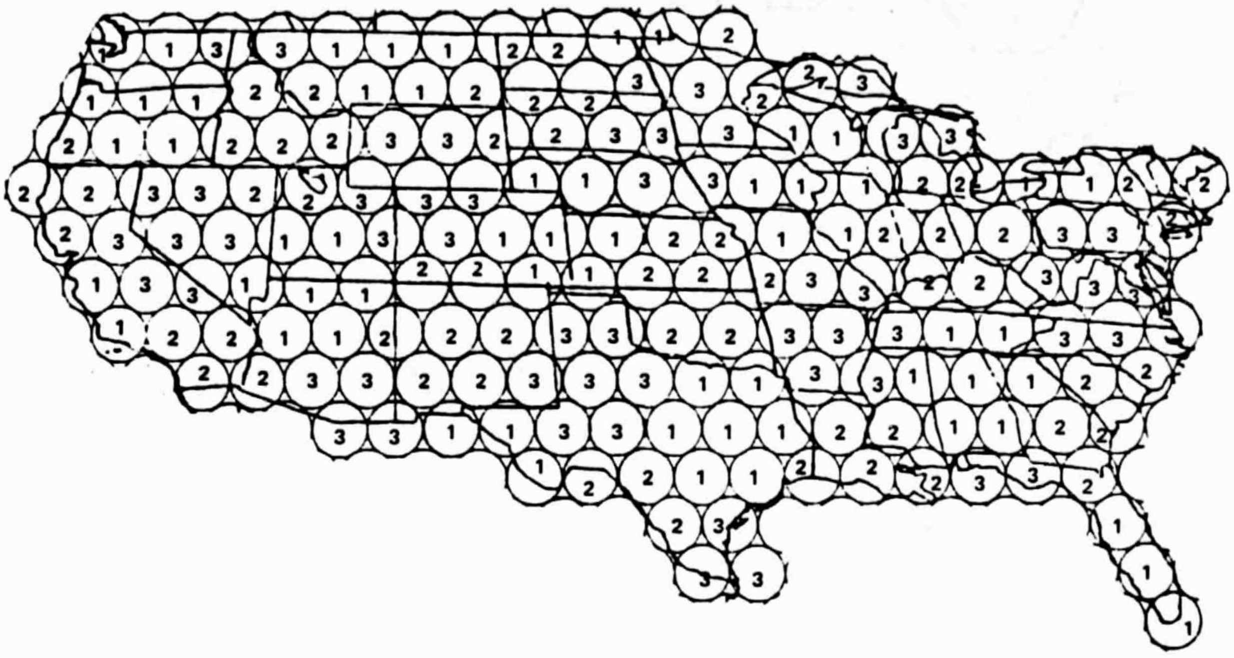


Figure 3-4c. Beam Plan 4: triangular layout of septets, single polarization. $\alpha = .3°$, $n = 178$, $N_B = 35$, $N = 11.67$.

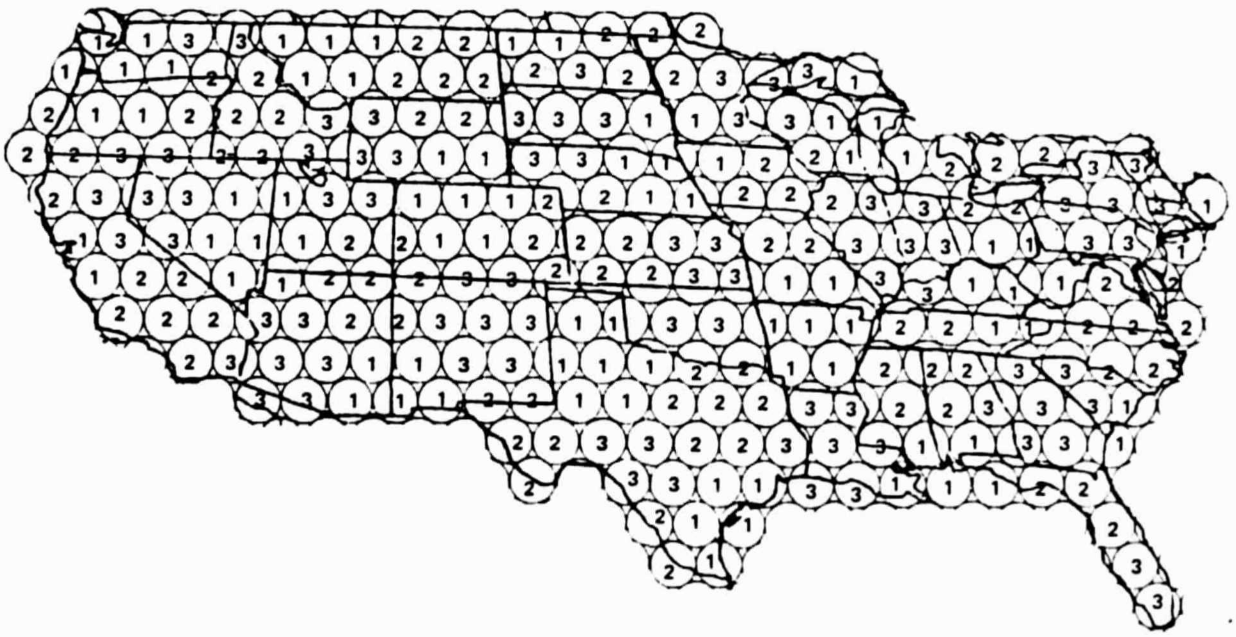


Figure 3-4d. Beam Plan 4: triangular layout of septets, single polarization. $\alpha = .25°$, $n = 251$, $N_B = 45$, $N = 15$.

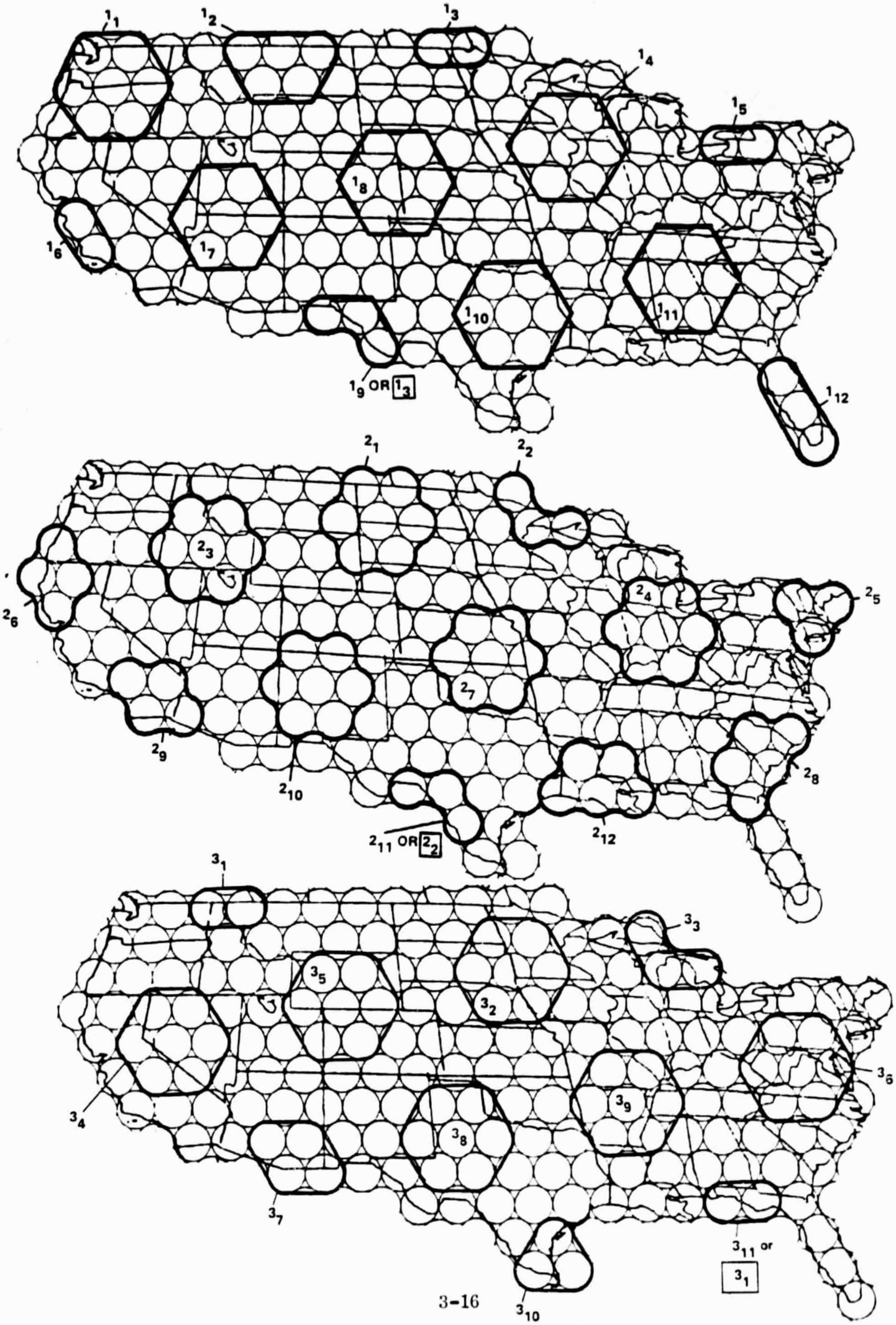


Figure 3-4e. Beam Plan 4: example to show combination of incomplete shaped beams in order to reduce system complexity. (Basic layout is similar to the one shown on Figure 3-4d.)
α = .3°, n = 178, N_R = 32, N = 10.67.

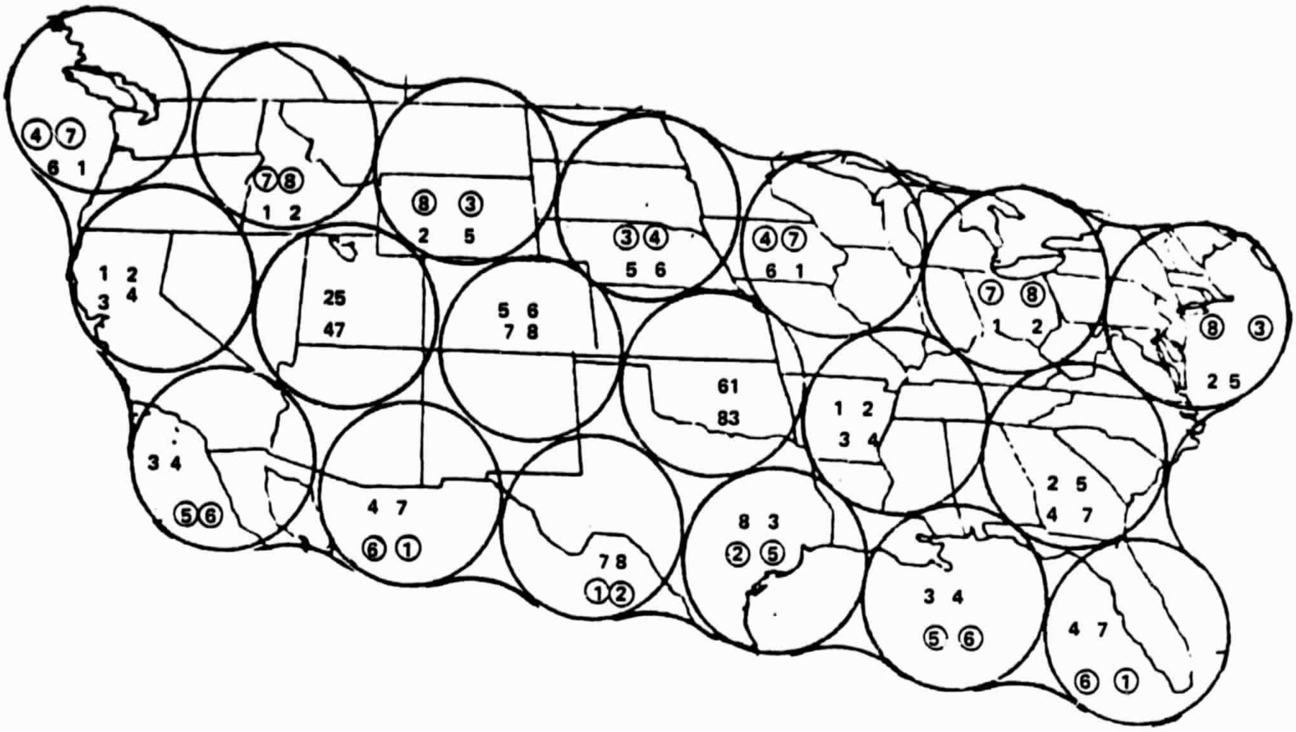


Figure 3-5a. **Beam Plan 5:** overlapping layout of rhombic quadruplets, dual polarization.
$\alpha = 1°$, $n = 19$, $N_B = 30$, $N = 3.75$.

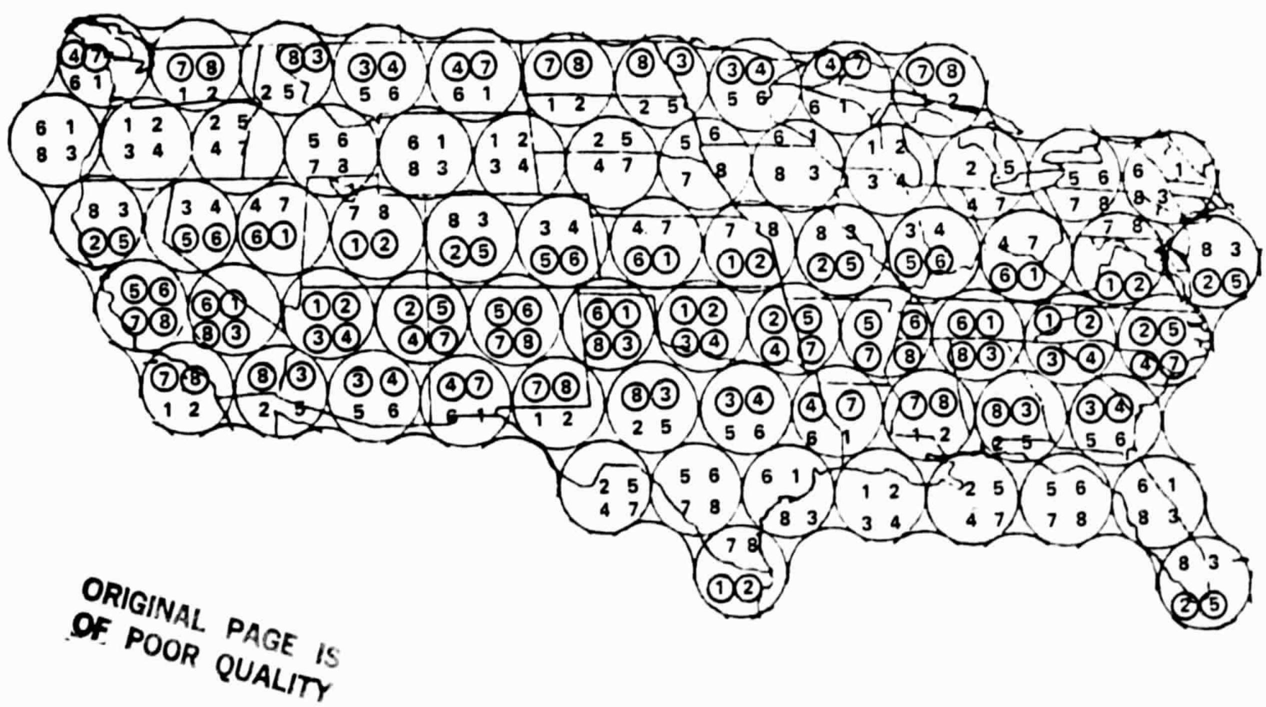


Figure 3-5b. **Beam Plan 5:** overlapping layout of rhombic quadruplets, dual polarization.
$\alpha = .5°$, $n = 68$, $N_B = 87$, $N = 10.87$.

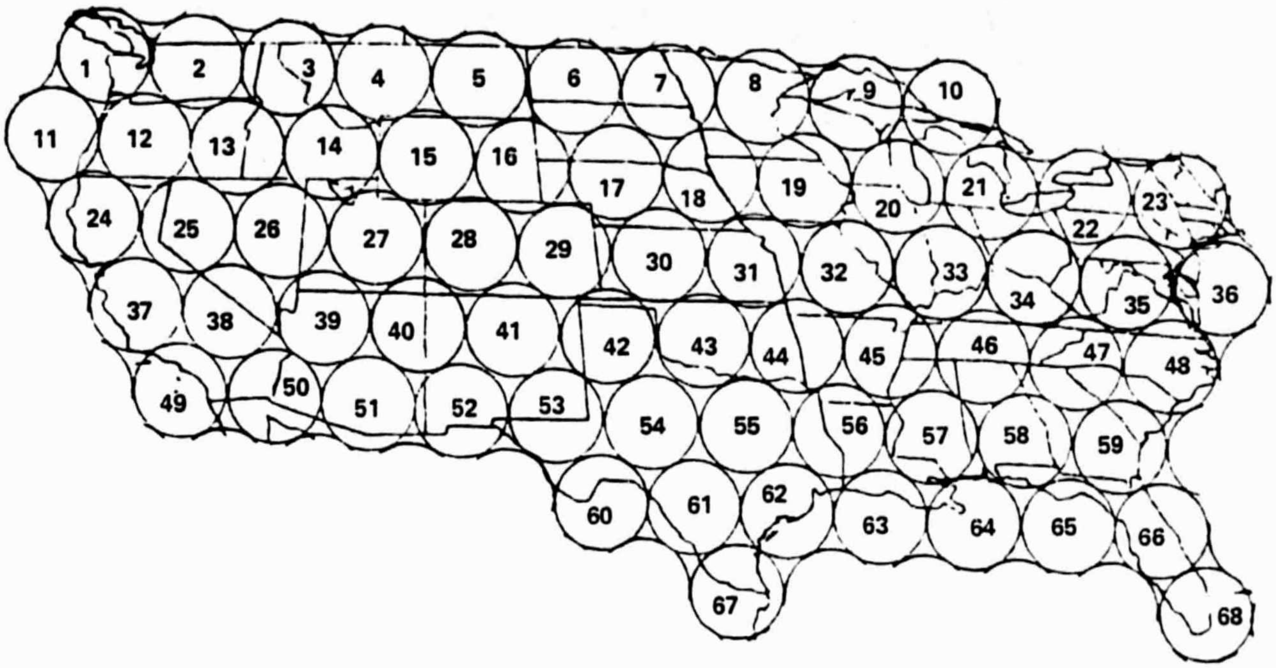


Figure 3-5c. Serial No. designation of component beams for α = .5° n = 68.

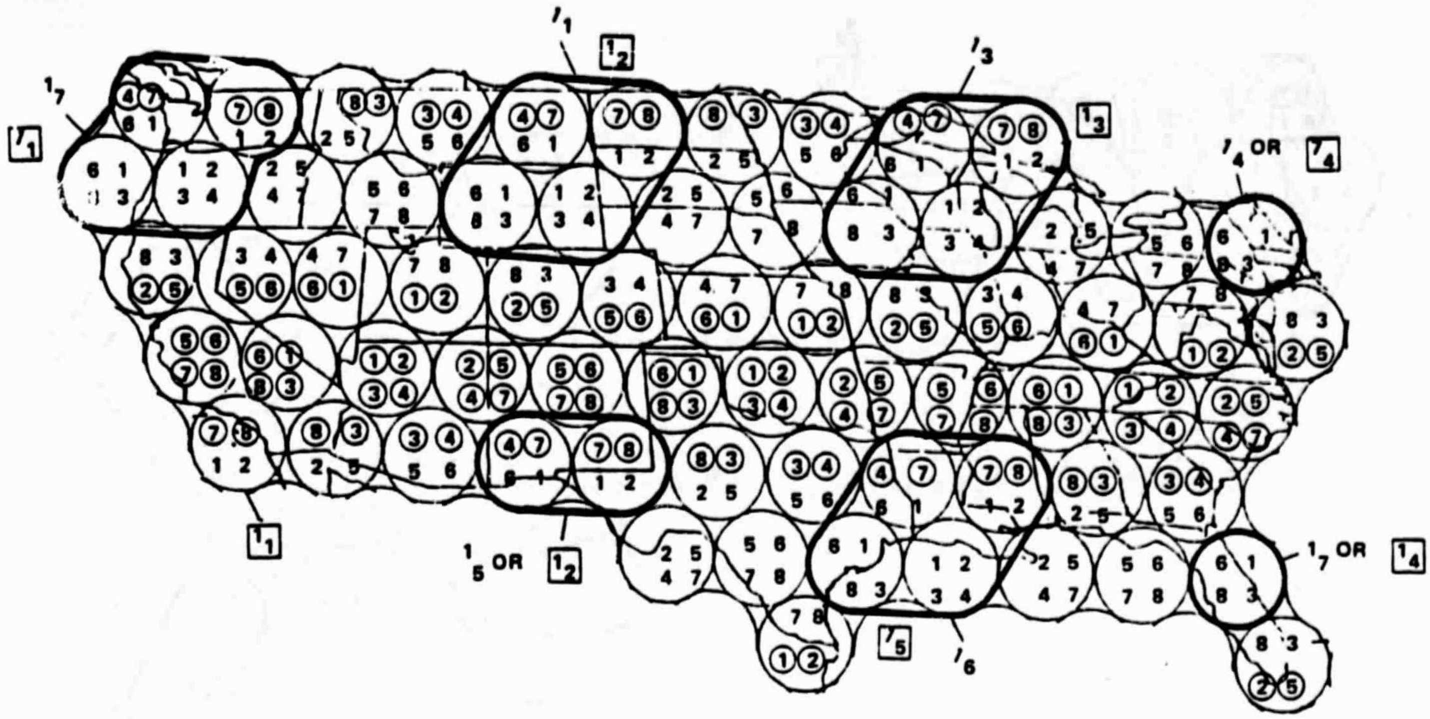


Figure 3-5d. Serial No. designation of shaped beams for α = .5° Plan 5 and for modified Plan 5 (in brackets). Modification is for reduction of system complexity, resulting in N_B = 64 and N = 8. Channel 1.

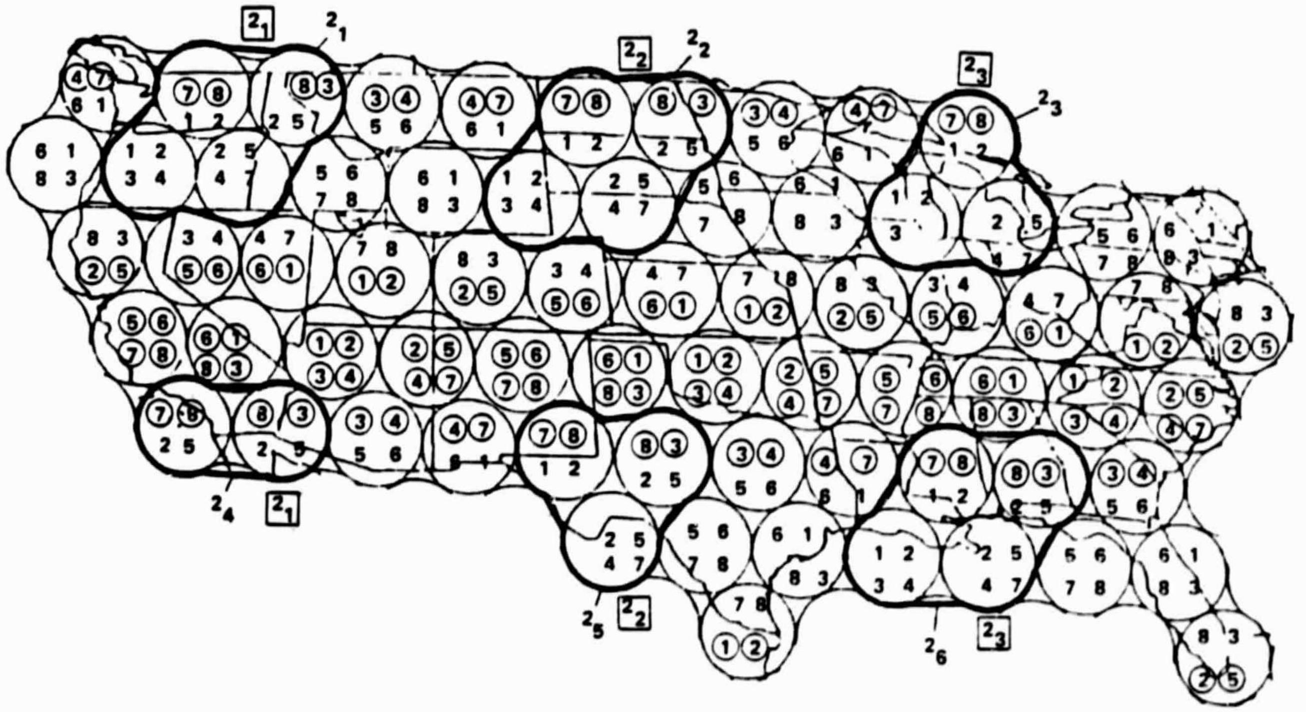


Figure 3-5e. Serial No. designation of shaped beams for α = .5° Plan 5 and for modified Plan 5 (in brackets). Modification is for reduction of system complexity, resulting in N_B = 64 and N = 8. Channel 2.

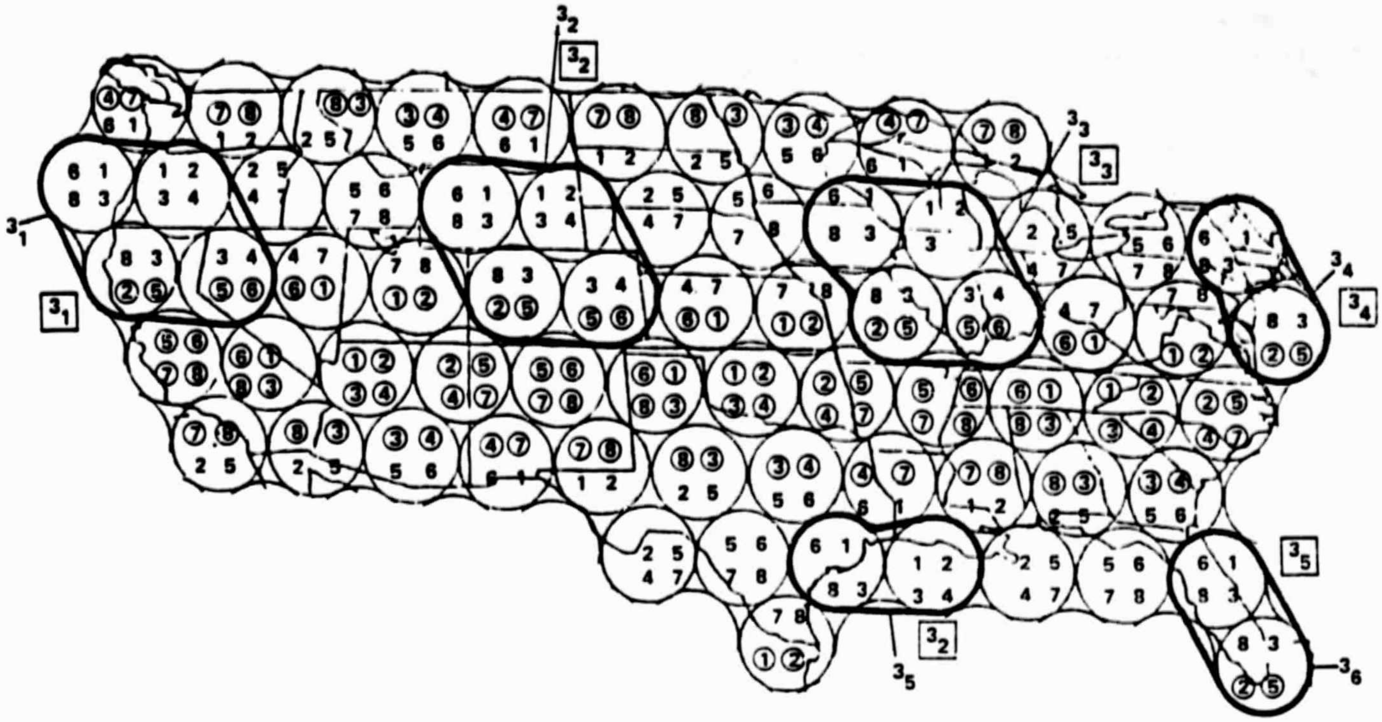


Figure 3-5f. Serial No. designation of shaped beams for α = .5° Plan 5 and for modified Plan 5 (in brackets). Modification is for reduction of system complexity, resulting in N_B = 64 and N = 8. Channel 3.

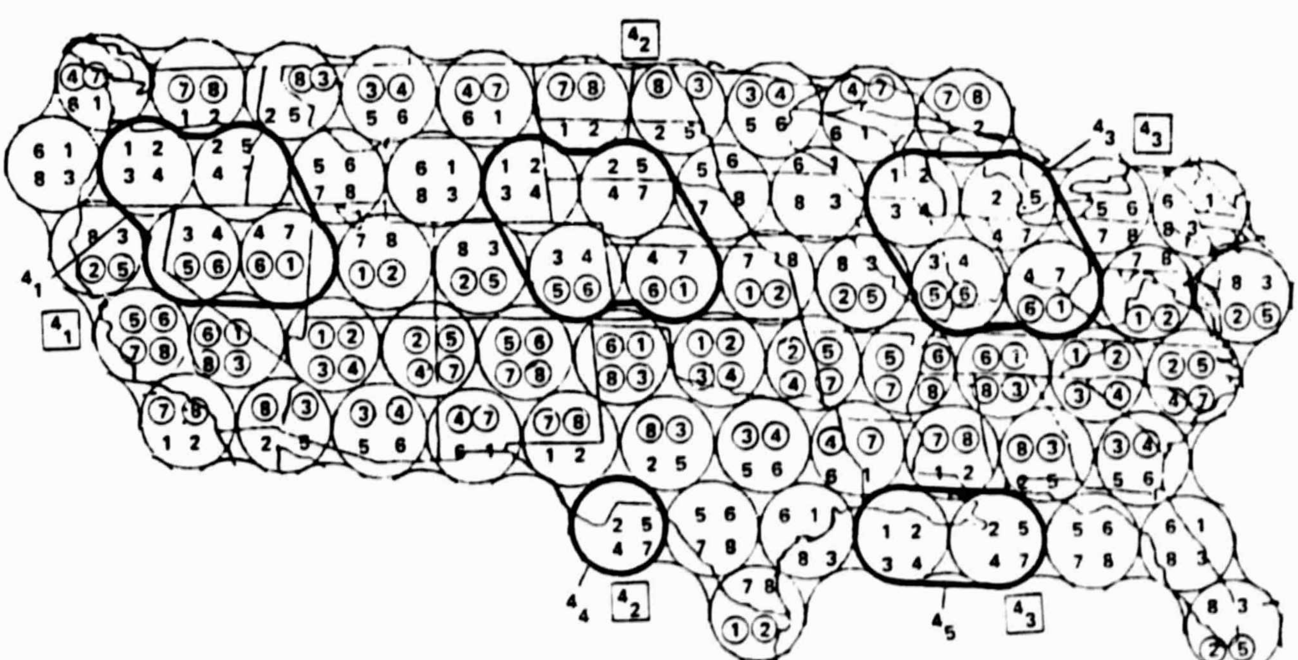


Figure 3-5g. Serial No. designation of shaped beams for α = .5° Plan 5 and for modified Plan 5 (in brackets). Modification is for reduction of system complexity, resulting in N_B = 64 and N = 8. Channel 4.

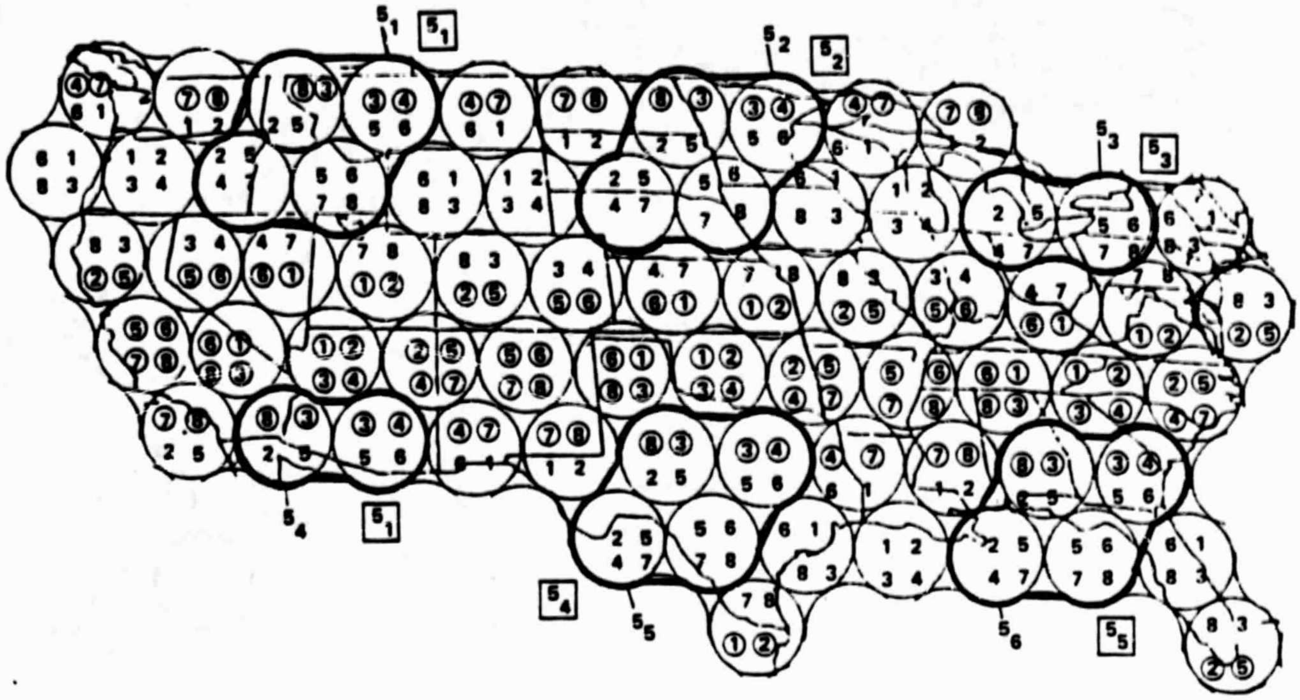


Figure 3-5h. Serial No. designation of shaped beams for $\alpha = .5°$ Plan 5 and for modified Plan 5 (in brackets). Modification is for reduction of system complexity, resulting in $N_B = 64$ and $N = 8$. Channel 5.

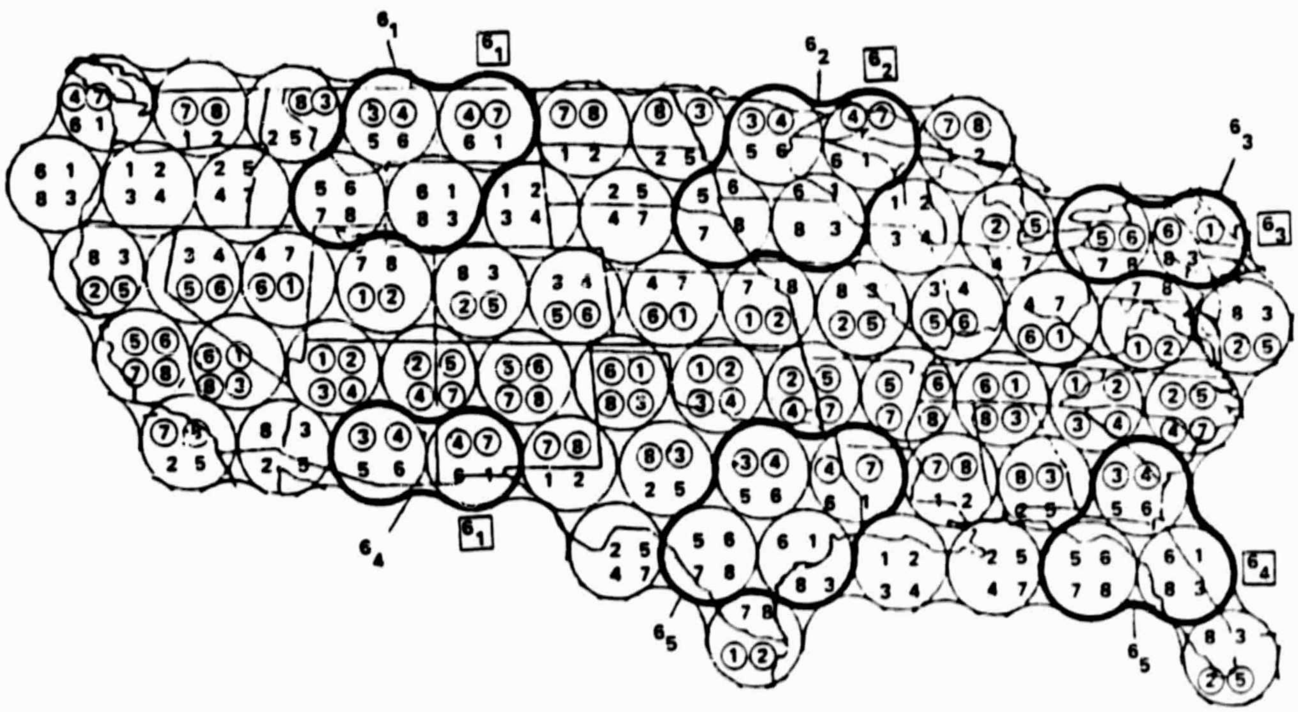


Figure 3-5i. Serial No. designation of shaped beams for $\alpha = .5°$ Plan 5 and for modified Plan 5 (in brackets). Modification is for reduction of system complexity, resulting in $N_B = 64$ and $N = 8$. Channel 6.

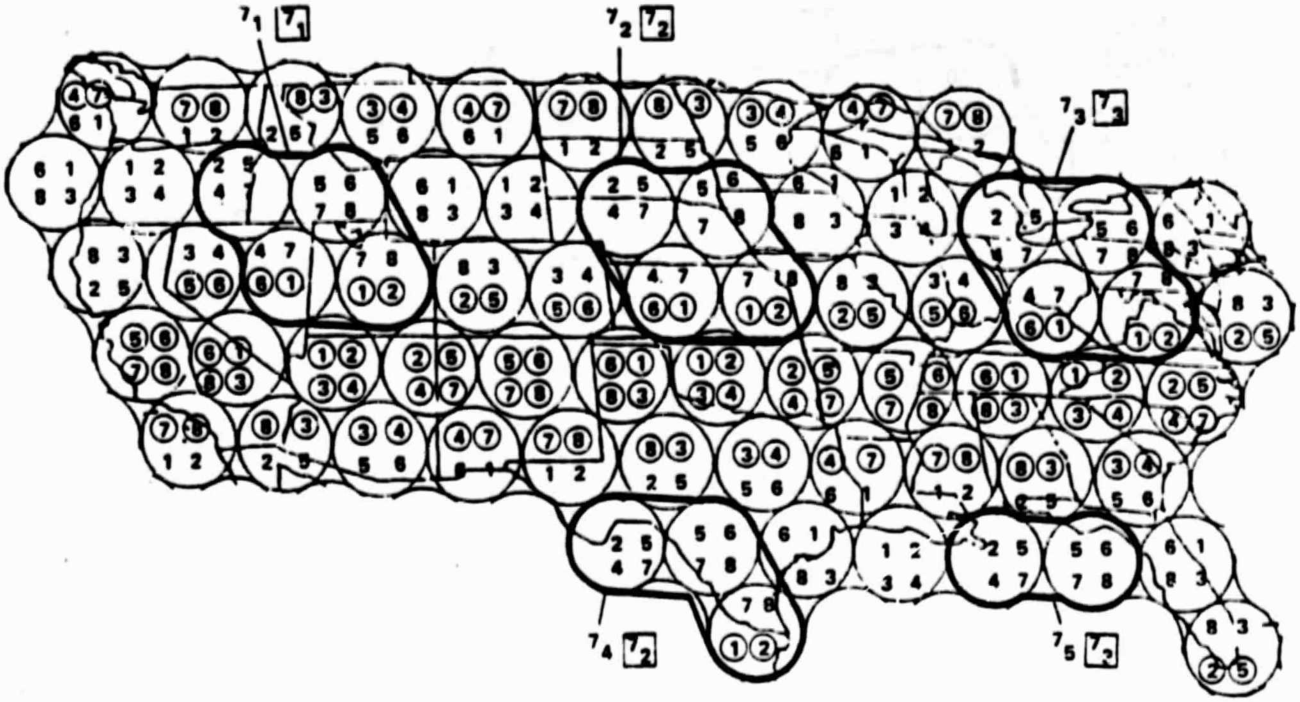


Figure 3-5j. Serial No. designation of shaped beams for α = .5° Plan 5 and for modified Plan 5 (in brackets). Modification is for reduction of system complexity, resulting in N_B = 64 and N = 8. Channel 7.

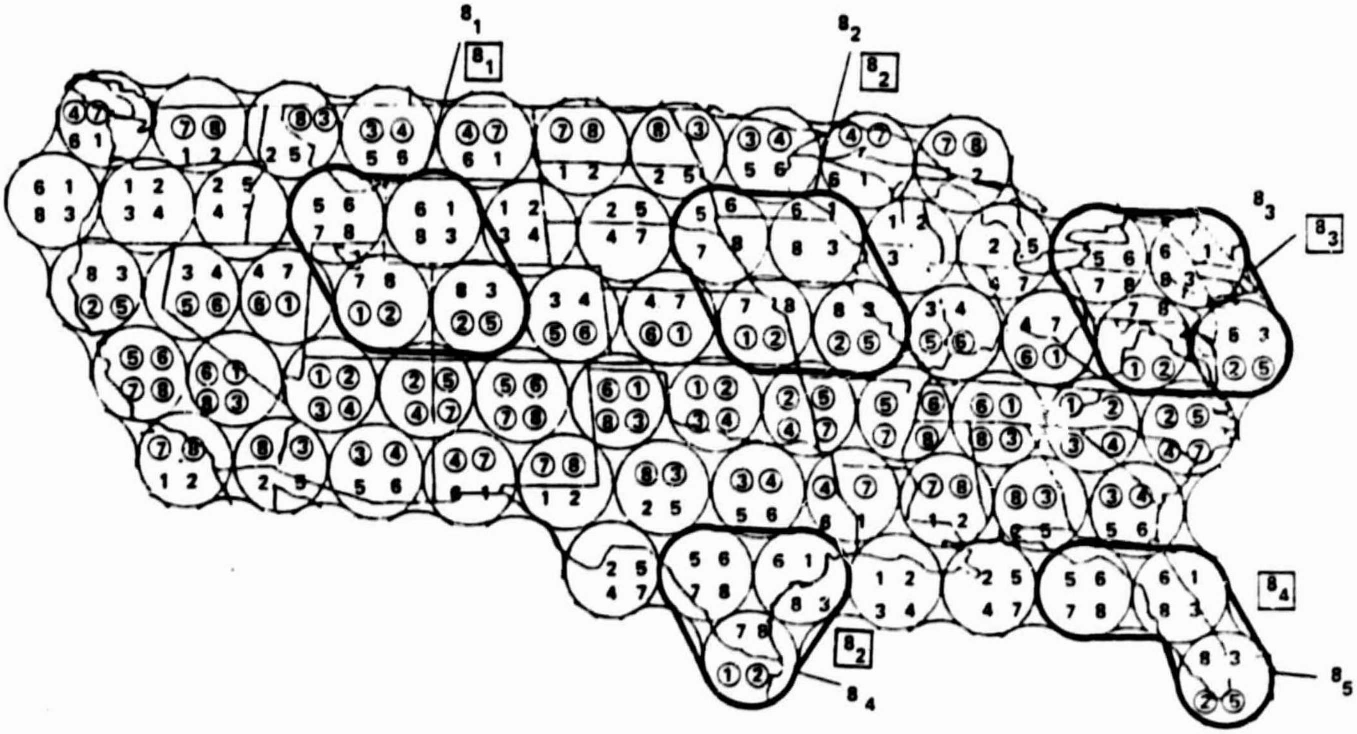


Figure 3-5k. Serial No. designation of shaped beams for α = .5° Plan 5 and for modified Plan 5 (in brackets). Modification is for reduction of system complexity, resulting in N_B = 64 and N = 8. Channel 8.

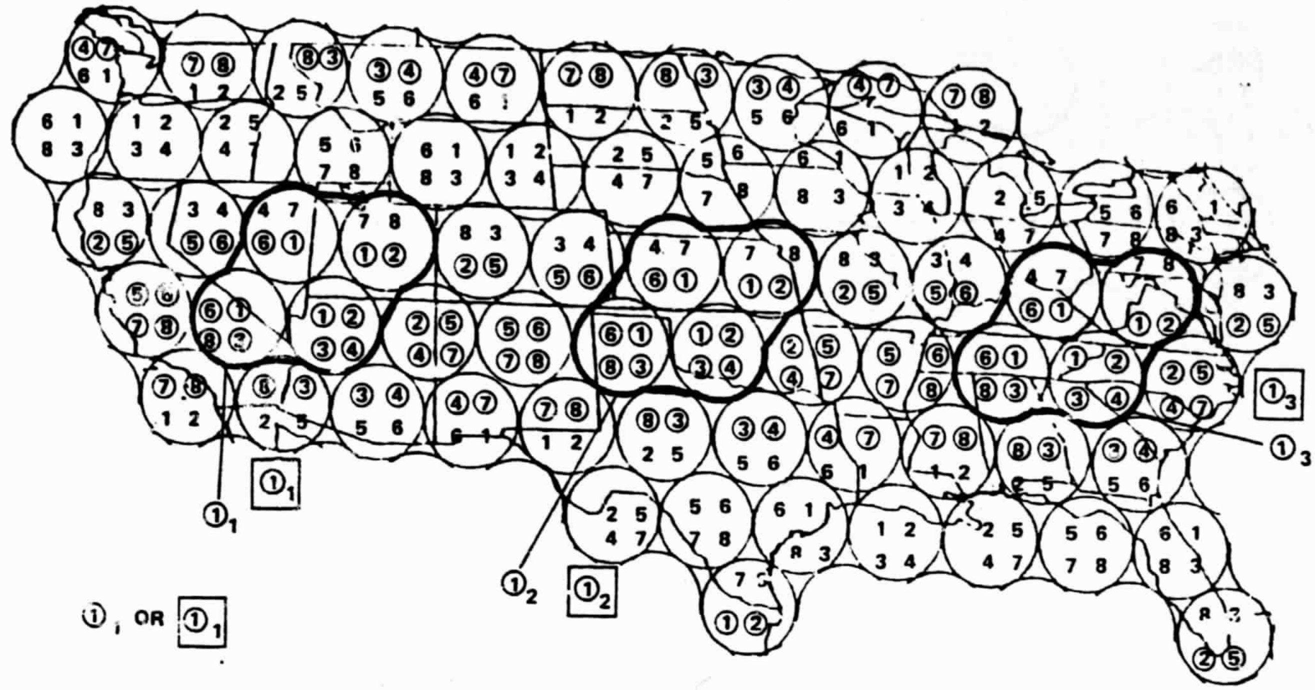


Figure 3-51. Serial No. designation of shaped beams for α = .5° Plan 5 and for modified Plan 5 (in brackets). Modification is for reduction of system complexity, resulting in N_B = 64 and N = 8. Channel ①.

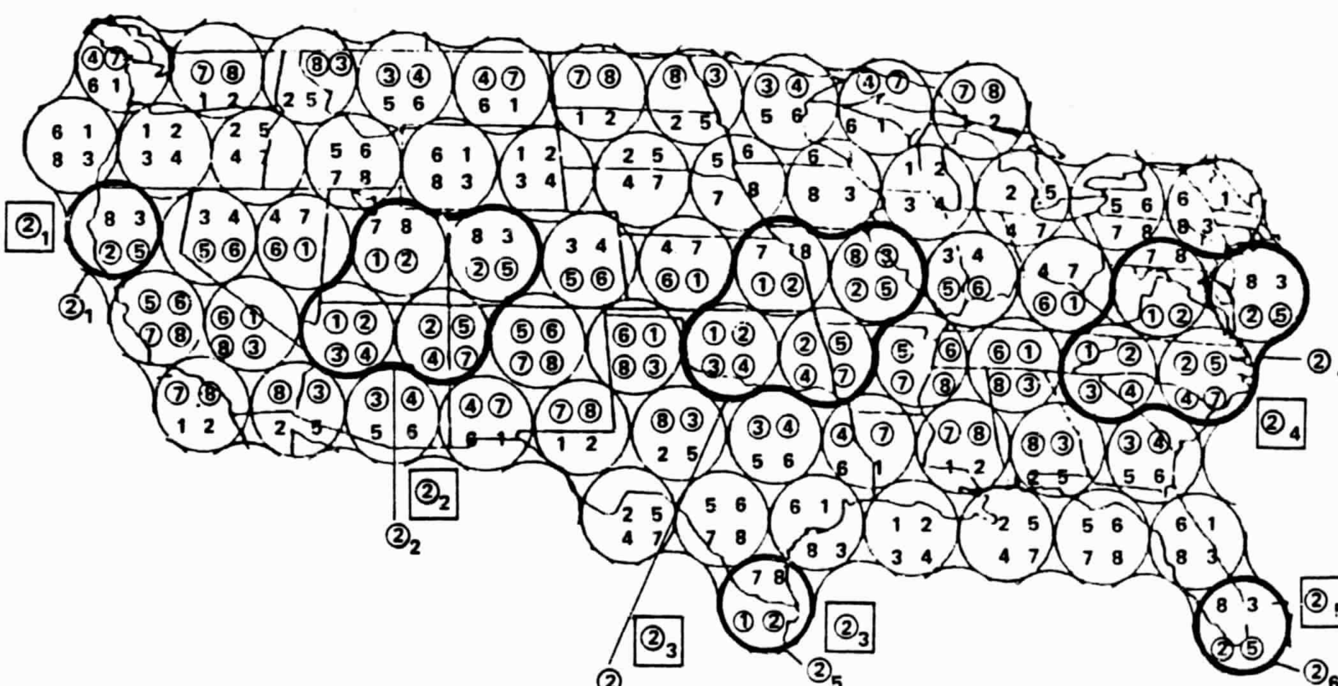


Figure 3-5m. Serial No. designation of shaped beams for α = .5° Plan 5 and for modified Plan 5 (in brackets). Modification is for reduction of system complexity, resulting in N_B = 64 and N = 8. Channel ②.

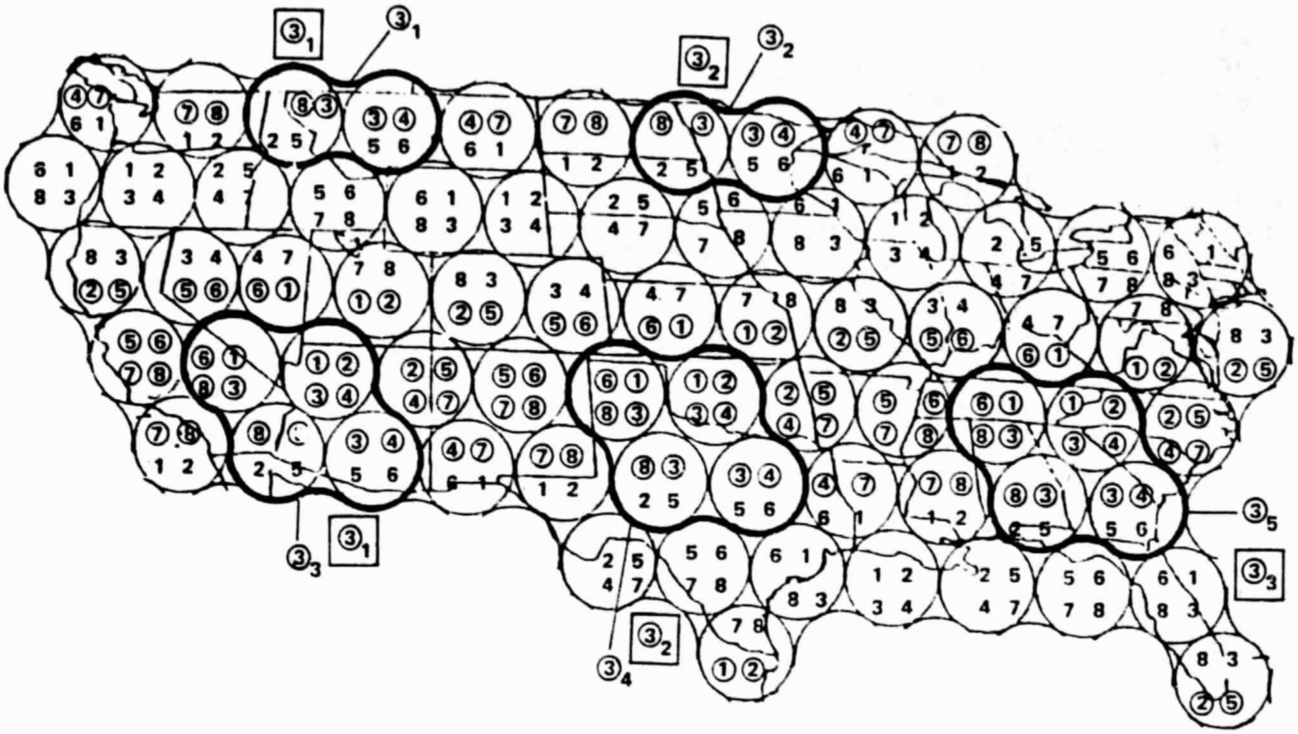


Figure 3-5n. Serial No. designation of shaped beams for α = .5° Plan 5 and for modified Plan 5 (in brackets). Modification is for reduction of system complexity, resulting in N_B = 64 and N = 8. Channel ③.

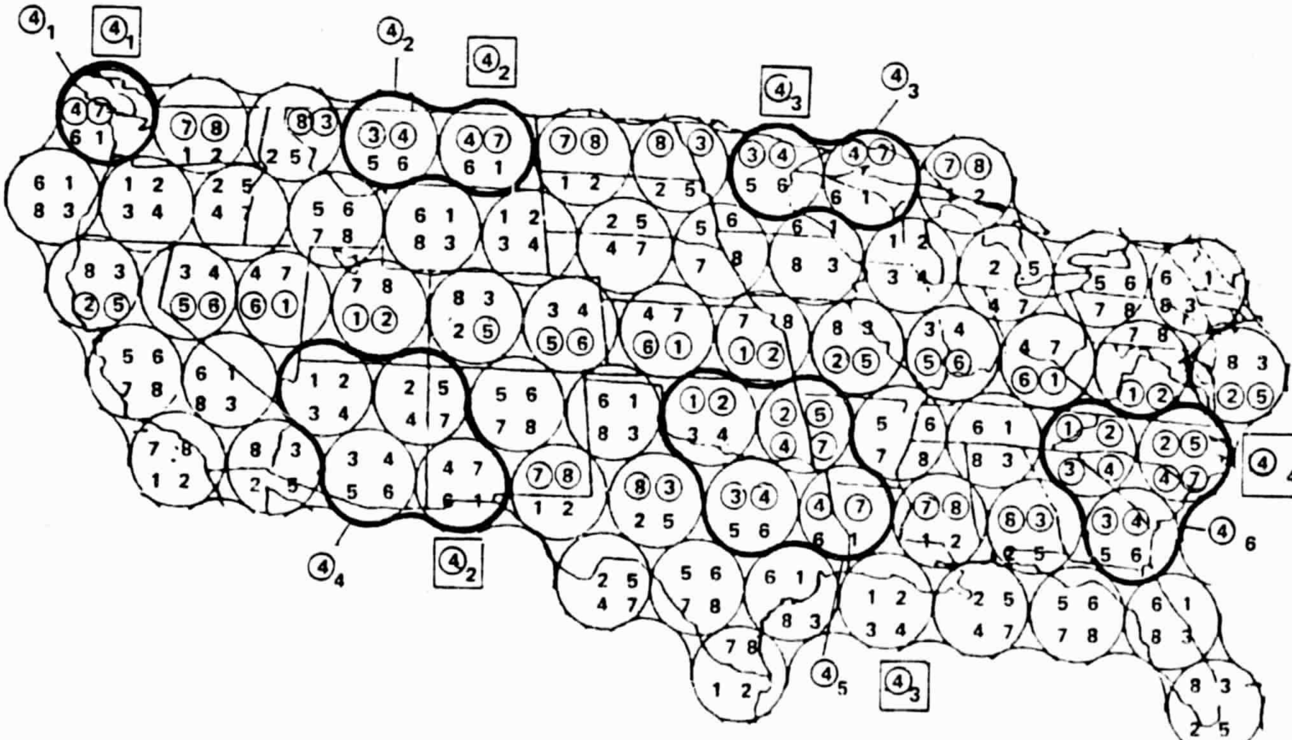


Figure 3-5o. Serial No. designation of shaped beams for α = .5° Plan 5 and for modified Plan 5 (in brackets). Modification is for reduction of system complexity, resulting in N_B = 64 and N = 8. Channel ④.

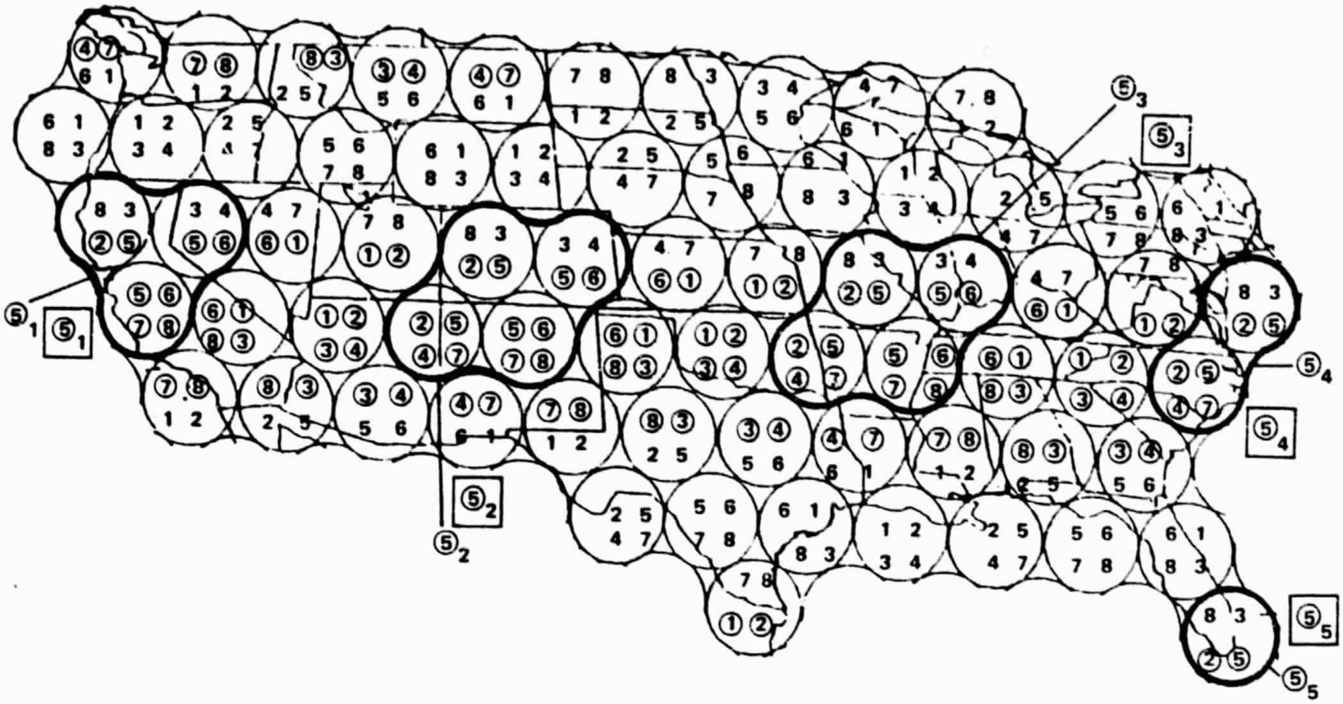


Figure 3-5p. Serial No. designation of shaped beams for α = .5° Plan 5 and for modified Plan 5 (in brackets). Modification is for reduction of system complexity, resulting in N_B = 64 and N = 8. Channel 5

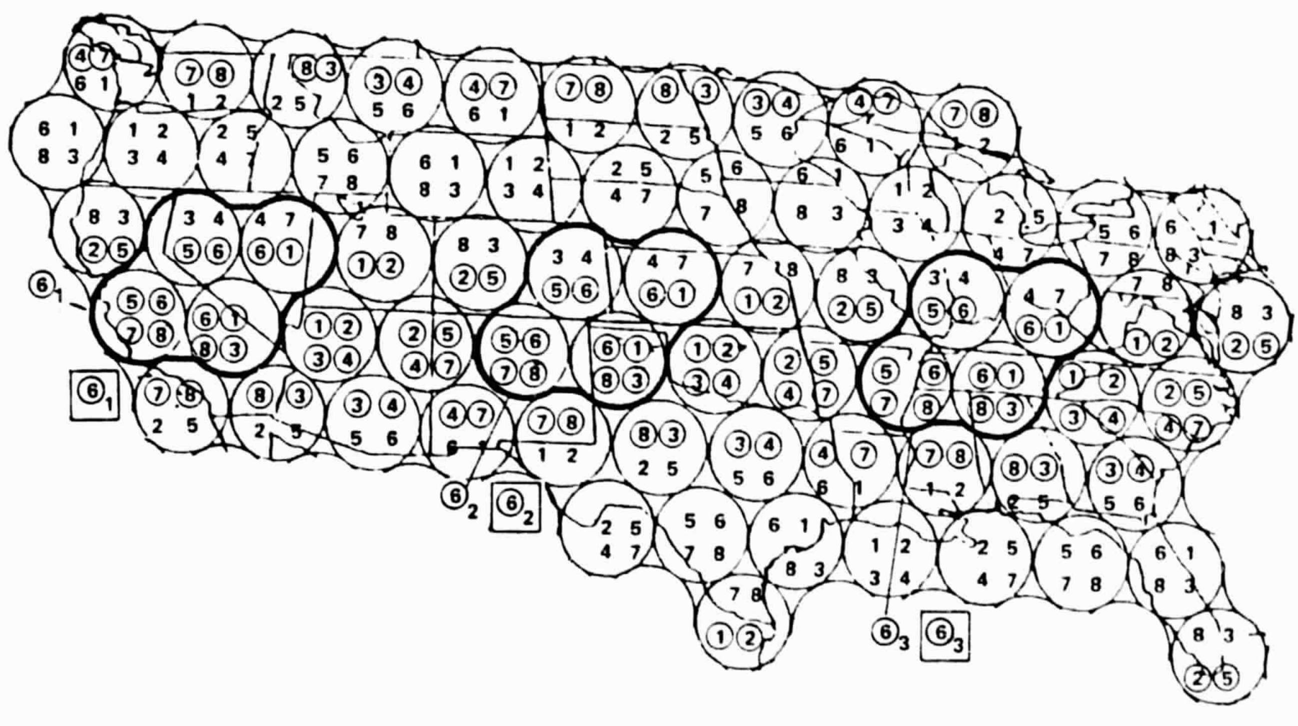


Figure 3-5q. Serial No. designation of shaped beams for α = .5° Plan 5 and for modified Plan 5 (in brackets). Modification is for reduction of system complexity, resulting in N_B = 64 and N = 8. Channel 6

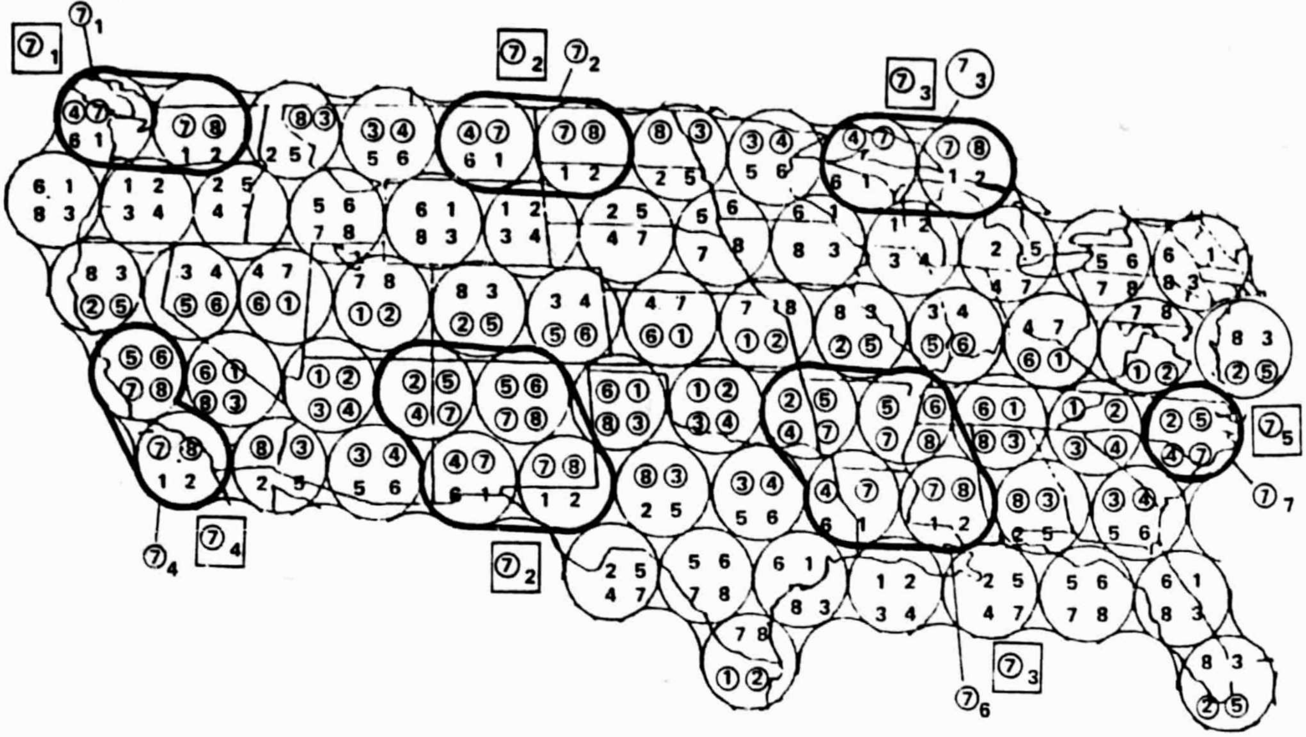


Figure 3-5r. Serial No. designation of shaped beams for α = .5° Plan 5 and for modified Plan 5 (in brackets). Modification is for reduction of system complexity, resulting in N_B = 64 and N = 8. Channel 7.

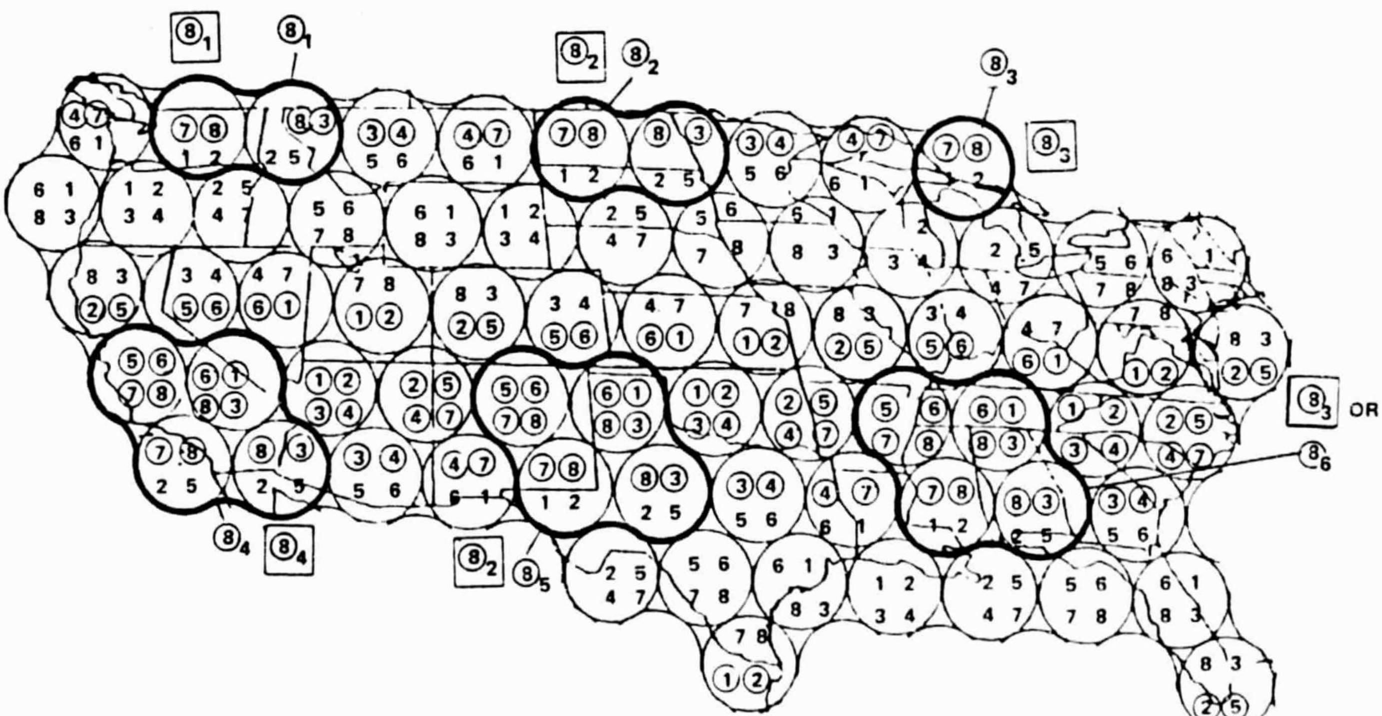


Figure 3-5s. Serial No. designation of shaped beams for α = .5° Plan 5 and for modified Plan 5 (in brackets). Modification is for reduction of system complexity, resulting in N_B = 64 and N = 8. Channel 8.

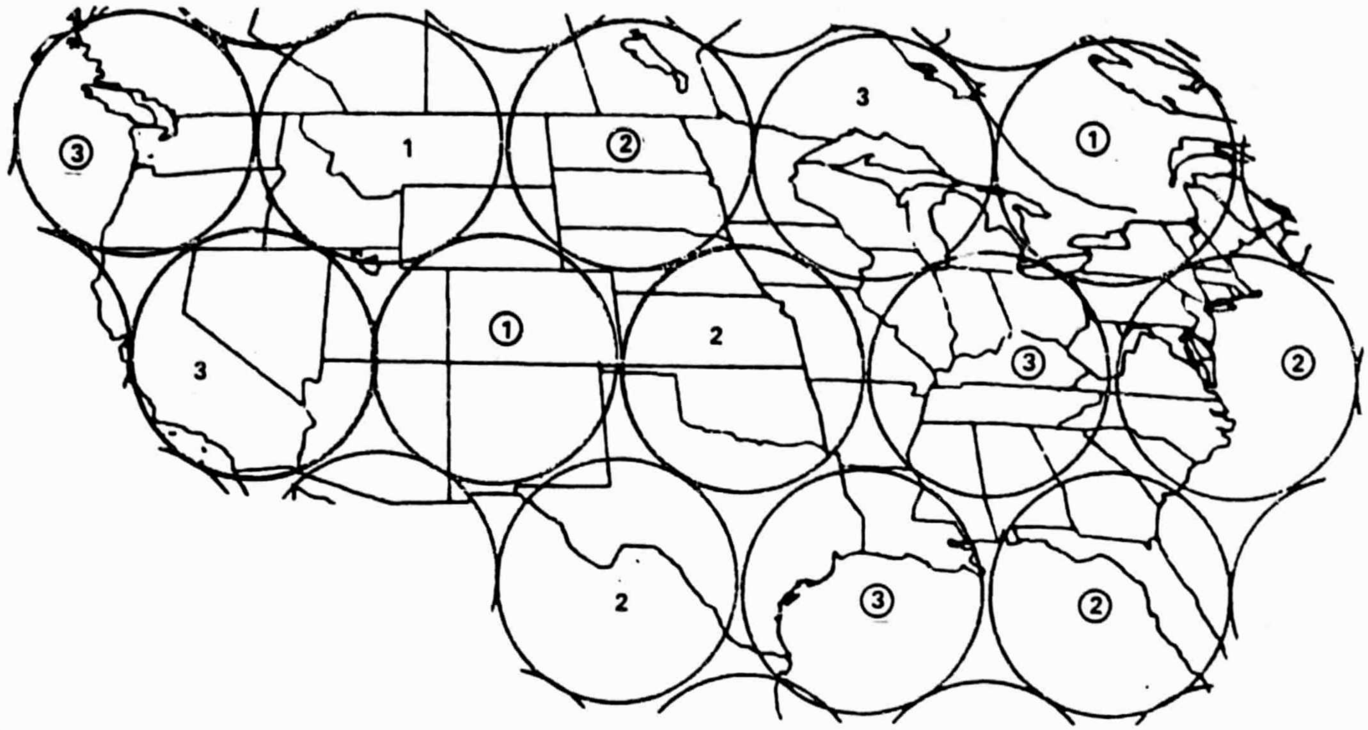


Figure 3-6a. **Beam Plan 6: triangular layout of doublets, dual polarization, polarization reuse spots.**
$\alpha = 1.3°$, $n = 14$, $N_B = 10$, $N = 3.3$.

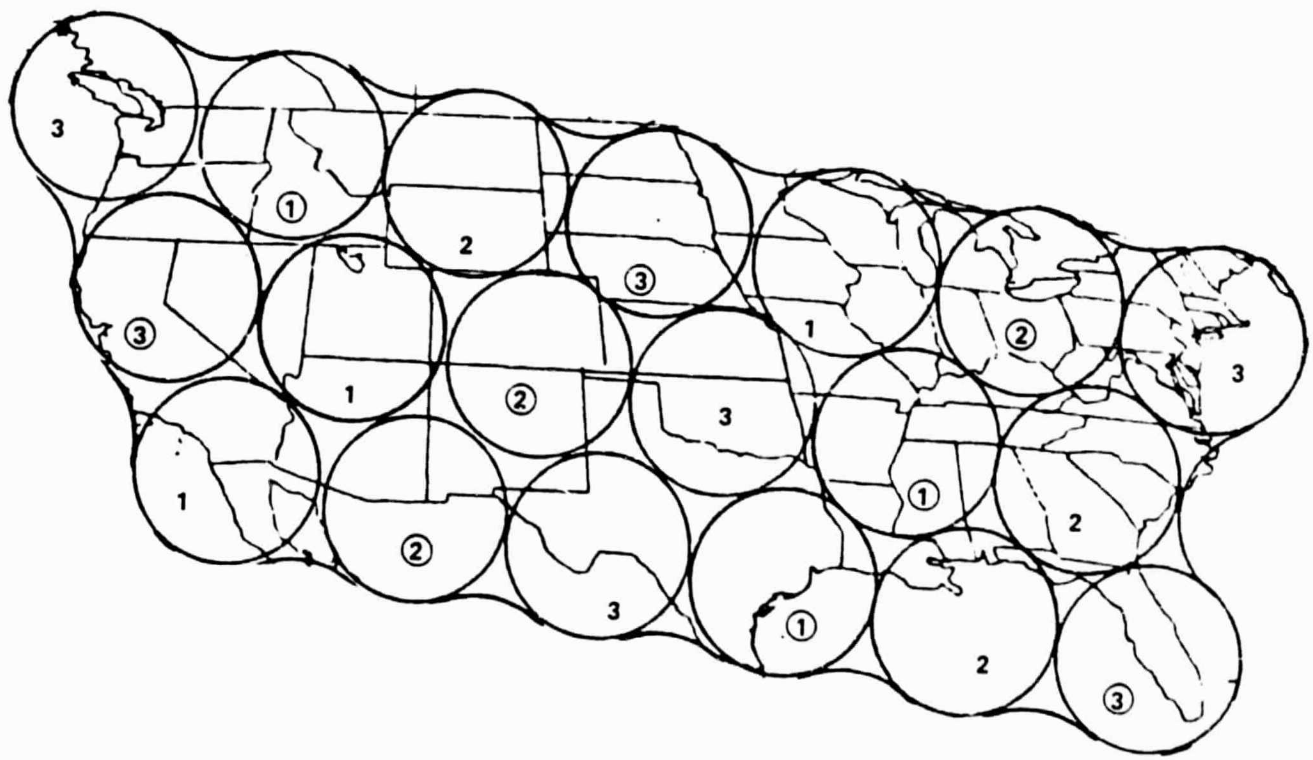


Figure 3-6b. **Beam Plan 6: triangular layout of doublets, dual polarization, polarization reuse spots.**
$\alpha = 1°$, $n = 19$, $N_B = 14$, $N = 4.67$.

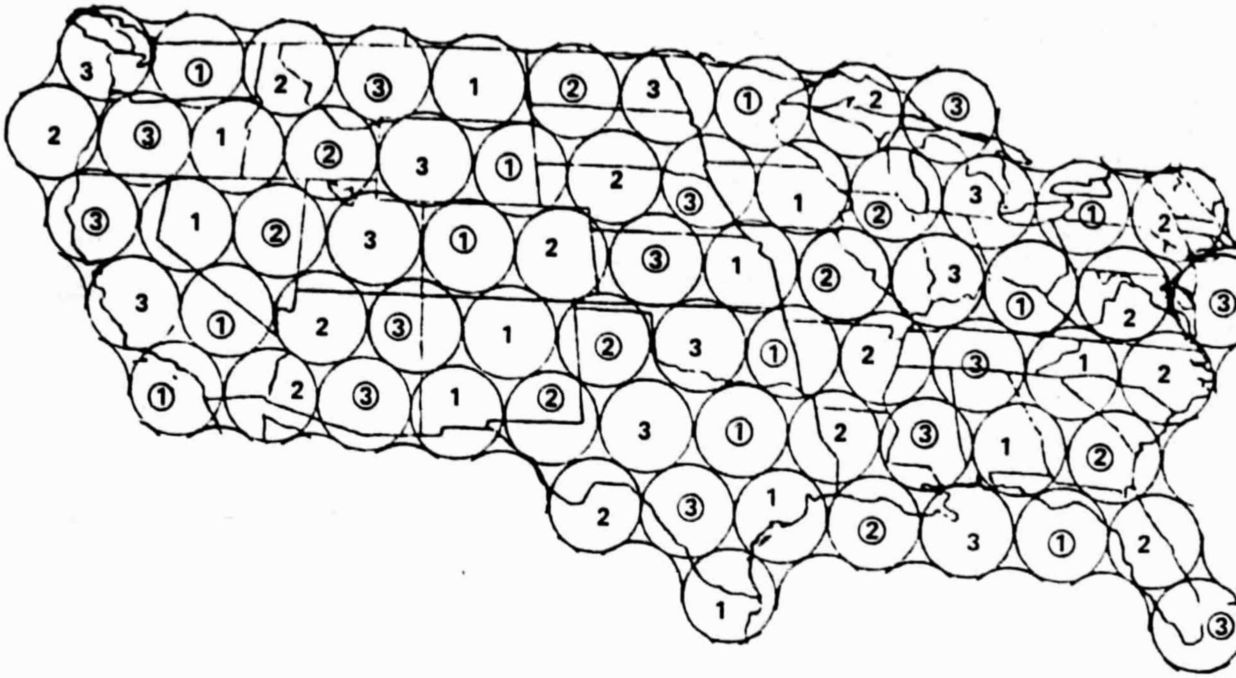


Figure 3-6c. Beam Plan 6: triangular layout of doublets, dual polarization, polarization reuse spots.
$\alpha = .5°$, $n = 68$, $N_B = 45$, $N = 15$.

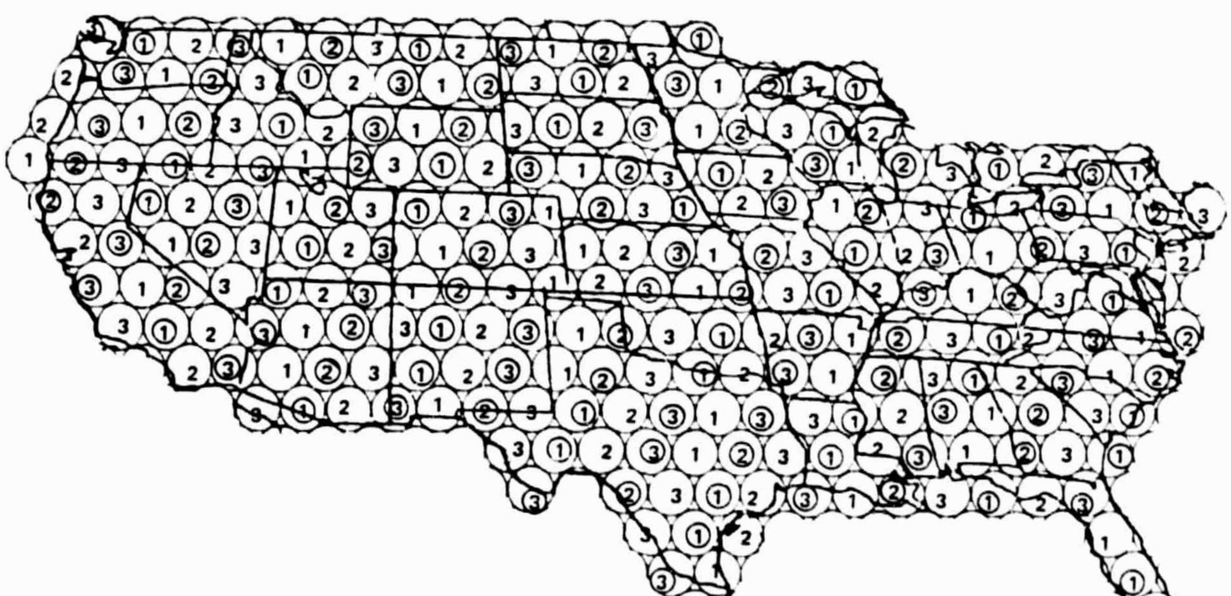


Figure 3-6d. Beam Plan 6: triangular layout of doublets, dual polarization, polarization reuse spots.
$\alpha = .25°$, $n = 251$, $N_B = 143$, $N = 47.67$.

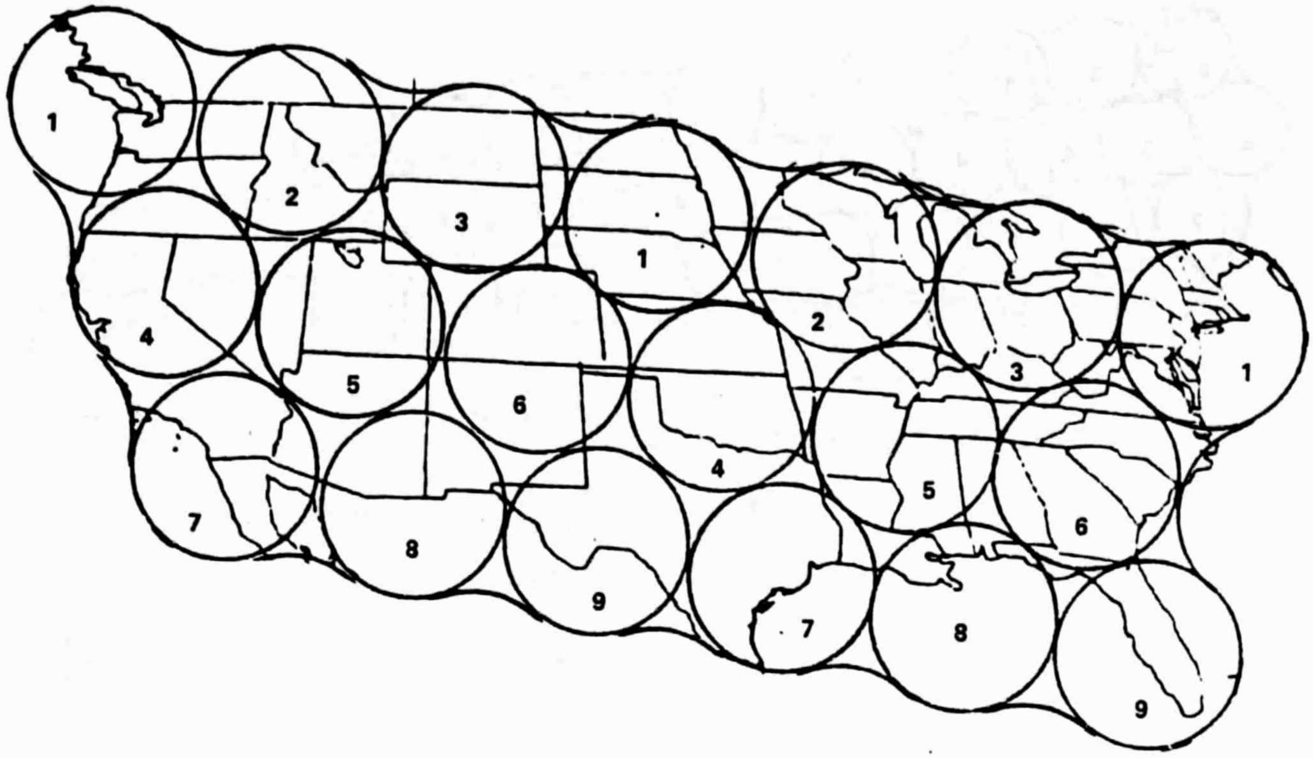


Figure 3-7a. **Beam** Plan 7: rhomboidal layout of **singlets**, single polarization. $\alpha = 1°$, $n = 19$, $N_B = 19$, $N = 2.11$.

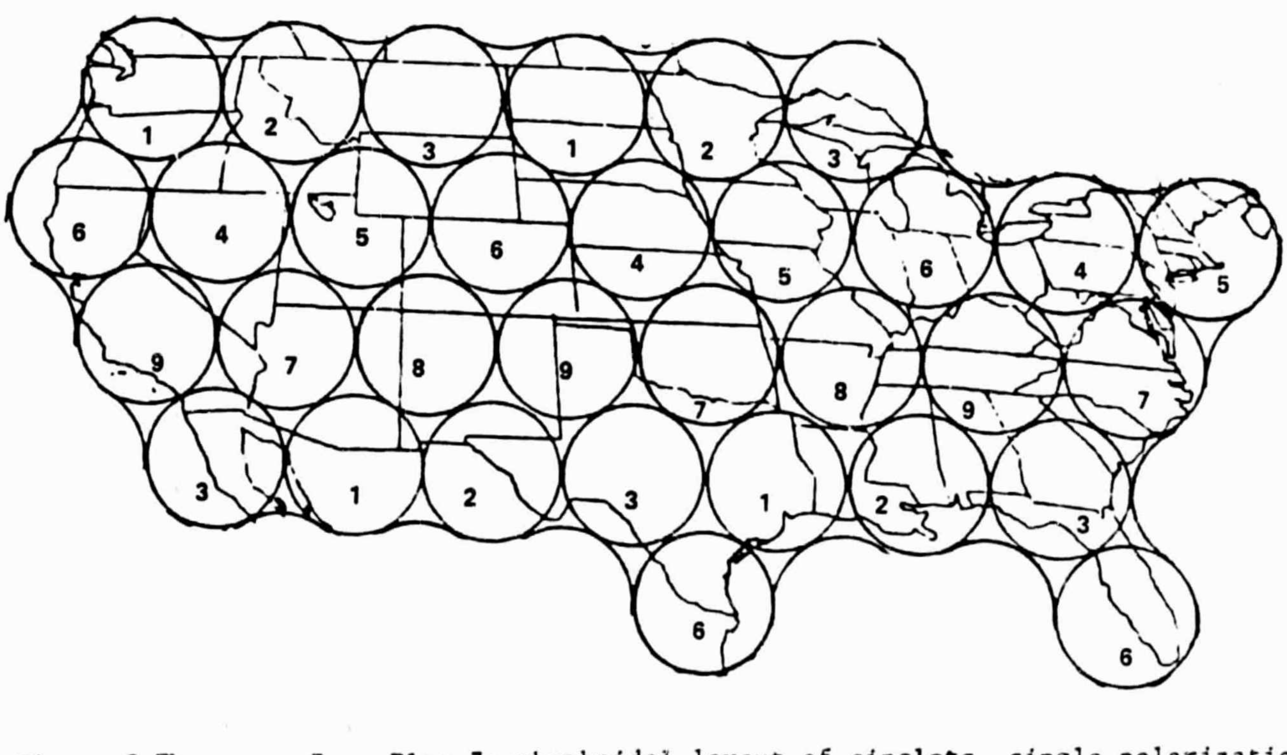


Figure 3-7b. Beam Plan 7: rhomboidal layout of **singlets**, single polarization. $\alpha = .75°$, $n = 32$, $N_B = 32$, $N = 3.55$.

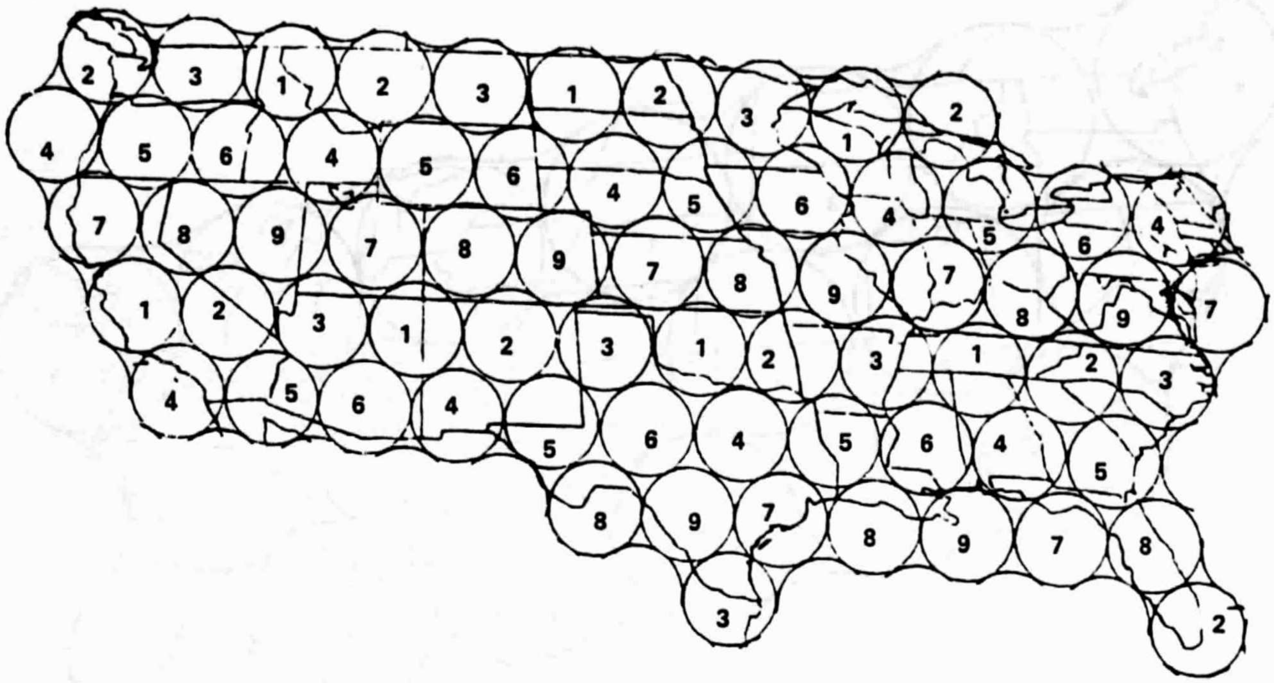


Figure 3-7c. **Beam Plan 7**: rhomboidal layout of singlets, single polarization. $\alpha = .50°$, $n = 68$, $N_B = 68$, $N = 7.55$.

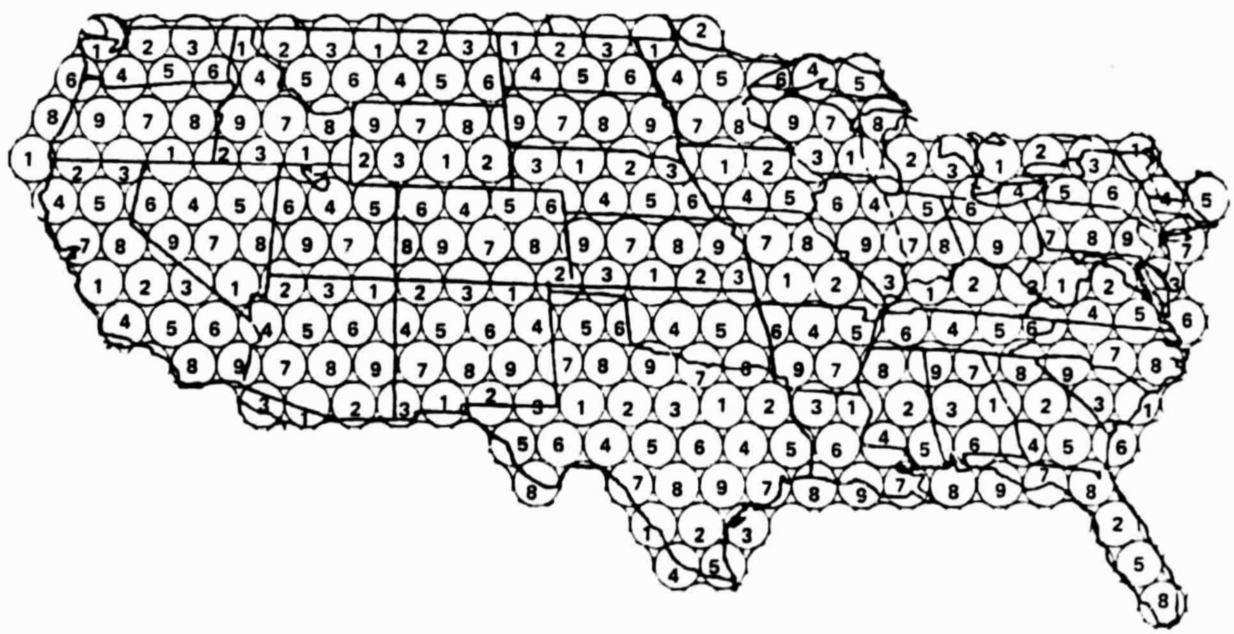


Figure 3-7d. **Beam Plan 7**: rhomboidal layout of singlets, single polarization. $\alpha = 25°$, $n = 251$, $N_B = 251$, $N = 27.89$.

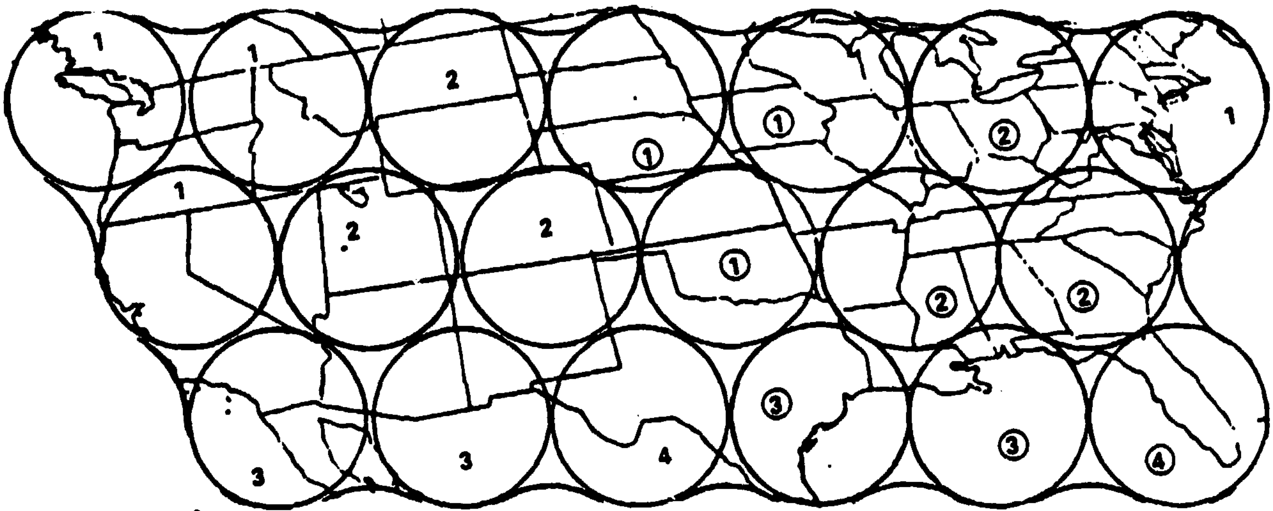


Figure 3-8a. Beam Plan 8: rhomboidal layout of triangular triplets, dual polarization.
α = .25°, n = 19, N_B = 9, N = 2.25.

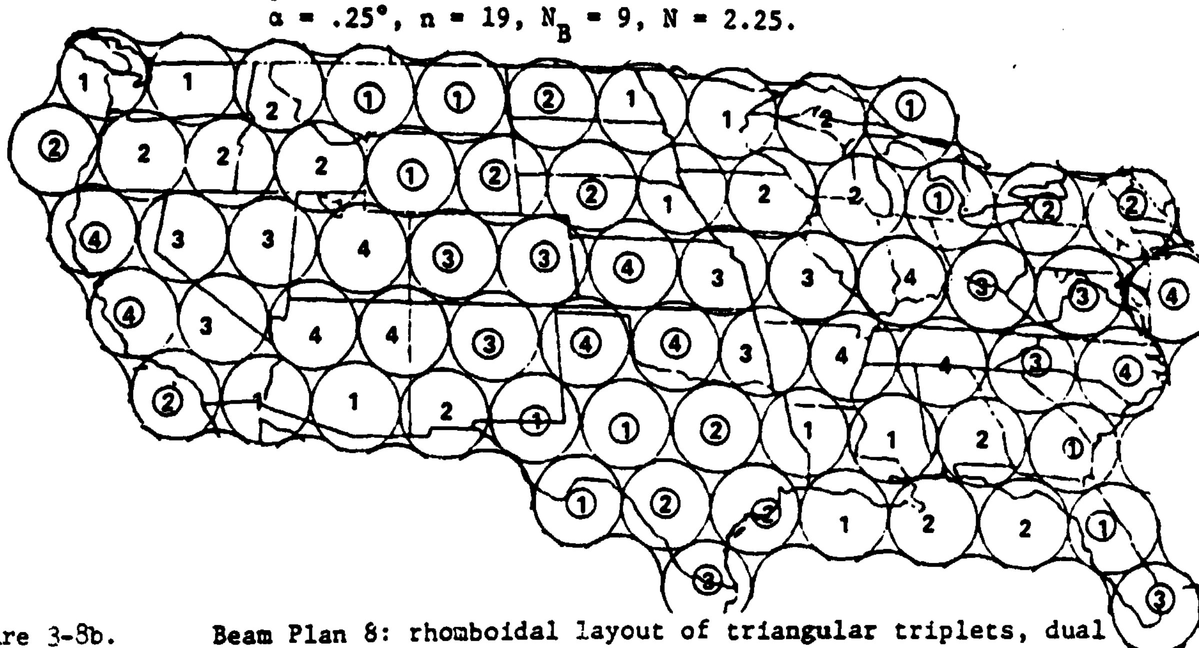


Figure 3-8b. Beam Plan 8: rhomboidal layout of triangular triplets, dual polarization.
α = .5°, n = 68, N_B = 28, N = 7.

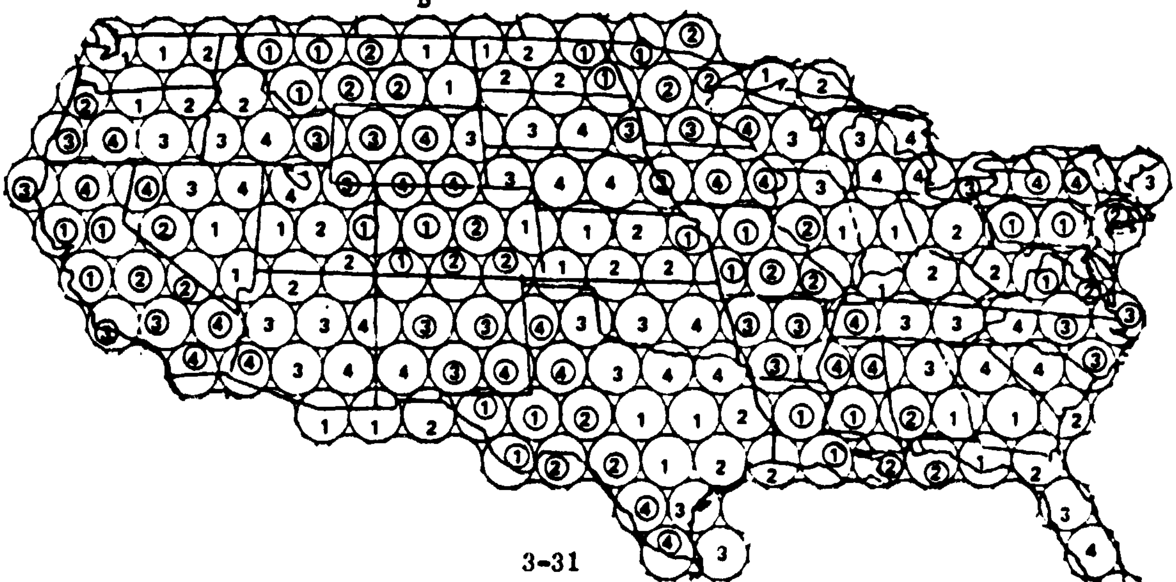


Figure 3-8c. Beam Plan 8: rhomboidal layout of triangular triplets, dual polarization.
α = .30°, n = 178, N_B = 67, N = 16.75.

$1^\circ \geq \alpha \geq .25^\circ$ range. In Table 3-2 and in the figures the bold and light or circled and uncircled numbers within a component beam footprint represent orthogonally polarized channels. For Plan 5, which has some particularly attractive characteristics all the derived shaped beams are also shown separately. In this plan the total spectrum of the communication system is divided into 8 channels, resulting in channel numbers 1, 2, . . . 8 for the vertically polarized and 1, 2, . . . 8 for the horizontally polarized channels. These channels are typically reused about 6 times, each such channel reuse is indexed to indicate the number of reuses. For instance the shaped beam providing the 3rd reuse of the horizontally polarized channel 2 is shown as 2_3.

All the plans listed in Table 3-2 provide contiguous coverage. Noncontiguous coverage for example, for off shore coverage, can be derived from these plans by simply omitting the unnecessary component beams. As will be shown later such noncontiguous plans allow the realization of larger beam isolation with the same plan than for contiguous coverage. Thus some plans may become acceptable for this case, while they are unacceptable for contiguous coverage.

Generally beam isolation can be improved by 1) increasing the antenna diameter and using increasing numbers of component beams in a shaped beam, 2) increasing the number of reflectors and thus reducing the scan angle requirement of the component beams, 3) increasing the feed diameter for a given reflector size. The last method leads to a rapid increase of antenna area. For instance an increase of feed diameter by 2 requires an increase of the number of antenna reflectors by 4. In noncontiguous systems feed diameter increase may be implemented with no or only modest increases in the number of required antenna reflectors. Thus noncontiguous coverage systems generally can be designed with much more flexibility.

Figure 3-9 shows examples for noncontiguous beam plans which provide 10 to 12 spot or shaped beams with the use of main and auxiliary horns. The figures also exhibit the percent of total frequency band available in the individual shaped beams. It is

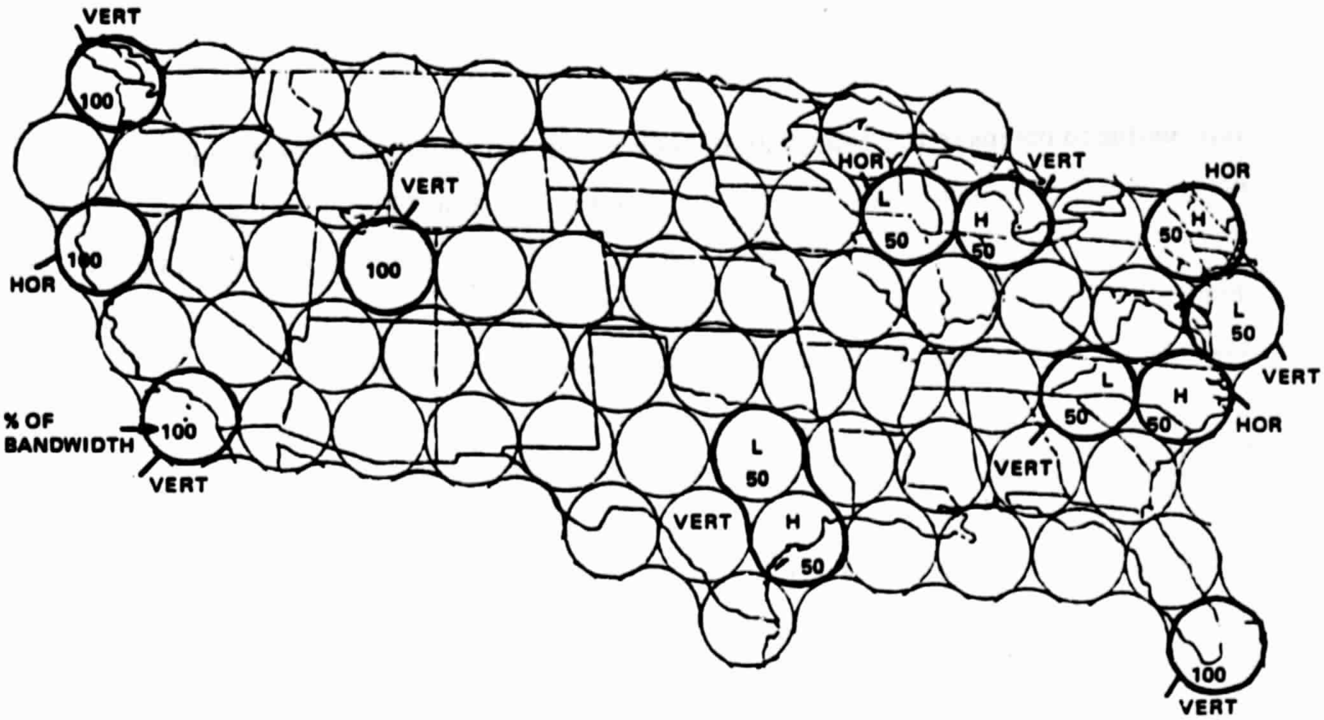


Figure 3-9a. Example for noncontiguous beam plan, assuming "large horn" implementation.
$\alpha = .5°$, $N_B = 12$, $N = 9$.

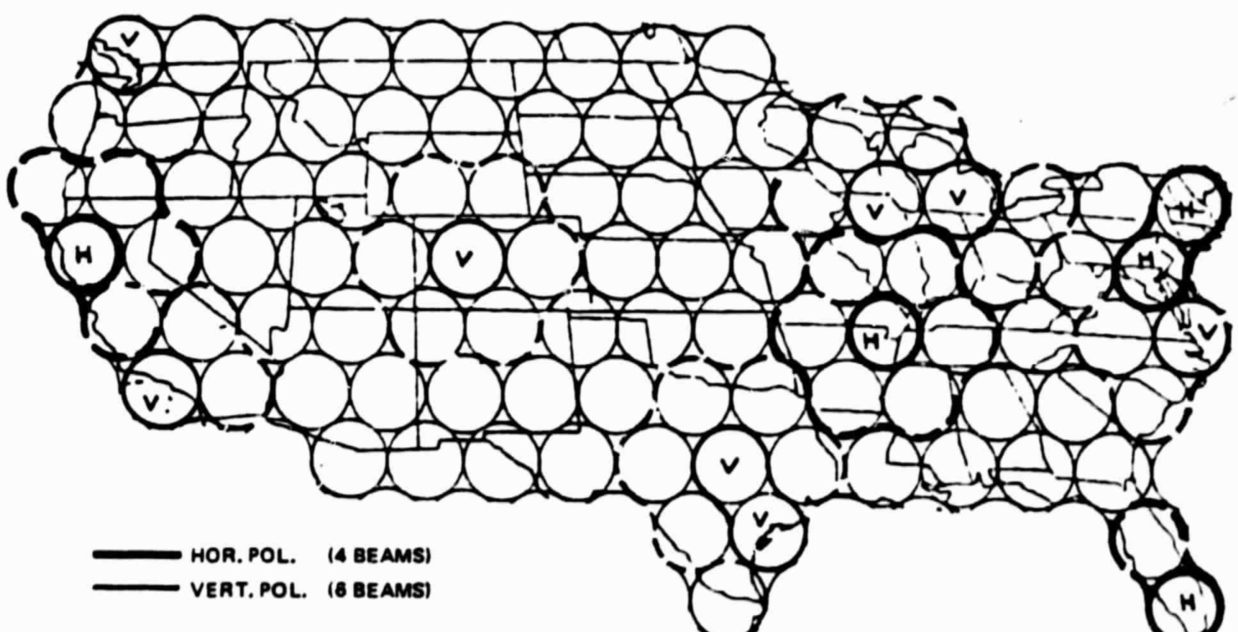


Figure 3-9b. Example for noncontiguous beam plan, assuming shaped beam implementation using a total of 13 main and 31 auxiliary horns.
$\alpha = .41°$, $n = 44$, $N_B = 10$, $N = 10$.

interesting to compare the capability of such a noncontiguous plan with a contiguous configuration with non uniform traffic distribution based on Plan 5, in Figure 3-10. It can be seen that the available bandwidth in the "heavy route" cells corresponding to Figure 3-9a is reduced by a factor of 0.37 to 0.88, but in the "thin route" cells an average of 12.3% of the total system bandwidth is still available.

3.4 COMPUTED GAIN CONTOURS FOR VARIOUS BEAM TOPOLOGIES

According to the data presented in the previous section the contiguous coverage of the lower 48 states with the use of cell size component beams $1^\circ \geq \alpha \geq .25^\circ$ requires $19 \geq n \geq 251$ component beams. From earlier studies it is evident that the offset fed paraboloid reflector with F/D near unity and illuminated by a multi-element feed array is a good candidate for this application. Indeed this optical configuration has such flexibility in the synthesis of its radiation field that its acknowledged disadvantage in asymmetry can be nearly compensated. Once the effect of asymmetry is compensated its advantages, like simplicity, low loss and lightweight compare very favorabley with other configurations like symmerrically fed lenses or phased arrays using high-directivity individual elements. The offset fed antenna also is capable of large bandwidth, and its mechanical configuration is usually convenient for satellite applications.

On the basis of these considerations the present study considers only offset fed optical structures, (the generalization of the results to symmetrical optics (lenses and arrays) is straightforward).

The assumed basic optical configuration is shown on Figure 3-11. In this arrangement the antenna is characterized by the following parameters:

- D = the projected aperture diameter of the paraboloid
- F/D = focal length to diameter ratio
- Q = offset or distance between axis of paraboloid and its closest point to the axis

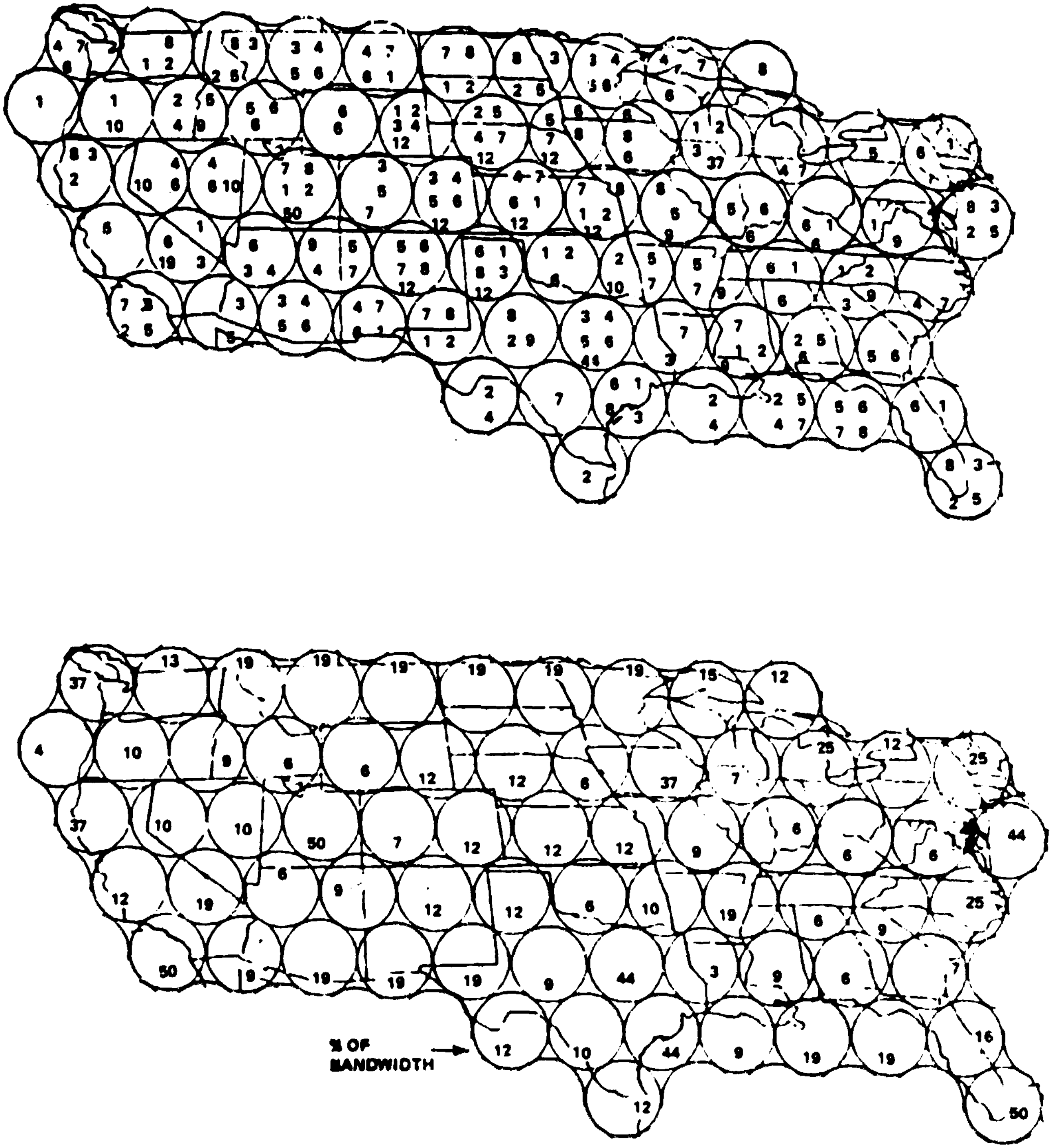


Figure 3-10. Example for contiguous beam plan providing traffic density matched bandwidth capability. Nonuniform channel distribution is obtained by modifying Plan 5.
$\alpha = .5°$, $n = 68$, $N_B = 87$, $N = 11.73$.
Average bandwidth in 13 heavy traffic route cells: $.382\Delta F$
Average bandwidth in 55 thin traffic route cells: $.123\Delta F$

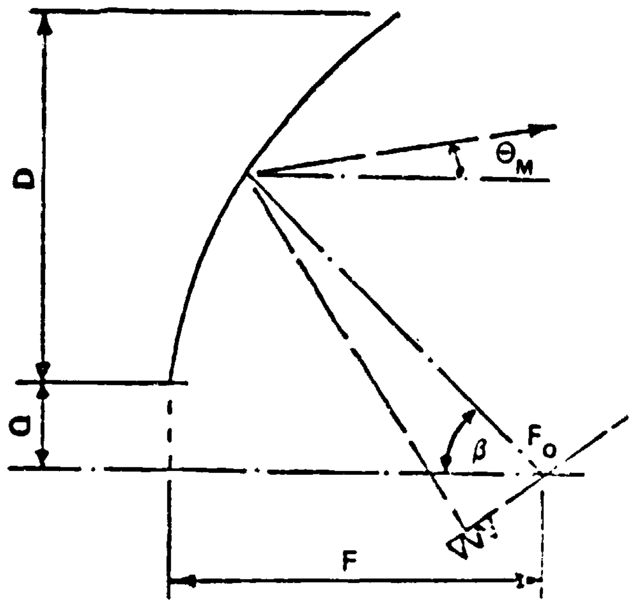


Figure 3-11. Basic Geometry of Antenna Configuration Using Offset Fed Paraboloid Optics.

d	=	diameter of feed horn
β	=	tilt angle between axis of feed horn and axis of paraboloid
n	=	number of feed horns
θ_M	=	angle between axis of paraboloid and axis of the component beam associated to a feed horn (scan angle).

Given the above geometrical parameters and the field excitement in the aperture of the horns the radiation performance of the antenna is uniquely determined. The field excitement in the horns usually is given by the type, polarization and relative complex amplitude of waveguide modes in the aperture. Generally more than one waveguide mode is necessary to characterize a desired secondary pattern. The most frequently used modes are the TE_{11} (basic) mode and the TM_{11}, TE_{21} and TM_{01} higher order modes.

The following discussions are restricted to $F/D \simeq 1$. In this case the use of higher order modes is not important up to about $d_\lambda \sim 1.8$ and $D_\lambda \sim 67$. For such conditions the peak crosspolarized power level in the secondary pattern will be better than -28 db for circular polarization (CP) and better than -25 db for linear polarization (LP) below the peak of the main beam. However, for larger antenna directivities or better cross-polarization levels the use of the TM_{11} mode is desirable for CP and additionally the use of the TE_{21} and TM_{01} mode is desirable for LP depending on whether the direct of the main polarization is in the plane perpendicular or parallel with the plane of the offset. Table 3-4 show some typical polarization characteristics achievable with solid surface reflectors.

A few results will be given for higher order mode excitations, but most of the calculations were carried out only for TE_{11} mode excitation. This restriction does not significantly influence the results for the main polarization but reduces the cost of computations. Furthermore it will be shown that for the type of beam topologies under consideration even the crosspolarized levels are acceptable with TE_{11} mode excited circular borns.

All the calculations were made using a GE developed vector field based computer program. This program neglects the phase error in the aperture of horns and treats them like open waveguides. Furthermore it neglects mutual coupling between horns or the effect of currents flowing on the outside wall of the horns, rear of the reflectors or supporting structures. However, all other effects were considered, resulting in a peak gain accuracy of better than 0.3 db and sidelobe level accuracy at the -30 db level of approximately 2 db.

Figure 3-12 shows the main and crosspolarized gain contour plots of a so-called low sidelobe singlet. This is achieved by using a single, large, multi-mode horn. (See the mode power ratios in Figure 3-12.) Note, that for the exhibited ideal (unscanned) case the first sidelobe level is about -32 db and the peak crosspolarized level is about

TABLE 3-4. PEAK CROSSPOLARIZATION LEVELS IN SYMMETRICAL AND OFFSET FED PARABOLOIDS USING CIRCULAR APERTURE FEEDS

Feed Diameter:	$d = 3.32$ in
Paraboloid Diameter	$D = 95$ in or 134 in ($F/D = 1$)
Paraboloid Offset Ratio:	$Q/D = .26$[illegible]
Aperture Angle of Paraboloid:	$\psi =$ [illegible]°

Primary Pattern Crosspolarization Below Main Polarization at $\psi = 25^\circ$ (db)

Waveguide Mode in Feed Aperture	f^{MHz}	Calculated LP	Calculated CP	Measured LP
TE_{11}	6325	22.1	22.2	--
$TE_{11} + .0172\ TM_{11}$	6325	47.1	47.9	45
TE_{11}	6725	21.4	21.5	--
$TE_{11} + .0172\ \angle 20^\circ\ TM_{11}$	6725	29.2	30.0	30
Hor. Pol. $TE_{11} + .0172\ TM_{11} + .021\ \angle 90^\circ\ TE_{21}$	6325	18.6		
Vert. Pol. $TE_{11} + .0172\ TM_{11} + .003\ \angle -90^\circ\ TM_{01}$	6325	18.8		
Hor. Pol. $TE_{11} + .0172\ \angle 20^\circ\ TM_{11} + .021\ \angle 110^\circ\ TE_{21}$	6725	19.1		
Vert. Pol. $TE_{11} + .0172\ \angle 20^\circ\ TM_{11} + .003\ \angle -70^\circ\ TM_{01}$	6725	18.7		

Secondary Pattern Crosspolarization Below Main Polarization (db) Ideal Polarizer

Waveguide Mode in Feed Aperture	f^{MHz}	D^{in}	$\theta_M = 0^\circ$ (Beam at Axis of Paraboloid) Sym. Paraboloid LP	$\theta_M = 0^\circ$ Offset Paraboloid LP	$\theta_M = 0^\circ$ Offset Paraboloid CP	$\theta_M = 6.4^\circ$ (Beam Scanned to Edge of Earth) Offset Parabolid) LP	$\theta_M = 6.4^\circ$ Offset Parabolid) CP
TE_{11}	6325	95	28.6				
		134	28.6		30.6		25.5
$TE_{11} + .0172\ TM_{11}$	6325	95	63.0	24.3			
		134		24.4	51.7		49.4
$TE_{11} + .0172\ \angle 20^\circ\ TM_{11}$	6725	95					
		134	39.0	24.1	39.0		34.3
Hor. Pol. $TE_{11} + .0172\ TM_{11} + .021\ \angle 90^\circ\ TE_{21}$	6325	134		45.5		41.5	
Vert. Pol. $TE_{11} + .0172\ \angle 20^\circ\ TM_{11} + .003\ \angle -90^\circ\ TM_{01}$	6325	134		48.0		45.5	
Hor. Pol. $TE_{11} + .0172\ \angle 20^\circ\ TM_{11} + .021\ \angle 110^\circ\ TE_{21}$	6725	134		36.7		31.1*	
Vert. Pol. $TE_{11} + .0172\ \angle 20^\circ\ TM_{11} + .003\ \angle -70^\circ\ TM_{01}$	6725	134		30.8		31.9*	

Secondary Pattern Crosspolarization Below Main Polarization (db) Measured Polarizer

Waveguide Mode in Feed Aperture	f^{MHz}	Axial Ratio of Polarizer db	$\theta_M = 0$	$\theta_M = 6.4^\circ$
$TE_{11} + .0172\ TM_{11}$	6325	.35	33.7	33.6*
$TE_{11} + .0172\ \angle 20^\circ\ TM_{11}$	6725	.25	31.9	29.5*
Hor. Pol. $TE_{11} + .0172\ \angle 20^\circ\ TM_{11} + .021\ \angle 110^\circ\ TE_{21}$	6725	.25	30.6	27.36
Ver. Pol. $TE_{11} + .0172\ \angle 20^\circ\ TM_{11} + .003\ \angle -70^\circ\ TM_{01}$	6725	.25	27.07	27.87

*A minimum of 2 db improvement is possible with multihorn cluster feed.

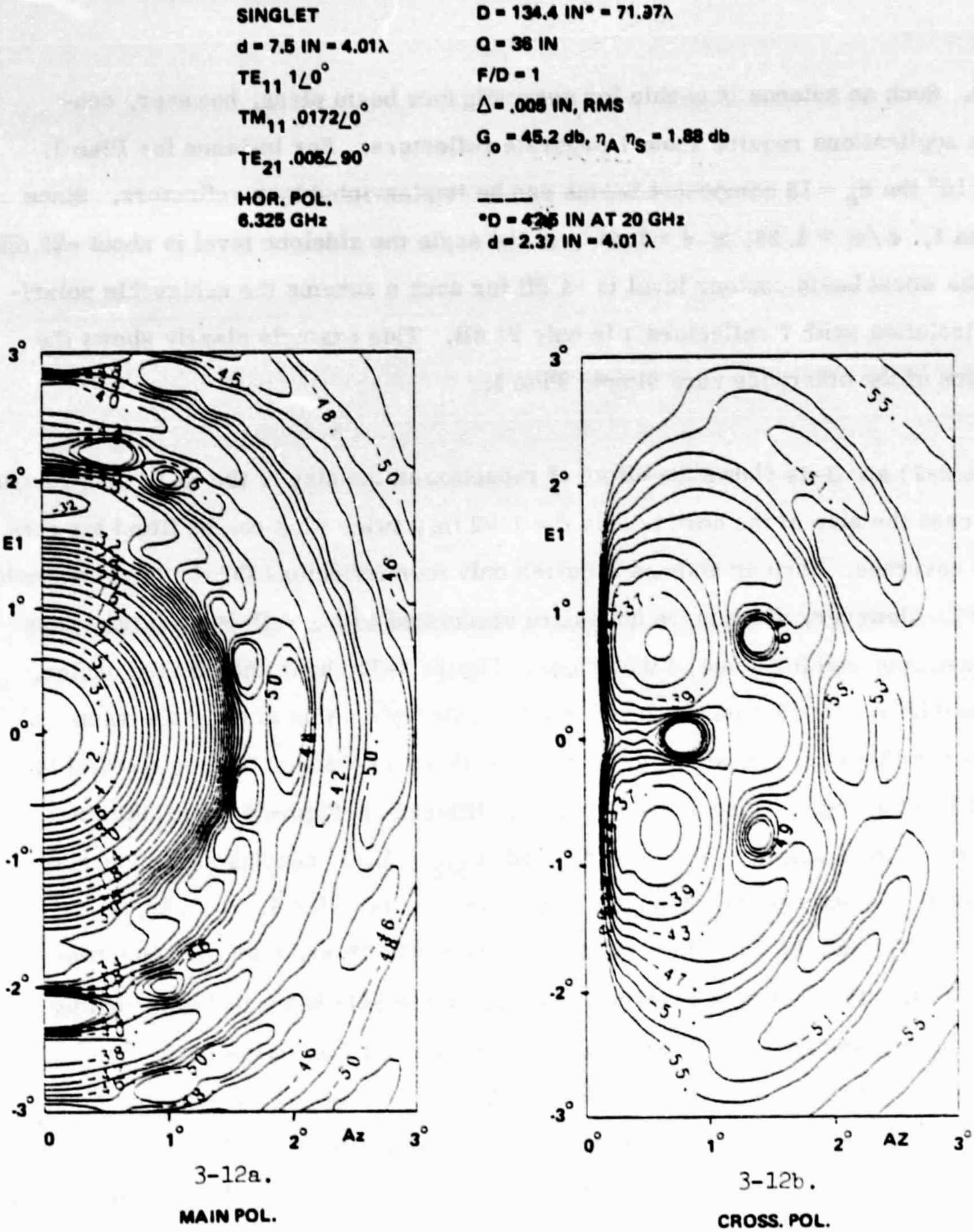


Figure 3-12a. Gain contour plot of an unscanned, low sidelobe level singlet achieved by a large diameter (d_λ = 4.01) multimode horn. Main polarization.

Figure 3-12b. Gain contour plot of an unscanned, low sidelobe level singlet achieved by a large diameter (d_λ = 4.01) multimode horn. Cross polarization.

-35 dB. Such an antenna is usable for noncontiguous beam plans, however, contiguous applications require about 7 separate reflectors. For instance for Plan 1, $\alpha = 1.15°$ the $n_c = 15$ component beams can be implemented by 7 reflectors. Since for Plan 1, $\epsilon/\alpha = 1.23$, at $\epsilon = 1.41°$, at the angle the sidelobe level is about -25 dB. Since the worst beam coutour level is -4 dB for such a scheme the achievable polarization isolation (with 7 reflectors!) is only 21 dB. This example clearly shows the limitation of the otherwise very simple Plan 1.

Figures 3-13 and 3-14 shows the effect of reduction in the size of the multi-mode horn. In this case the size of the horn is only d = 1.82 in., twice what can be fitted for contiguous coverage. Such an antenna requires only four reflectors for the implementation of Plan 1. However, since more horns are associated with a reflector, larger scan angles are required for some of the beams. Figure 3-13 shows that for the center, unscanned beam -32 dB sidelobe level is still schievable, (this is about the same value as for Figure 3-12), although the polarization was changed from horizontal to vertical and the required mode distribution is different.) Figure 2-14 exhibits a singlet which is scanned to $\theta_{MA} = 2.95°$ and $\theta_{ME} = 1.30°$ causing a considerable distortion of the main beam. If such a singlet is used for Plan 1, $\alpha = .5°$, $\epsilon = 1.23\ \alpha = .615°$ and the sidelobe level is only about -13 dB and the beam isolation is only .9 dB. This example illustrates that the achievable beam isolation rapidly deteriorates as the number of reflectors are reduced for the "large horn" implementation of Plan 1. It may be noted that the situation is somewhat better for the "multiple horn cluster" implementation of Plan 1, but the basic limitation remains because of the small ϵ/α ratio, (for the multiple horn cluster implementation, i.e., one main and several auxiliary horns, see Reference 1.)

Up to now the horn generating the singlet beam was excited by an ideal combination of higher order modes. In the following TE_{11} mode excitation will be assumed. Figure 3-1a shows the topology layout for Plan 1 using $\alpha = 1.5°$. For such a large α angle the required maximum beam scan is only $\theta_M/\alpha \sim 2$ thus the beam distortion is

SINGLET
$d = 1.82$ IN $= 3.08\lambda$
$TE_{11}\ 1\angle 0^\circ$
$TM_{11}\ .05\angle 0^\circ$
$TM_{01}\ .001\angle 0^\circ$
VERT. POL. MAIN
20 GHz

$D = 96$ IN $= 162.96\lambda$ ($G^x = 54.16$ db)
$Q = 6.7$ IN
$F/D = 1$
$G_o = 53.3$ db, $\eta_A\ \eta_S = .86$ db
$\alpha = .5^\circ$ (4 REFLECTORS FOR CONT. COVERAGE)
$\Theta_{MA} = \Theta_{ME} = 0$
$\Theta_3 = .425^\circ$

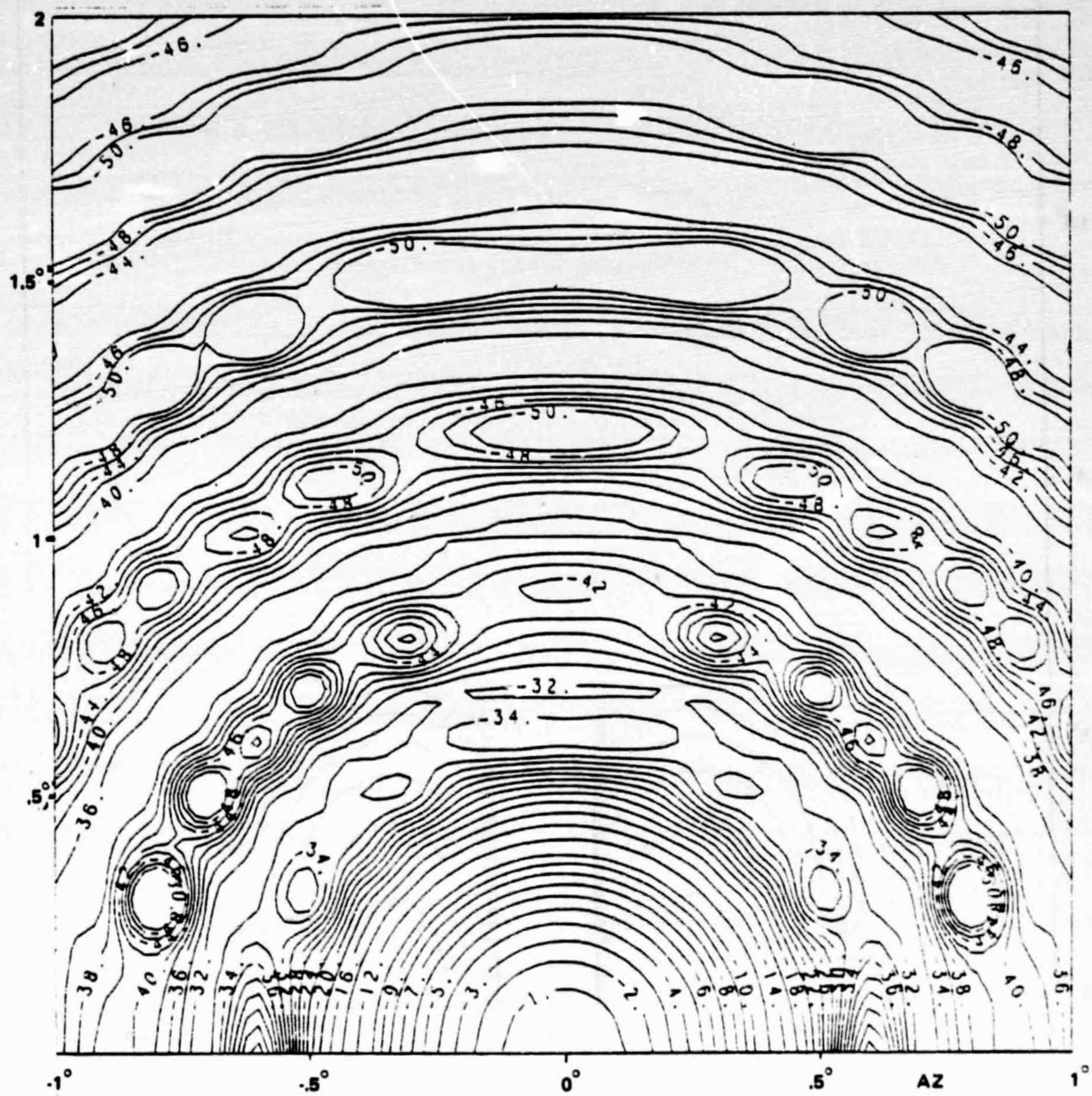


Figure 3-13a. Gain contour plot of an unscanned, low sidelobe level singlet, achieved by a large diameter ($d_\lambda = 3.08$) multimode horn. Main polarization.

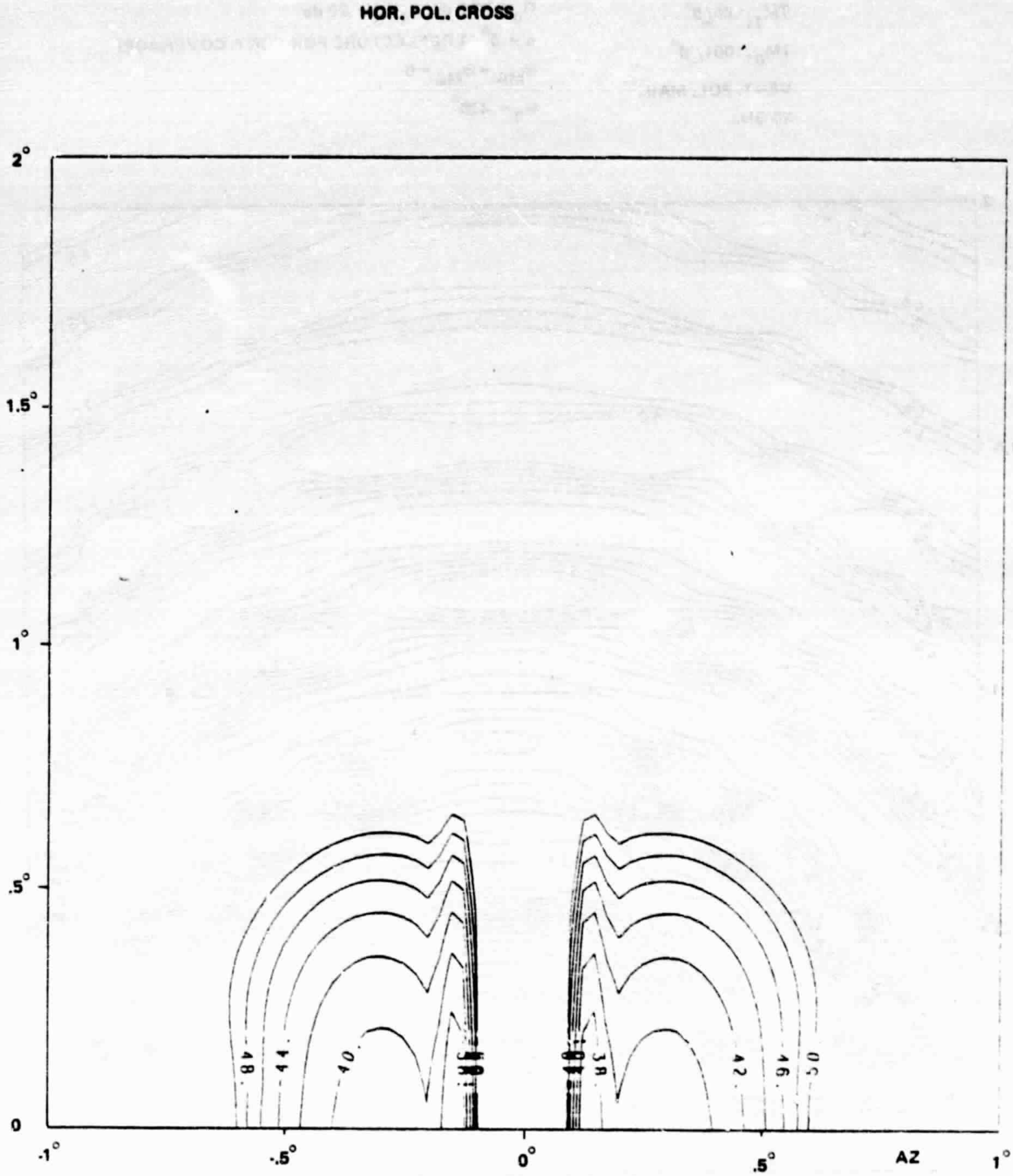


Figure 3-13b. Gain contour plot of an unscanned, low sidelobe level singlet, achieved by a large diameter (d_λ = 3.08) multimode horn. Cross polarization.

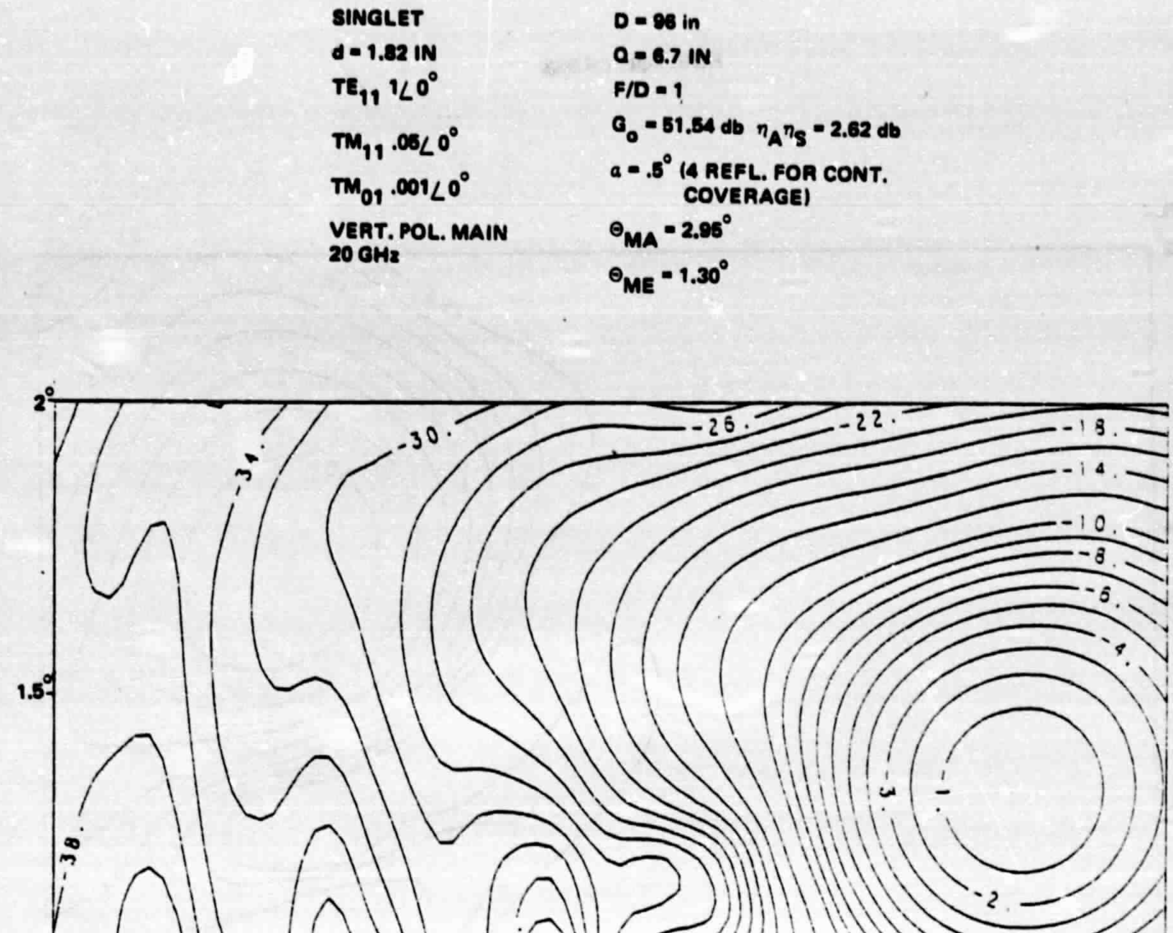


Figure 3-14a. Gain contour plot of a scanned, low sidelobe level singlet, achieved by a large diameter (d_λ = 3.08) multimode horn. Main polarization.

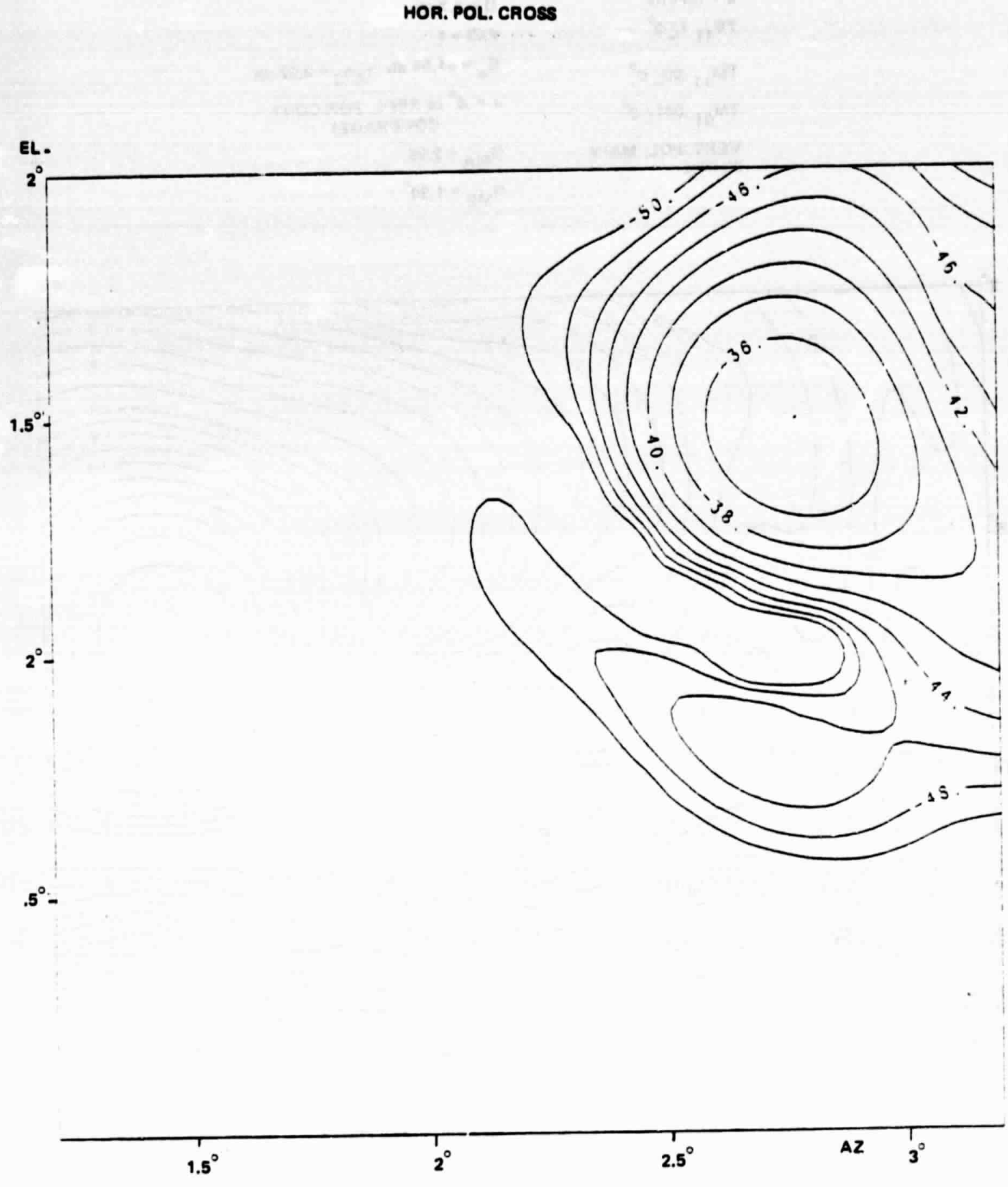


Figure 3-14b. Gain contour plot of a scanned, low sidelobe level singlet, achieved by a large diameter (d_λ = 3.08) multimode horn. Cross polarization.

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relatively moderate. The necessary component beamwidth can be realized with D = 33 in. When d = .94 in. all the beams can be realized with a single reflector. Figures 3-15 and 3-16 show the calculated main and crosspolarized gain contour plots. It can be seen that for the relatively unscanned beam 1, the edge of contour is at -8 dB (at the triple crossover point of three adjacent beams) while the sidelobe level of this beam at the -6 dB contour of the 1_2 beam is at 20 dB. Thus the achievable adjacent beam isolation between these two beams is about 14 dB. The crosspolarized level within the 1_1 beam is -28 dB and within the 1_2 beam it is -40 dB. Since Plan 1 uses only one polarization the crosspolarized power level is irrelevant.

Figures 3-17 and 3-18 shows the sidelobe levels caused by 1_3 and 1_4 into the coverage of 1_2. The limitation is from the most scanned 1_3 beam resulting in -17.5 dB sidelobe level and 11.5 dB adjacent beam isolation. The effect of the 1_4 beam in comparison is nearly negligible resulting in about 24 dB beam isolation.

Figure 3-19 shows the resultant interference power of 1_1, 1_3 and 1_4 into the coverage area of 1_2. The calculations were done in two different manners. Figure 3-19a shows the resultant level on the basis of power addition of the interferences. The resultant level reaches -17.5 dB within the -8 dB contour of the beam, yeilding 9.5 dB resultant beam isolation. Figure 3-19b exhibits the results if the fields are added with a random phase attached to the three interfering channels. The result for this model is -16 dB peak sidelobe level or 8 dB beam isolation. It can be seen that the resultant beam isolation, whether it is calculated on the basis of power or randomly phased field addition is in the range of 8 dB to 9.5 dB even for a $\alpha = 1.5°$ when only one reflector and a single horn per singlet beam is used. This proves that Plan 1 cannot be realized with acceptable beam isolation in such a simple manner.

Figures 3-20, 3-21 and 3-22 exhibit the results of additional singlet calculations for $\alpha = 5°$. The worst edge of contour gain is -7 dB and -6 dB relative to the maximum of the singlet for the minimally scanned and maximally scanned beams respectively.

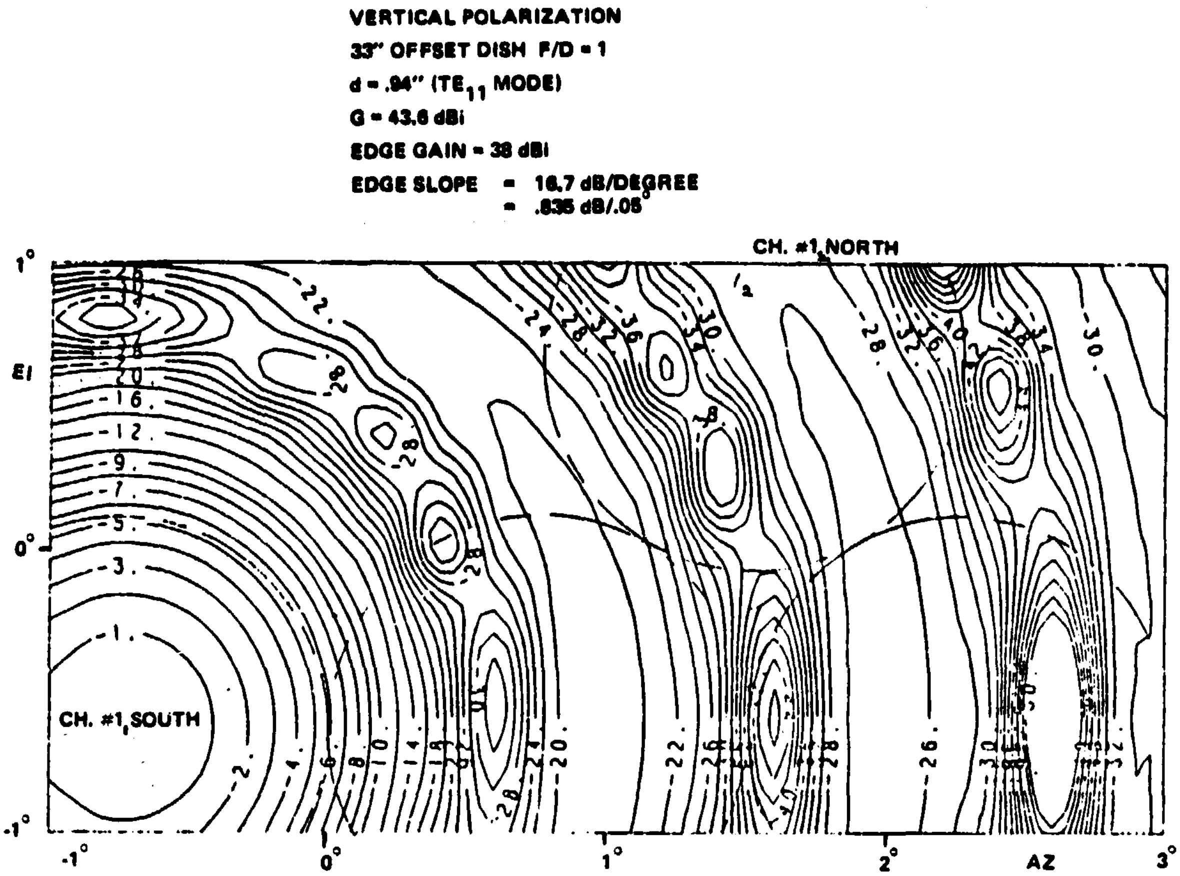


Figure 3-15. Gain contour plot of beam l_1 in main beam region of beam l_2 for Plan 1, $\alpha = 1.5°$. Antenna employs D = 55.88λ reflector diameter, F/D = 1, d = 1.54λ horn diameter, TE_{11} mode excitation. Horn diameter allows contiguous coverage with 1 reflector. Main polarization.

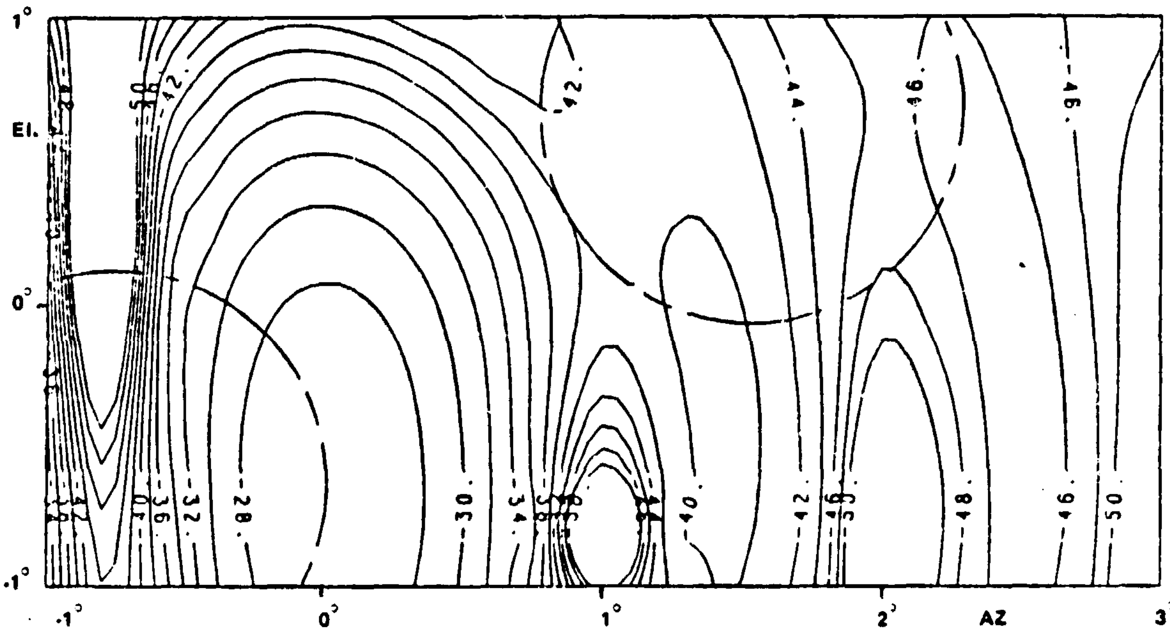


Figure 3-16. Gain contour plot of beam l_1 in main beam region of beam l_2 for Plan 1, $\alpha = 1.5°$. Antenna employs D = 55.88λ reflector diameter, F/D = 1, d = 1.54λ horn diameter, TE_{11} mode excitation. Horn diameter allows contiguous coverage with 1 reflector. Cross polarization.

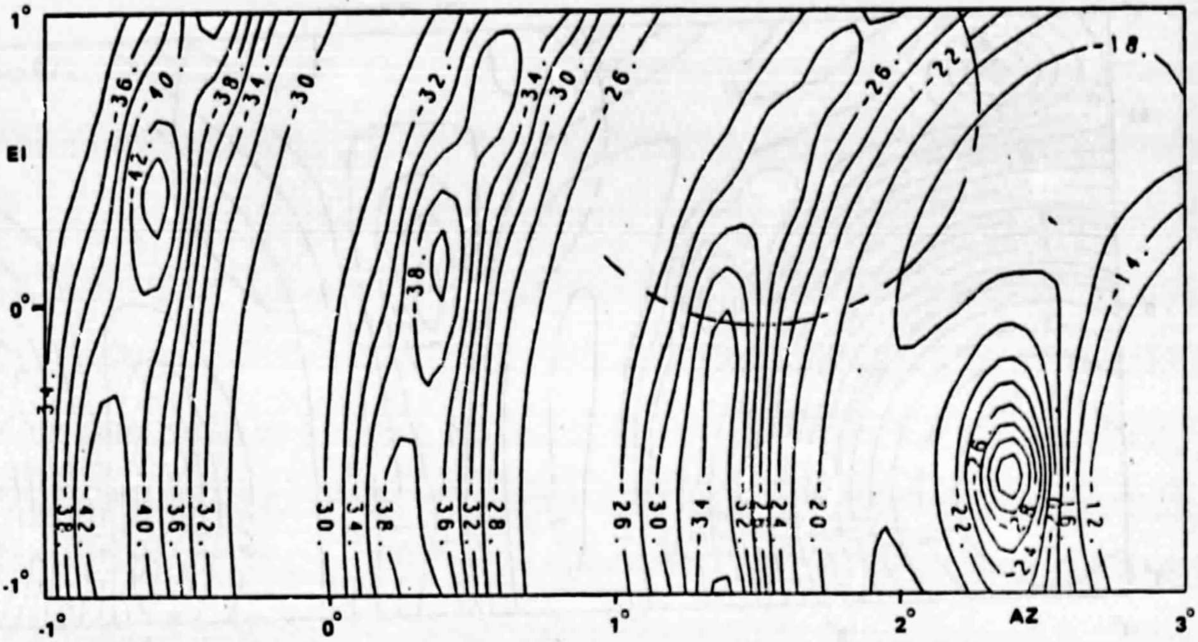


Figure 3-17. Same as Figure 3-15 for beam 1_3 in main beam region of beam 1_2. Main polarization.

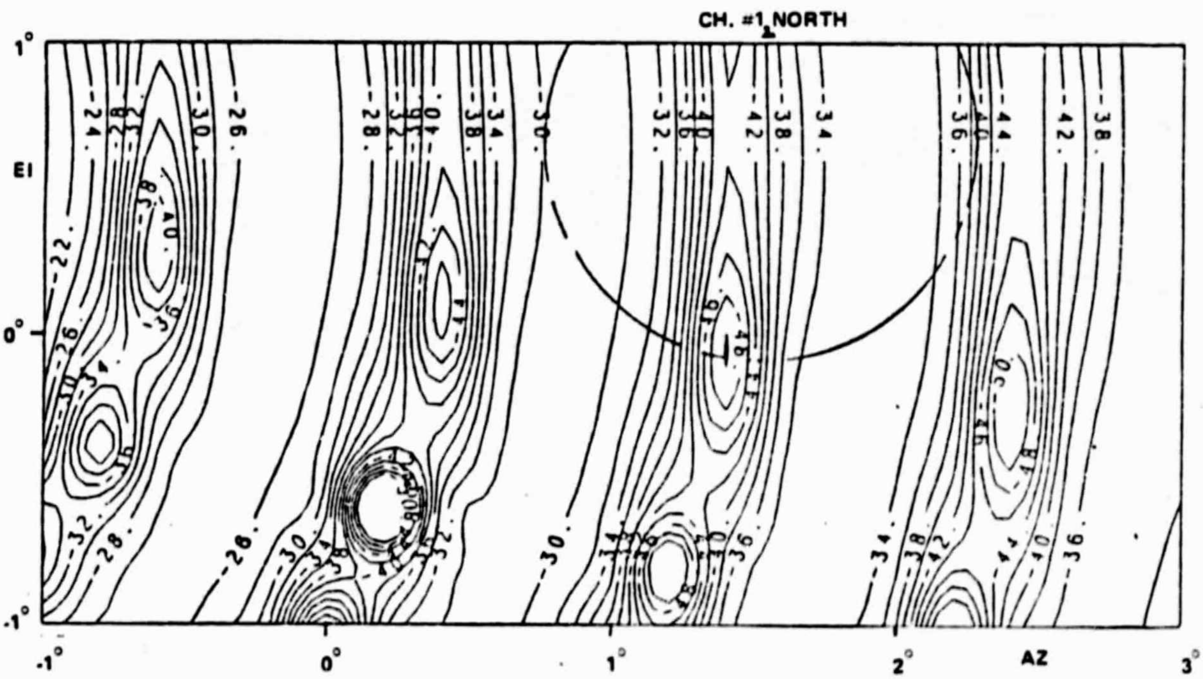


Figure 3-18. Same as Figure 3-15 for beam 1_4 in main beam region of beam 1_2. Main polarization.

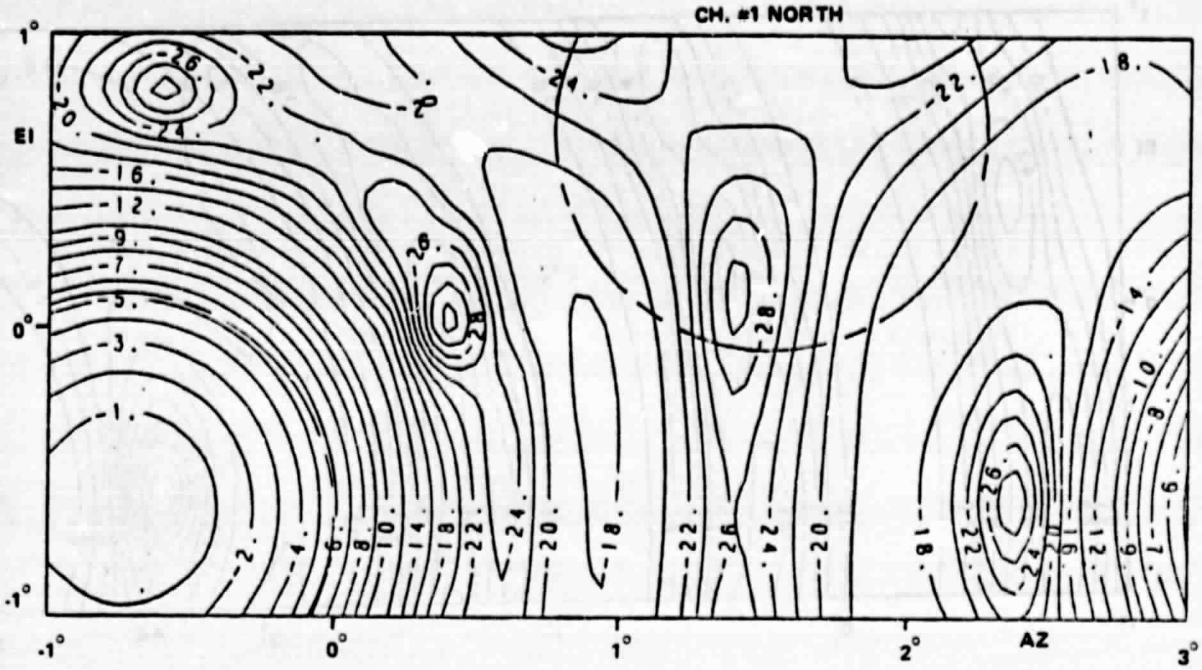


Figure 3-19a. Combined interference level caused by beam l_1, l_3 and l_4 in main beam region of l_2 using gain contours shown on Figures 3-15 - 3-18. Calculation is based on power addition.

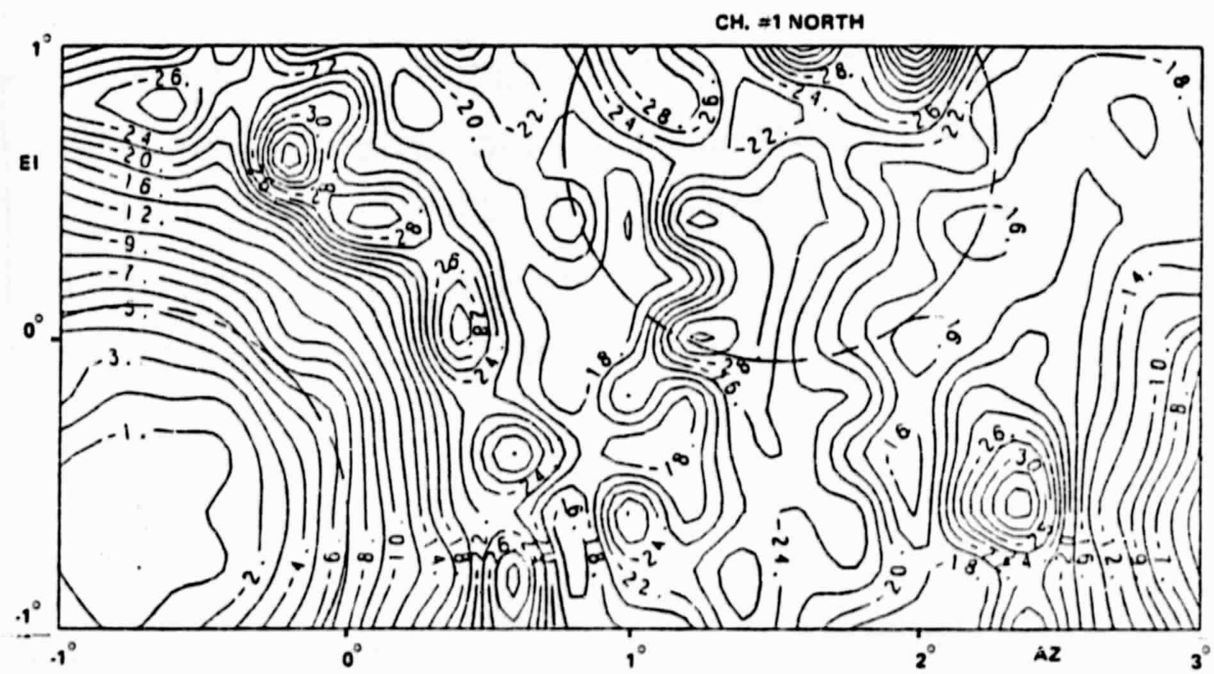


Figure 3-19b. Combined interference level caused by beam l_1, l_3 and l_4 in main beam region of l_2 using gain contours shown on Figures 3-15 - 3-18. Calculation is based on addition of fields with random phases.

SINGLET (PLAN 1,7)
d = .91 IN
TE_{11}

20 GHz

D = 96 IN
Q = 6.7 IN
F/D = 1
G_o = 52.6 db, $\eta_A \eta_S$ = 1.56 db
$\alpha = .5^\circ$
$\Theta_{MA} = .5^\circ$
$\Theta_{ME} = .5^\circ$

VERT. POL. MAIN

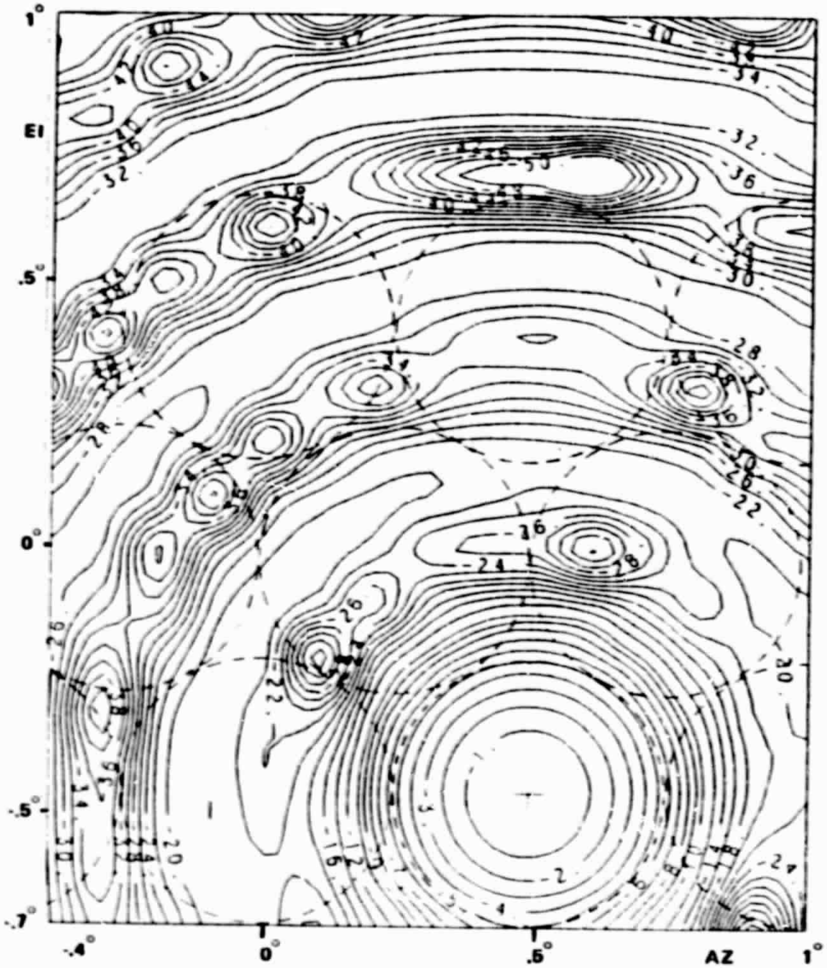


Figure 3-20a.

Gain contours of singlet beam. $D_\lambda = 162.56$, F/D = 1, $d_\lambda = 1.54$, $\alpha = .5^\circ$.
$\theta_{MA} = .5^\circ$, $\theta_{ME} = .5^\circ$, main polarization.

HOR. POL. CROSS

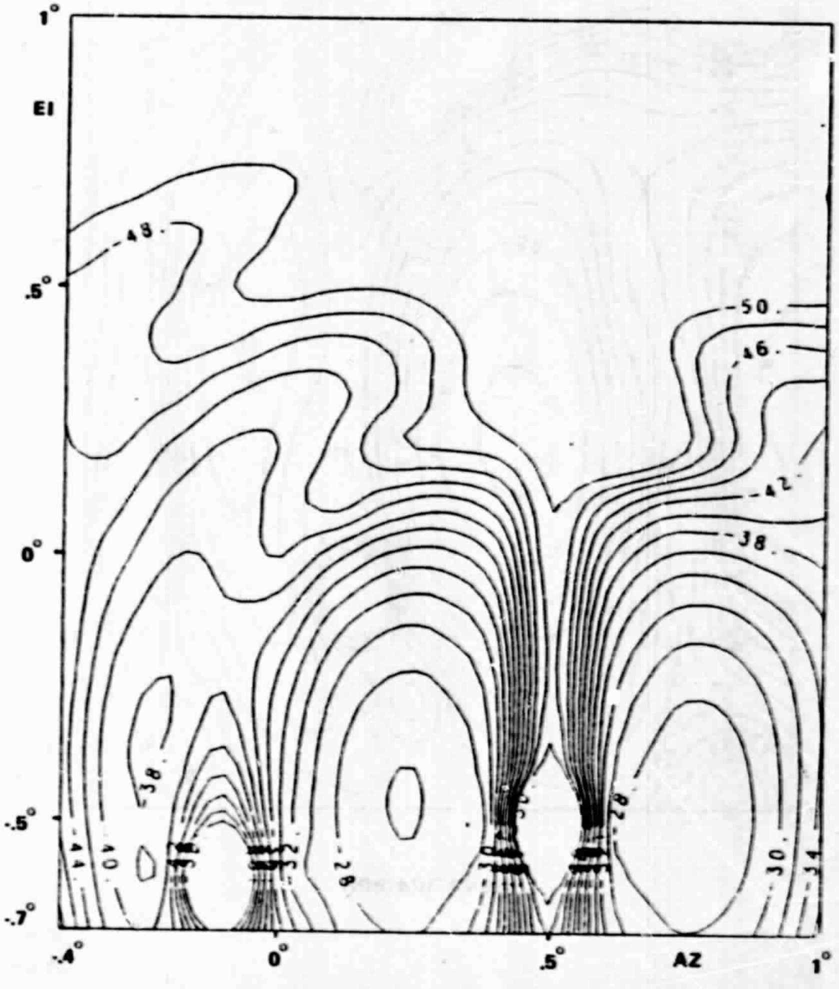


Figure 3-20b.

Gain contours of singlet beam. $D_\lambda = 162.56$, F/D = 1, $d_\lambda = 1.54$, $\alpha = .5^\circ$.
$\theta_{MA} = .5^\circ$, $\theta_{ME} = .5^\circ$, cross polarization.

SINGLET (PLAN 1,7)
d = .91 IN
TE_{11}

20 GHz

D = 96 IN
Q = 6.7 IN
F/D = 1
G_o = 52.2 db $\eta_A \eta_S$ = 1.96 db
α = .5°
Θ_{MA} = 1.45°
Θ_{ME} = 1.55°

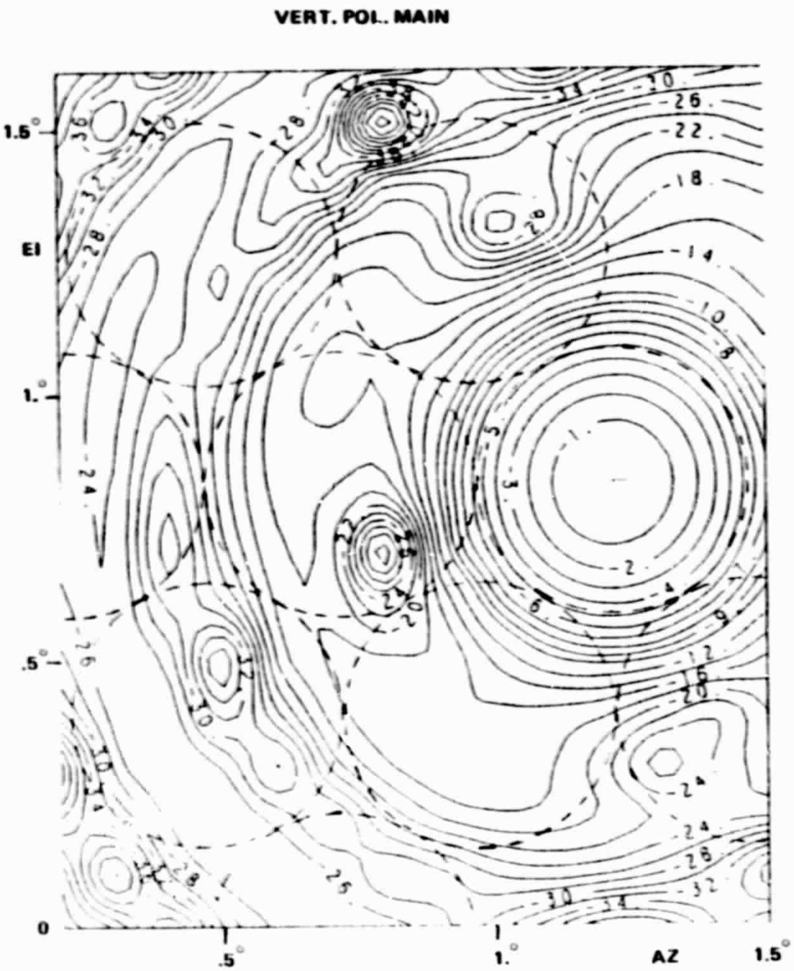


Figure 3-21a.

Gain contours of singlet beam. D_λ = 162.56, F/D = 1, d_λ = 1.54 , α = .5°.
θ_{MA} = 1.45°, θ_{ME} = 1.55°, main polarization.

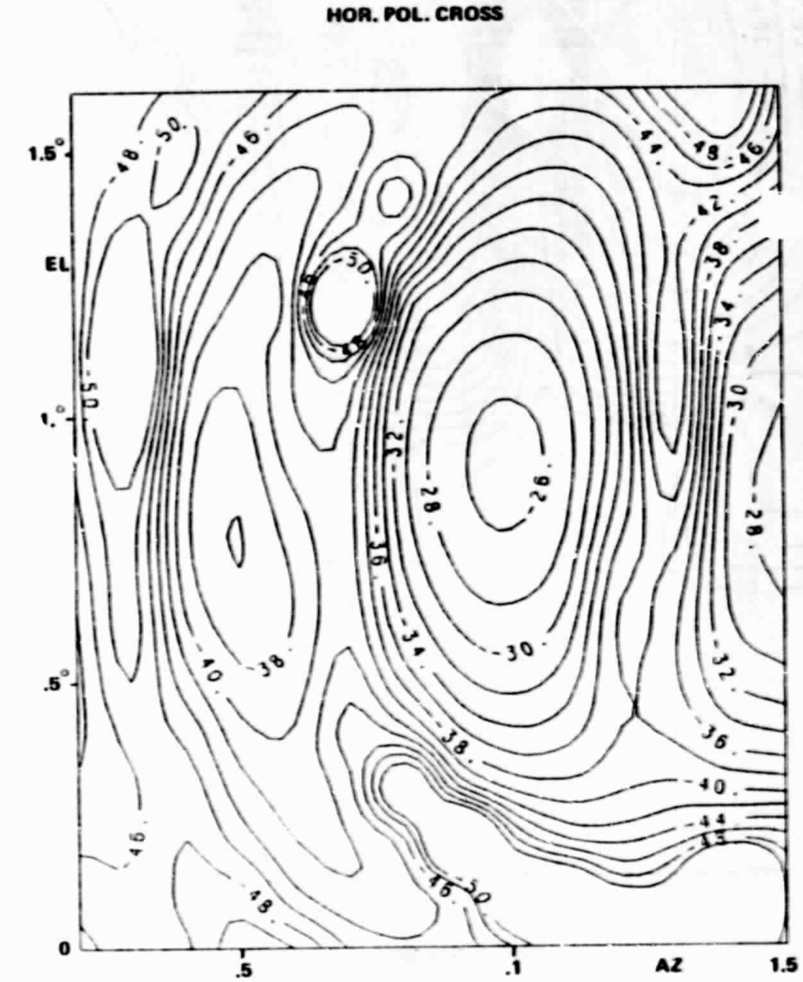


Figure 3-21b.

Gain contours of singlet beam. D_λ = 162.56, F/D = 1, d_λ = 1.54 , α = .5°.
θ_{MA} = 1.45°, θ_{ME} = 1.55°, cross polarization.

SINGLET (PLAN 1,7)
d = .91 IN
TE_{11}

20 GHz

D = 96 IN
a = 6.7 IN
F/D = 1
G_o = 50.4 db, $\eta_A \eta_S$ = 3.76 db
α = .5° (MAX. SCANNED BEAM)
Θ_{MA} = 2.9°
Θ_{ME} = 1.3°

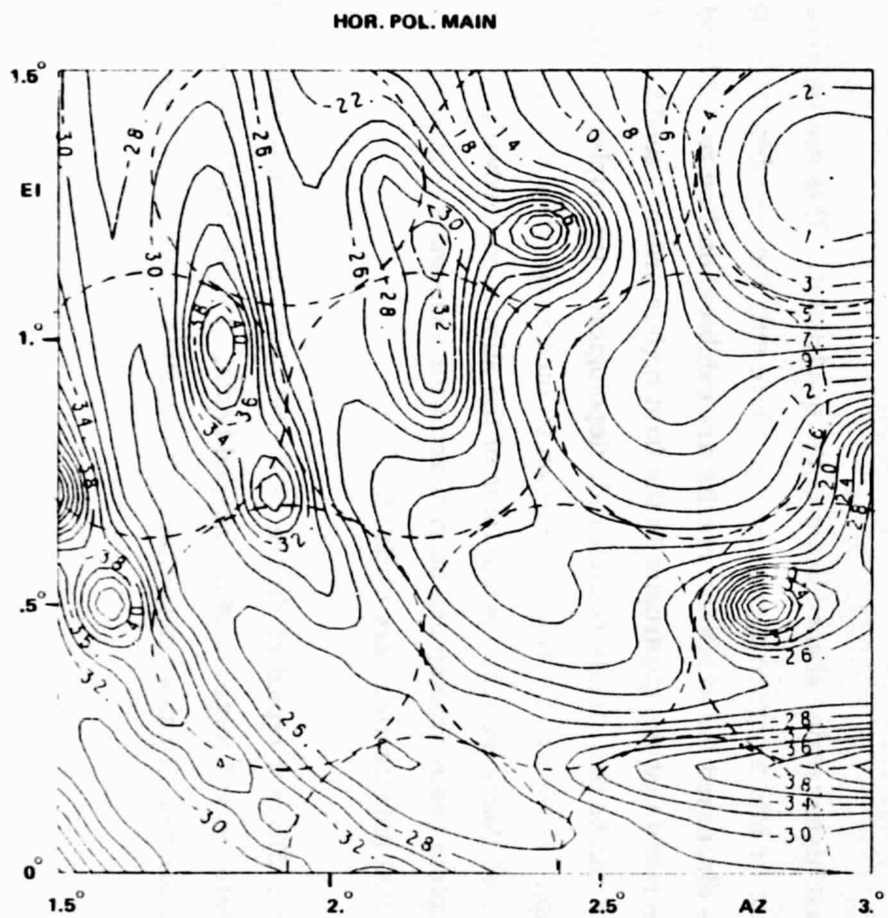


Figure 3-22a.

Gain contours of singlet beam. D_λ = 162.56, F/D = 1, d_λ = 1.54 , α = . 5°. θ_{MA} = 2.9°, θ_{ME} = 1.3°, main polarization.

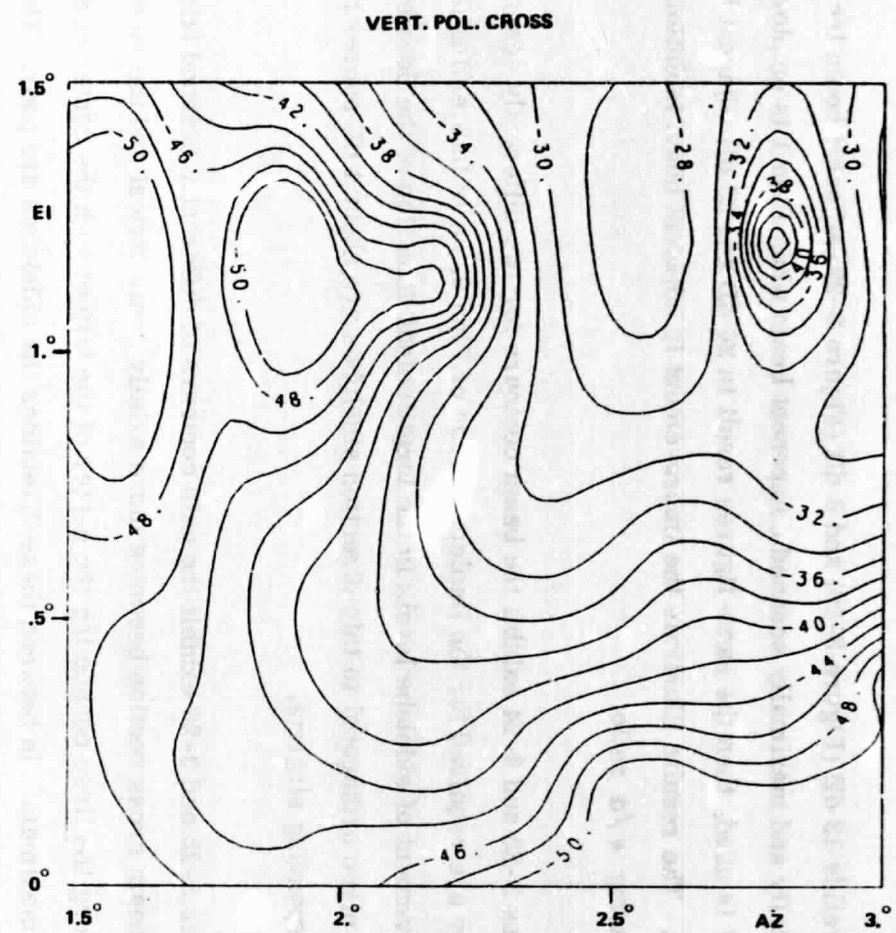


Figure 3-22b.

Gain contours of singlet beam. D_λ = 162.56, F/D = 1, d_λ = 1.54 , α = . 5°. θ_{MA} = 2.9°, θ_{ME} = 1.3°, cross polarization.

This yeilds 13 dB (Figure 3-20) and 8 dB (Figure 3-22) adjacent beam isolation for the minimally and maximally scanned component beam when Plan 1 is employed. When Plan 7 is used, then the same figures result in 26 dB and 23 dB adjacent beam isolation. The results illustrate the improvement in adjacent beam isolation with an increasing ϵ/α ratio.

Figures 3-23 and 3-24 exhibit the beam contours for two differently scanned doublets as they are required for the implementation of Plan 6. It is interesting to note the improvement of sidelobe levels in the longtitudinal direction of the doublet. However, in the plane orthogonal to this direction sidelobes are about 4 dB poorer than for the corresponding singlets.

Figures 3-25 and 3-26 exhibit the gain contours of differently scanned triplets. The main beam cross section becomes more axially symmetrical and the sidelobe level is low along the lines connecting the center of the triplet with the center of a component beam maximum. In between these directions the sidelobes are poor. This feature of the triplet is utilized advantageously when it is employed for Plan 6.

The next higher order shaped beam is the quadruplet. This can be arranged into a rhombic or into a square configuration. Here only the rhombic layout quadruplet will be discussed for which the results are exhibited on Figure 3-27 and 3-28. It may be noted that the quadruplet is the first higher order shaped beam for which beam overlap can be practically considered, (topologically the triplet already allows beam overlap). However, the implementation of such an overlap requires a very large number of channels, i.e., subdivision of the allocation bandwidth.) In the case of quadruplets beam overlap can be realized with 8 channels provided that both orthogonal polarizations are utilized.

Figure 3-27b is a repeat of Figure 3-27a but shows the relationship between beam cell contours and contours of the shaped beam. Figure 3-22d shows 4 adjacent shaped beams. It can ve seen that such a system assures contiguous coverage with

DOUBLET (PLAN 6)
d = .91 IN
TE_{11}

20 GHz

D = 96 IN
Q = 6.7 IN
F/D = 1
G_o = 50 db
α = .5°
Θ_{MA} = .75°
Θ_{ME} = -.48°

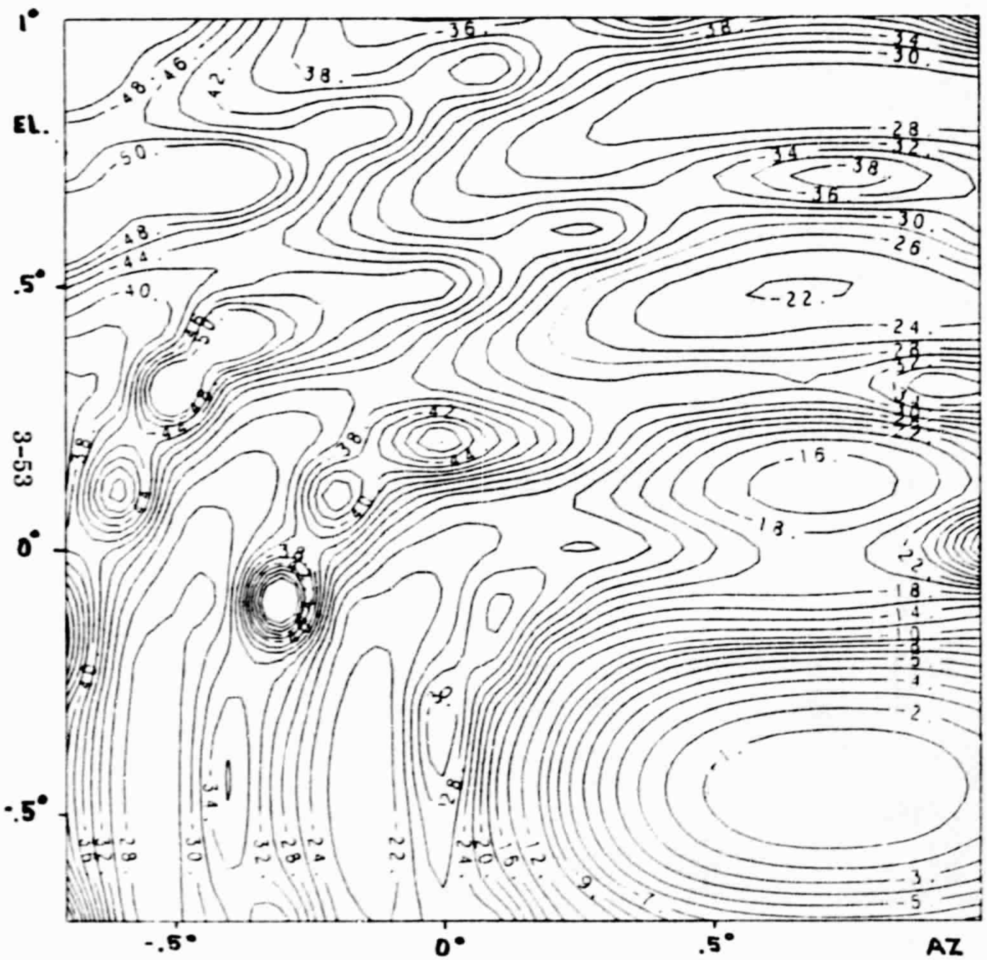


Figure 3-23a.

Gain contours of doublet beam. D_λ = 162.56, F/D = 1, d_λ = 1.54 , α = . 5°.
θ_{MA} = .75°, θ_{ME} = - .48°, Vertical Pol. Main.

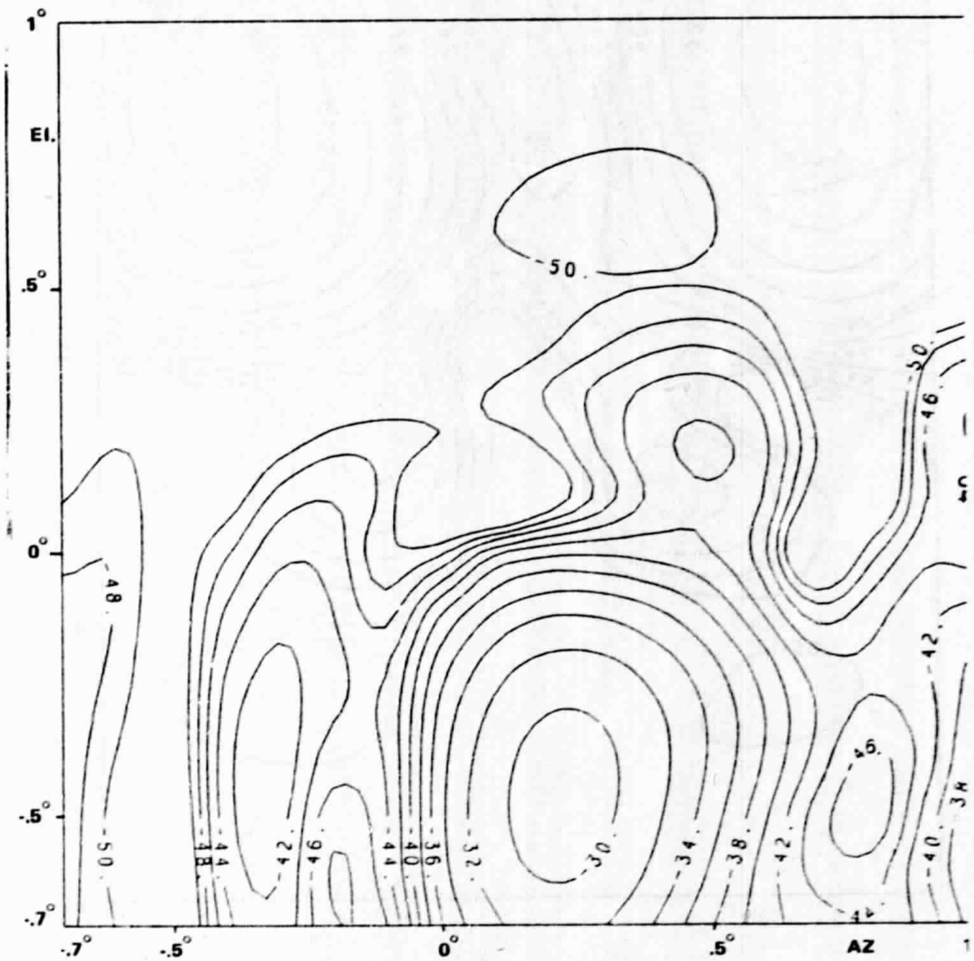


Figure 3-23b.

Gain contours of doublet beam. D_λ = 162.56, F/D = 1, d_λ = 1.54 , α = . 5°.
θ_{MA} = .75°, θ_{ME} = - .48°, Horizontal Pol. Cross.

DOUBLET (PLAN 6)

20 GHz

D = 96 IN, Q = 6.7 IN

G_o = 50.6 db

α = .5°

HOR. POL. MAIN

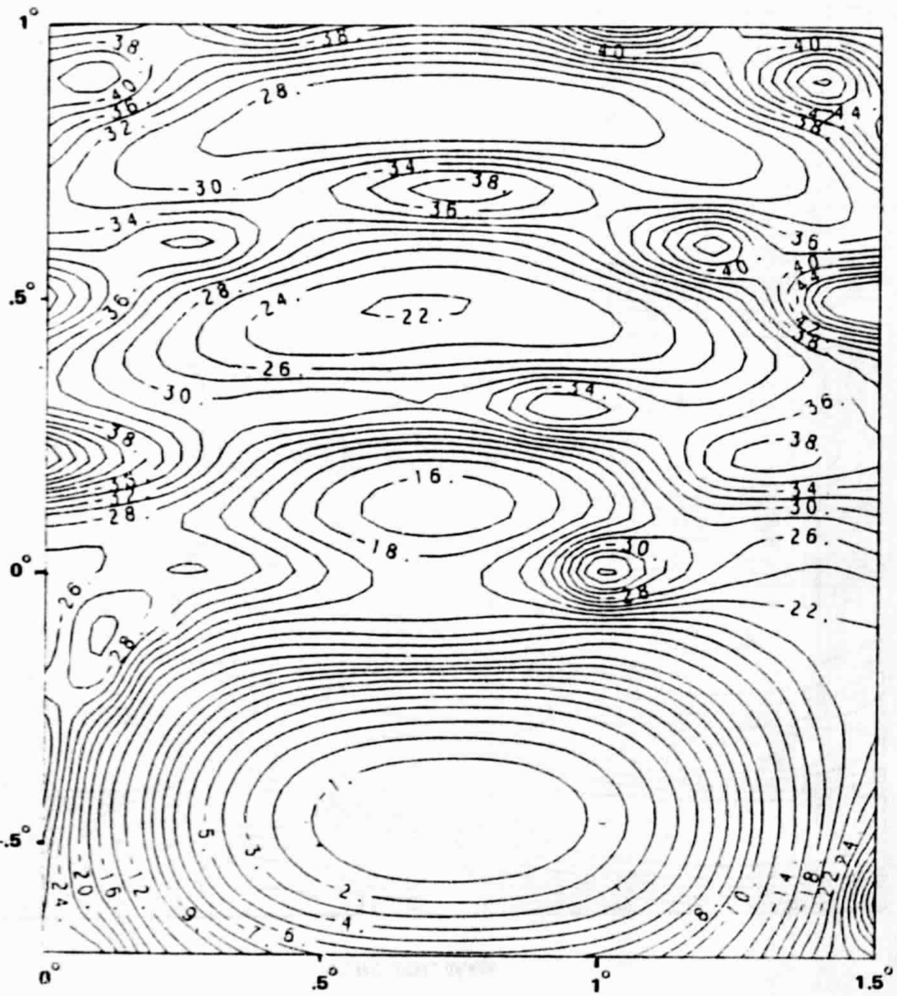


Figure 3-23c.

Gain contours of doublet beam. D_λ = 162.56, F/D = 1.
d_λ = 1.54 , α = . 5.
θ_{MA} = .75°, θ_{ME} = - .48°, Horizontal Pol. Main.

VERT. POL. CROSS

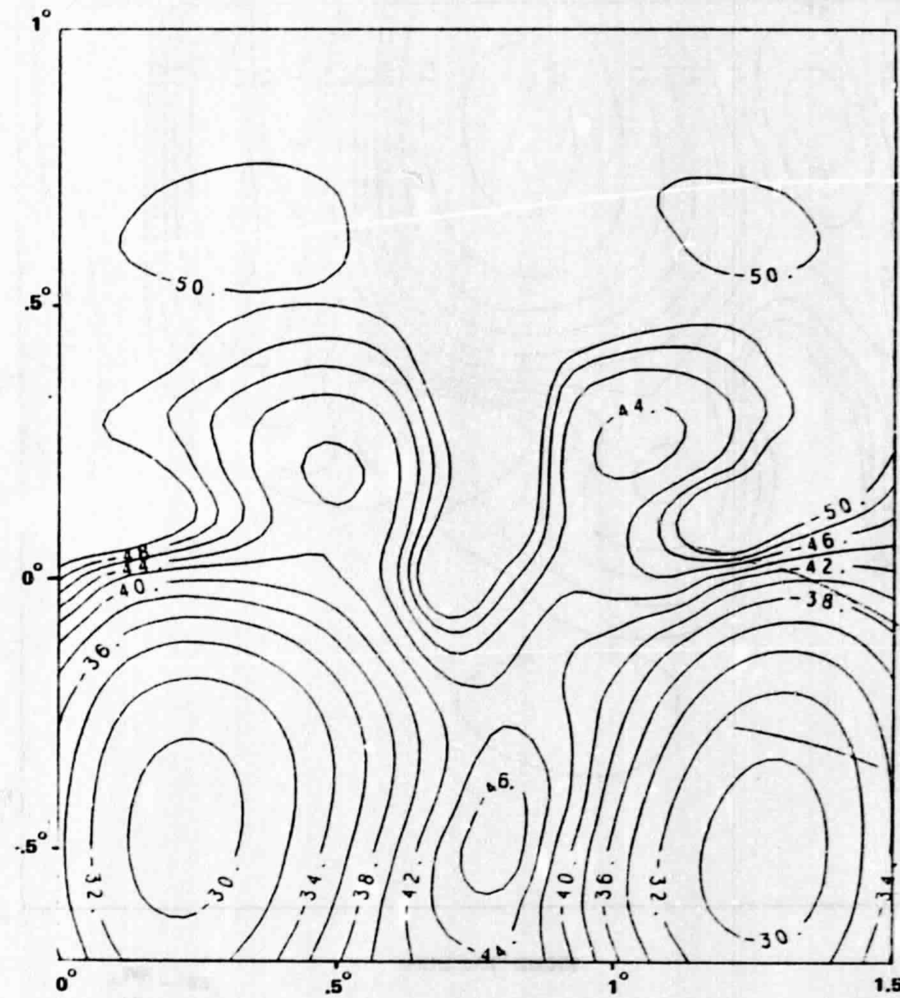


Figure 3-23d.

Gain contours of doublet beam. D_λ = 162.56, F/D = 1,
d_λ = 1.54 , α = . 5.
θ_{MA} = .75°, θ_{ME} = - .48°, Vertical Pol. Cross.

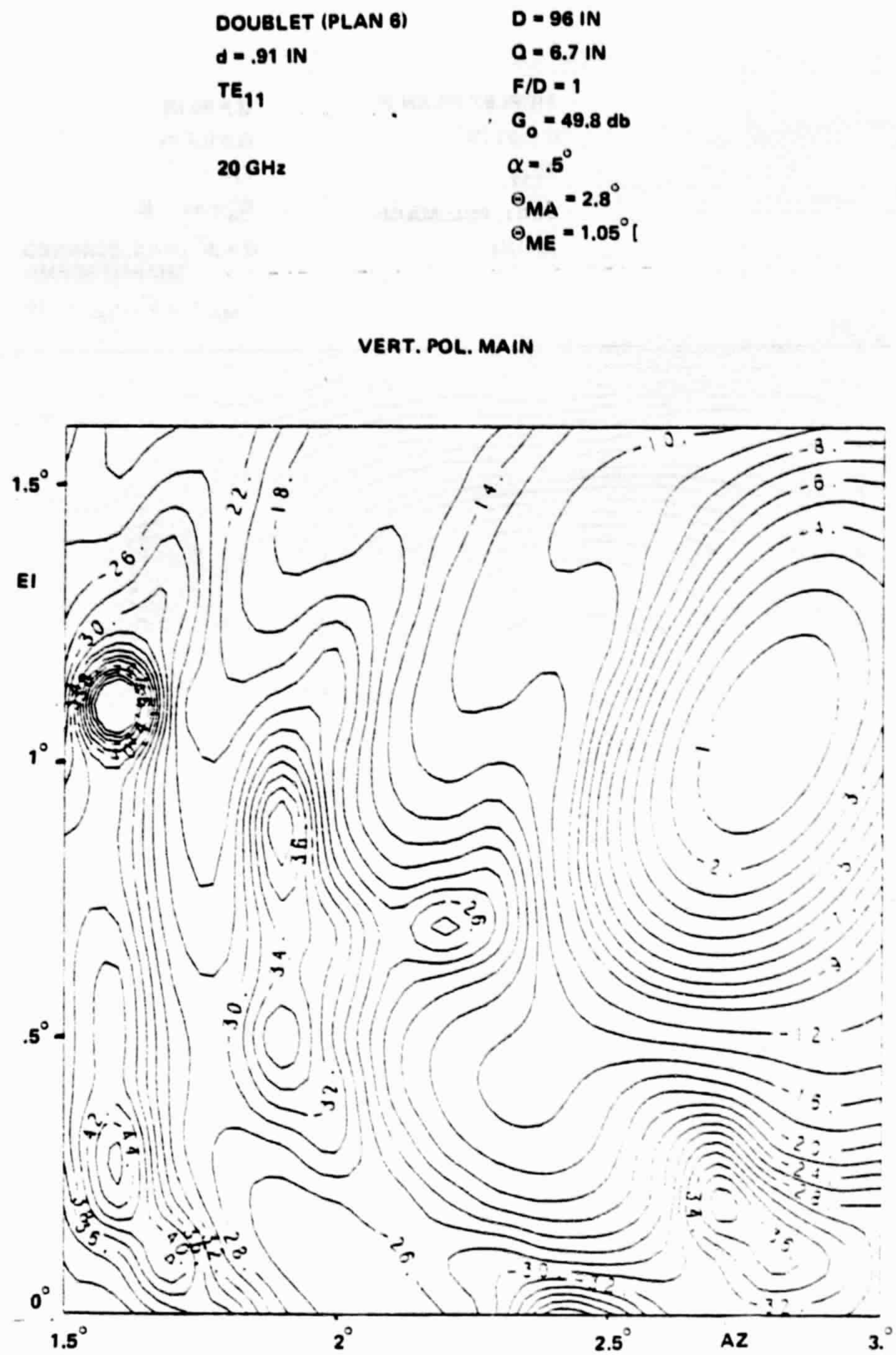


Figure 3-24. Gain contours of doublet beam. D_λ = 162.56, F/D = 1, d_λ = 1.54 , α = . 5. θ_{MA} = 2.8°, θ_{ME} = 1.05°, Vertical Pol. Main.

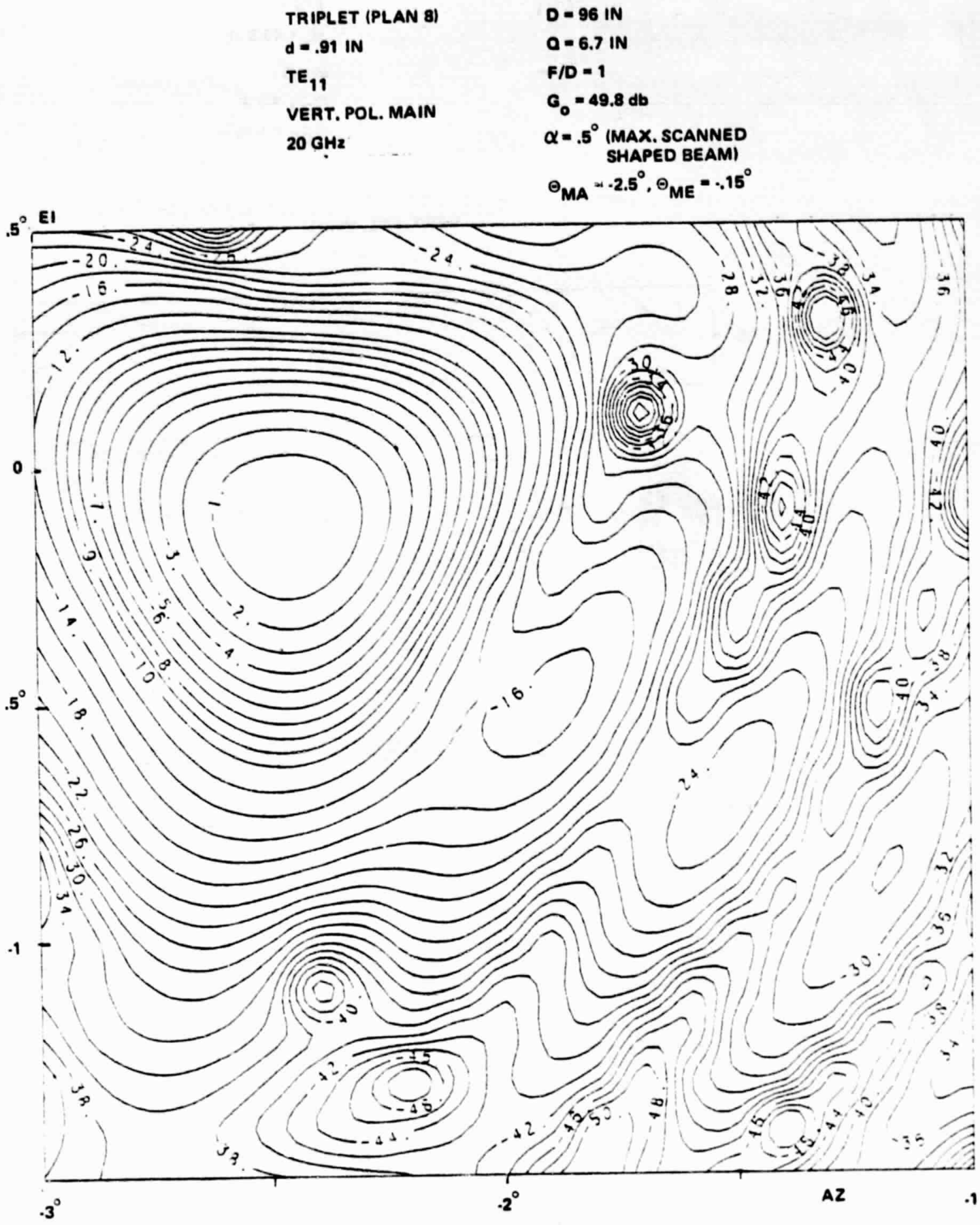


Figure 3-25a. Gain contours of triplet beam. D_λ = 163.56, F/D = 1, d_λ = 1.54, α = . 5°.
θ_{MA} = -2.5°, θ_{ME} = - .15°, Vertical Pol. Main.

HOR. POL. CROSS

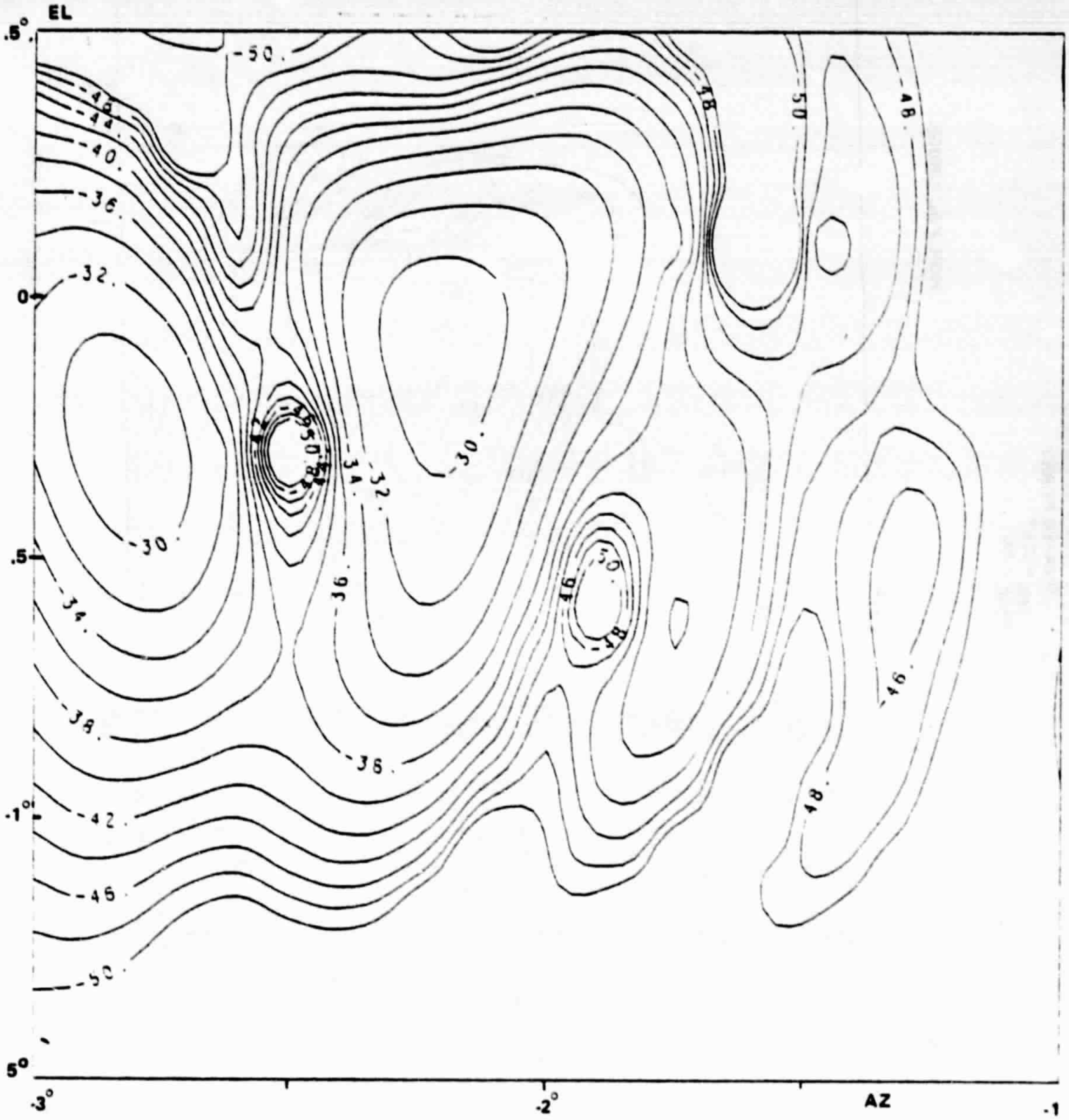


Figure 3-25b. Gain contours of triplet beam. D_λ = 163.56, F/D = 1, d_λ = 1.54, α = . 5°.
θ_{MA} = - 2.5°, θ_{ME} = - .15°, Horizontal Pol. Cross.

TRIPLET (PLAN 8)
d = 1.2 IN.
TE_{11}

20 GHz

D = 160 IN.
Q = 15.2 IN.
F/D = 1.335
G_o = 54.8 db
α = .3° (MAX. SCANNED SHAPED BEAM)
Θ_{MA} = -2.8°
Θ_{ME} = .58°

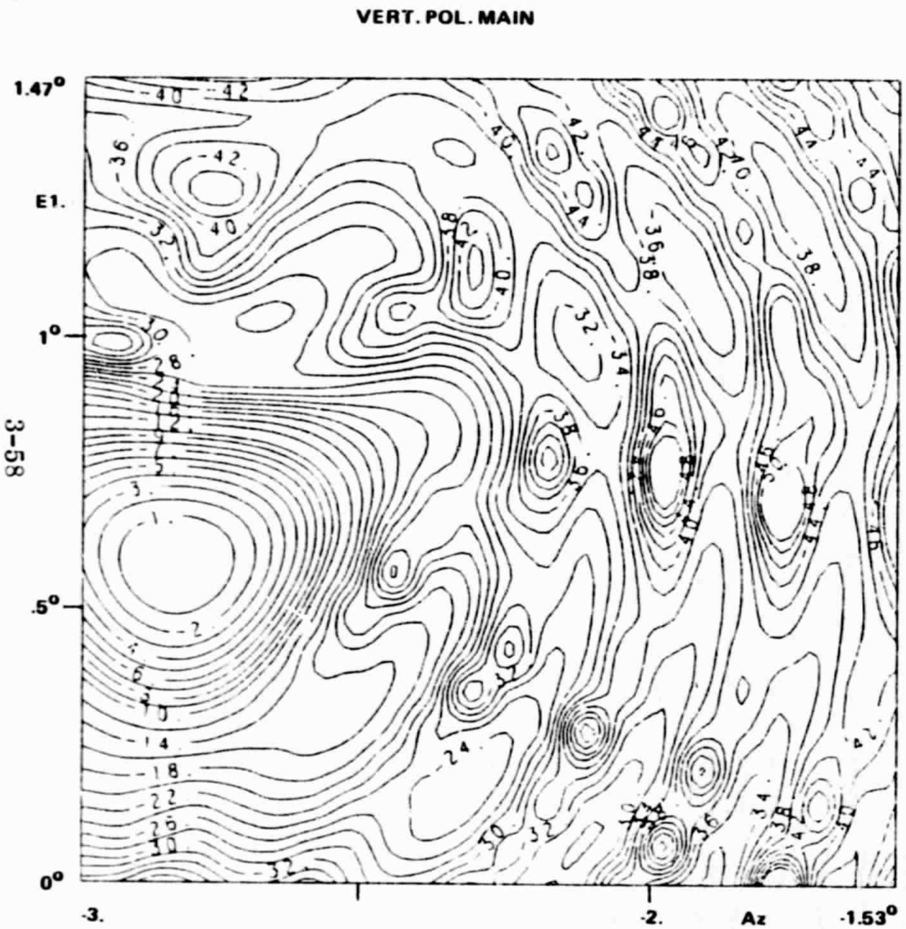


Figure 3-26a.

Gain contours of triplet beam. D_λ = 270.93, F/D = 1.335, d_λ = 2.032, α = .3°.
θ_{MA} = - 2.8°, θ_{ME} = .58°, Vertical Pol. Main.

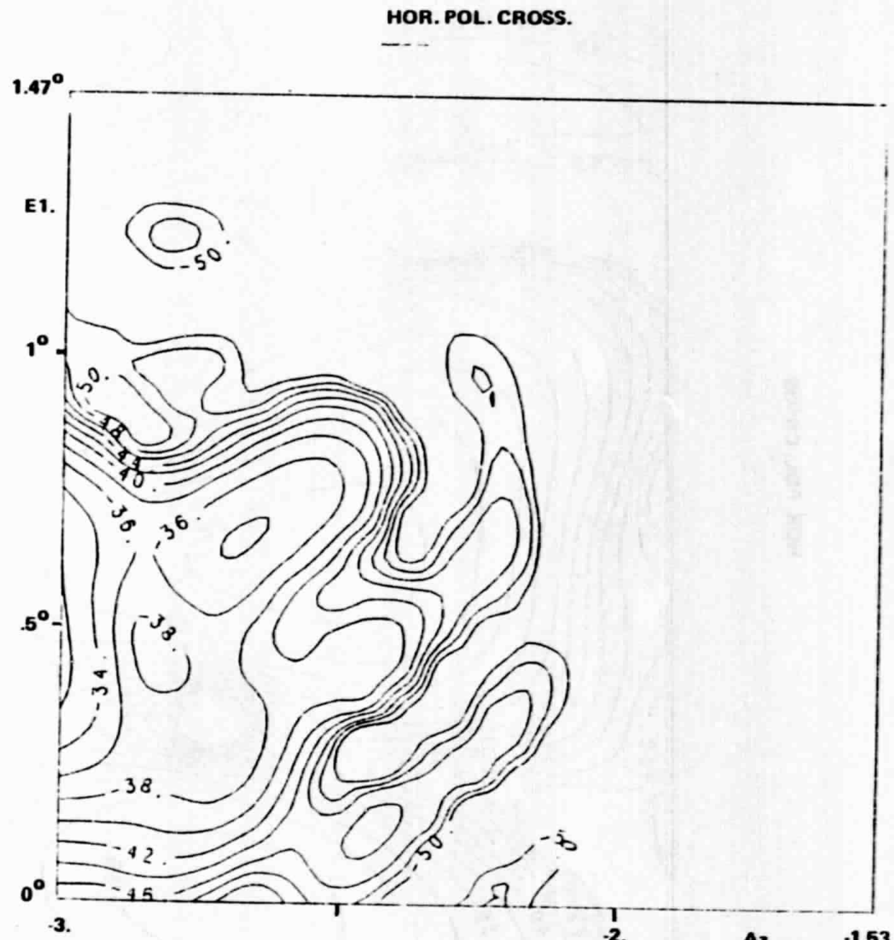


Figure 3-26b.

Gain contours of triplet beam. D_λ = 270.93, F/D = 1.335
d_λ = 2.032, α = .3°.
θ_{MA} = - 2.8°, θ_{ME} = .58°, Horizontal Pol. Cross

QUADRUPLET (PLAN 5)
d = .91 IN.
TE_{11}

20 GHz

D = 96 IN.
Q = 6.7 IN.
F/D = 1
G_o = 47.7 db
α = .5° (BEAM 2)
$\left.\begin{array}{l}\theta_{MA} = .95^\circ \\ \theta_{ME} = 1^\circ\end{array}\right\}$ CENTER OF RESULTANT BEAM.

VERT. POL. MAIN

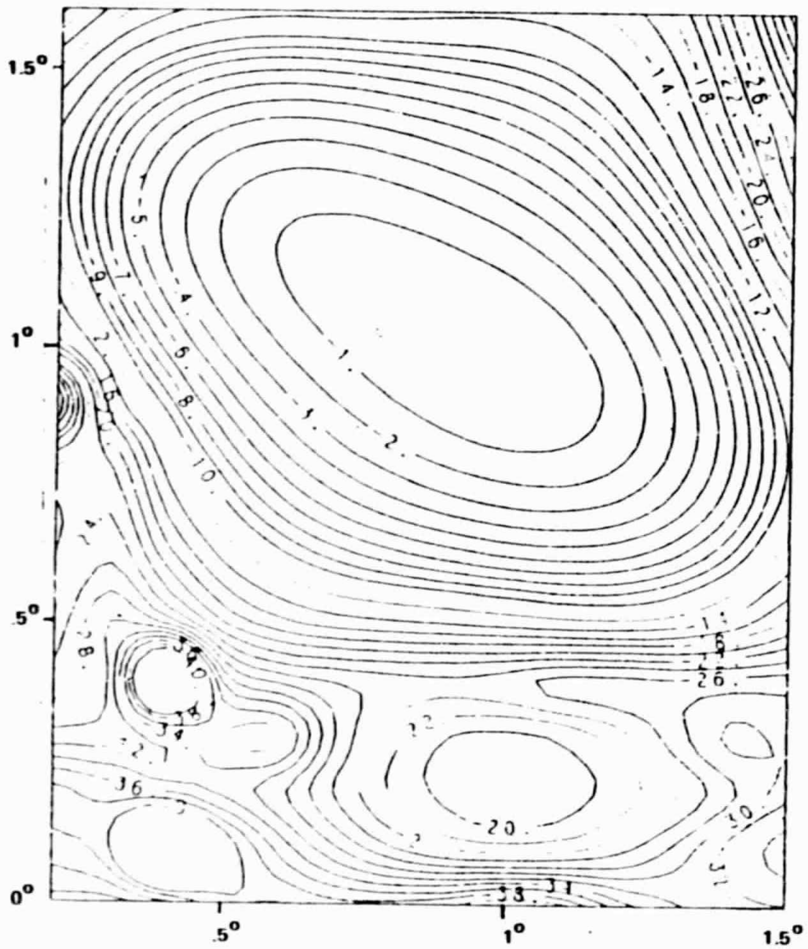


Figure 3-27a.

Gain contours of quadruplet beam. D_λ = 162.56, F/D = 1, d_λ = 1.54, α = .5°.
θ_{MA} = .95°, θ_{ME} = 1°, Vertical Pol. Main.

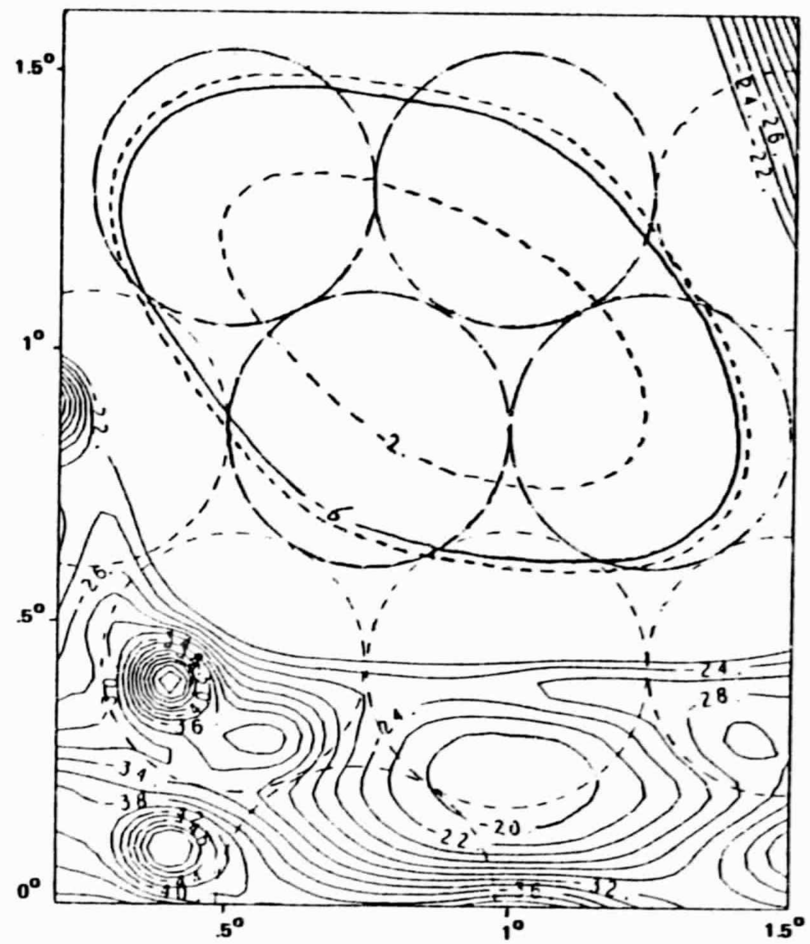


Figure 3-27b.

Same as Figure 4.17a, but showing relationship between component beam cells and contours of resultant quadruplet beam

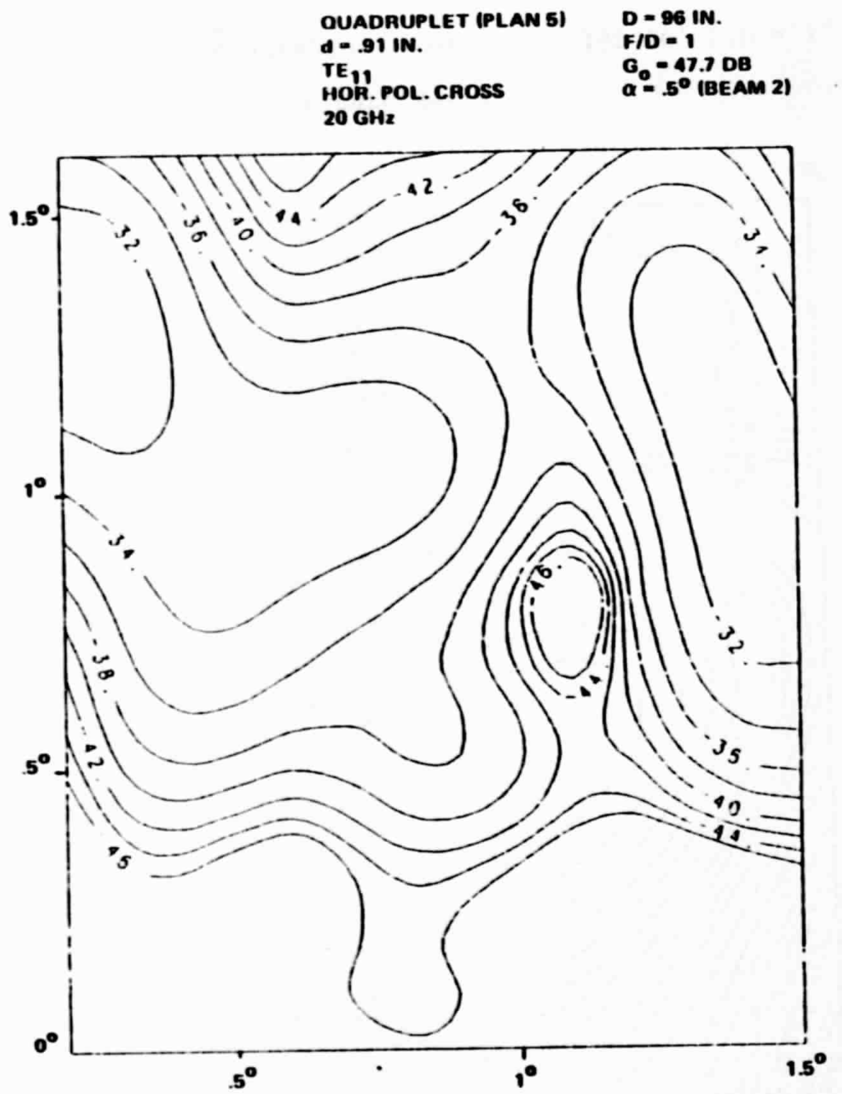


Figure 3-27c.

Gain contours of quadruplet beam, $D\lambda$ = 162.56, F/D = 1, d_λ = 1.54, α = . 5°.
θ_{MA} = .95°, θ_{ME} = 1°, Horizontal Pol. Cross.

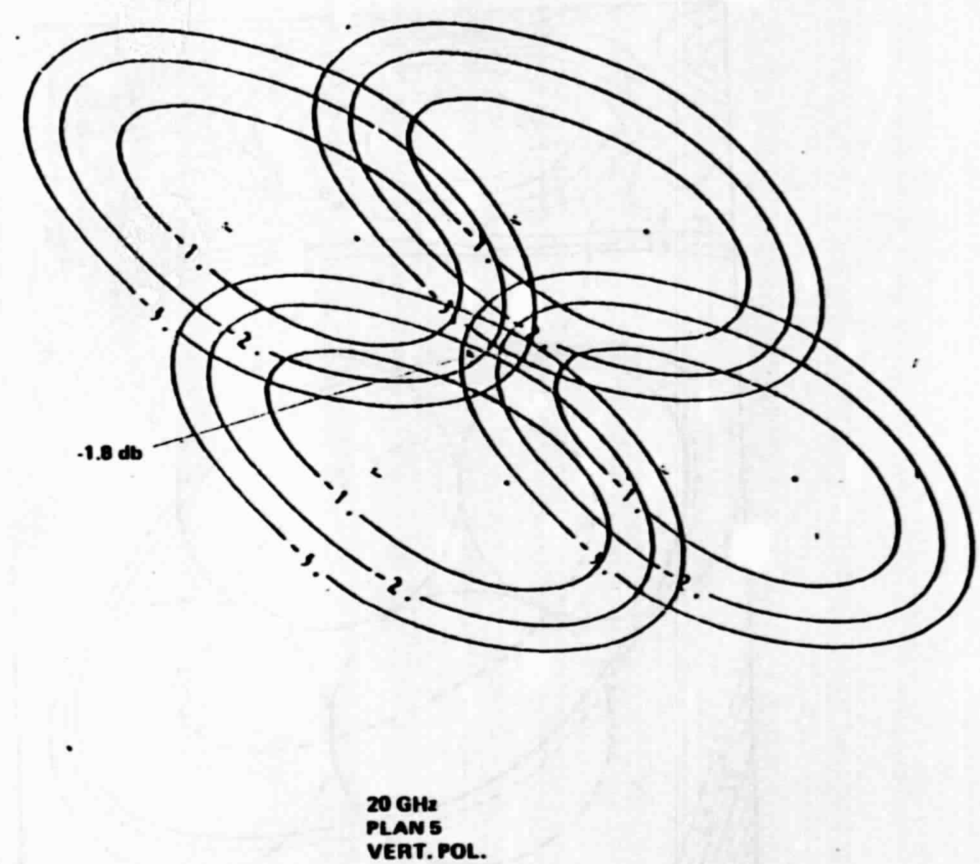


Figure 3-27d.

Gain contours of 4 adjacent quadruplets of the type shown on Figure 4.17a. Each quadruplet is operated at a different frequency channel. Lowest level at quadruple crossover is appr. - 2 db relative to peak of beams.

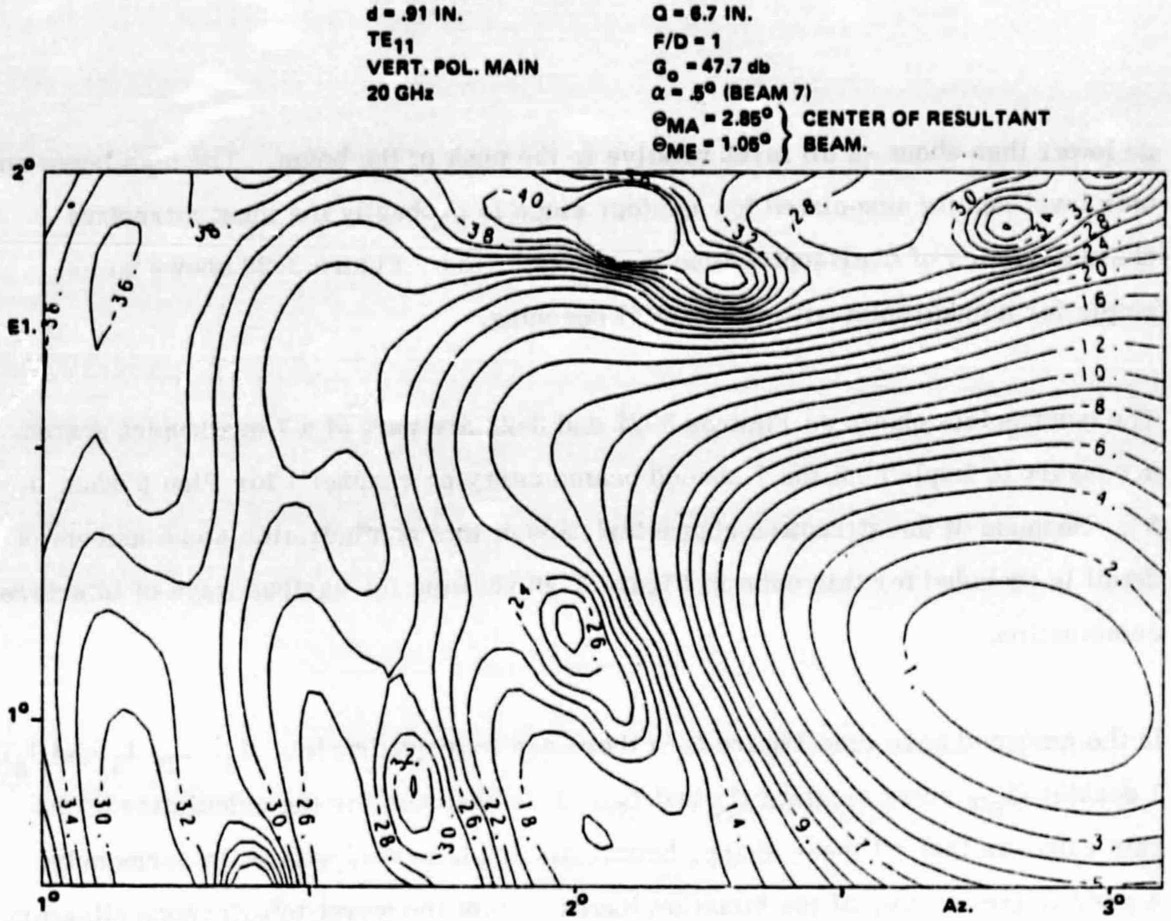


Figure 3-28a. Gain contours of quadruplet beam, D_λ = 162.56, F/D = 1, d_λ = 1.54, α = .5°. θ_{MA} = 2.85°, θ_{ME} = 1.05°, Vertical Pol. Main.

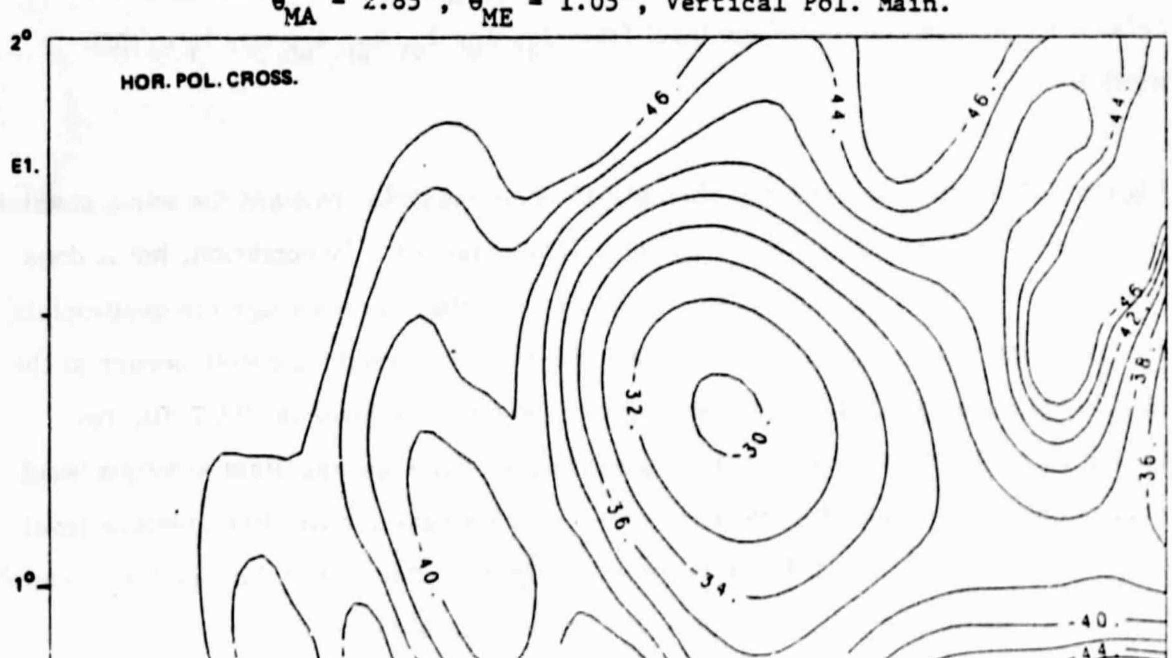


Figure 3-28b. Gain contours of quadruplet beam, D_λ = 162.56, F/D = 1, d_λ = 1.54, α = . 5°. θ_{MA} = 2.85°, θ_{ME} = 1.05°, Horizontal Pol. Cross.

no lower than about -2 dB level relative to the peak of the beam. The high beam contour level and its associated low contour slope is probably the most attractive characteristics of overlapping type beam topologies. Figure 3-28 shows an example for a quadruplet with substantial scanning.

The quadruplets shown on Figures 3-27 and 3-28 are part of a 7 quadruplet system necessary to implement the 7 shaped beams carrying channel 1 for Plan 5 when α = 5°. Because of the attractive characteristics of this configuration some amount of detail is included for this case in Figure 3-29 showing the various ways of interference combination.

In the analyzed case (see Figure 3-5) there are four quadruplets (1_1, 1_2, 1_3 and 1_6), 1 doublet (1_5), and 2 singlets (1_4 and 1_7). It is assumed for the calculation of the gain contours that all these shaped beams are implemented with main horns only. A preliminary survey of the situation indicates that the worst interference situation occurs in the $0 \leq \theta_A \leq 1.5^\circ$ azimuth and $0 \leq \theta_E \leq 1.65^\circ$ elevation angles window, which contains the shaped beam carrying channel 1_3. Thus the interference is calculated as the resultant sidelobe level from 1_1, 1_2, 1_4, 1_5, 1_6, and 1_7 in this window.

Figure 3-29a shows the results when all of these channels transmit the same absolute power and the fields are added in phase. This is not a likely condition, but it does represent the worst possible case. Since for a contiguous coverage the quadruplets must be used down to their -2 dB contour the resultant sidelobe power occurs at the -2 dB contour of beam 1_2. At this contour, the sidelobe level is -22.7 dB, resulting in 20.7 dB beam isolation. Figure 3-29b shows the resultant sidelobe level when the fields are added in random phase. In this case the limiting sidelobe level is at -22 dB yeilding 20 dB beam isolation. Figure 3-28c shows the resultant sidelobe level when the beams are added on power bases. In this case the resultant limiting sidelobe power is about -23.3 dB and the beam isolation is 21.3 dB. When the

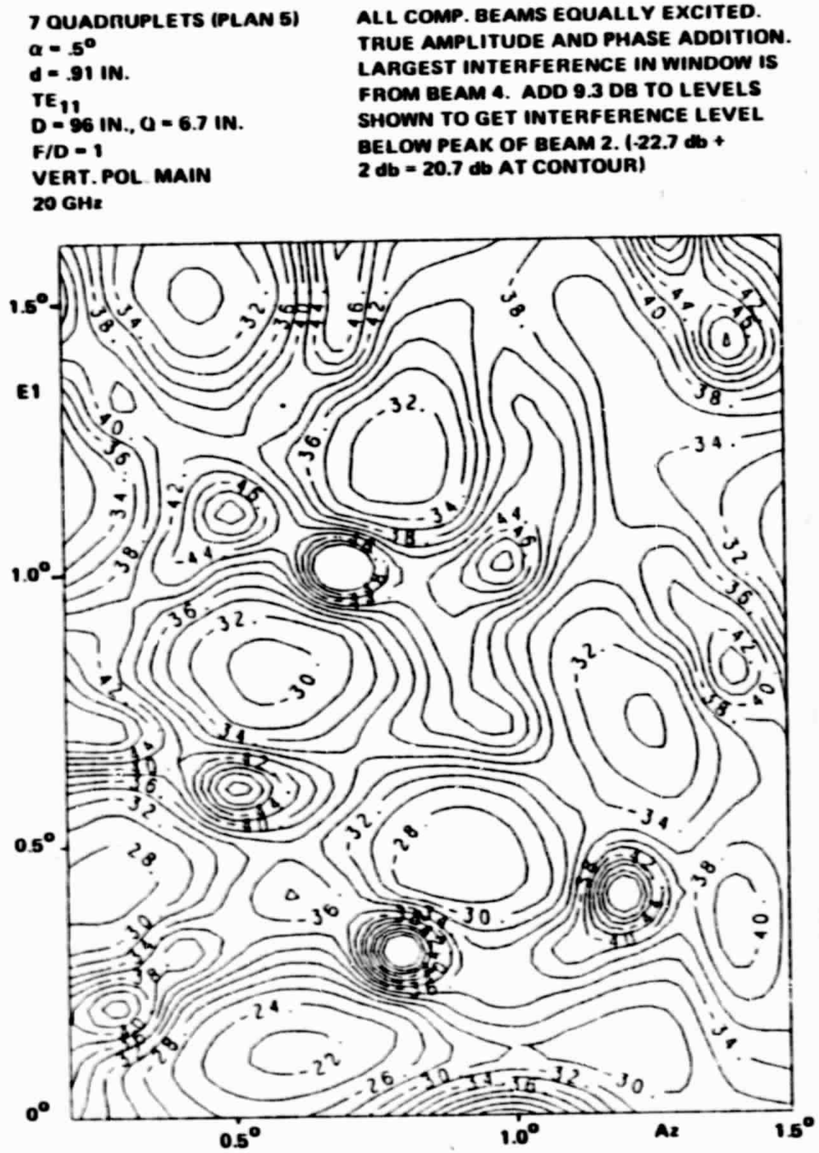


Figure 3-29a.

Contour plots of total interfering power for Plan 1, $\alpha = .5°$, vertical main polarization, $D_\lambda = 162.56$, $F/D = 1$, $d_\lambda = 1.54$. Plot shows interference in beam 1_2 from beams 1_1, 1_3, 1_4, 1_5, 1_6 and 1_7.

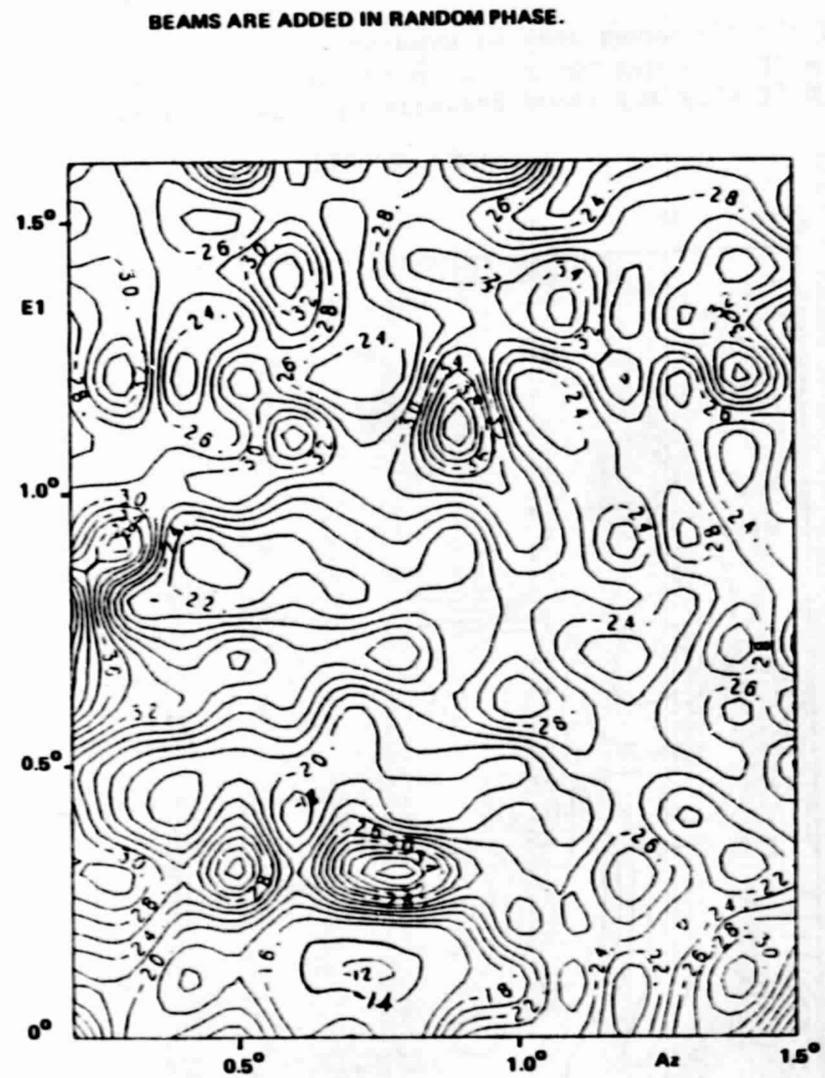


Figure 3-29b.

Contour plots of total interfering power for Plan 1, $\alpha = .5°$, vertical main polarization, $D_\lambda = 162.56$, $F/D = 1$, $d_\lambda = 1.54$. Plot shows interference in beam 1_2 from beams 1_1, 1_3, 1_4, 1_5, 1_6 and 1_7.
PA output powers are equal, fields are added in random phase.

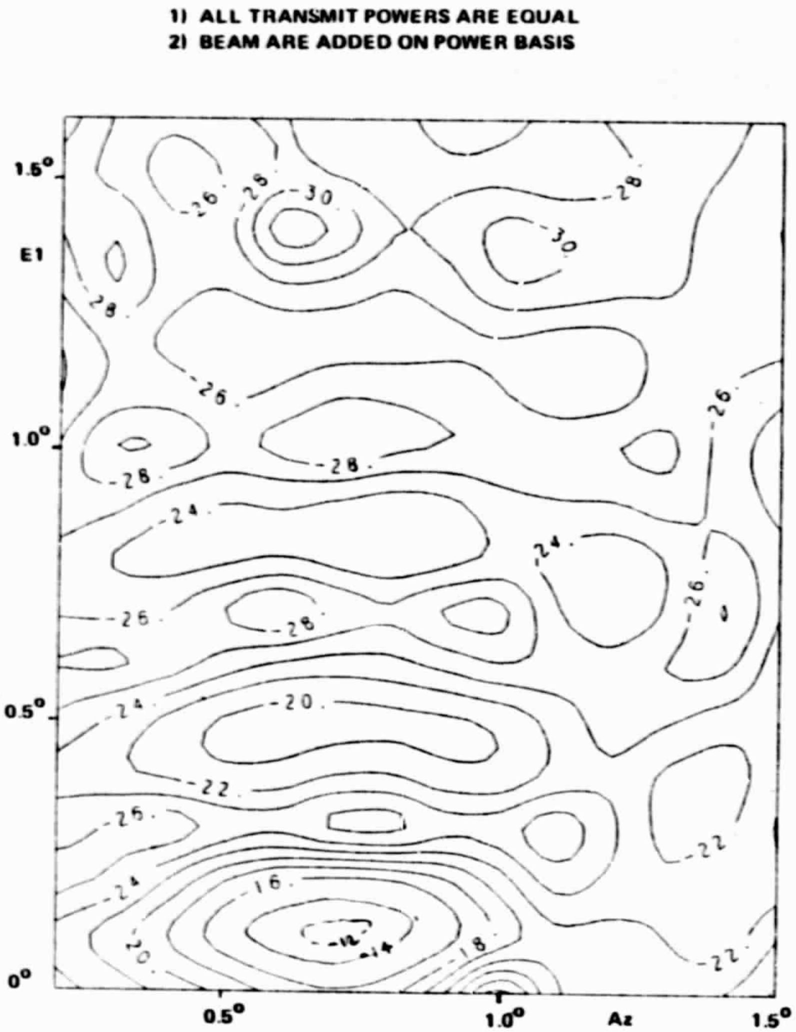


Figure 3-29c.

Contour plots of total interfering power for Plan 1, $\alpha = .5°$, vertical main polarization, $D_\lambda = 162.56$, $F/D = 1$, $d_\lambda = 1.54$. Plot shows interference in beam 1_2 from beams 1_1, 1_3, 1_4, 1_5, 1_6 and 1_7.
PA output powers are equal, interference is added on power basis.

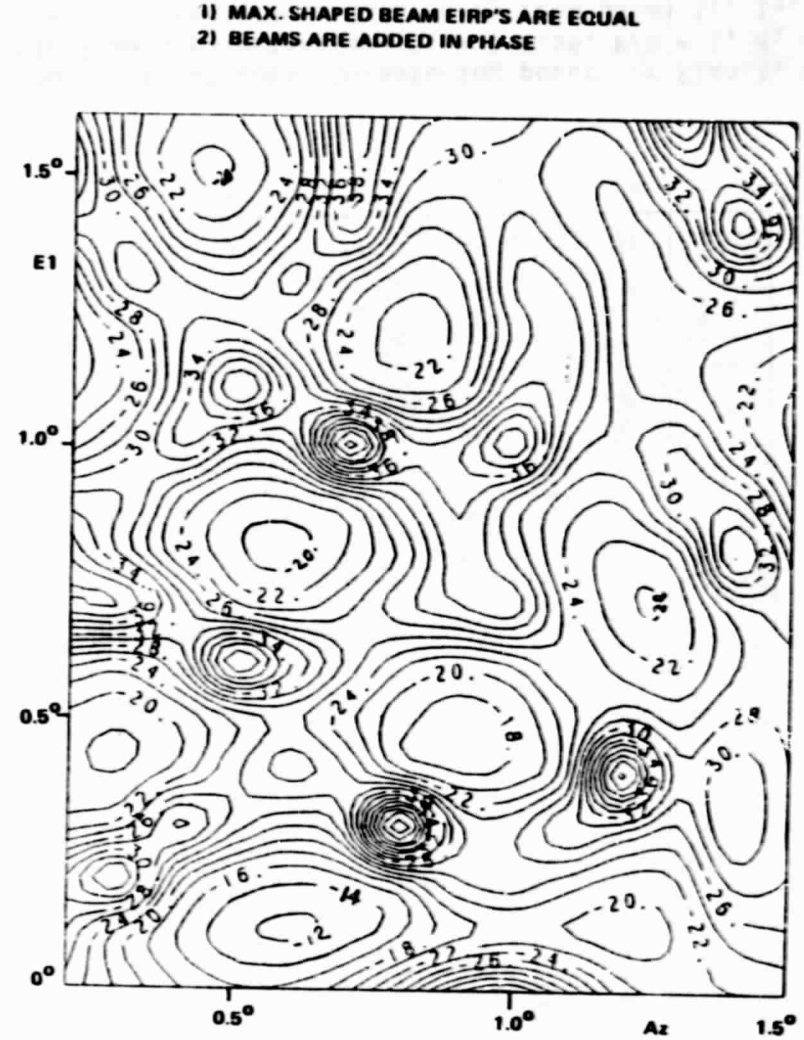


Figure 3-29d.

Contour plots of total interfering power for Plan 1, $\alpha = .5°$, vertical main polarization, $D_\lambda = 162.56$, $F/D = 1$, $d_\lambda = 1.54$. Plot shows interference in beam 1_2 from beams 1_1, 1_3, 1_4, 1_5, 1_6 and 1_7.
Maximum EIRP of beams are equal, fields are added in phase.

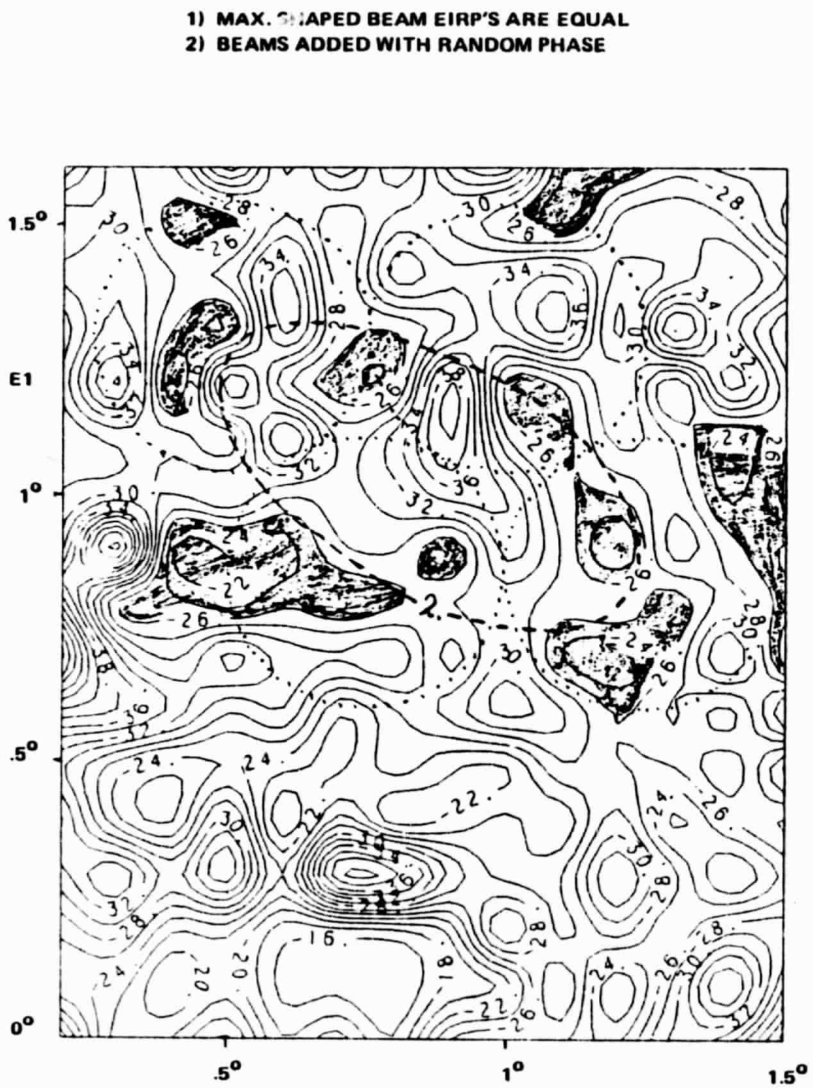


Figure 3-29e.

Contour plots of total interfering power for Plan 1, $\alpha = .5°$, vertical main polarization, $D_\lambda = 162.56$, $F/D = 1$, $d_\lambda = 1.54$. Plot shows interference in beam 1_2 from beams 1_1, 1_3, 1_4, 1_5, 1_6 and 1_7.
Maximum EIRP of beams are equal, field are added in random phase.

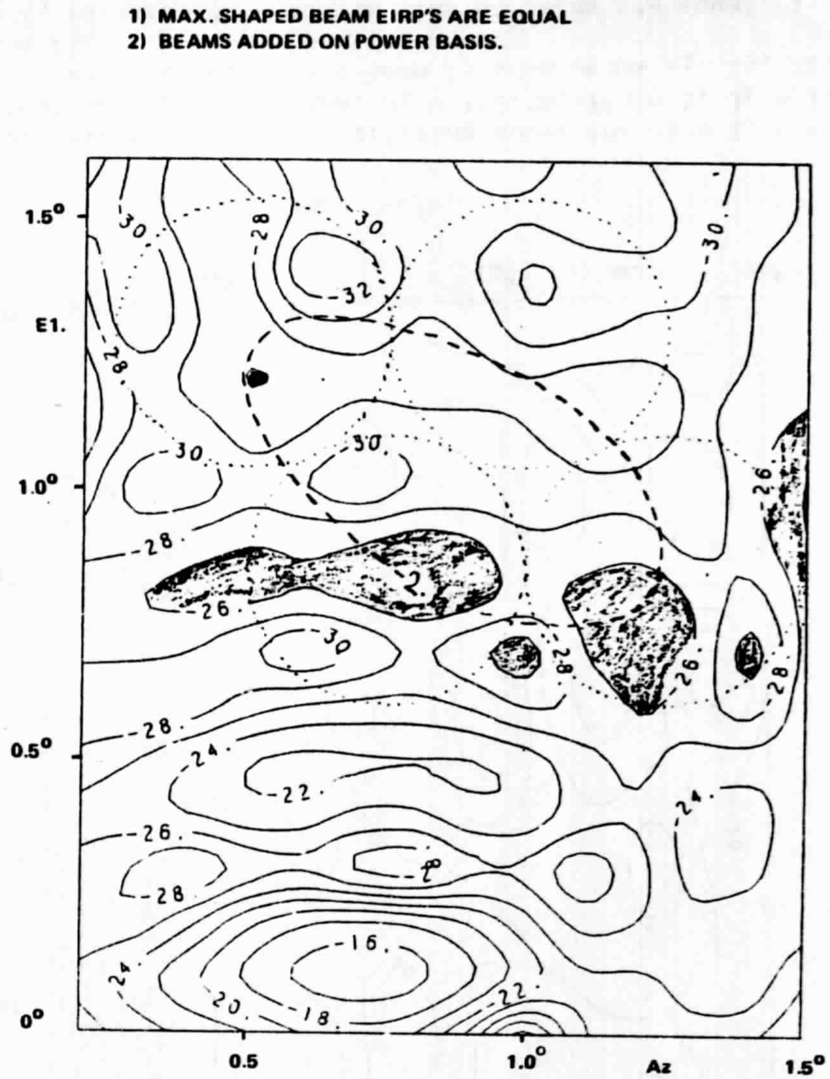


Figure 3-29f.

Contour plots of total interfering power for Plan 1, $\alpha = .5°$, vertical main polarization, $D_\lambda = 162.56$, $F/D = 1$, $d_\lambda = 1.54$. Plot shows interference in beam 1_2 from beams 1_1, 1_3, 1_4, 1_5, 1_6 and 1_7.
Maximum EIRP of beams are equal, interference is added on power basis.

1) MAX. SHAPED BEAM EIRP'S ARE EQUAL
2) BEAMS ARE ADDED WITH RANDOM PHASE
3) BEAM 5 (DOUBLET) IS OMITTED.

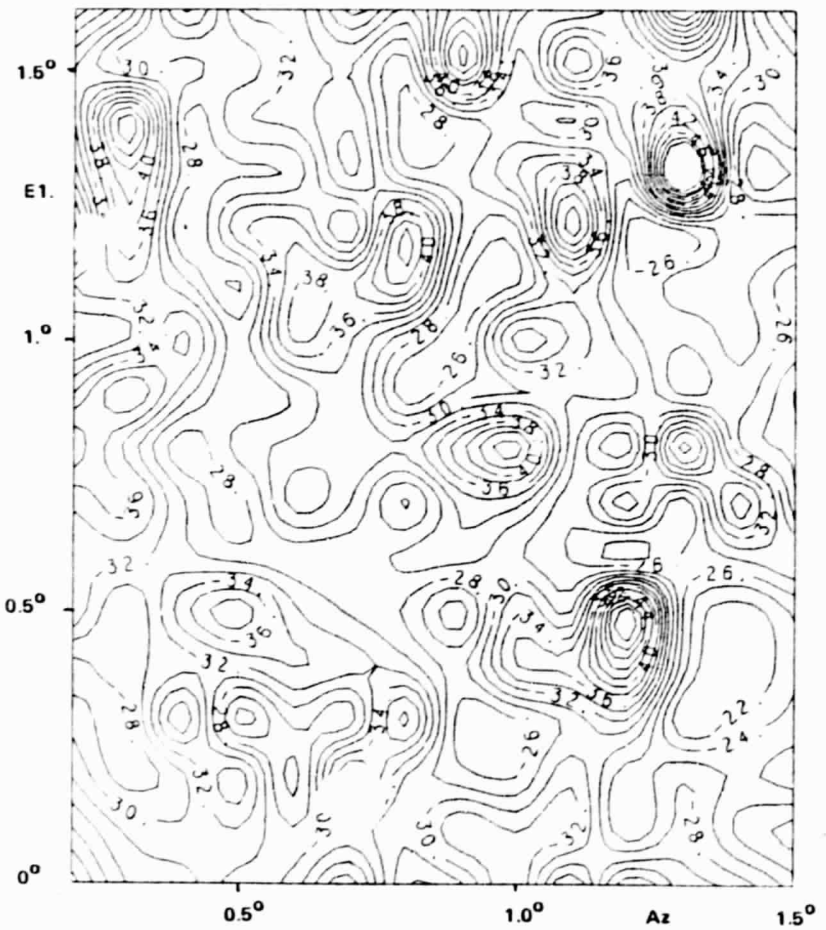


Figure 3-29g.

Contour plots of total interfering power for Plan 1, $\alpha = .5°$, vertical main polarization, $D_\lambda = 162.56$, $F/D = 1$, $d_\lambda = 1.54$. Plot shows interference in beam l_2 from beams l_1, l_3, l_4, l_6 and l_7.
Beam l_5 is omitted.

Maximum EIRP of beams are equal, field are added in random phase.

1) MAX SHAPED BEAM EIRP'S ARE EQUAL
2) BEAMS ARE ADDED ON POWER BASIS
3) BEAM 5 (DOUBLET) IS OMITTED.

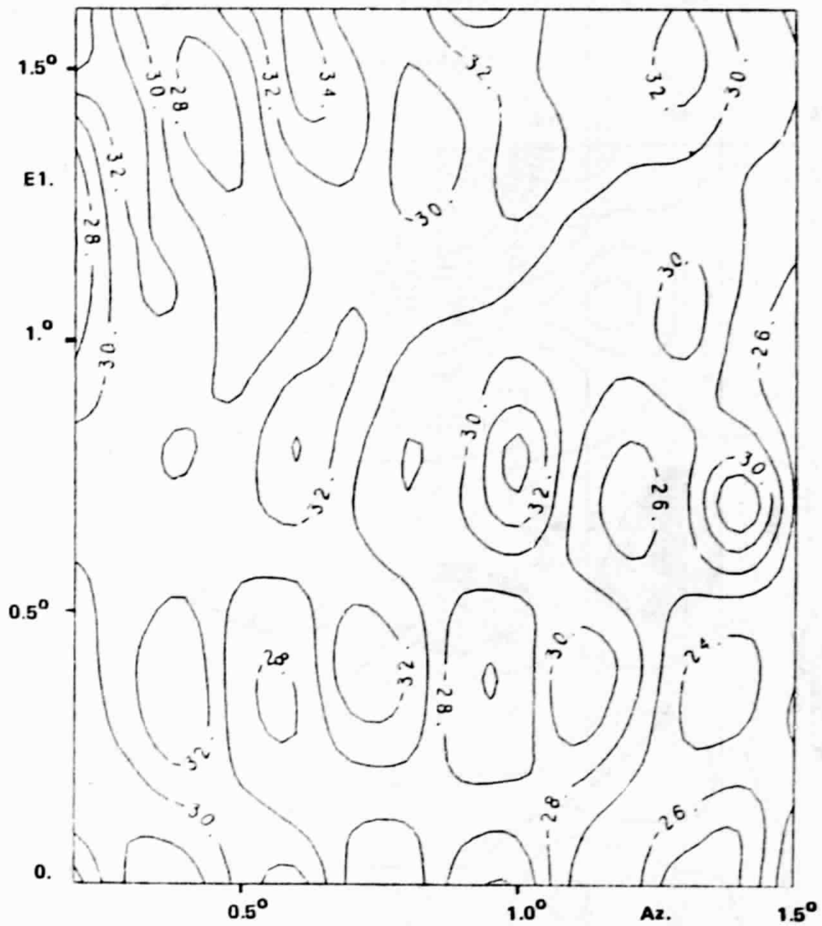


Figure 3-29h.

Contour plots of total interfering power for Plan 1, $\alpha = .5°$, vertical main polarization, $D_\lambda = 162.56$, **F/D** = 1, $d_\lambda = 1.54$. Plot shows interference in beam l_2 from beams l_1, l_3, l_4, l_6 and l_7.
Beam l_5 is omitted. Maximum EIRP of beams are equal, interference is added on power basis.

transmitters are adjusted to produce the same maximum EIRP as beam 1_2 and the phases are equal the resultant limiting sidelobe level is -23 dB and the beam isolation is 21 dB, (see Figure 3-29d). The same results are obtained when the fields are added with random phase, (see Figure 3-29c). For the same condition but using power addition the limiting sidelove level is -25 dB and the beam isolation 23 dB, (see Figure 3-29f).

Figure 3-29g shows the resultant sidelobe level when the most interfering beam, the 1_5 doublet is omitted, the transmit EIRP's are equalized and the fields are added with random phase. In this case the peak sidelobe level is about -25.8 dB and the beam isolation is 23.8 dB, (in practice the 1_5 beam can be "omitted" by uniting it with another quadruplet, for instance with 1_2).

The following conclusions can be drawn from the above charts:

a. For the relatively large shaped beam numbers (6) the in phase, random phase and power addition give about the same results, 20.7 dB, 20 dB and 21.3 dB respectively.

b. Equalization of the transmit EIRP improves isolation from 21.3 dB to 23 dB.

c. Omission of the most interfering beam (doublet) improves beam isolation from 23 dB to 23.8 dB.

The next higher order shaped beam worth considering is the septet, which can be used for Plan 4. Figures 3-30, -31, and -32 show the gain contours for $\alpha = .3^\circ$ of a substantially scanned shaped beam ($\theta_{MA} = 2.5^\circ$). It is evident from the plots that the decay of sidelobe level with this configuration is quite good, but the system requires a relatively large reflector. For Plan 4 $\frac{\epsilon}{\alpha} = 2.5$, thus $\epsilon = .75^\circ$, the sidelobe level is around -24 dB and the beam isolation is about 14 dB.

3.5 ANTENNA RELATED SYSTEM CHARACTERISTICS

The most important characteristics derived from the calculations presented in Section 4 are the variations in sidelobe level as a function of angle for the various topology plans.

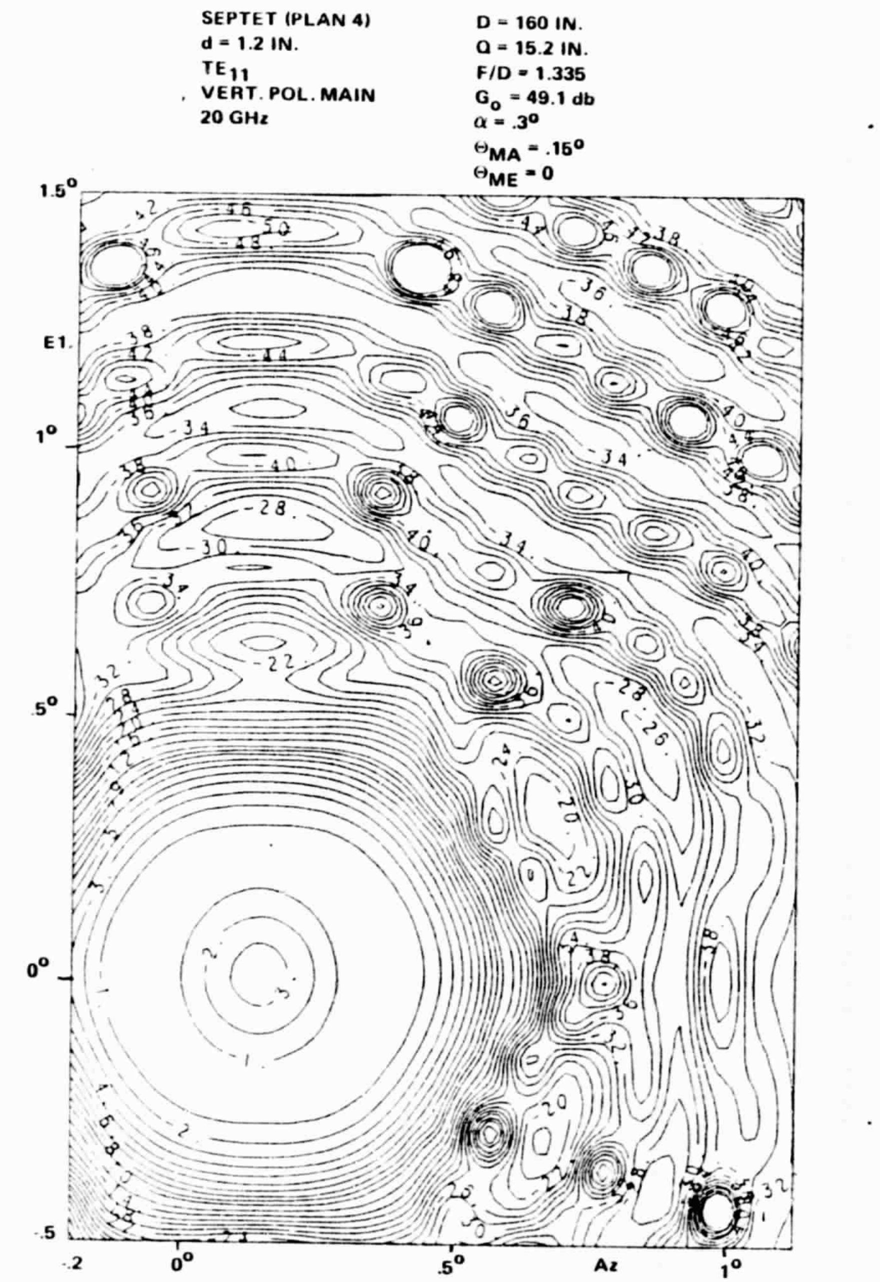


Figure 3-30.

Gain contours of septet beam, D_λ = 270.93, F/D = 1.335, d_λ = 2.032, α = .3°. θ_{MA} = .15°, θ_{ME} = 0, Vertical pol. main.

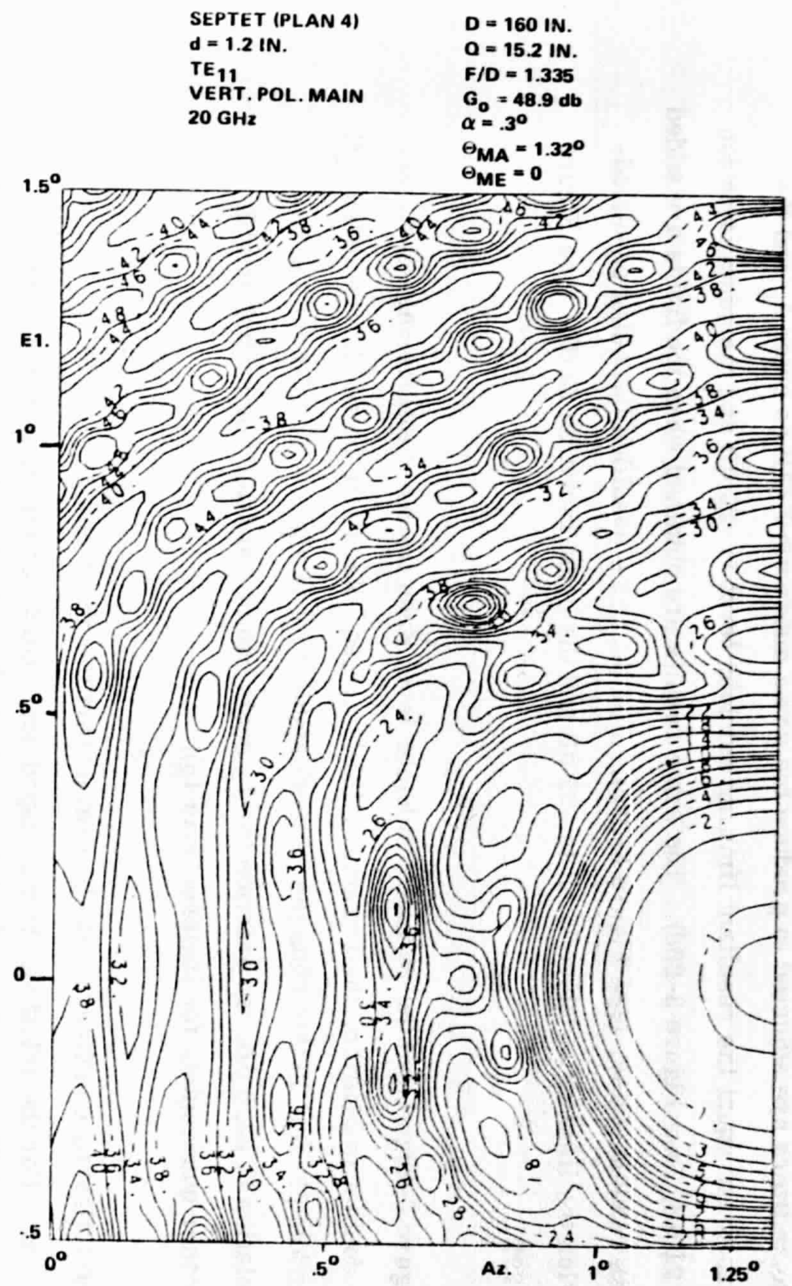


Figure 3-31.

Gain contours of septet beam, D_λ = 270.93, F/D = 1.335, d_λ = 2.032, α = . 3°. θ_{MA} = 1.32°, θ_{ME} = 0, Vertical Pol. Main.

SEPTET (PLAN 4)
d = 1.2 in
TE_{11}

20 GHz

D = 160 in
Q = 15.2 in
F/D = 1.335
G_o = 50.5 db
α = .3° (MAX. SCANNED SHAPED BEAM)
Θ_{MA} = -2.5°
Θ_{ME} = 0

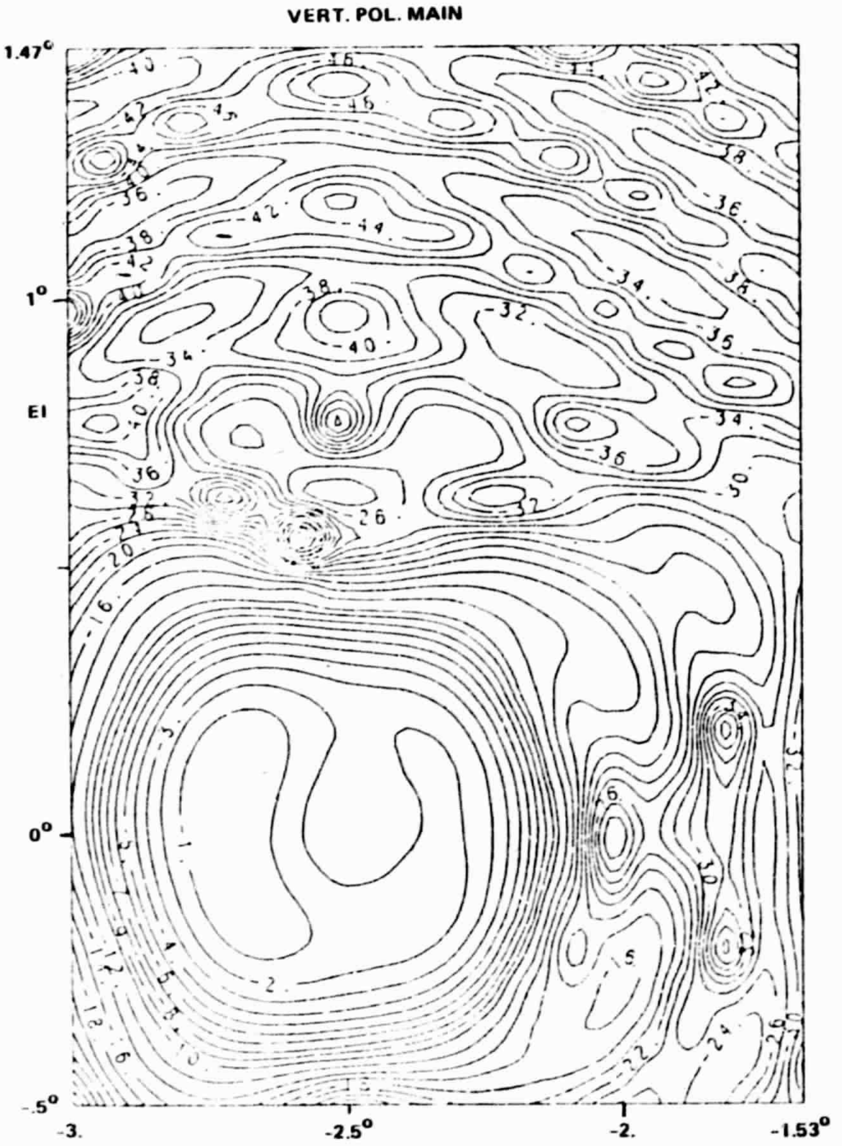


Figure 3-32a.

Gain contours of septet beam, D_λ = 270.93, F/D = 1.335, d_λ = 2.032, α = .3°. θ_{MA} = 2.5°, θ_{ME} = 0, Vertical Pol. Main.

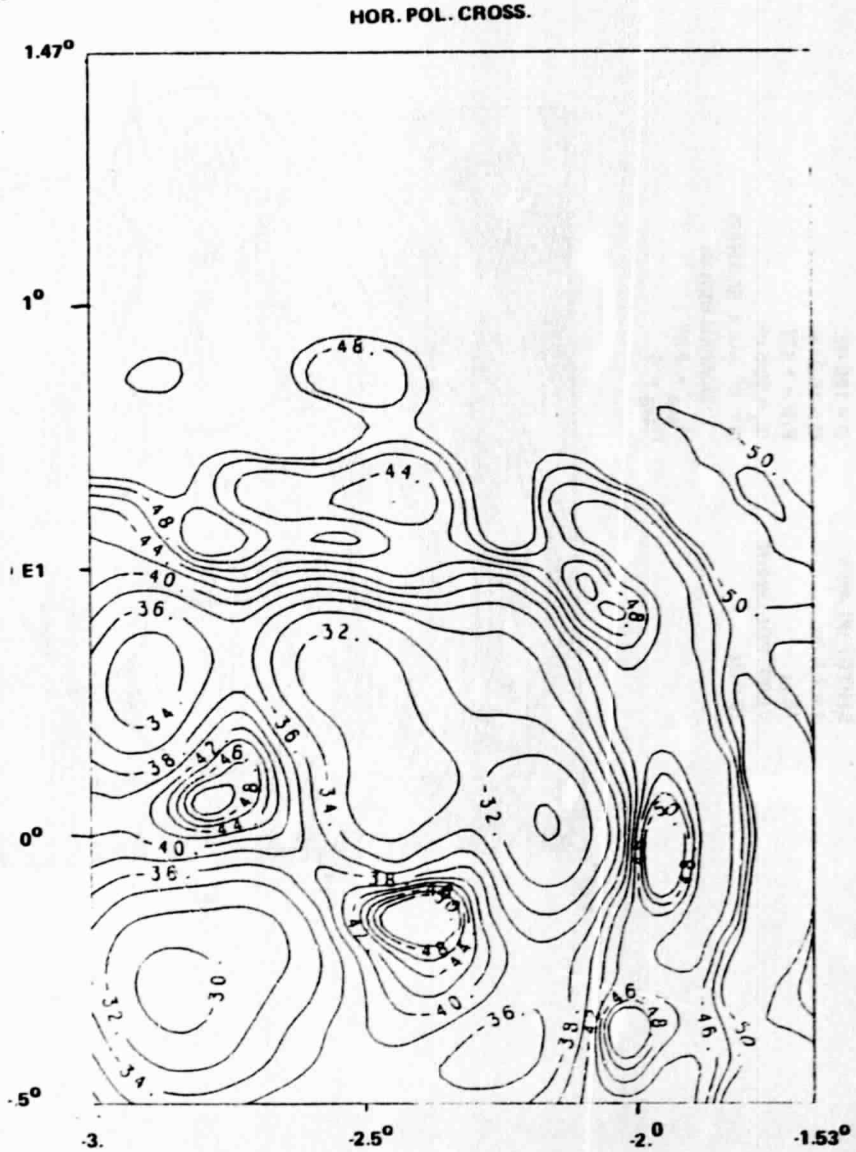


Figure 3-32b.

Gain contours of septet beam, D_λ = 270.93, F/D = 1.335, d_λ = 2.032, α = .3°. θ_{MA} = - 2.5°, θ_{ME} = 0, Horizontal Pol. Cross.

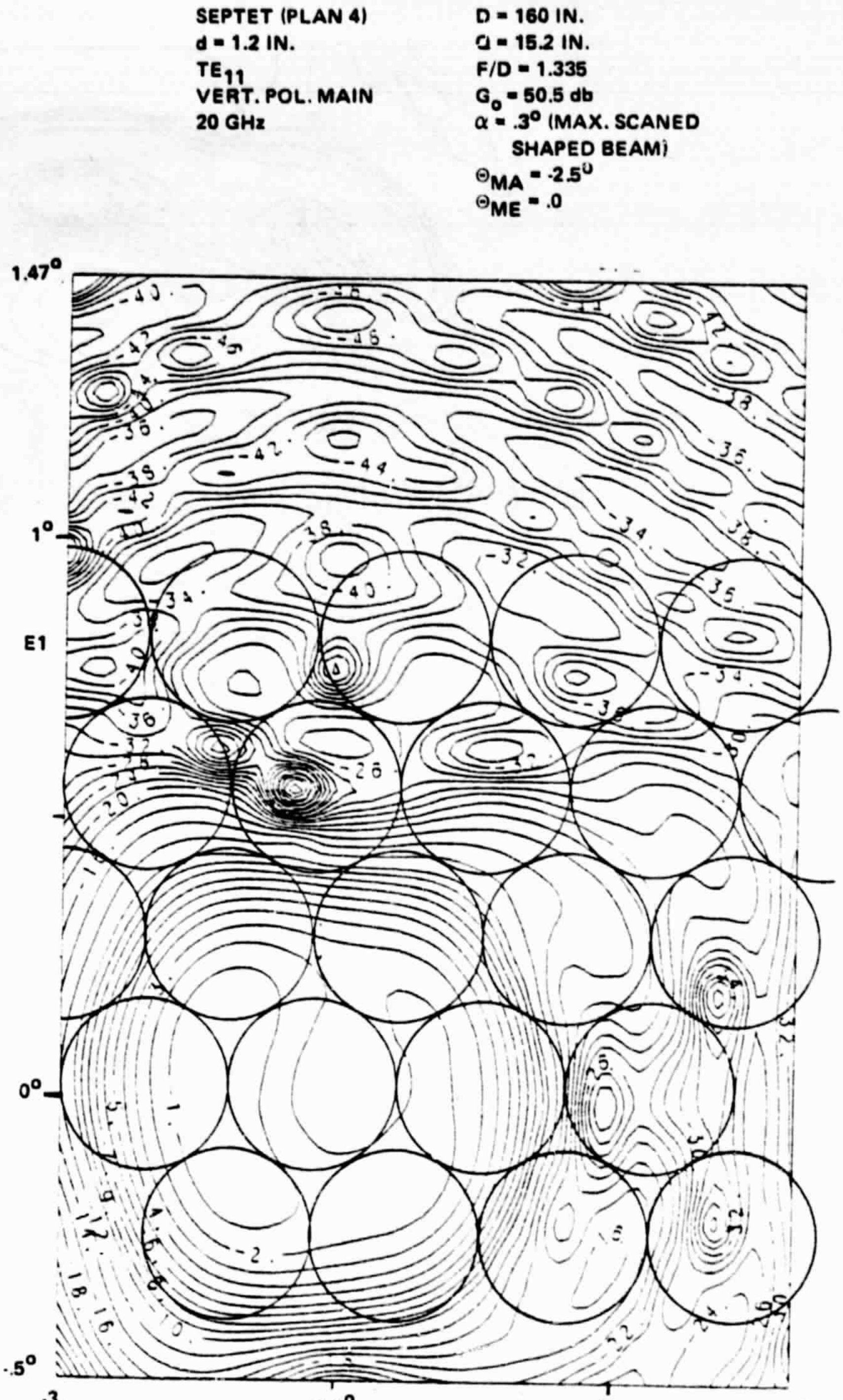


Figure 3-32c. Same as Figure 4.22a, but showing relationship between component beam cells and contours of resultart septet beam.

It is convenient to express this angle, ϵ as measured in the θ_A azimuth and θ_E elevation angle coordinate system between the maximum of the outermost main component beam to a typical direction. If the "shaped" beam is using one component beam, as in the case for Plans 1 and 7 then the angle ϵ is identical to the conventional angular argument of the antenna pattern function. For a higher order shaped beam, for instance a septet, which may contain 7 main beams and 12 auxiliary beams around them, ϵ is measured from the center of an outer main component beam in the plane presenting the slowest fall off of the sidelove level. In order to make the results generally usable it is convenient to normalize ϵ to the cell size α. In the following the ϵ/α ration will be presented for a given sidelobe level .

It is obvious that the sidelobe level of a realizable shaped beam is a function of the angular position or scan relative to the axis of the optical system (axis of paraboloid). This angular position, θ_M is defined as the angle between the axis of the paraboloid and the direction of maximum field. Again the normalized θ_M/α angle will be introduced and ϵ/α will be plotted as a function of θ_M/α .

Figure 3-33 shows the θ_M/α function for singlet beam in the 10 dB to 30 dB sidelobe level range. Results are given for four different kinds of singlet implementations: a) 1 main horn, b) 1 main horn + 6 auxiliary horns, c) 1 main horn + 11 auxiliary horns, d) 1 large multimode horn per singlet and 4 reflectors for the antenna system.

- For a given $\frac{\epsilon}{\alpha}$ the sidelobe level deteriorates with the scan angle.
- For a given scan angle, θ_M/α the sidelove level can be improved by increasing the number of auxiliary horns.
- Decreasing cell zize, α (increasing spectrum reuse) increases the normalized scan angle, thus it deteriorates the sidelobe level.
- The sidelobe level can be decreased by decreasing the scan angle, which is achievable by using an increasing number of reflectors.
- The 4 reflector implementation produces comparable sidelobe level to the 1 reflector implementation with 1 main and 11 auxiliary horns.

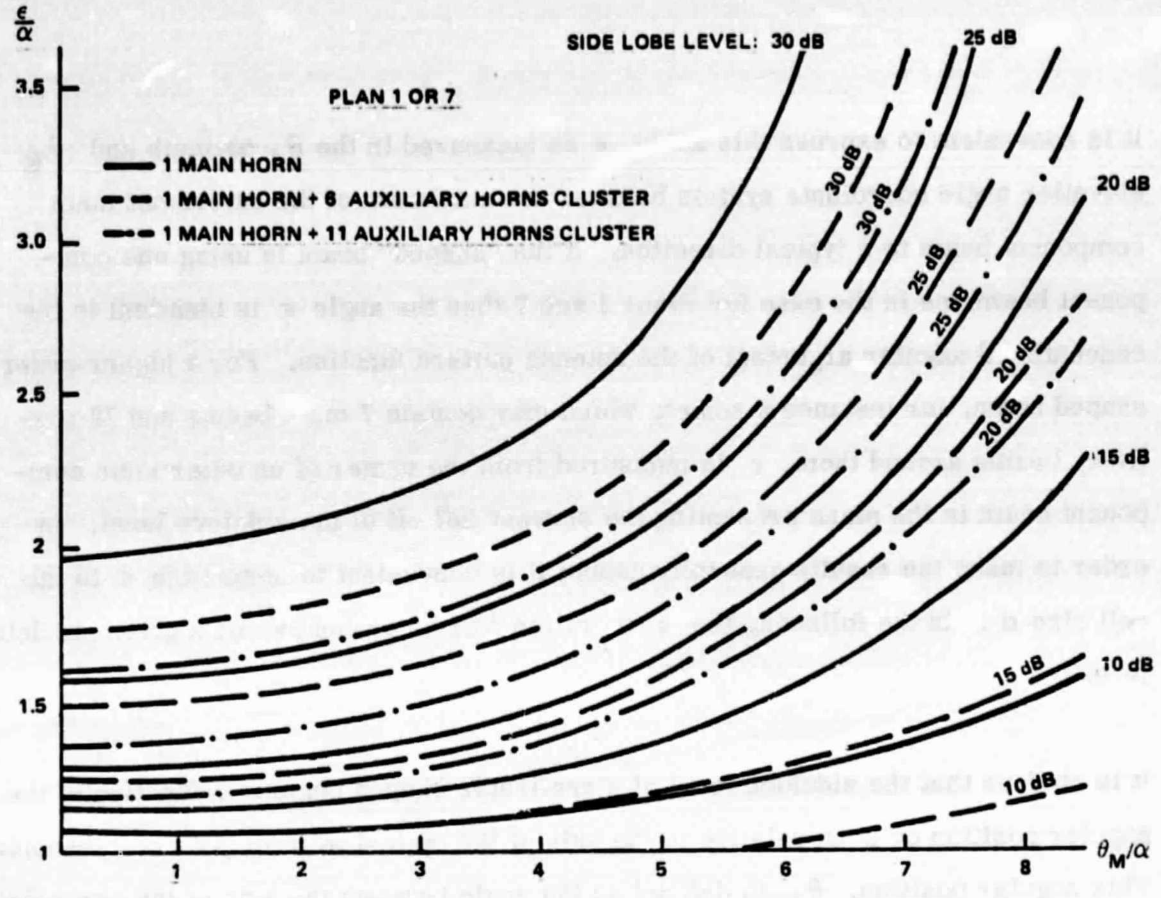


Figure 3-33a. ϵ/α vs θ_M/α for various sidelobe levels at or larger than ϵ/α angles from center of main beam. Beam shape: Singlet.

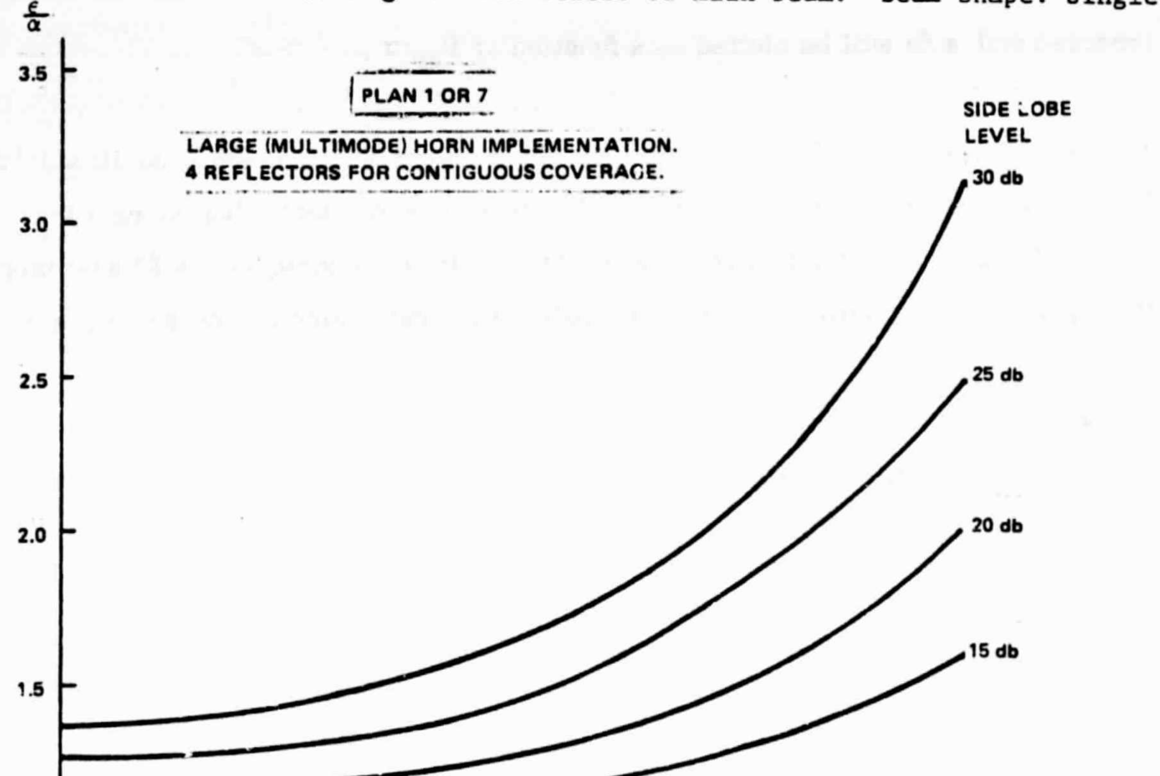


Figure 3-33b. ϵ/α vs Θ_M/α for various sidelobe levels at or larger than ϵ/α angles from center of main beam. Beam shape: Singlet.

Note, that Figure 3-33a was prepared for horn sizes which allow contiguous coverage with one reflector. When the number of reflectors are increased to allow the use of larger horns, then the performance becomes comparable to that achievable with larger horn clusters. The performance for this case is shown in Figure 3-33b.

Table 3-5 shows some examples obtained from Figure 3-33 for $\theta/_M/\alpha = 7$ (as necessary to implement $\alpha \sim .5°$) for $\epsilon/\alpha = 1.23$ (Plan 1) and $\epsilon/\alpha = 2.5$ (Plan 7). It can be seen from Table 3-5 that the sidelobe levels at $\epsilon/\alpha = 1.23$ are very poor even if a very large number of small horns or one large horn and 4 reflectors are used. The sidelobe performance is probably unacceptable even with 8 reflectors, which is hardly a practical proposition. Another conclusion is that a single reflector with 1 main + 11 auxiliary horns yields nearly the same sidelobe performance as the 4 reflector antenna system with large horns. It is interesting to observe from Figure 3-1b, that using the large multimode horn and separate reflector for each singlet ($\theta_M/\alpha = 0$) the sidelobe level at $\epsilon/\alpha = 1.23$ (Plan 1) is 22.5 dB. This value yields 18.5 dB beam isolation at the -4 dB points. This is the best implementable case with Plan 1.

TABLE 3-5. COMPARISON BETWEEN VARIOUS SINGLET IMPLEMENTATIONS

Implementation	Sidelobe Level (db) $\frac{\epsilon}{\alpha} = 1.23$ (Plan 1)	Sidelobe Level (db) $\frac{\epsilon}{\alpha} = 2.5$ (Plan 7)
1 main horn	8.6	21.5
1 main horn + 6 auxiliary horns	12.5	23.5
1 main horn + 11 auxiliary horns	13.5	25
1 (double dia.) horn + 4 reflectors	11	27.5
1 (double dia.) horn + 8 reflectors	20	35

Figures 3-34 through 3-37 exhibit the $\frac{\epsilon}{\alpha}$ vs $\frac{\theta_M}{\alpha}$ charts for the doublet, triplet quadruplet and septet beams. All these beams are in the so-called higher order shaped beam categories, which means that they use more than 1 main component beam to build up the shaped beam. All these higher order beams may contain in practice auxiliary component beams but the calculation of radiation characteristics is increasingly more expensive and these data are not available at the present.

A study of these figures reveals that the sidelobe level characteristics improve with increasing numbers of main component beams (then is not shown in the figures, but further improvement is possible by increasing the number of horns per shaped beam by using additional auxiliary horns). For instance at $\theta_M/\alpha = 7$ the 25 dB sidelobe level is achieved for the

Doublet	at	$\epsilon/\alpha = 3.25$
Triplet	at	$\epsilon/\alpha = 2.12$
Quadruplet	at	$\epsilon/\alpha = 2.34$
Septet	at	$\epsilon/\alpha = 2.10$

In the above series the triplet represents a small anomally. It behaves better than the number of horns suggests. This is caused by the fact that in Plan 8 the interfering triplet beams can be arranged in such a manner that they produce a low sidelobe level region toward the adjacent beams. (Such regions are in the direction the apexes of the triplets are pointed.) Thus Plan 8 utilizes the triplet beams in a very advantageous manner. This feature of the triplet makes the applicable sidelobe level practically independent of the scan angle θ_M.

On the basis of the topology plans presented the relationship between the cell size α and spectrum reuse N can be established for all presented 8 topology plans. This yields an $\alpha = \alpha$ (N) function for each plan, (see Figure 3-42.) The applicable $\frac{\epsilon}{\alpha}$ value characterizing adjacent beam separation can be determined from simple geometry and is presented in Table 3-2. The edge of contour to peak gain difference

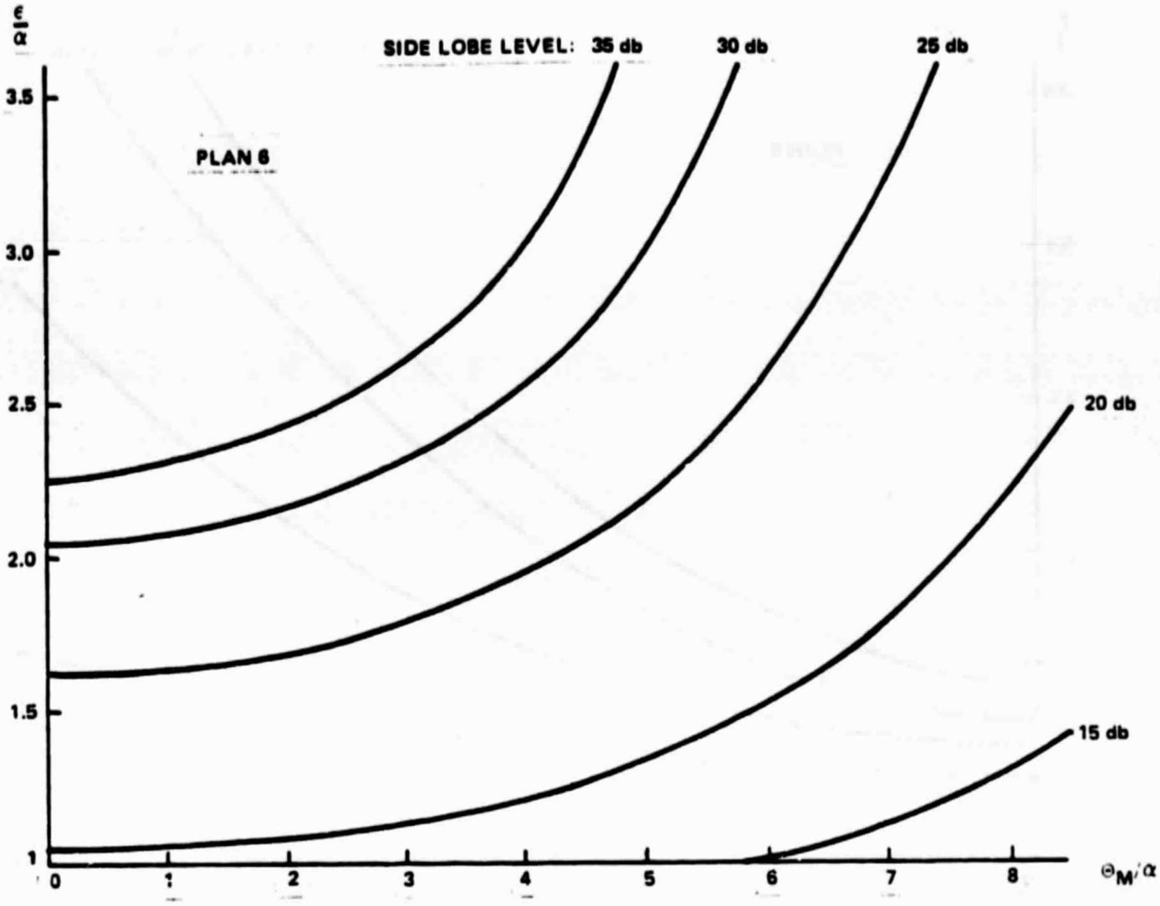


Figure 3-34. ϵ/α vs Θ_M/α for various sidelobe levels at or larger than ϵ/α angles from center of main beam. Beam shape: Doublet (East/West).

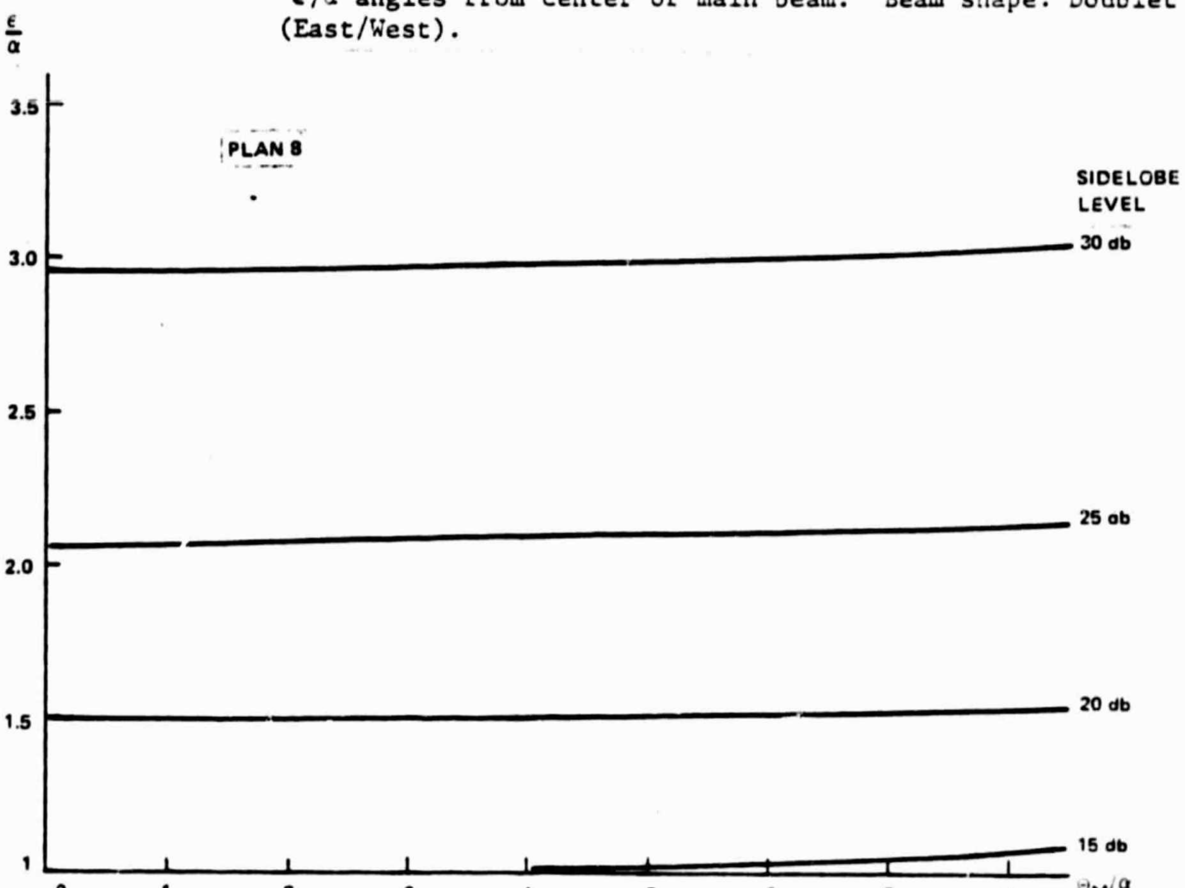


Figure 3-35. ϵ/α vs Θ_M/α for various sidelobe levels at or larger than ϵ/α angles from center of main beam. Beam shape: Triplet.

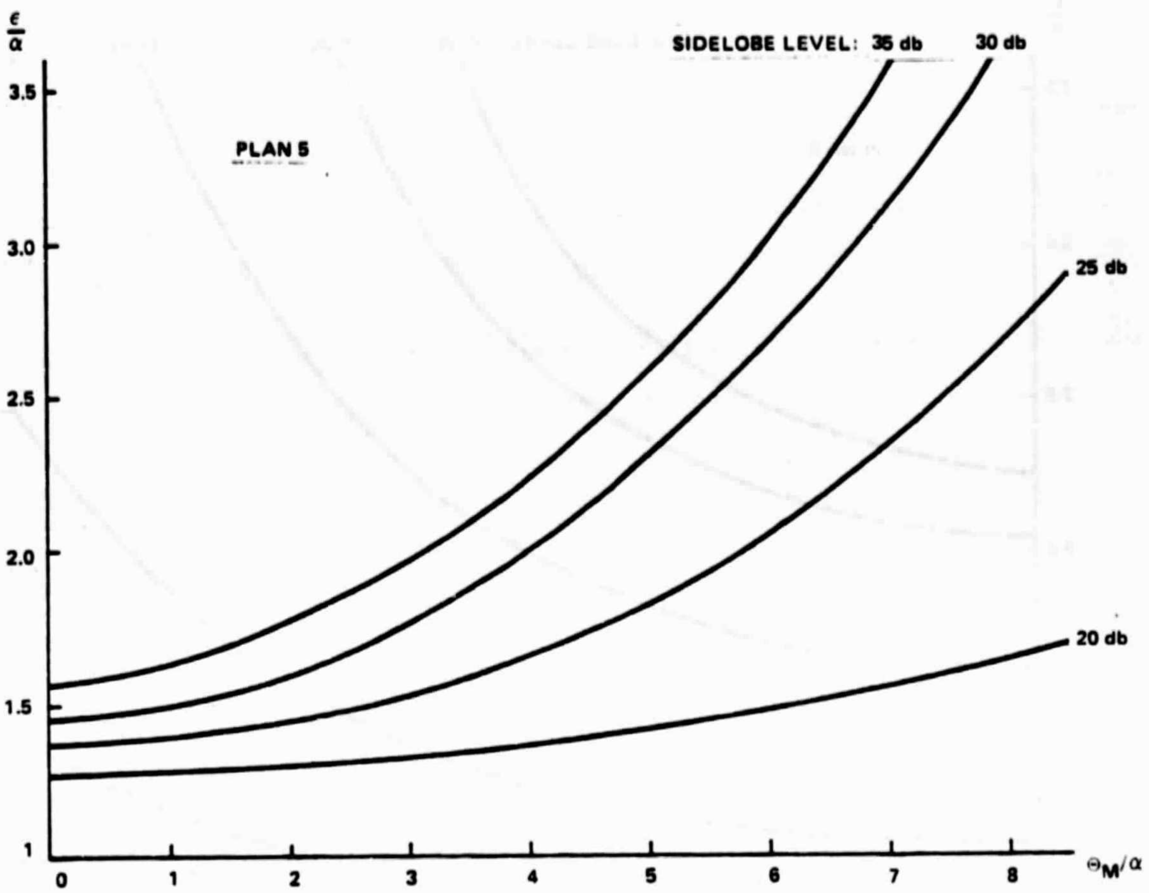


Figure 3-36. ϵ/α vs θ_M/α for various sidelobe levels at or larger than ϵ/α angles from center of main beam. Beam shape: Quadruplet.

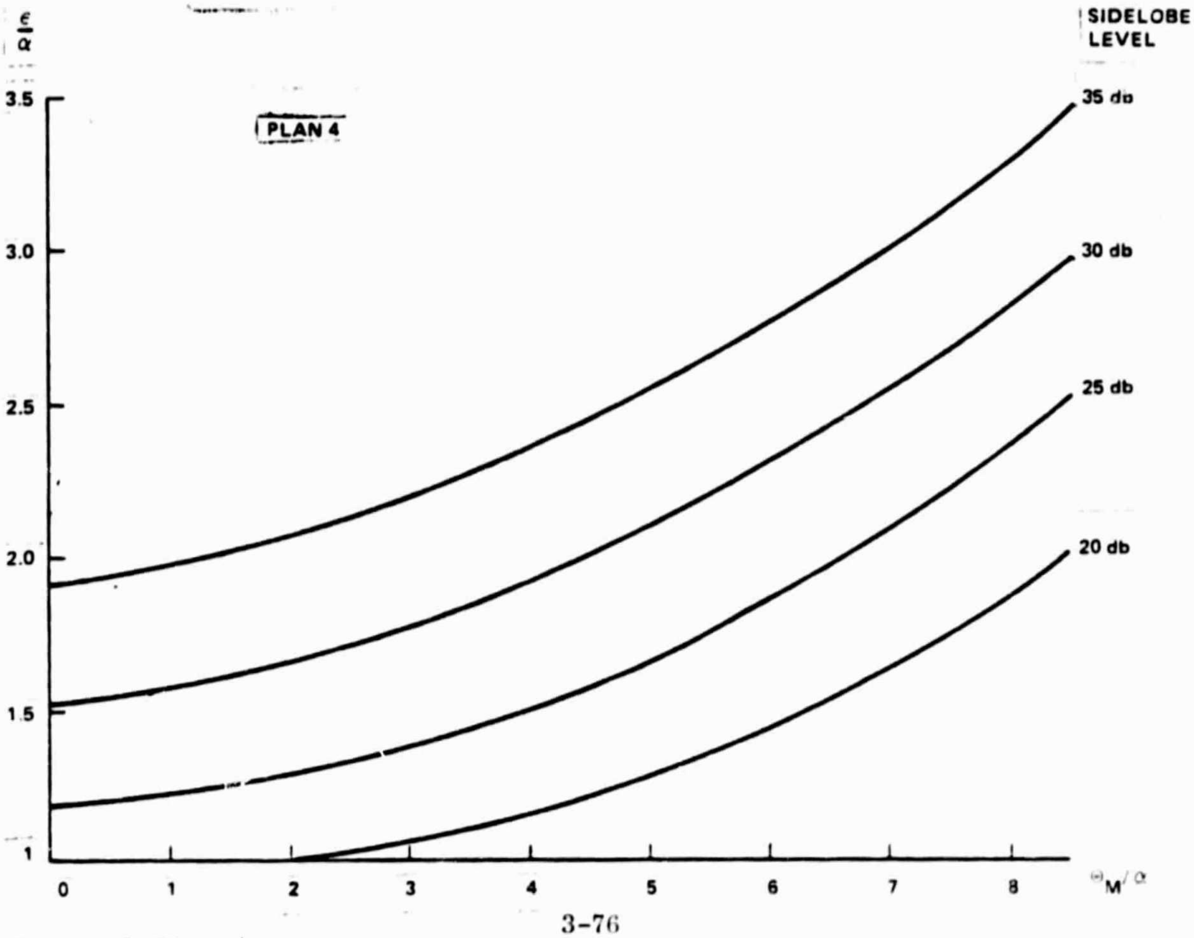


Figure 3-37. ϵ/α vs θ_M α for various sidelobe levels at or larger than ϵ/α angles from center of main beam. Beam shape: Septet.

ΔG_c, can be obtained for each type of shaped beam as exhibited by the pattern calculation results of Section 4. These values are shown for each plan in Table 3-5. Finally using Figures 3-33 through 3-37 and noting that for the lower 48 states $\theta_M \sim 3.5°$ the sidelobe level, $\varkappa$ can be determined for each topology plan for the limiting $\theta_M/\alpha = \theta_M/\alpha$ (N). The $\varkappa - \Delta G_c$ difference then gives the most important antenna characteristics, the adjacent beam isolation, I_{Bsdj} as a function of spectrum reuse, N.

In Figures 3-38 through 3-41 I_{Badj} is plotted as a function of N for the various topology plans. The curves are indexed to show the applicable conditions. The first index shows the Plan, the second index the number of horns forming the shaped beam (including auxiliary horns if applicable), the third index shows the number of reflectors. The 2 reflector configuration was calculated on the basis of dividing the country into an Eastern and a Western half and using 1 reflector for each half.

The figures cover a total of 20 different configurations. Each configuration represents a different type of feed configuration, but each uses an offset fed paraboloid. Since for the beam numbers presently under consideration the beam scan is the limiting contributor in adjacent beam isolation and the offset nature of the optics has only secondary importance it can be expected that symmetrical optics, like lenses will produce essentially the same results (their added complexity, weight, frequency dependency, loss and cost are additional considerations). Thus the restriction of the present investigation to offset fed paraboloids does not represent a major limitation of generality.

It can be seen from the figures, that the best adjacent beam isolation among the studied 20 cases is 5.4.2, i.e., Plan 5 using 4 horns for each quadruplet beams and 2 reflectors. For the 1 reflector realizations and for N $\sim$ 8, the case characterized by 5.4.1 is quite promising ($I_{Badj} \sim 25$ dB). It may be pointed out that I_{Badj} is not the resultant beam isolation in a multiple shaped beam system where several shaped beams

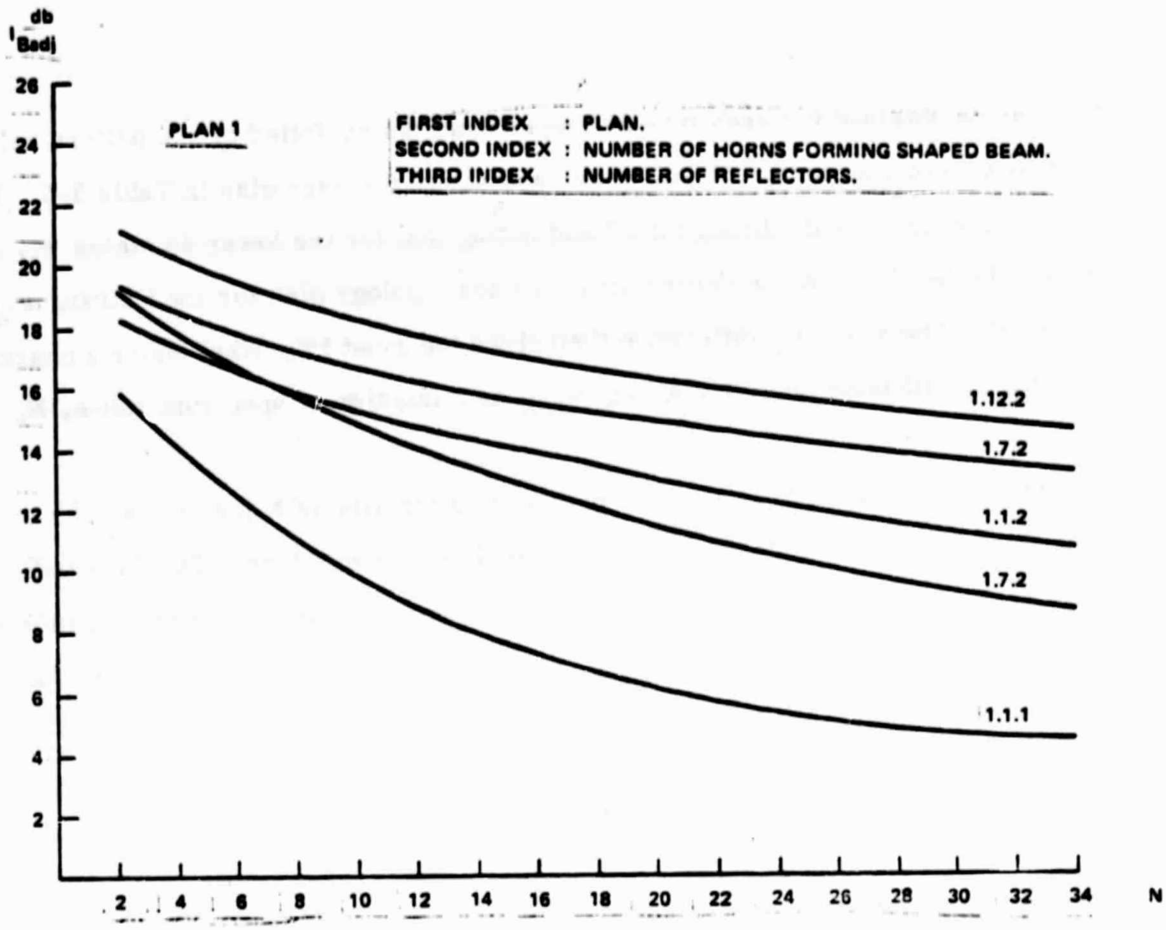


Figure 3-38. Adjacent beam isolation vs number of spectrum reuse. Plan 1.

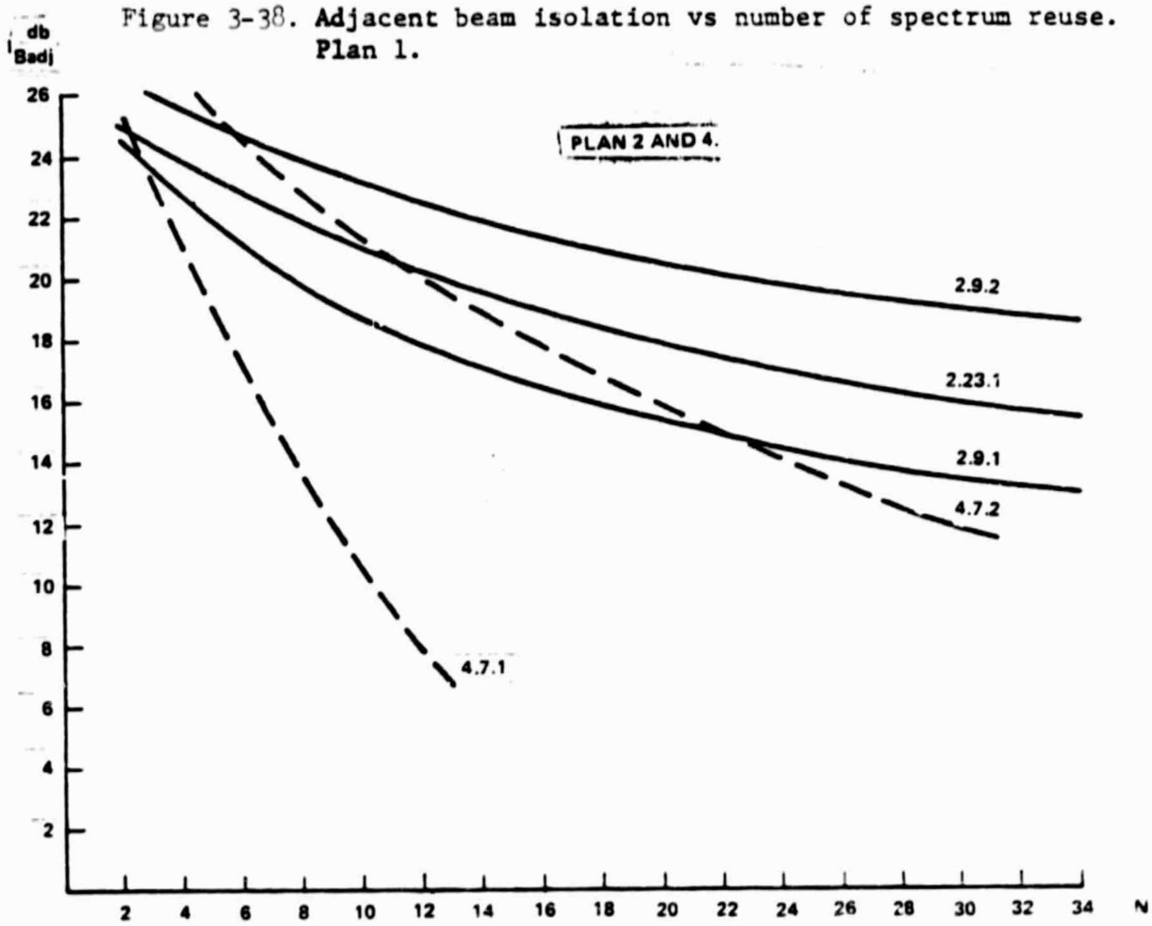


Figure 3-39. Adjacent beam isolation vs number of spectrum reuse. Plan 2 and 4.

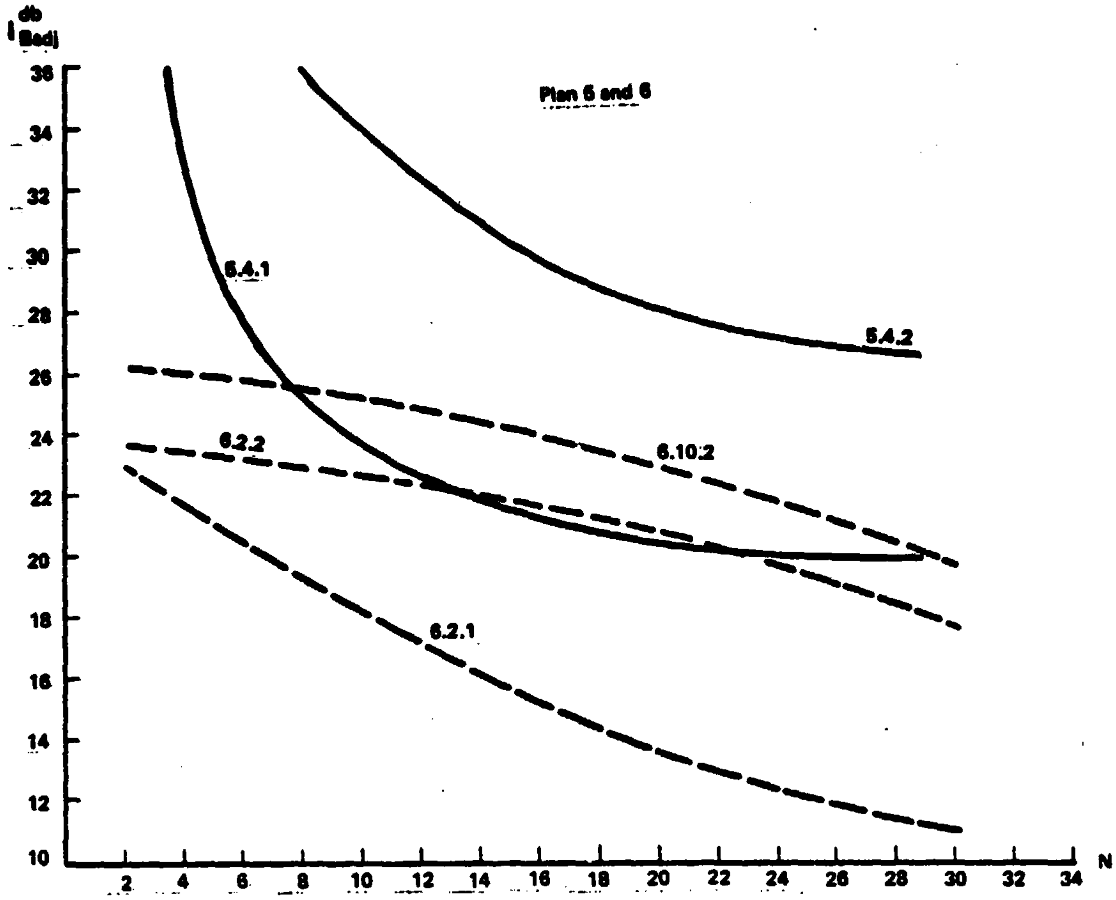
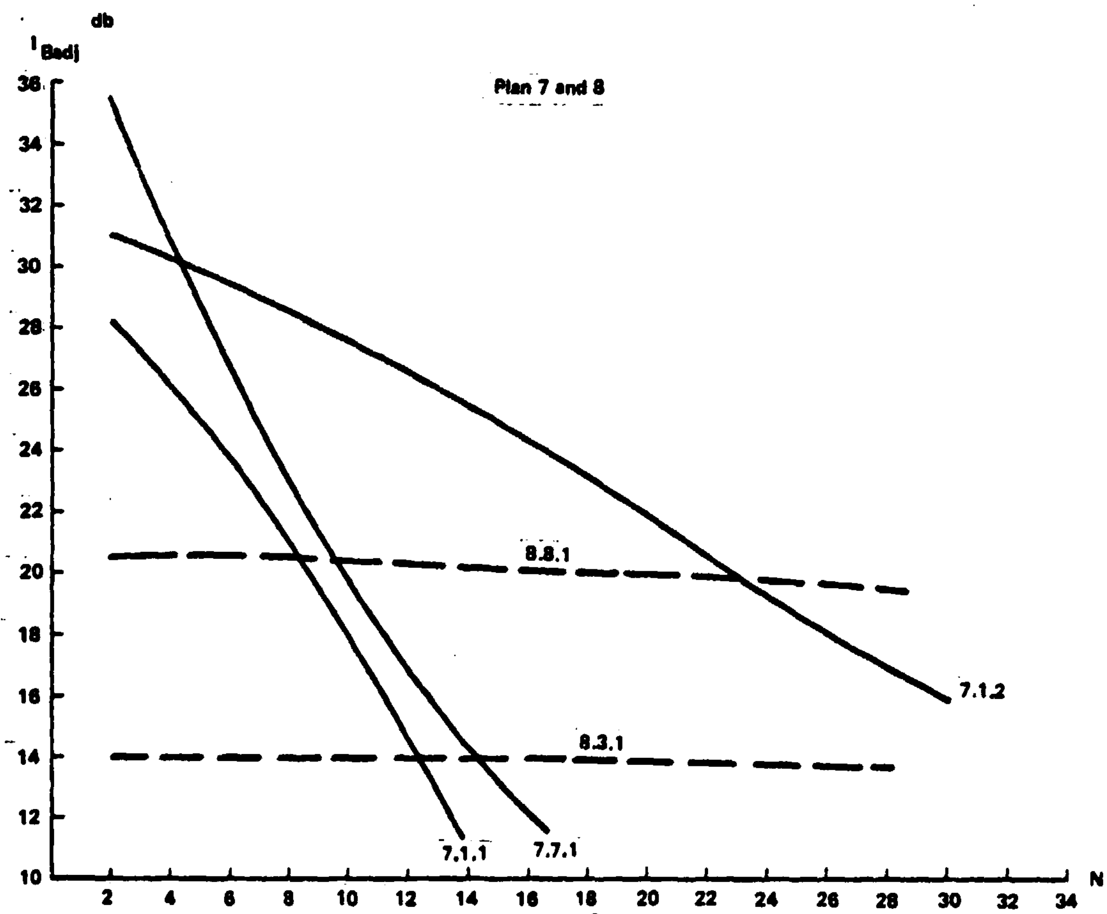


Figure 3-40. Adjacent beam isolation vs number of spectrum reuse. Plan 5 and 6.



C-3

Figure 3-41. Adjacent beam isolation vs number of spectrum reuse.

use the same spectrum. However, I_{Badj} gives a good indication since the more distant interfering shaped beams typically have a much smaller effect. (The resultant beam isolation, I_B for Plan 5, $\alpha = .05°$, for instance is $\sim$ 2 dB poorer than I_{Badj}).

Figure 3-42 presents the relationship between the cell diameter in degrees and the number of spectrum reuse for the different plans. One obvious basic conclusion from the figure is that increasing spectrum reuse requires inversely decreasing cell diameters. Again in the $8 \leq N \leq 16$ range $.87° \geq \alpha \geq .22°$ for the various plans and $.6° \geq \alpha \geq .4°$ for Plan 5.

Figure 3-43 shows the available contour gain as a function of spectrum reuse for the various topology plans. The contour gain of course is a very vital system characteristic because for a given transmitter power and receive G/T it determines the link carrier to noise ratio and thus the available margin. It can be seen that G_c increases with N for all topology plans. Plan 7 is the leader, Plan 5 is a strong second. For the range of spectrum reuse numbers of $8 \leq N \leq 16$, G_c is between 38 dB and 50 dB if all studied plans are considered, for Plan 5 43.6 dB $\leq G_c \leq$ 46.8 dB.

Figure 3-44 exhibits the antenna diameter in wavelengths as a function of spectrum reuse. It can be seen that in terms of antenna surface area utilization the various plans are vastly different. For instance if $8 \leq N \leq 16$ then $96 \leq D_\lambda \leq 370$ for the various plans. At f = 30 GHz this results in 1.44 m $\leq D \leq$ 5.55 m. For Plan 5, which is in the middle of the pack from this point of view the range is only $136 \leq D_\lambda \leq 205$ or 2.04 m $\leq D \leq$ 3.07 m at 20 GHz.

On the basis of the previously presented data the most important characteristics of various system implementations can be approximately calculated. In order to make the comparison of the various implementations realistic a given effective (reused) system bandwidth was selected and the calculations were carried out for this total capacity in each case. Table 3-6 shows a summary of some of the calculated data utilizing the assumptions of Table 3-1.

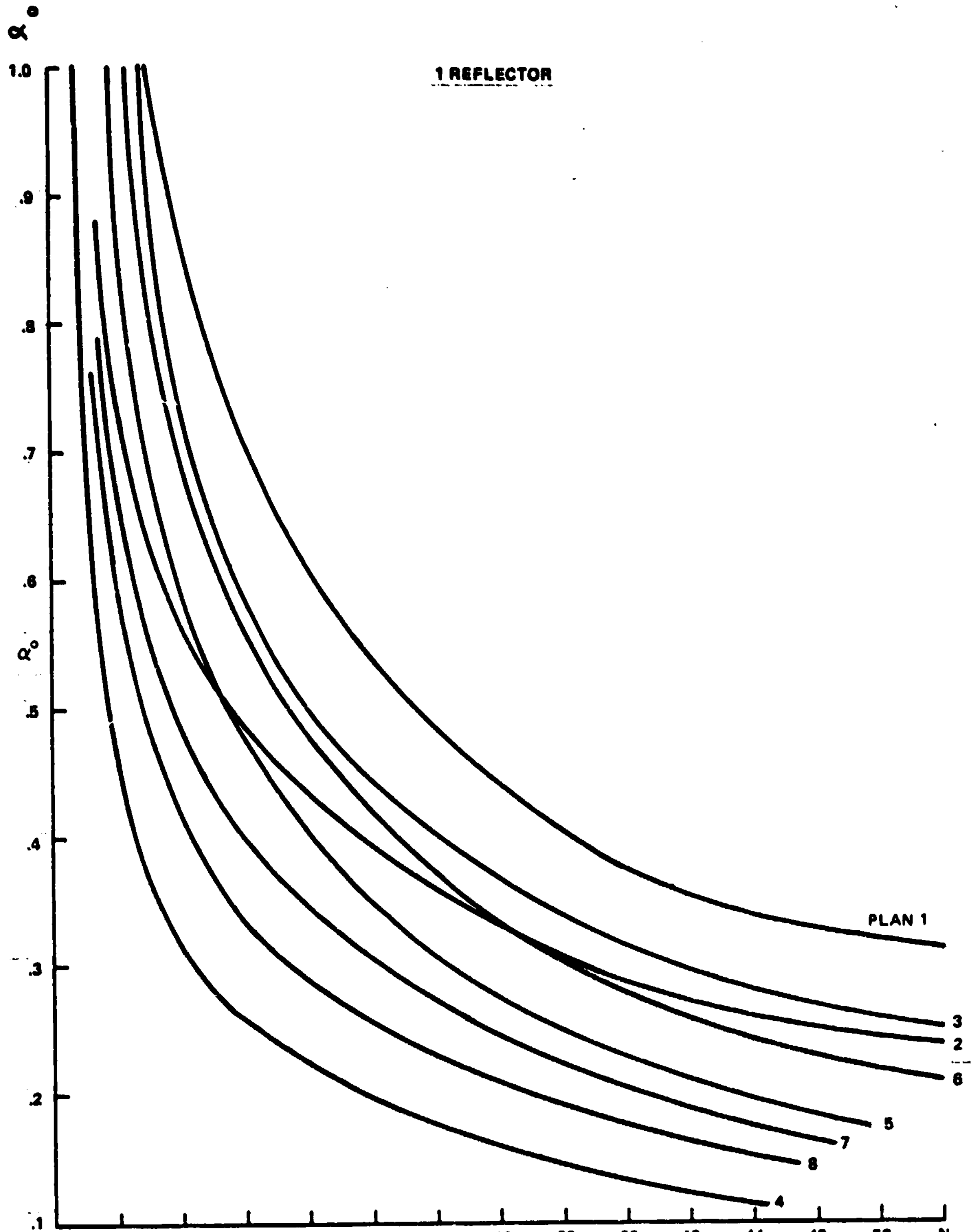


Figure 3-42. Component beam center to center separation vs number of spectrum reuse for various beam plans using 1 reflector.

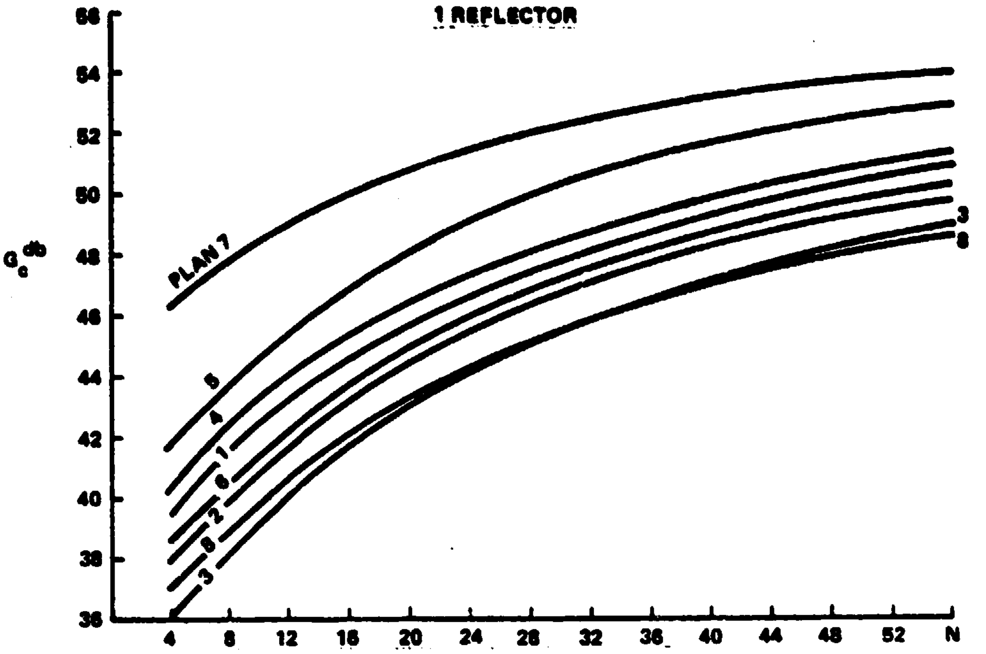


Figure 3-43.

Contour gain vs. number of spectrum reuse for various beam plans using 1 reflector

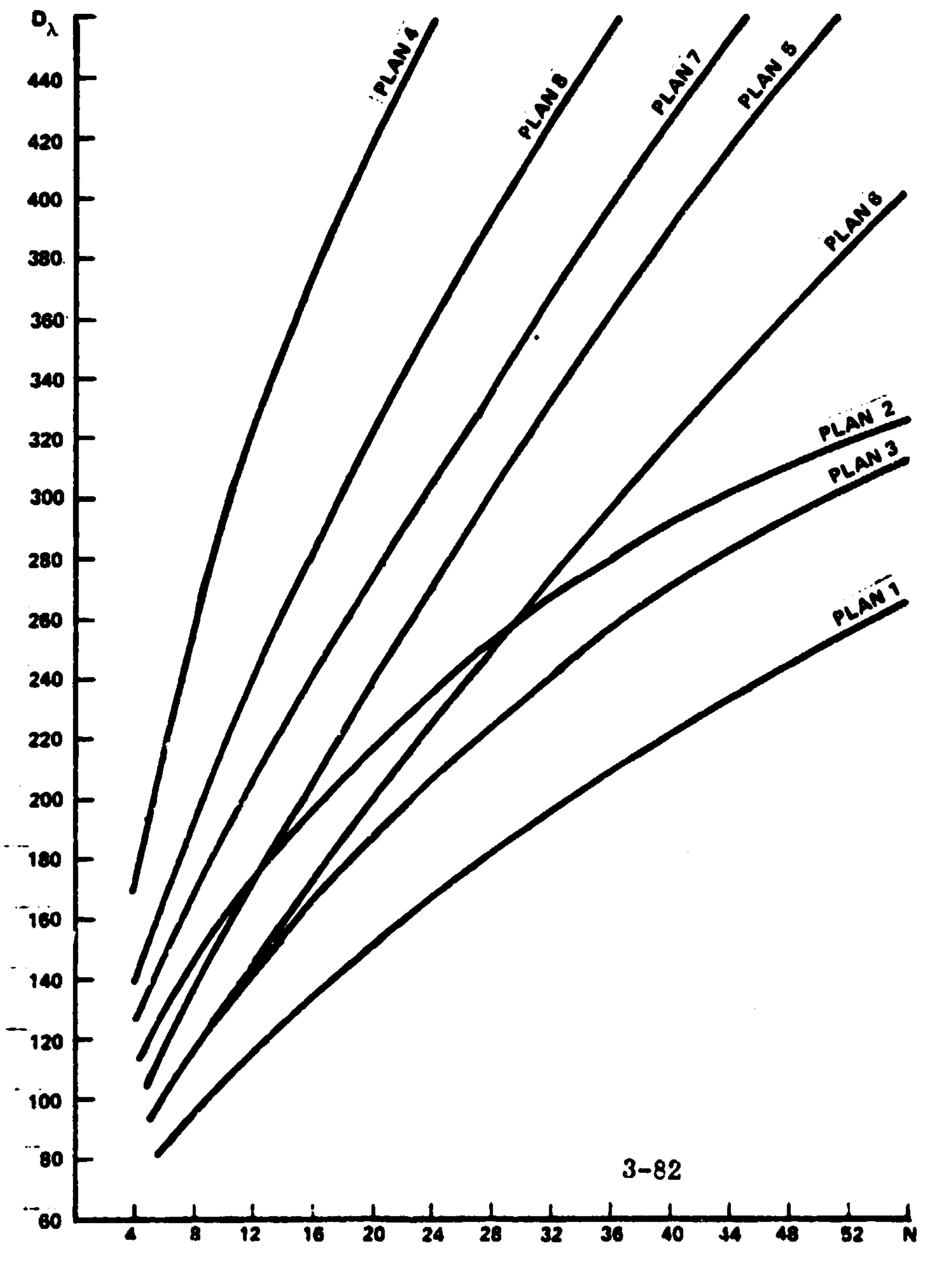


Figure 3-44.

Antenna reflector diameter vs number of spectrum reuse for F/D = 1 for various beam plans.

TABLE 3-6. SUMMARY OF CHARACTERISTICS FOR C_o = 16.384 GHz EFFECTIVE COMMUNICATION SYSTEM BANDWIDTH (α_{atm} = 1.2 dB, α_{ES} = .5 dB)

Plan Identification Number	Cell Diameter or Component Beam Center to Center Separation (Degrees)	Number of Feed Horns or Component Beams of Satellite Antenna	Number of Spot Beams	Total Up or Downlink Frequency Band (Allocation Frequency Bandwidth)	Maximum Contiguous Frequency Band Available in One Shaped Beam (Channel)	Total Maximum Frequency Bandwidth Available in One Component Beam	Number of Spectrum Reuse	Diameter of Satellite Antenna (in meters), at 20 GHz	Peak Gain of Satellite Antenna Shaped Beam	Contour Level, Below Peak Gain	Slope of the Pattern at its Contour Coverage 1.05°	Attenuation of the Feed Circuit of Satellite Beam	Total Attenuation From the Peak Gain, Including Contour Level and Earth Station Receive Losses (1.2 dB Atmospheric Loss and 0.5 dB Earth Station Loss)	Effective Gain at the Coverage Contour Lines ($G_o - \alpha_{total}$)	Carrier to Interference Noise Ratio
Plan	$\alpha°$	n	N_B	ΔF^{MHz}	B^{MHz}	F_B^{MHz}	N	$D_{20\ GHz}^{m}$	G_o^{db}	ΔG_c^{db}	$S_{.05°}^{db}$	α_{sat}^{db}	α_{Total}^{db}	G_{eff}^{db}	C/I^{db}
1	.75	32	32	1536	512	512	10.67	1.625	50.2	-4	1.2	-1.0	- 6.7	43.5	18.5
2	.30	178	44	744.7	372	372	22	4.063	48.6	-2 (EW)	1.8	-2.5	- 6.2	42.4	17.5
4	.30	178	32	1536	512	512	10.67	4.063	50.5	-8	7.1	-2.0	-11.7	38.8	21
5	.50	68	64	2048	256	1024	8	2.438	47.7	-2	1.0	-2.0	- 5.7	42.0	23
6	.50	68	45	1092.3	364	364	15	2.438	50.6	-7	3.5	-1.5	-10.2	40.8	24
7	.60	46	46	3205	356.2	356.2	5.11	2.032	52.1	-4	1.5	-1.0	- 6.7	45.4	22
8	.30	178	67	978	244.5	244.5	16.75	4.063	54.2	-12	10.7	-2.0	-15.7	38.5	20

The cell size, α characterizes the resolution of the multibeam antenna system. The largest resolution and smallest α is desirable. In nonuniformly distributed traffic situations, the small α provides flexibility of planning, since it allows concentration of traffic to small geographical areas.

The number of component beams, n, characterizes the complexity of the antenna feed, since for each component beam a separate feed horn and associated circuit is necessary. Thus the smallest n is desirable.

The number of shaped beams, N_B (see Figure 3-45) is near the number of points or equipment terminals for the antenna, thus it relates to the number of up and downlink passes and to the size of the on-board routing switch.

ΔF is the allocation bandwidth necessary for the communication system. In an early implementation of a K_a band system the potentially available bandwidth probably will

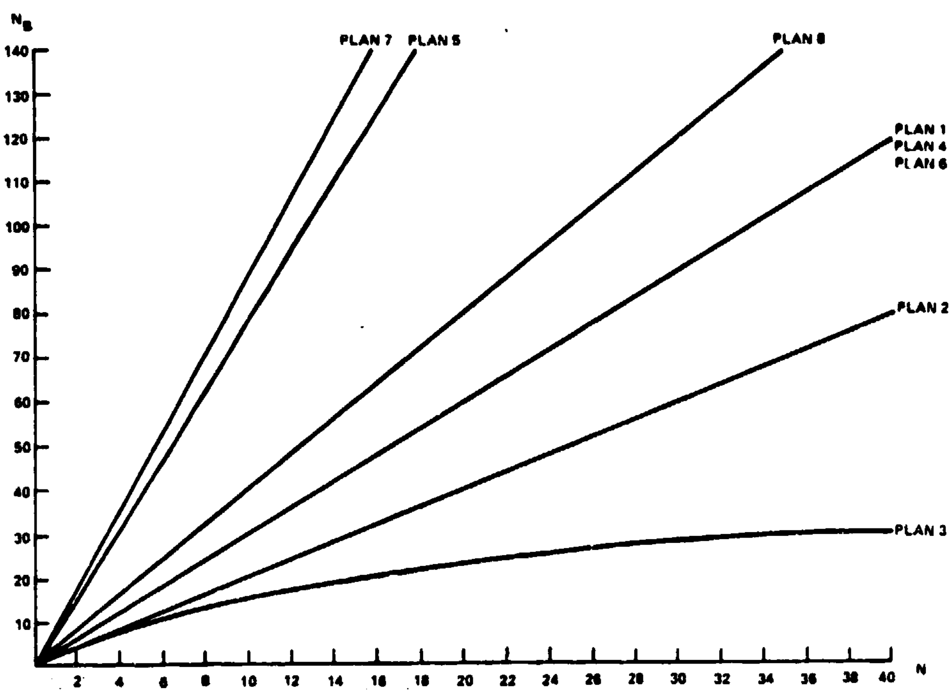


Figure 3-45. Number of beams vs spectrum reuse for various beam plans.

not be exhausted, thus ΔF may not be a limiting characteristics. However, eventually ΔF must be minimized for best utilization of natural resources.

B is the bandwidth available in a shaped beam. For this parameter the largest value is desirable.

F_B is the maximum potentially available bandwidth in a cell; the largest value is desirable.

N is the spectrum reuse number. A high value for this characteristic is desirable, but again its actual value is not a system "driver" for early implementations.

$D_{20\ GHz}$, the antenna diameter at 20 GHz is a major system complexity characteristic. It is questionable in the 1980-85 time period whether a deployable antenna is practical because of the required surface accuracies. The largest of the antenna diameters shown in Table 3-6 are about at this limit.

G_o is the antenna gain at the maximum gain point of a shaped beam without any losses. The actual available gain at the contour of the coverage, G_{eff} will be less. The most important reduction factor, ΔG_c is the difference between maximum and contour gain values. ΔG_c can vary widely in a contiguous type coverage system. It can be improved by the fine details of the antenna design (varying F/D for a given horn size and excitation) but only at the expense of overall antenna efficiency. Its value is really good only for overlapping type topology plans. An obvious way to improve ΔG_c is to give up complete contiguity of coverage by the multibeam system. The missing coverage gaps then can be covered by a single, high power CONUS beam operated in a dedicated frequency band.

Related to ΔG_c is the slope, S. For ideally pointed antennas the value of the slope is irrelevant, however in practice it is a significant contributor to total fading, thus it influences the necessary transmit power.

Satellite and earth station antenna losses and ΔG_c determines the total available effective satellite antenna contour gain.

Once the pointing accuracy of the system is determined, system fading due to pointing errors, can be determined from Table 3-6.

The G_{eff} - .05° (S) value then determines the necessary transmit power.

Finally Table 3-6 shows the carrier to interference ratio, C/I. The interference is defined as the maximum level of the sum of all main polarized sidelobe power in a given shaped beam coverage coming from the remaining cofrequency beams of the system.

3.6 ANTENNA CIRCUIT IMPLEMENTATIONS FOR VARIOUS BEAM TOPOLOGIES

The antenna circuits necessary to implement the beam topology plans presented in Table 3-2 will be discussed briefly.

The antenna beam/forming networks (BFN) for these implementations typically contain the following components:

- Radiating elements (horns).
- Orthogonal couplers.
- Fixed or variable power dividers.
- Fixed or variable phase shifters.
- Couplers.
- Filters.
- Hybrids.
- Connecting lines.

The complexity of the overall BFN is a function of two variables: a) The number of components in one BFN, associated with the formation of one shaped beam, b) The number of BFN's in the overall antenna system.

The number of components in one BFN is a function of the beam plan and the desirable beam isolation, which determines the number of necessary auxiliary horns.

Figure 3-46 shows a part of the overall BFN for Plan 1. In this plan the allocated bandwidth is divided into three subbands, called channels 1, 2 and 3. Most of the output power from 1_1 travels to horn No. 1 via a coupler C, and filter F_1, which is part of a triplexer. Horn 1 serves as a main horn for transmitter 1_1 and its beam provides the coverage regions for this transmitter. For sidelobe level control a part of the 1_1 transmit power is fed into the six horns around horn No. 1. Horn Nos. 2 and 3 are among these six horns, which serve as auxiliary horns insofar as transmitter No. 1 is concerned. The necessary feed network consists of the C coupler, followed by a six-way divider and the F_1 filter arms of the triplexers of the corresponding horns. It can be seen from Figure 3-46 that the following components are required per beam:

- One horn
- Three filters with a common junction
- Two three-way combiners
- One six-way divider
- One coupler

Table 3-7 shows the total number of BFN components necessary for the implementation of the Plan 1 antenna of Table 3-6.

TABLE 3-7. NO. OF BFN COMPONENTS FOR PLAN 1, C_0 = 16.384 GHz

Component	Qty
Horn	32
Filter	96
Three-way combiner	64
Six-way divider	32
Coupler	32

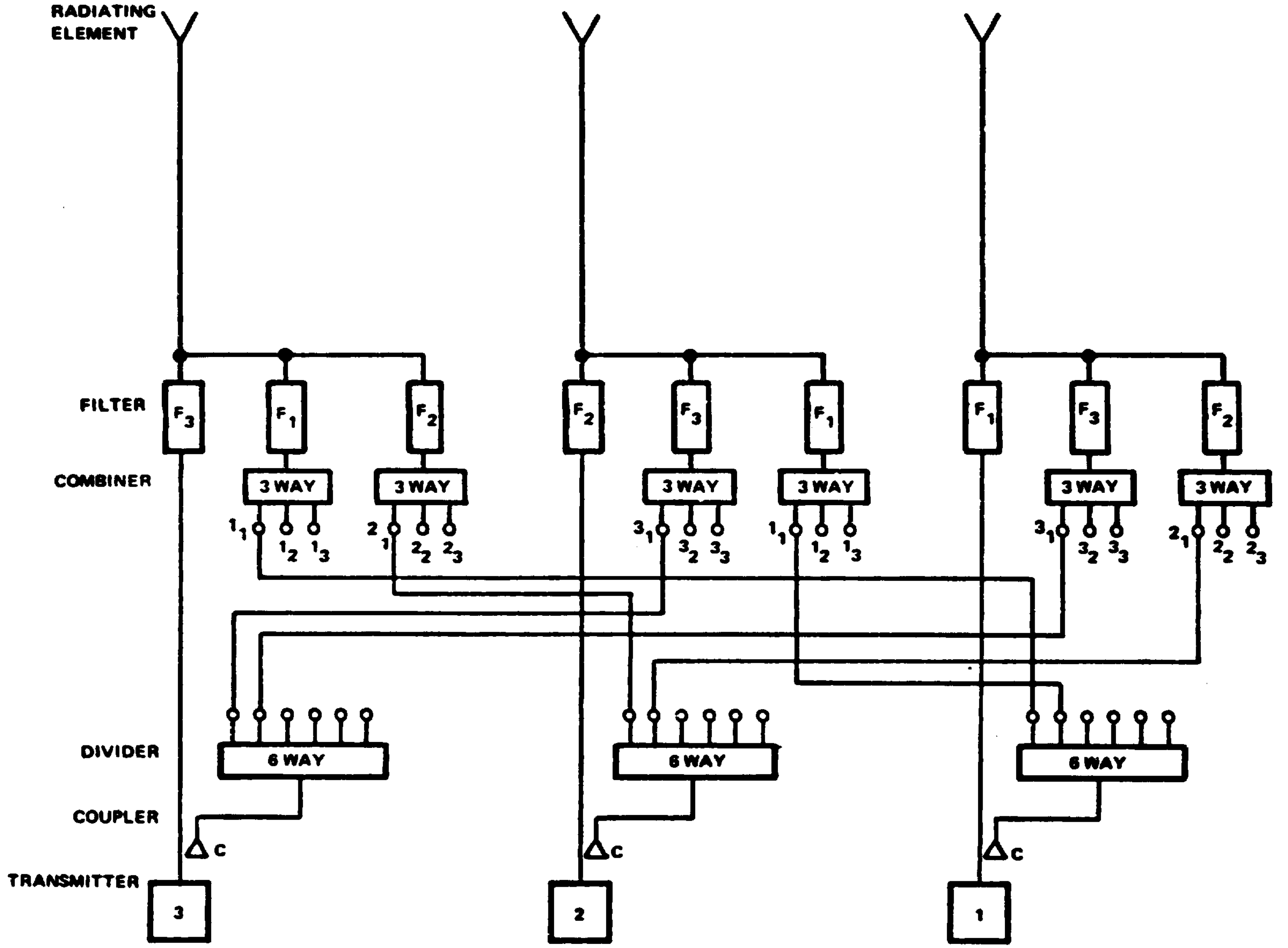


Figure 3-46. **Antenna BFN for beam plan 1.**

The coupler and the filters must be implemented in waveguide because they are in the main signal path. However, the dividers and combiners can be stripline because their power level is low.

Figure 3-47 exhibits a part of the BFN necessary for the implementation of Plan 2. This network assumes that each transmitter is feeding 9 main and 12 auxiliary horns. The main horns are fed via a coupler and a nine-way power divider. The auxiliary horns are fed via the coupler and a 12-way power divider. Generally each horn must have an orthocoupler. In this plan for each horn one orthocoupler, 2-2/3 filters and 1/2 two-way divider is necessary. In addition, for each shaped beam one coupler, one nine-way divider and one 12-way divider is needed. Table 3-8 lists the BFN components needed for the entire antenna. Notice the drastic increase in BFN complexity relative to Table 3-7. It is worth emphasizing, that this increased complexity did not improve the C/I of the antenna, however, it cut the necessary allocation bandwidth to 48.5% relative to Plan 1.

For Plan 2 the filters, couplers, the nine-way and two-way dividers must be waveguide, the 12-way divider can be stripline.

TABLE 3-8. NO. OF BFN COMPONENTS FOR PLAN 2, C_0 = 16.384 GHz

Component	Qty
Horn	178
Orthocoupler	178
Filter	474
Two-way divider	59
Coupler	44
Nine-way divider	44
12-way divider	44

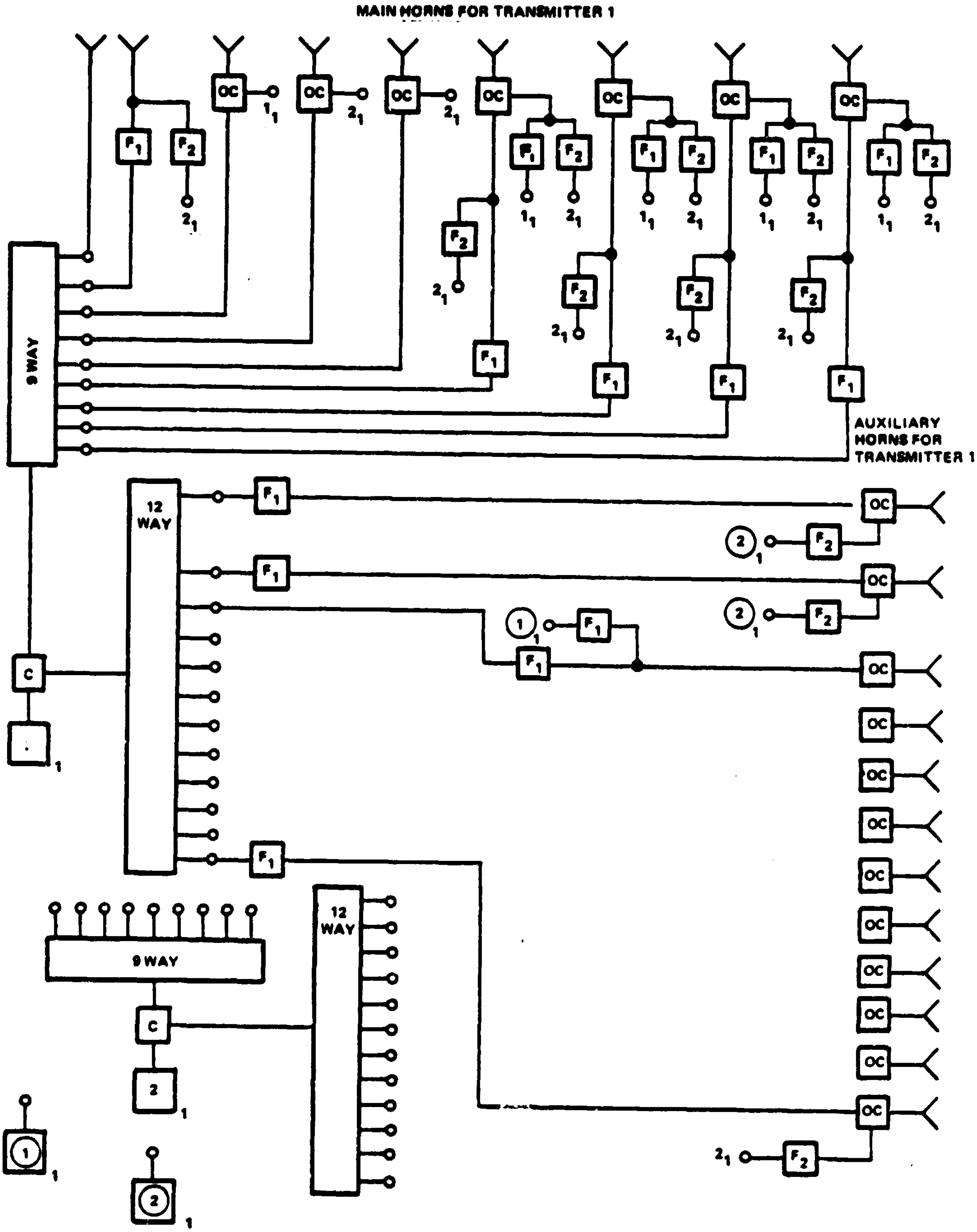


Figure 3-47. Antenna BFN for beam plan 2.

Figure 3-48 shows the block diagram of the BFN for Plan 3. The principle of this network is similar to the one shown on Figure 3-47, but the network complexity is less. Notice that this simplification increases the system sensitivity to rain, because it introduces East-West coverage strips in which the communication is dependent on polarization isolation. In this system one orthocoupler, 1-1/2 two-way divider, and two filters are needed per horn and an additional one three-way divider, one six-way divider and one coupler is necessary per beam. The total component count of the resultant BFN is shown in Table 3-9.

The BFN for Plan 4 is fairly simple, it requires only one seven-way divider per beam (see Figure 3-49).

The simplification occurs because no auxiliary horns are used and the system is singly polarized. The BFN component count is exhibited in Table 3-10.

TABLE 3-9. NO. OF BFN COMPONENTS FOR PLAN 3 (EAST-WEST LINEAR TRIPLETS), C_o = 16.384 GHz

Component	Qty
Horn	178
Orthocoupler	178
Filter	356
Two-way divider	267
Coupler	88
Three-way divider	88
Six-way divider	88
Total	1243

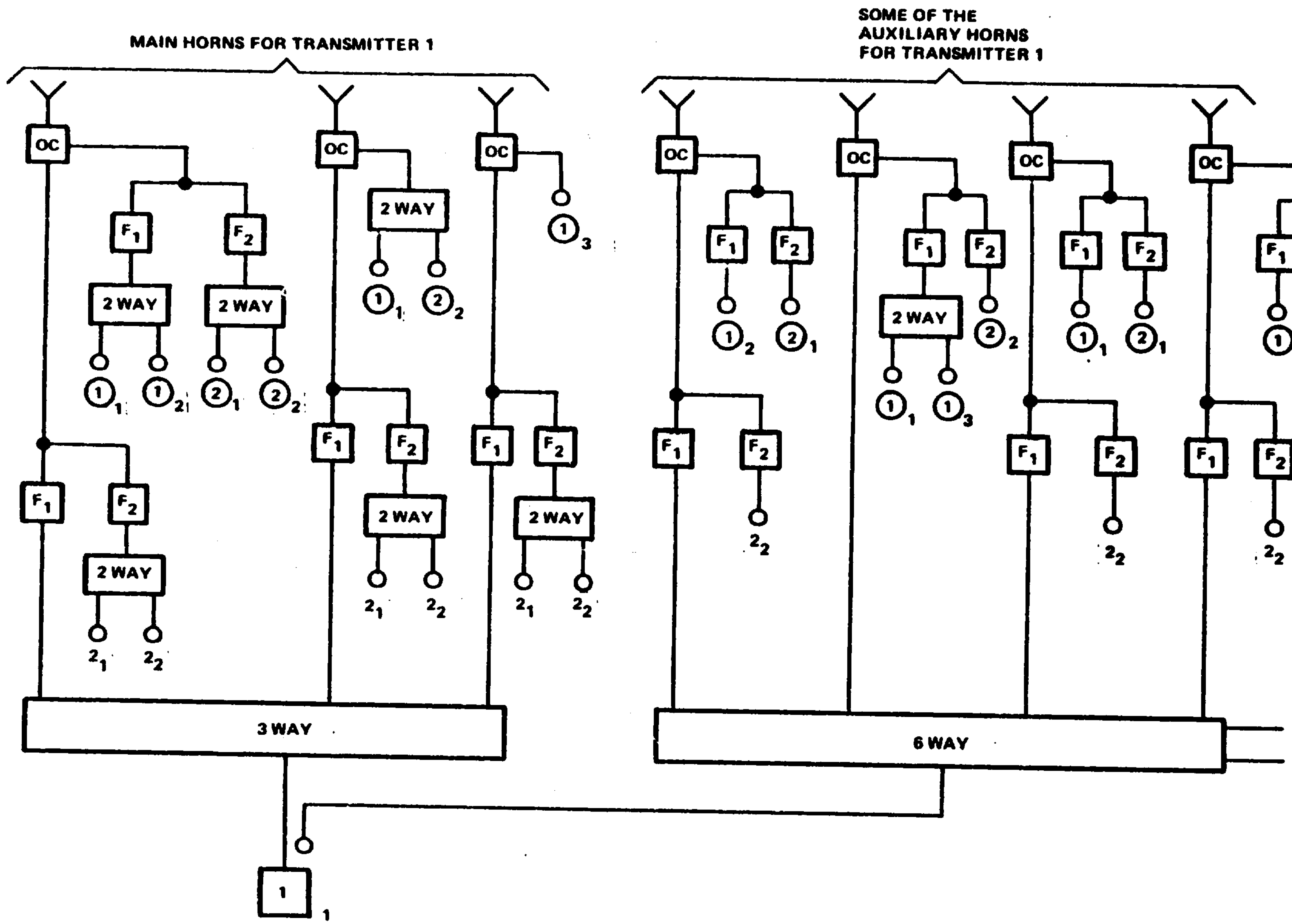


Figure 3-48. Antenna BFN for beam plan 3.

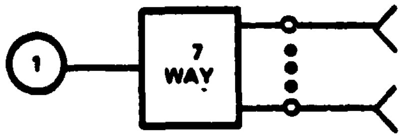


Figure 3-49. Antenna BFN for beam plan 4.

Figure 3-50a shows the BFN for Plan 5 when no auxiliary horns are used. In this case, a 1/2 orthocoupler is used per horn and four filters and 3/4 variable power dividers per beam. Since a variable power divider contains two hybrids and two phase shifters, one VPD can be taken as equivalent in complexity to four components. Table 3-11 shows the BFN component count for Plan 5.

TABLE 3-10. NO. OF BFN COMPONENTS FOR PLAN 4, C_o = 16.384 GHz

Component	Qty
Horn	178
Seven-way power divider	32
Total	210

TABLE 3-11. NO. OF BFN COMPONENTS FOR PLAN 5, C_o = 16.384 GHz

Component	Qty
Horn	68
Orthocoupler	34
Filter	256
VPD (48 Actual)	192 (Equivalent)
Total	550

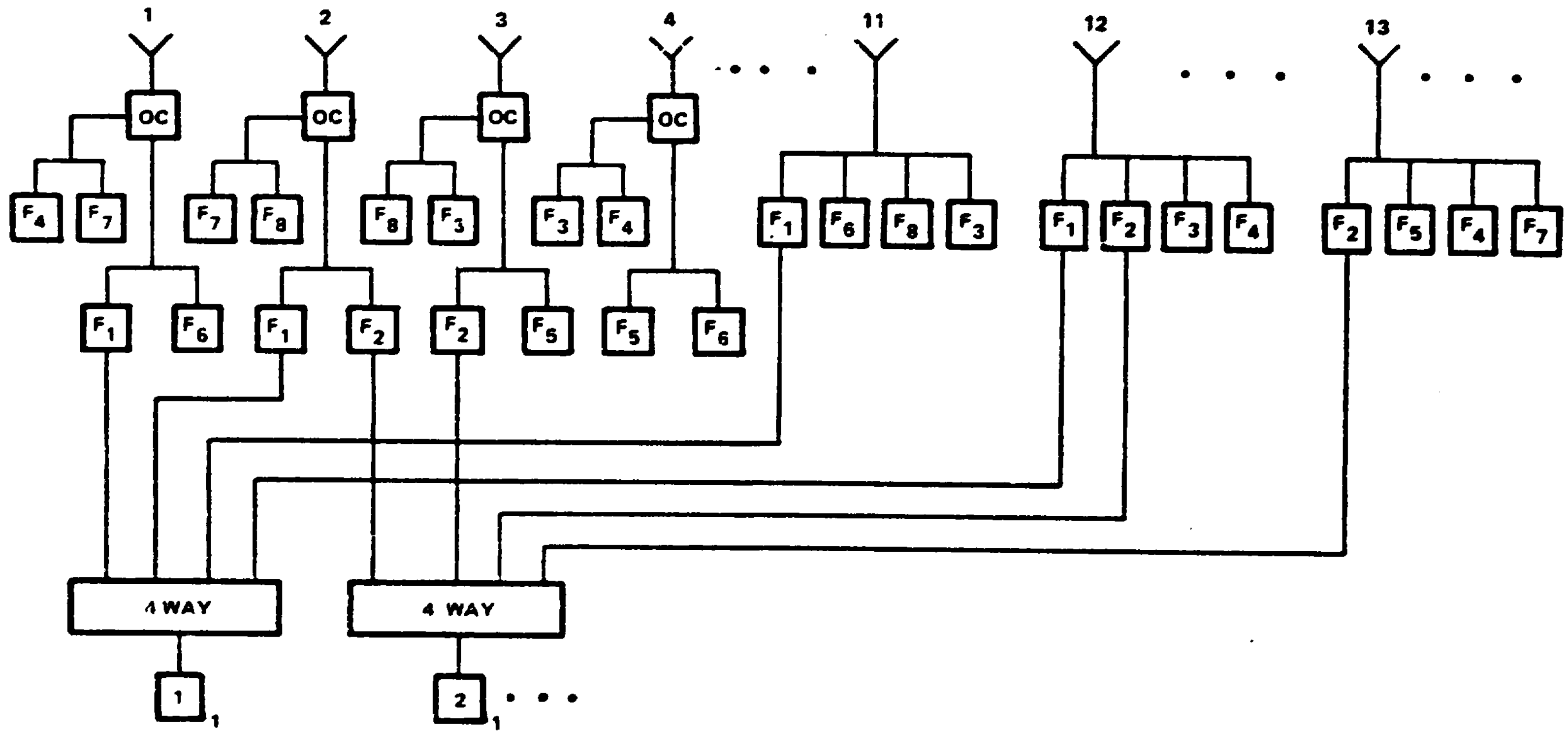


Figure 3-59a. Antenna BFN for beam plan 5 without auxiliary horns.

When the beam isolation must be better than the 23 db value achievable with the relatively simple BFN shown on Figure 3-50a, then auxiliary horns must be introduced and the component count increases rapidly. Figure 3-50b shows an example for such a BFN.

Plan 6 requires a very simple BFN, consisting of a single 90° hybrid (Figure 3-51). No auxiliary horns and orthocouplers are needed.

Table 3-12 lists the quantity of required components.

The simplest BFN of all plans is for Plan 7. In this case a transmitter is directly connected to a single horn and the total feed has only 46 components.

The BFN for Plan 8 is also very simple; it requires one three-way divider per beam. The component count is shown in Table 3-13.

TABLE 3-12. NO. OF BFN COMPONENTS FOR PLAN 6, C_o = 16.384 GHz

Component	Qty
Horn	68
Hybrids	23
Total	91

TABLE 3-13. NO. OF BFN COMPONENTS FOR PLAN 8, C_o = 16.384 GHz

Component	Qty
Horn	178
Three-way dividers	67
Total	245

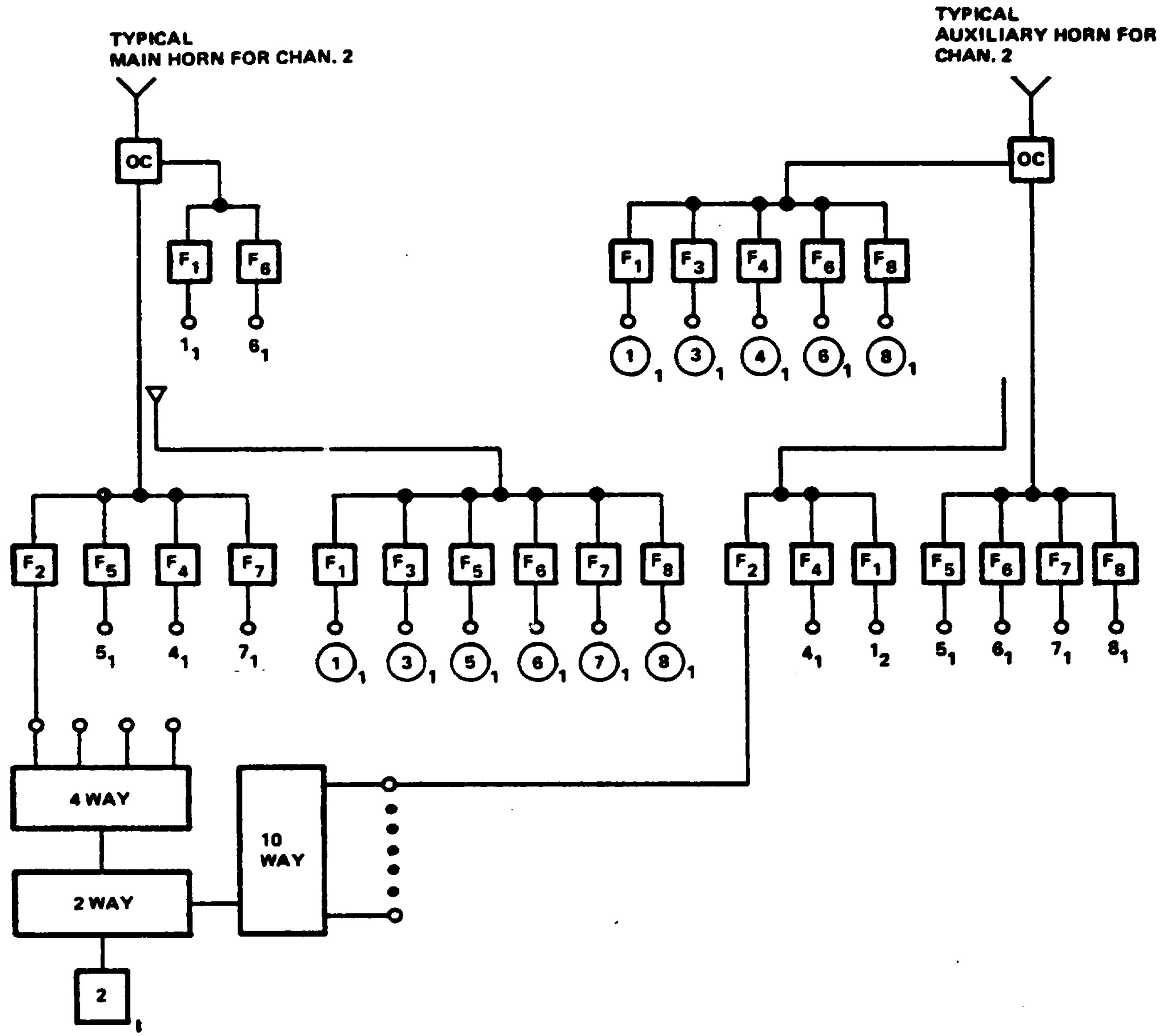


Figure 3-50b. Antenna BFN for beam plan 5 with auxiliary horns.

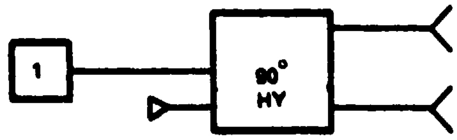


Figure 3-51. Antenna BFN for beam plan 6.

Table 3-14 gives a summary of the total BFN component count for the analyzed plans. Notice that the total equivalent bandwidth for all these cases is constant, C_o = 16.384 GHz.

Figure 3-52 shows the achieved number of spectrum reuses vs. number of BFN components for C_o = 16.383 GHz effective communication bandwidth. It can be seen that generally the component count increases with increasing spectrum utilization efficiency and Plans 6 and 8 are the best from this point of view.

TABLE 3-14. TOTAL BFN COMPONENT COUNT CHARACTERISTICS NECESSARY TO IMPLEMENT C_o = 16.384 GHz EQUIVALENT COMMUNICATION BANDWIDTH WITH VARIOUS BEAM PLANS

Plan	No. of Spectrum Reuses	Total BFN Components	Note
1	10.67	256	
2	22	1021	
3		1243	
4	10.67	210	
5	8	550	Reconfigurable
6	15	91	
7	5.11	46	
8	16.75	245	

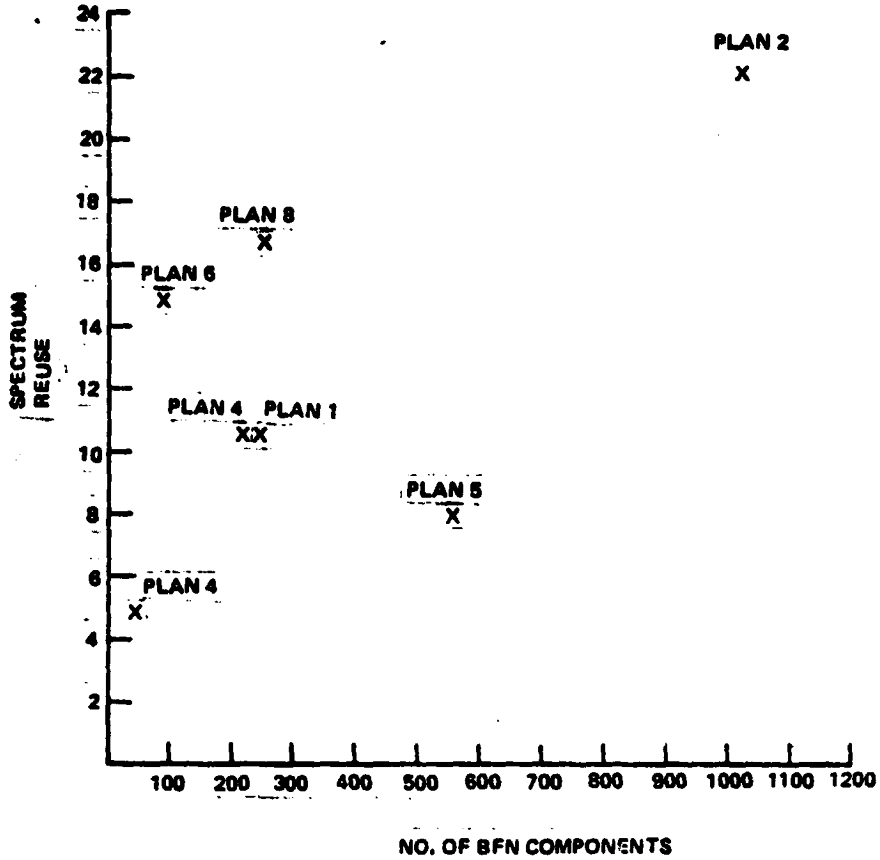


Figure 3-52. Number of spectrum reuse vs number of BFN components for C_o = 16.384 GHz effective communication bandwidth.

3.7 COMPARISON AND RATING OF THE VARIOUS SYSTEMS

Table 3-15 is an attempt to score the various system characteristics shown in Table 3-6. The score is based on a vertical column weighted system in which 1 is the best and 10 is the worst score. Additionally, characteristics falling into unacceptable ranges are marked by an asterisk. It can be seen that for the 16 GHz effective communication bandwidth Plan 5 and 1 have about the same score. However, all but Plan 5 have at least one unacceptable characteristic. It may be noted that some of the objections may be eliminated by reducing the communication system capacity or by refining the details of the antenna design.

TABLE 3-15. SCORE SHEET OF VARIOUS ANALYZED SYSTEMS (C_o = 16.384 GHz)

Plan	α	D	n	N_B	G_{eff}	S	ΔF	B	F_B	N	C/I	Total
1	5	1	1	1	2	2	4	1	4	3	10*	34
2	1	6	6	2	3	4	1	2	6	1	10*	42
4	1	6	6	1	9*	9*	4	1	4	3	6	50
5	2	3	3	3	3	1	5	5	1	4	2	32
6	2	3	3	2	6	7*	3	3	6	2	11	38
7	3	2	2	2	1	33	8*	4	6	7	4	42
8	1	6	6	3	10*	10*	2	6	8	2	8	62

Note: *Indicates unacceptable performance for indicated C_o system bandwidth.

Table 3-16 provides a summary of the best and worst features of the analyzed frequency plans.

3.8 TYPICAL OPERATIONAL FEATURES OF A POSSIBLE CONTIGUOUS BEAM SYSTEM

In the following a description of one possible system will be given. Plan 5 with α = 0.5° and N_B = 64 was chosen for the example. This plan has a reasonably balanced set of characteristics and considerable flexibility for adaptation to nonuniform traffic distribution and changing propagation conditions

Figure 3-53 shows a simplified block diagram of this configuration. The diagram shows only one typical path through the system, but on the left side of the figure the component count for the entire system is also shown at various levels of signal routing.

TABLE 3-16. SUMMARY OF BEST AND WORST FEATURES OF STUDIED BEAM PLANS

Plan	Best Features	Worst Features
1	Small reflector. Small feed cluster. Large channel bandwidth.	Poor C/I
2	Small system bandwidth. Large spectrum reuse.	High average total system power. Poor C/I. Complex BFN.
4	Small switch matrix. Large channel bandwidth.	High average PA power. Large slope. Large reflector.
5	Large bandwidth potential in component beam.	Non contiguous channels.
6	Small system bandwidth. Small average total system power. Small average P_A power.	
7	Small average total system power. Small average P_A power.	
8	Small system bandwidth. Large spectrum reuse.	Large slope. Large average power.

The component beam layout showing the serial numbering of the component beams (horns) and the channel assignments to the various horns is given in Figure 3.5d. It can be seen from this figure and from Figure 3-5e. that the first use of the vertically polarized channel 1 occurs in the Northeast of the country and it involves component beams (horns) No. 1, 2, 11 and 12. Horns 1 and 2 handle vertical and horizontal polarized channels, thus requiring orthocouplers. The vertical line output on Figure 3-53 from these orthocouplers signifies the vertically polarized outputs via uplink filters F_{U1}. Horns 11 and 12 handle only vertical polarization, thus no orthocouplers are needed in their circuits. It can be seen from Figures 3-5e. and 3-53 that each horn handles 4 channels. (Horn 1: Channels 1, 6, 4, 7; Horn 2: Channels 1, 2, 7, 8; Horn 11: Channels 1, 6, 8, 3; Horn 12: 1, 2, 3, 4). Consequently for the orthocoupler equipped

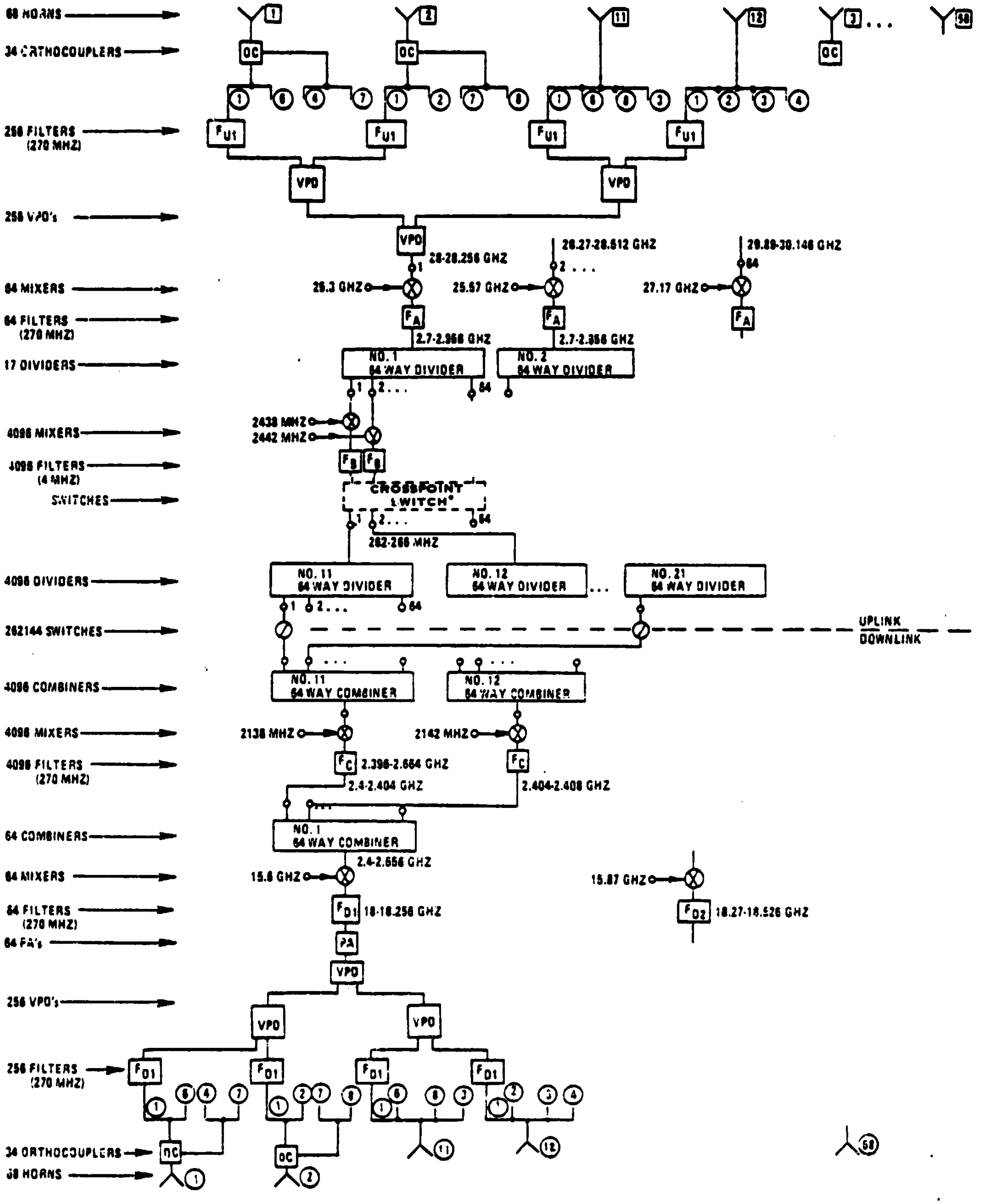


Figure 3-53. Block diagram of an uplink-downlink contiguous coverage multibeam communication system using plan 5.

horns two diplexers, and for the other horns 1 quadruplexer, must be provided. A total of 68 horns and 34 orthocouplers are associated with this part of the circuit. If these circuit elements are fully utilized then it can be seen from Figures 3-5e - 3-5s that a total of 87 beams can be generated. However, by combining some of these beams as indicated on the figures the number of beams are reduced to 64. The assumed channel bandwidth for the present example is 256 MHz (28 to 28.256 GHz), while the center to center frequency separation between channels is 270 MHz. (This requires a contiguous MUX for certain diplexers and quadruplexers, which may not be economical due to filter losses. However, the problem can be reduced by increasing the center frequency between channels resulting in a corresponding increase of the allocation bandwidth, ΔF.)

An additional feature of this system is its ability to adaptively couple the beams to favor fading earth stations. Channel 1 signals from horns 1, 2, 11 and 12 are combined by a four-way variable power combiner network. Three variable power dividers (VPD) make up such a network. During ideal weather the power combiner network is set for equal power conditions taking 25% power from each horn. Assume that a 9.6 db rain attentuation occurs in beam 1. In this case the power combiner network can be reset to take 75% of the power from horn 1 and 8.33% from each of the other horns. Under these conditions a 4.8 db fading will appear in all four beams, and the peak fading is reduced by 4.8 db for the rain influenced beam.

Figure 3-54 shows a typical four-way variable power combiner developed by GE for the 5.925 - 6.725 GHz frequency band.

The output of the VPD network produces the channel 1, signal. Since the spectrum is reused several times in this example, there are four more received vertical polarization channels in the same frequency band, producing signals 1_2, 1_3, 1_4 and 1_5 as shown in Figure 3-5e. Altogether 64 output channels are generated in eight subbands.

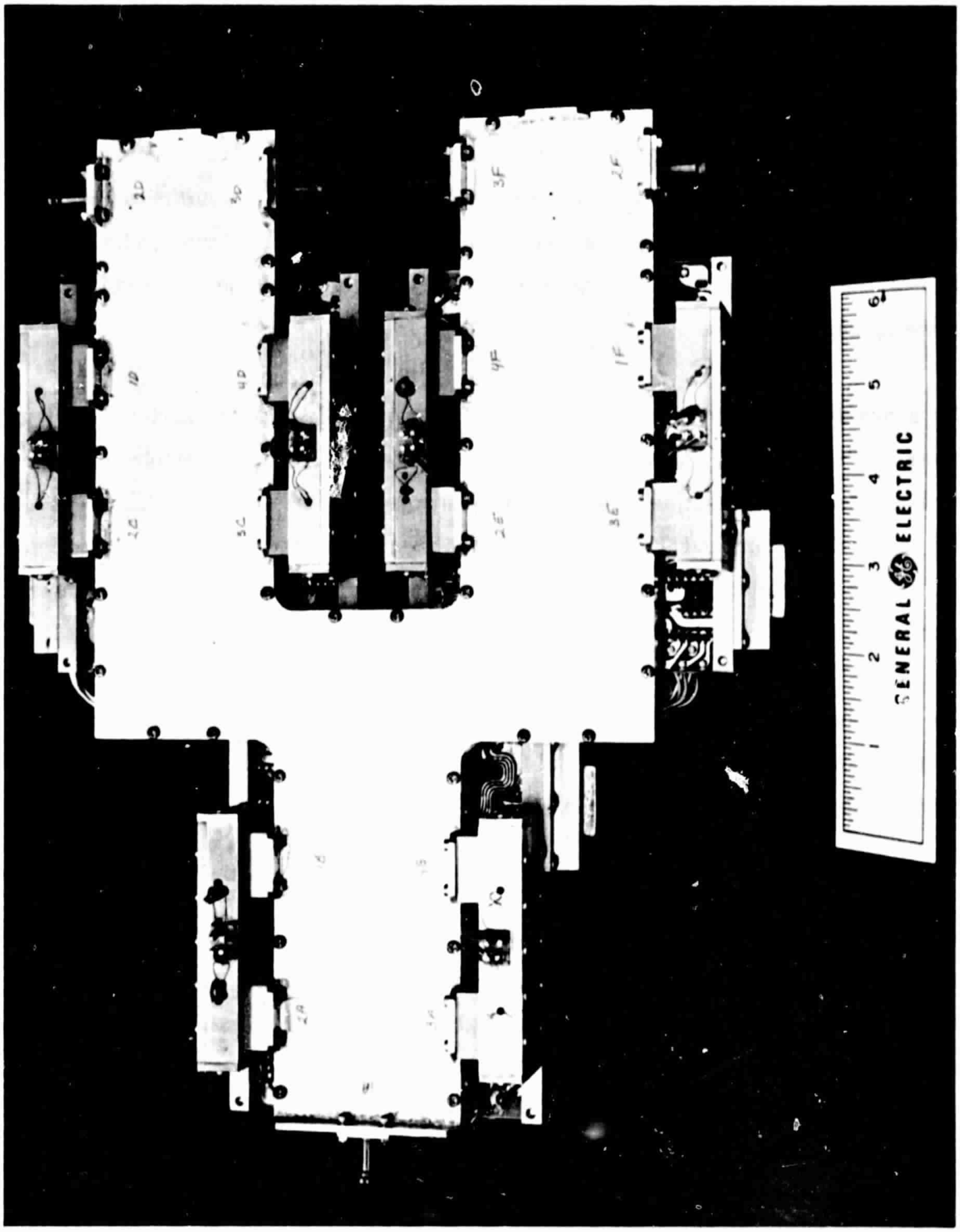


Figure 3-54. Four-way variable power divider developed by GE for 5.925 - 6.725 GHz frequency band.

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In order to provide connectivity among these channels a fairly complicated switching network is necessary, which is discussed elsewhere in this report. The block diagram shown on Figure 3-53 exhibits only the principle of one possible layout of this connectivity network.

From the cross point switch the signal reassembly takes place which is a mirror image process of the signal division. The output BFN of the antenna is similar to the input, except it is operated in the 18 GHz frequency band. The VPD settings can be linked to the input VPD settings so the fading reduction process is simultaneously implemented on the up and downlinks.

The above described system provides an available communication bandwidth distribution shown on Figure 3-10. In this system the largest bandwidths are available in component beam 49 (LA), component beam 68 (Miami) and component beam 36 (NY) where 50%, and 44% of the allocation bandwidth is available. The smallest bandwidth (3%) is available in component beam 56. Thus the maximum and minimum bandwidth ratio in the system is about 15.

If it is desirable to further increase this ratio and at the same time reduce the complexity of switching then the number of independent shaped beams can be further reduced, without reducing the size of the antenna.

Advantage can be taken of the unequal traffic distribution typical of US usage to combine the lesser used beams. This has the effect of achieving the maximum frequency reuse and maximum antenna gain for those areas having the most traffic congestion. At the same time, the geographical areas having light traffic are illuminated with "super" beams having more total traffic (and less antenna gain). The net result is reduced number of separate geographical areas requiring access - essentially reducing the on-board switch requirements. Figure 3-55 is a nominal 10 beam pattern reduced, by beam combination, to five geographic areas. Note that spot beams still illuminate the Northeastern

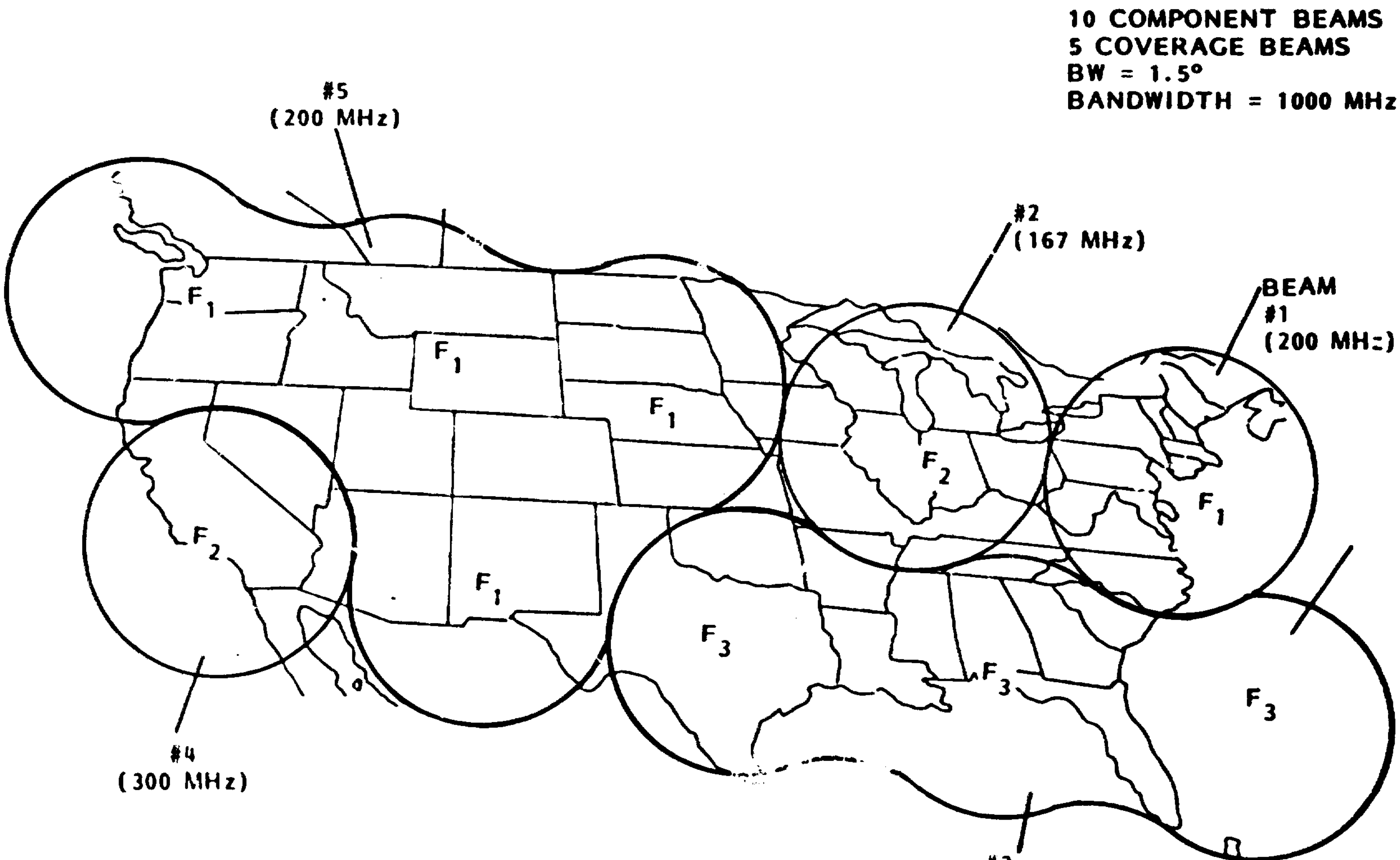


Figure 3-55. Example 10 Beam Pattern Reduced to 5 Geographical Areas by Beam Combination

seaboard, the upper mid-west and the LA-San Francisco areas whereas the South and Northwest are illuminated by combined beams. The on-board switch by reducing the geographic areas from 10 to 5 is considerably simplified - by approximately a factor of four. Figure 3-56 is a variant of the 10 beam case showing only four geographical areas for an approximate six fold reduction in on-board switch complexity. Figure 3-57 depicts similar arrangements, in this case a 68 beam system reduced to only 19 geographical regions. The shaped beams illuminate heavy traffic areas with full frequency reuse afforded by the resolution of the 68 beam antennas. Lighter traffic areas make use of combined beams.

In a practical system two or possibly three different slot bandwidths are desirable to accomodate different type of traffic.

3.9 MECHANICAL CONSIDERATIONS

The most important mechanical characteristics of the antenna system are its weight and volume. The weight of the antenna system for the downlink (20 GHz band) is determined by the weight of its two major subsystems: reflector and feed.

The reflector weight may be calculated on the basis of Figure 3-58, which shows the weight of a number of spacecraft antennas as a function of their aperture area. Assuming aluminum honeycomb with carbon fiber composite face sheet construction and 0.008 inch, rms surface accuracy the reflector weight is approximately 2.5 kg/m^2. This weight includes the reflector and its backup structure but not the weight of the reflector and feed support. For these components a 50% increase relative to the reflector weight itself is assumed.

The feed weight can be calculated from the weight of the BFN components.

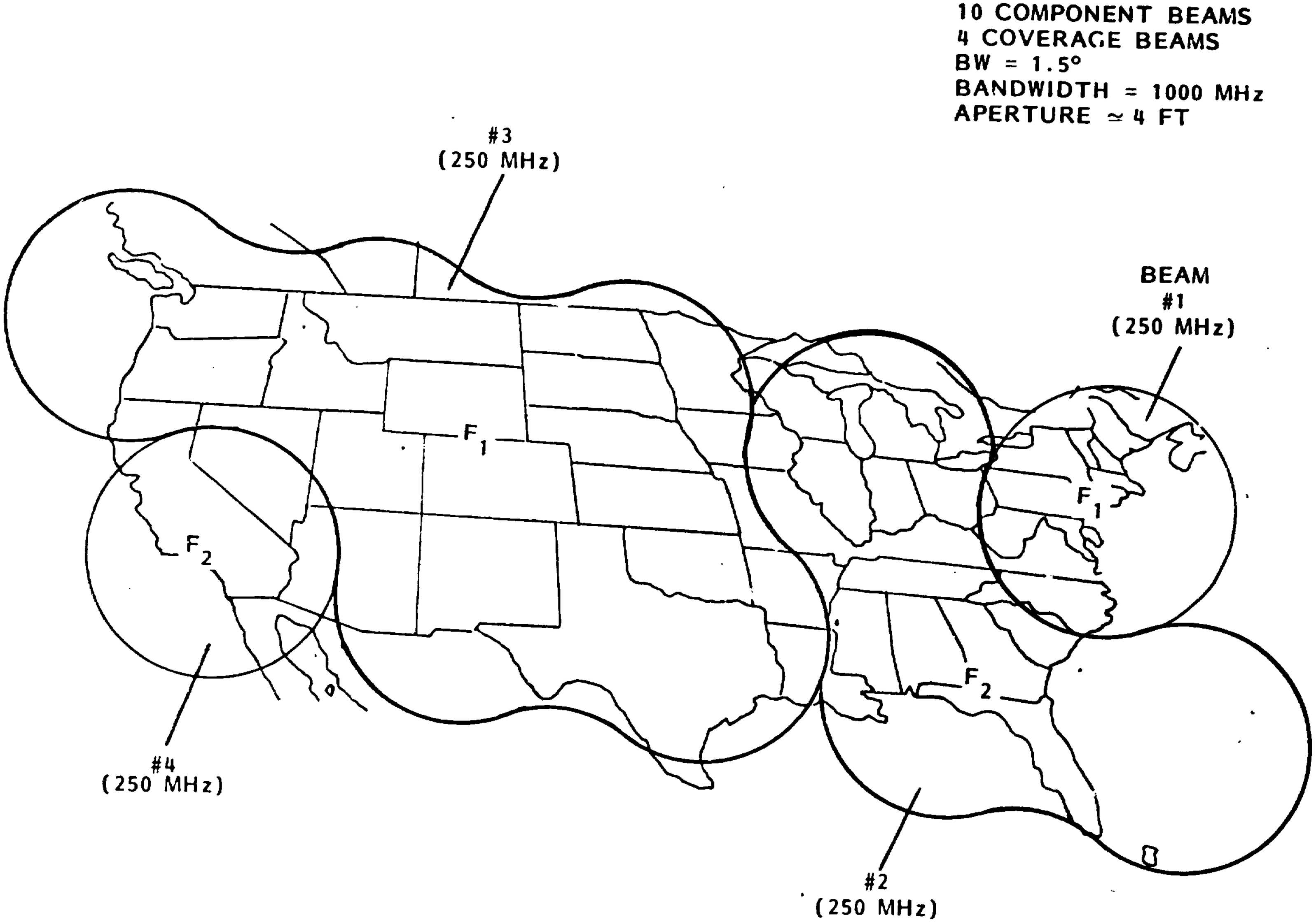


Figure 3-56. Example 10 Beam Pattern Reduced to 4 Geographical Areas

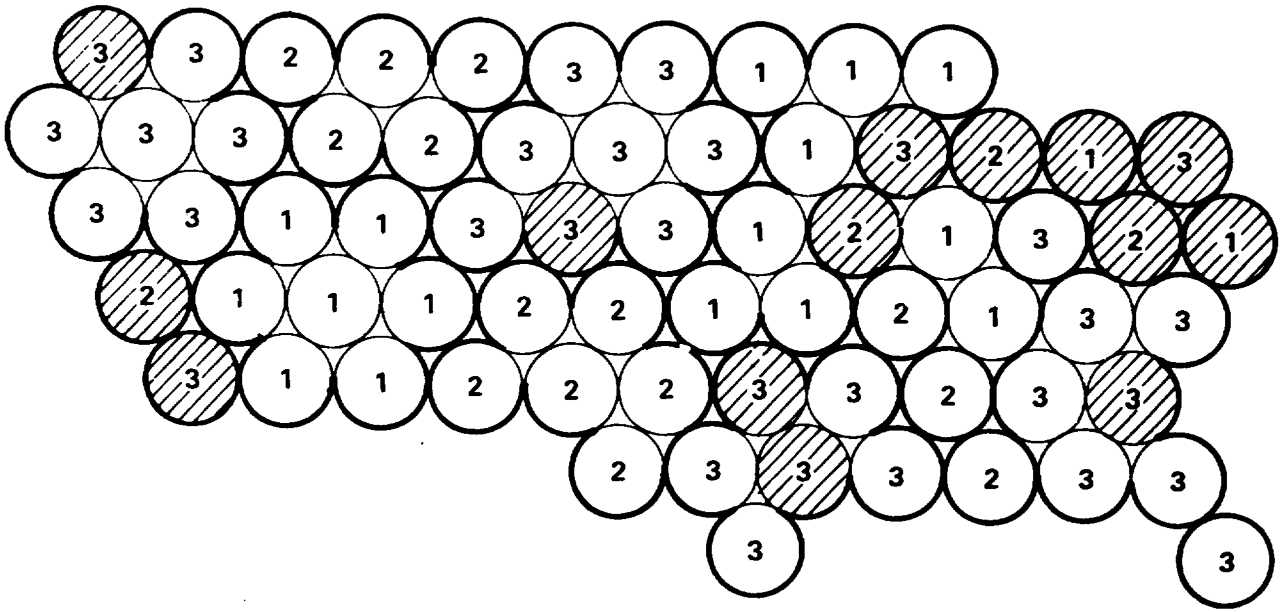


Figure 3-57. Example for Triangular, Single Polarization Beam Topology Plan; 19 Beams, 68 Component Beams.

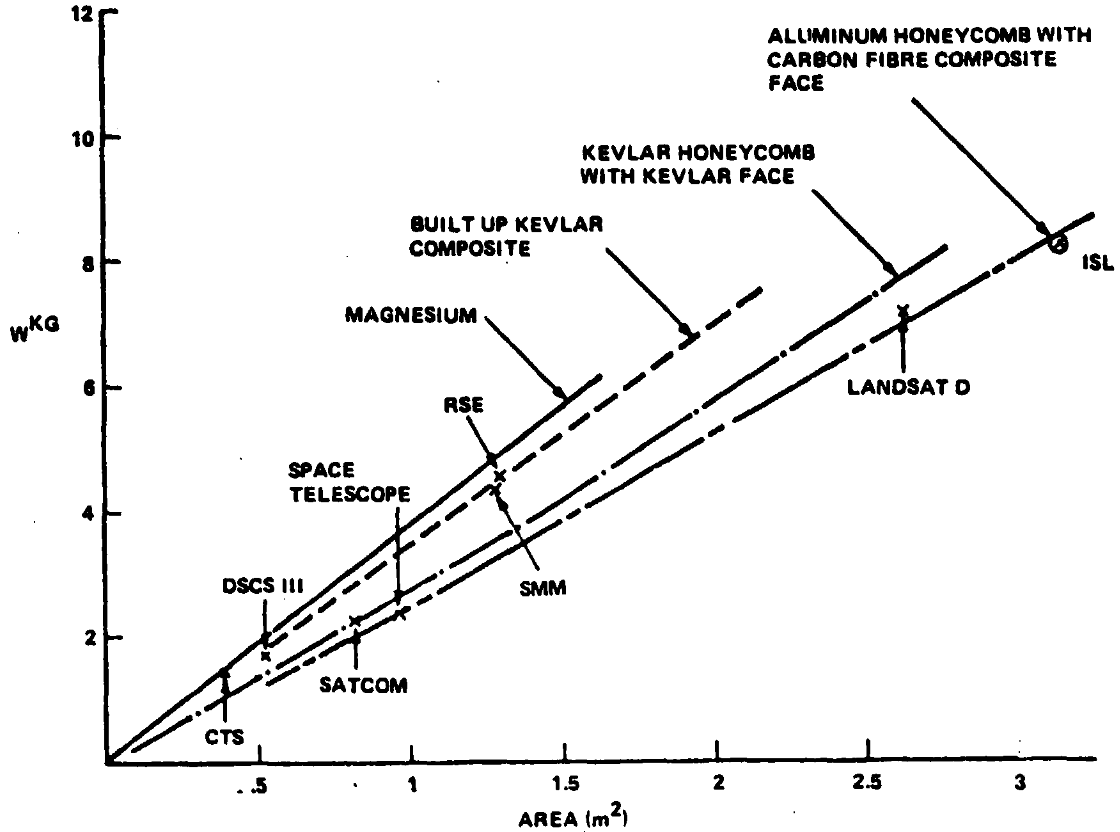


Figure 3-58. Weight of various reflector constructions.

Table 3-17 shows a breakdown of weight calculated for the 18 to 20 GHz band. It can be seen that the weight ratio of the heaviest and lightest of these antennas is about 6 even though the communication performance of the corresponding systems is normalized to unity. (All systems provide approximately 16 GHz equivalent bandwidth capability.)

Similar calculations can be conducted for the 28 to 30 GHz band. On the basis of experience with similar antenna systems the frequency scaling into the higher band will result in factor of 2 weight reduction (approximately).

TABLE 3-17. CALCULATED WEIGHT FOR 20 GHz BAND ANTENNA SYSTEMS (kg)

Plan	D^m	Reflector System: Reflector	Reflector System: Supports	BFN	Feed System Connecting Lines and Support	Total
1	1.625	5.35	2.68	4.03	2.01	14.07
2	4.063	32.41	16.20	19.07	9.85	78.16
3	4.063	32.41	16.20	20.76	10.38	79.75
4	4.063	32.41	16.20	5.42	2.71	56.74
5	2.438	11.67	5.83	5.16	2.58	25.24
6	2.438	11.67	5.83	.73	1.00	19.23
7	2.032	8.10	4.05	.23	1.00	13.38
8	4.063	32.41	16.20	5.48	2.74	56.83

On the basis of Table 3-17, the total up and downlink antenna system weight is approximately

$$20 \text{ kg} \leq w \leq 120 \text{ kg}$$

with a mean of w = 64.4 kg. Plan 5, which has a number of desirable characteristics has a total antenna system weight of 37.9 kg, considerably below the mean value of the considered frequency plans for the 16 GHz effective communication bandwidth capability.

3.10 EARTH STATION ANTENNA CONFIGURATIONS

3.10.1 INTRODUCTION

Use of small antennas does not necessarily imply wide satellite spaceings in order to control interference. At Ka Band, even small antennas (approximately 2 meters) are electrically large and are capable of good sidelobe level control. However, since the antennas are physically small consideration can be given to other antenna types such as horns, and offset fed parabolas which are ordinarily not good candidates for earth station antennas; the horn is expensive and the offset fed antenna is awkward mechanically. This section considers several methods for achieving good compromise between antenna gain and sidelobe level control.

3.10.2 MULTI-HORN FEED

An investigation of earth station sidelobe suppression techniques highlights important differences between earth station and satellite antennas. The most basic involves the need for satellite antennas to cover angular regions with the highest possible minimum gain, while ground station antennas are designed to yield high aperture efficiencies (i.e., to maximize the boresight gain).

Although no standards for sidelobe levels are presently in force for antennas in the 20 to 30 GHz band, it was decided to use the C-Band ground station antenna criterion of 32 to 25 $\log_{10}(\theta)$, but to modify it to 32 to 25 $\log_{10}(N\theta)$. Thus, if C-Band satellites can operate with orbital separations of 4 degrees or greater, with the original criterion, then the modified criterion will allow for millimeter wave satellites to operate with orbital separations down to 4/N degrees.

Sidelobe suppression for ground station antennas, is needed along the entire orbital arc. This leads to designs that have east-west symmetry and forces the use of at least four horns at 20 GHz in the east-west plane. To provide good aperture distribution in the north-south plane of the dish at both 20 and 30 GHz and east-west at 30 GHz. To insure that the horns are well above cutoff at 20 GHz, while a good aperture distribution in the east-west plane is maintained in both frequency bands, an F/D near unity is required.

For aperture diameters of 4 feet, or less, sidelobe suppression is not productive because at 20 GHz the criterion limits on the skirts of the main lobe rather than on the sidelobes. For that situation, a single horn feed (with fins to reduce the effective width in the east-west plane at 30 GHz) is a good choice. Figure 3-59 shows the performance of a well designed four-foot aperture. It can be seen that the value of N in the sidelobe criterion can exceed 1.5 for this design, and satellite orbital spacing could be as small as 2.5 degrees. The figure demonstrates that the lowest sidelobe design at 20 GHz does not meet the criterion as well as the design that balances the sidelobe level against the width of the major lobe. It is clear that for these modest apertures not much advantage can be obtained by employing sophisticated techniques, such as Horn-Reflector antennas or multiple feed horns to reduce sidelobes when the main lobes are contributing substantially to the interference levels. Under conditions where antenna noise temperature is critical, or if there is a local source of high level interference, they may become useful. If additional performance is required, a larger aperture providing main beam shaping - e.g., factor pattern roll off is needed.

At aperture diameters of about 90 wavelengths (54 inches at 20 GHz) the first pattern null occurs at 1 degree off boresight. Unless orbital separations of less than 1 degree are desired, the sidelobe level can be traded off against the aperture efficiency and antenna cost to determine the best design for antennas that are 5 feet, or more, in diameter. Horn-Reflector antennas, such as the one that first measured the cosmic background noise temperature, can be designed with sidelobes 50 db below the peak of the major lobe. They are very awkward, however, having a length in one plane of more than twice the aperture diameter. Such an antenna in a two-axis mount is expensive,

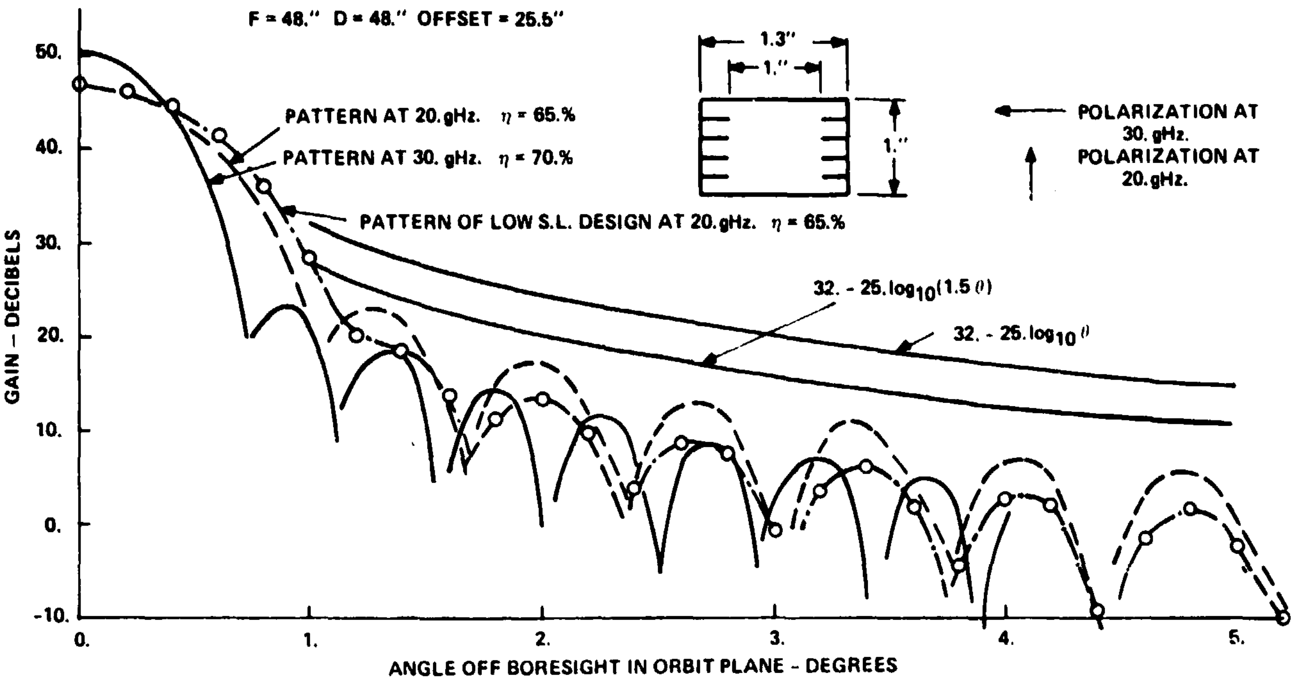


Figure 3-59. Single Horn Fed Offset Aperture

but since the satellite stationkeeping required for satellites that are about 1 degree apart in orbit would be required to be very accurate in any event, its use might be justified if a fixed mount can be used.

For apertures of five feet, or greater, it was decided to maintain the F/D value at unity, and to scale the offset, so that the same horn array can be used with the same tilt angle (29 degrees) for all aperture diameters. The horn arrangement is shown in Figure 3-60. In each case, only the two central horns are fed at 30 GHz. They are fed with equal power and in phase. All four horns are fed in phase at 20 GHz, but the relative power levels in the two outer horns is adjusted as a function of the aperture size, to balance the sidelobe level against the major lobe beamwidth, so that the criterion limits on the major lobe and the first sidelobe at the same time.

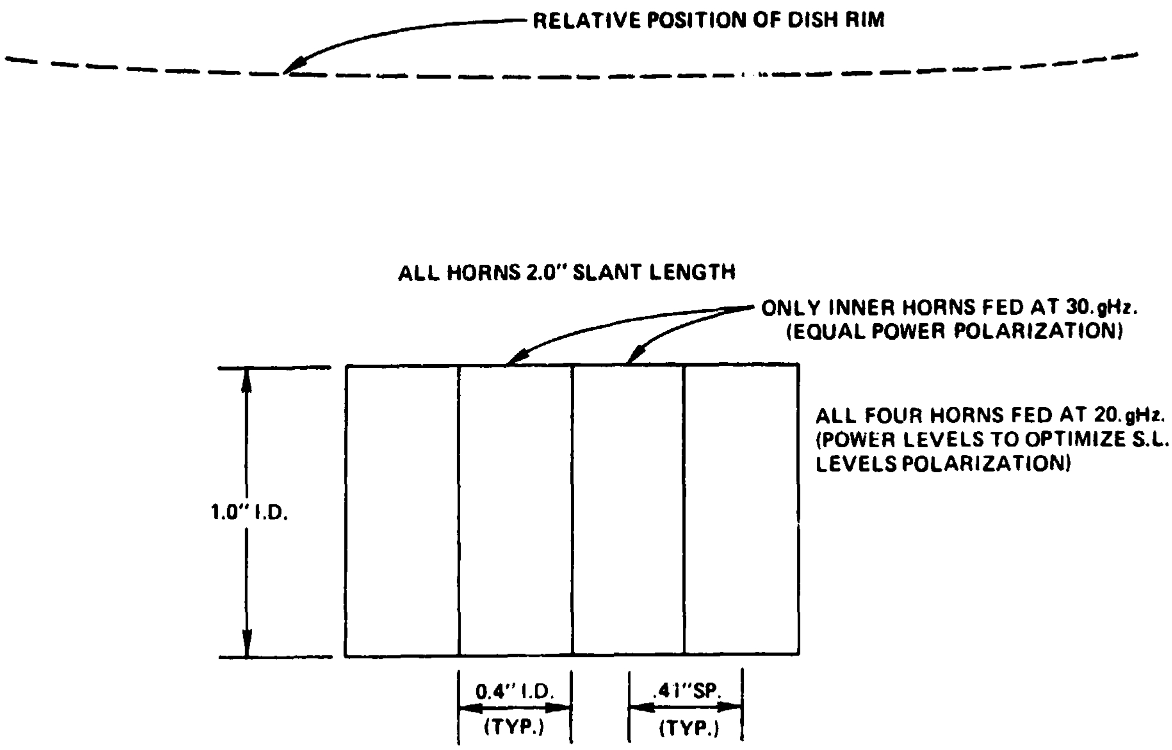


Figure 3-60. Horns For Sidelobe Suppression

Figure 3-61 shows how well this can be done for a five-foot diameter aperture. It also demonstrates that the performance at 20 GHz and at 30 GHz can be matched in both aperture efficiency and sidelobe performance. For this case, N can reach a value of 3.5, implying that the orbital separation of satellites can be less than 1.15 degrees.

Figure 3-62 shows patterns optimized with respect to the criterion for a six-foot diameter aperture. The horns are slightly wider in the east-west plane than for the five-foot aperture (0.41 inch vs. 0.40 inch), and the relative power in the outer horns is reduced from 0.3 to 0.2 at 20 GHz. At this diameter the criterion can be met with N = 4, and satellites can be located 1.0 degree apart in the orbit plane. Once again an almost perfect balance is seen between the performance at 20 GHz and at 30 GHz.

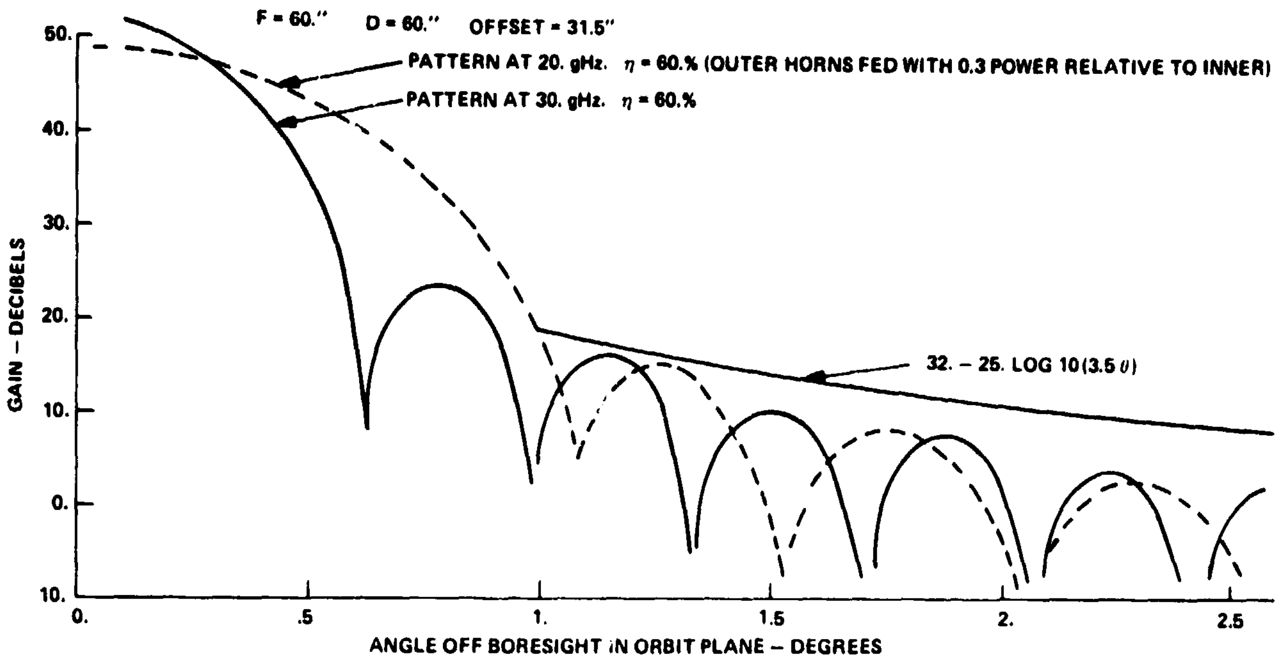


Figure 3-61. Five Foot Aperture Suppressed Sidelobe Patterns

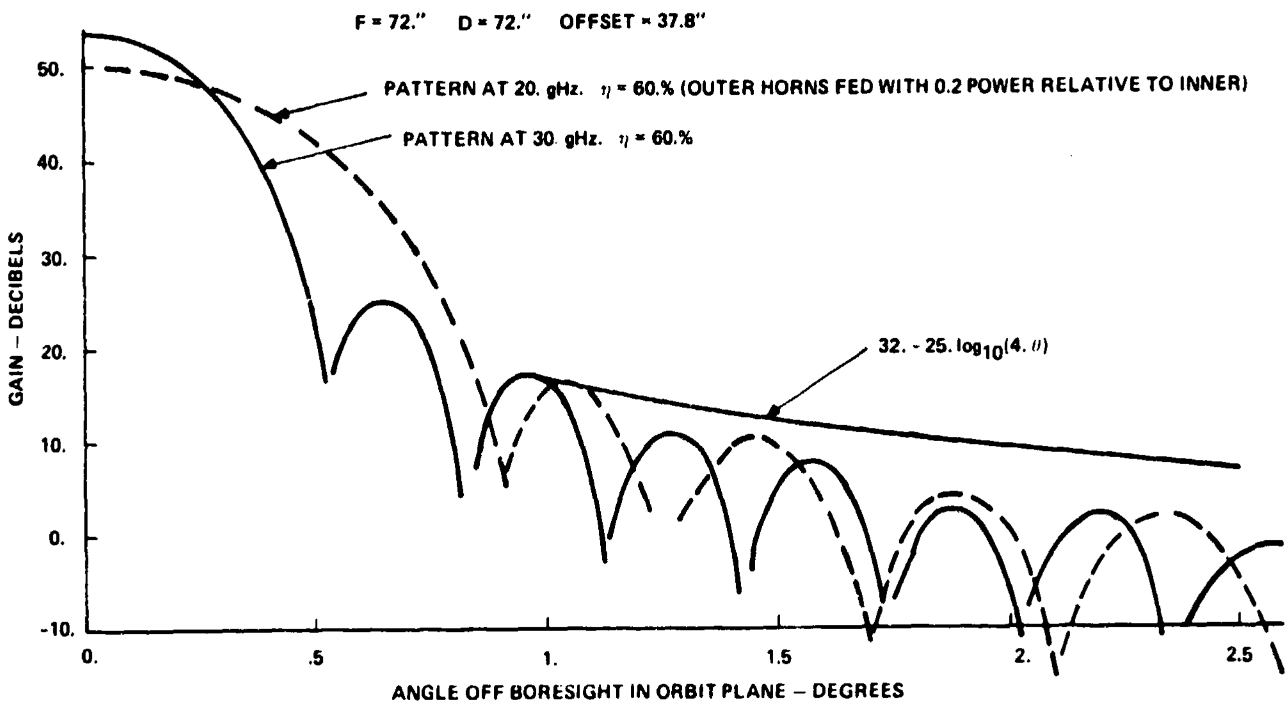


Figure 3-62. Six Foot Aperture Suppressed Sidelobe Patterns

3.10.3 OFFSET FED PARABOLIC ANTENNA

A principle advantage of this antenna type is the absence of blockage and deffraction from obstacles (such as subreflector, feed, struts, etc.), in the aperture path.

A possible disadvantage of the proposed system is that the F/D required for its implementation implies that the signals be subjected to the loss of six feet of waveguide to reach the focus, or it is necessary to place the high power transmitter and/or the low noise receiver near the focal point. Both of these approaches are practical; however, if a method can be found to bring the feed point below the aperture without sacrificing the sidelobe performance or substantially reducing the gain, it is worth evaluating in terms of cost-effectiveness against the designs described in the reference.

Blockage by the subdish of a conventional Cassegrainian antenna of this size can cause a large sidelobe to occur at about one degree from the peak of the major lobe. In 1965 "The Bell Technical Journal" showed the design of an "Open Cassegrain Antenna" for use as a millimeter wave ground station antenna. The design as shown does not quite meet the present requirements; however, a means of modifying the design is developed.

The geometry of the open cassegrain is quite complex, with many parameters to be considered. The parameters of interest are listed below:

D_m = The projected diameter of the large reflector

D_s = The diameter of the subreflector

F_m = The focal length of the large reflector

θ_t = The tilt angle from the focal point of the large reflector with respect to the vertex-focal point line that makes equal angles with the near and far rims of the large reflector

ψ_m = The angle to the near or far rim with respect to θ_t

ψ_s = The angle from the feed phase center between the θ_t line and the rim of the subreflector

F_s = The distance from the focal point along the θ_t line to the feed phase center

O_p = The offset of the focal point from the near rim of the large reflector to allow the sub-reflector to be clear of aperture blockage

P = The distance from the focal point along the θ_t line to the vertex of the hyperboloidal sub-dish

e = The eccentricity of the hyperboloid

a, b = Two additional parameters that define the hyperboloid and are derived from F_s and e.

F_e = The focal length of the equivalent paraboloid (i.e., a paraboloid with a diameter D_m that would be illuminated by the actual feed in the same way as the actual paraboloid if the sub-dish were removed and it were located at a distance F_e along the θ_t line)

Of these 13 parameters only four are independent and the others can be calculated from them. Unfortunately, simultaneous transcendental equations must be solved. A methodology has been developed, along with some desirable objectives, that allow a good geometry to be found in a relatively short time.

In conventional Cassegrain antennas (i.e., symmetrical systems with the sub-reflector blocking the aperture), the on-axis feed design can be accomplished by using the equivalent parabola and assuming that the sub-dish is transparent between the feed and the equivalent paraboloidal surface. The patterns in the region of the major lobe can also be computed by assuming the sub-dish to be opaque between the equivalent dish and the far field (the boresight must also be reversed). Recent investigations have shown that this technique is also useful for feeds that are moderately displaced from the focal point. No known body of work exists that extends these concepts to open Cassegrain antennas; however, certain observations can be made:

1. The geometric optics principles by which the equivalent parabola determines the feed design are as valid for an offset as for a symmetrical reflector system.
2. A properly designed open Cassegrain antenna does not allow blockage or feed spillover sidelobes to occur in, or near, the orbit plane for a ground station antenna, and therefore sidelobes in the orbit plane are only a function of the aperture distribution on the large reflector in that plane.

3. From 1 and 2 above, it can be deduced that, as far as sidelobe suppression is concerned, the techniques demonstrated in the referenced report can be employed in exactly the same way with the same type of results using the equivalent parabola. Wide angle sidelobes in the plane perpendicular to the orbit plane will occur due to spillover past the sub-dish, however, and the aperture efficiency will not be the same as for the focal point fed dish (the efficiency can probably be made slightly better for the open Cassegrain than for the focal point fed dish).

A rationale for the geometric design of the open Cassegrain is given below. First, the four independent parameters must be chosen. Of them, D_m will be assumed to be fixed at 72 inches (the diameter required to meet the one degree orbital spacing requirement for the focal point fed antenna described in the referenced report). Analysis reveals that D_s must be quite large to avoid a long waveguide run to the feed which involves both sidelobe level increases due to the blockage of the feed and feed supports, and losses in the waveguide run. The value chosen for D_s should be close to 40% of D_m. Next, ψ_s should be more than 10 degrees (to avoid an excessively long feed horn), and less than 20 degrees to achieve a reasonable balance between θ_t (if θ_t is too large the path loss to the far rim of the main reflector gets too large and the aperture efficiency becomes too low) and F_s (if F_s is too small the waveguide run to the feed is too long). Finally, ψ_m should be chosen greater than ψ_s (to avoid an infinite value for e), and less than twice ψ_s (to help keep F_s large and the space attenuation small. F_s can be calculated by:

$$F_s = \frac{D_s}{2}\left(\frac{1}{\tan\psi_m} + \frac{1}{\tan\psi_s}\right) \tag{1}$$

The formula for e is:

$$e = \frac{\sin\frac{1}{2}(\psi_m + \psi_s)}{\sin\frac{1}{2}(\psi_m - \psi_s)} \tag{2}$$

P is given by:

$$P = (F_s/2)(1-1/e) \tag{3}$$

The equation of the sub-dish hyperbola with its vertex centered on the Y-Z axis intersection is:

$$Z = a(\sqrt{1 + (Y/b)^2} - 1) \tag{4}$$

Where a and b are:

$$a = F_s/2e \qquad (5)$$

$$b = a\sqrt{e^2 - 1}$$

The following transcendental equation is solved for θ_t:

$$\frac{\tan[(\theta_t + \psi_m)/2]}{\tan[(\theta_t - \psi_m)/2]} = \frac{D_m + (D_s/2)\cos\theta_t + P\sin\theta_t}{(D_s/2)\cos\theta_t + P\sin\theta_t} \qquad (6)$$

O_p is obtained from:

$$O_p = (D_s/2)\cos\theta_t + P\sin\theta_t \qquad (7)$$

F_m from:

$$F_m = \frac{O_p}{2\tan[(\theta_t - \psi_m)/2]} \qquad (8)$$

And finally, the space attenuation (A) is given by:

$$A = 40\log_{10}\cos[(\theta_t + \psi_m)/2] \qquad (9)$$

Figure 3-63 shows a cross section of a nearly optimum design. The feed phase center lies in the dish surface, the required feed length is less than eight inches, and the maximum space attenuation is slightly over 3 dB.

Figure 3-64 is a view of the open Cassegrain antenna as seen from the boresight axis direction.

The feed parameters are calculated by first obtaining F_e from:

$$F_e = D_m/4\tan(\psi_s/2) \qquad (10)$$

The sidelobe suppression techniques previously described are then used to define a feed array for a center fed dish of diameter D_m and focal length F_e with no blockage.

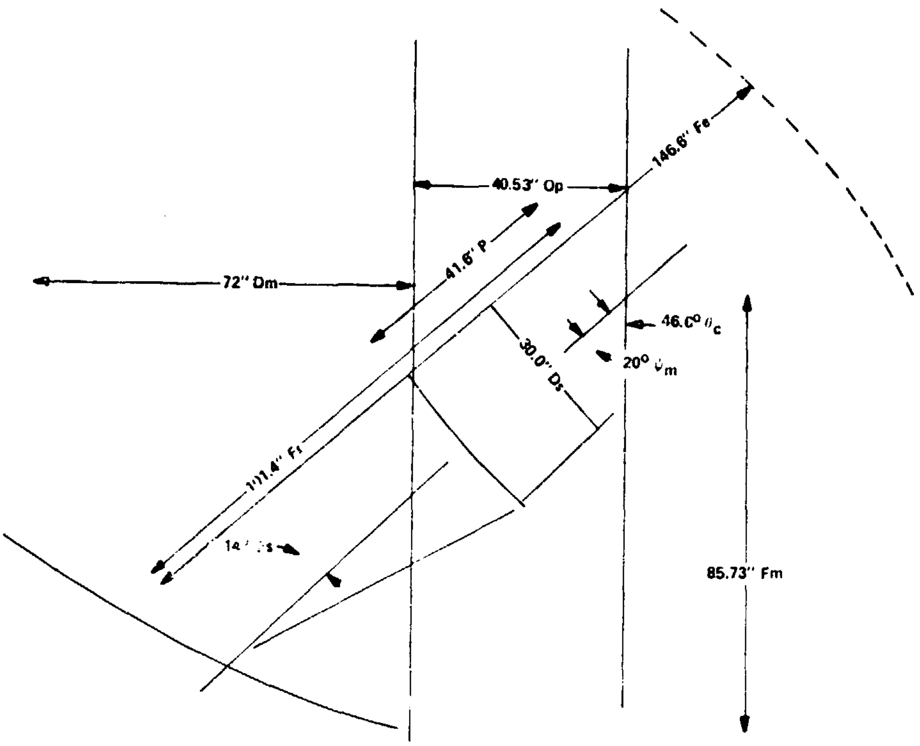


Figure 3-63. Cross-Section Of Open Cassegrain Antenna

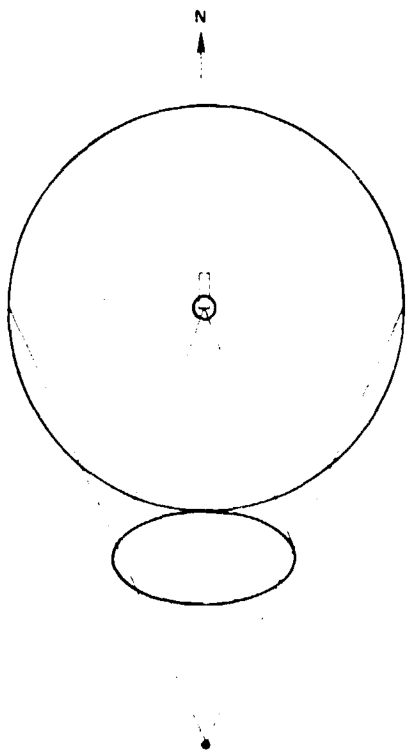


Figure 3-64. Boresight View Of Open Cassegrain

Figure 3-65 shows both the feed configuration (fed with equal phases and the same power distribution as in the referenced report) and the patterns computed by the equivalent parabola method. It can be seen that the sidelobe suppression is not quite as effective as in the referenced report. This is a result of the much larger value of the effective F/D for the present case. The effect of space attenuation in providing aperture taper in the orbit plane of the aperture is greatly reduced at large F/D values. This cannot be recovered by increased feed taper in that plane because the horn size is fixed by the sidelobe suppression method. If the computation were exact the criterion might be met for 1.1 degree orbital spacing with a six-foot dish or for 1.0 degree orbital spacing for a six and one-half foot dish.

An alternative method of computing the patterns is to employ the virtual focal point of the real aperture D_m. This involves a focal length F_m and a feed offset O_p. The feed aperture dimensions must be scaled such that the relative power level at ψ_m for the

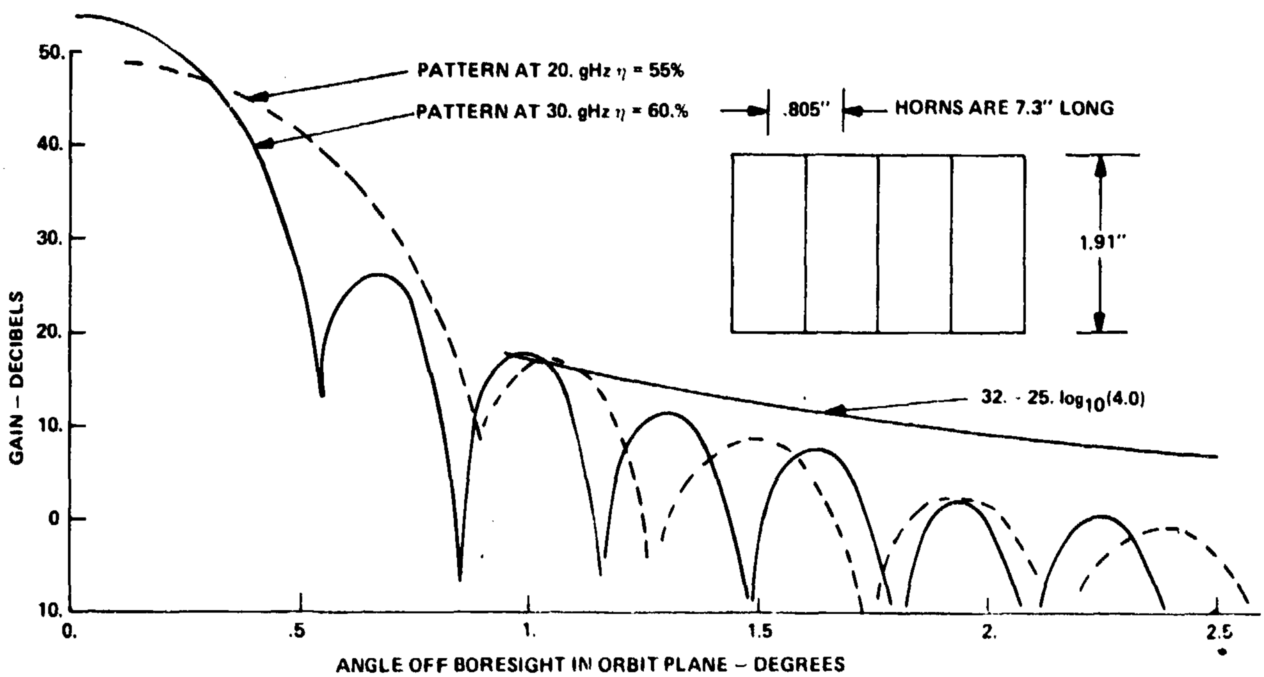


Figure 3-65. Six Foot Aperture Open Cassegrain Suppressed Sidelobes Equivalent Parabola Calculation

virtual feed equals that at ψ_s for the equivalent feed. A good approximation, if ψ_m and ψ_s are both small, is to use $\frac{\sin\psi_s}{\sin\psi_m}$ as the scale factor. For the present case, the error in the scale factor is less than 2%. The square of the scale factor is used to calculate the horn lengths. Figure 3-66 shows the scaled feed dimensions and the patterns computed by the virtual focal point method. Here the performance is very close to that in the referenced report and further horn size optimization would allow for the criterion to be met exactly for the six-foot dish at 1.0 degree orbital spacing.

Thus, no matter which computation method is more accurate, a significant sidelobe reduction for a given dish size is possible at good aperture efficiency.

Further work is necessary to increase confidence in sidelobe reduction with open Cassegrains. An engineering model of the design should be built and tested to give confidence in the low sidelobe performance.

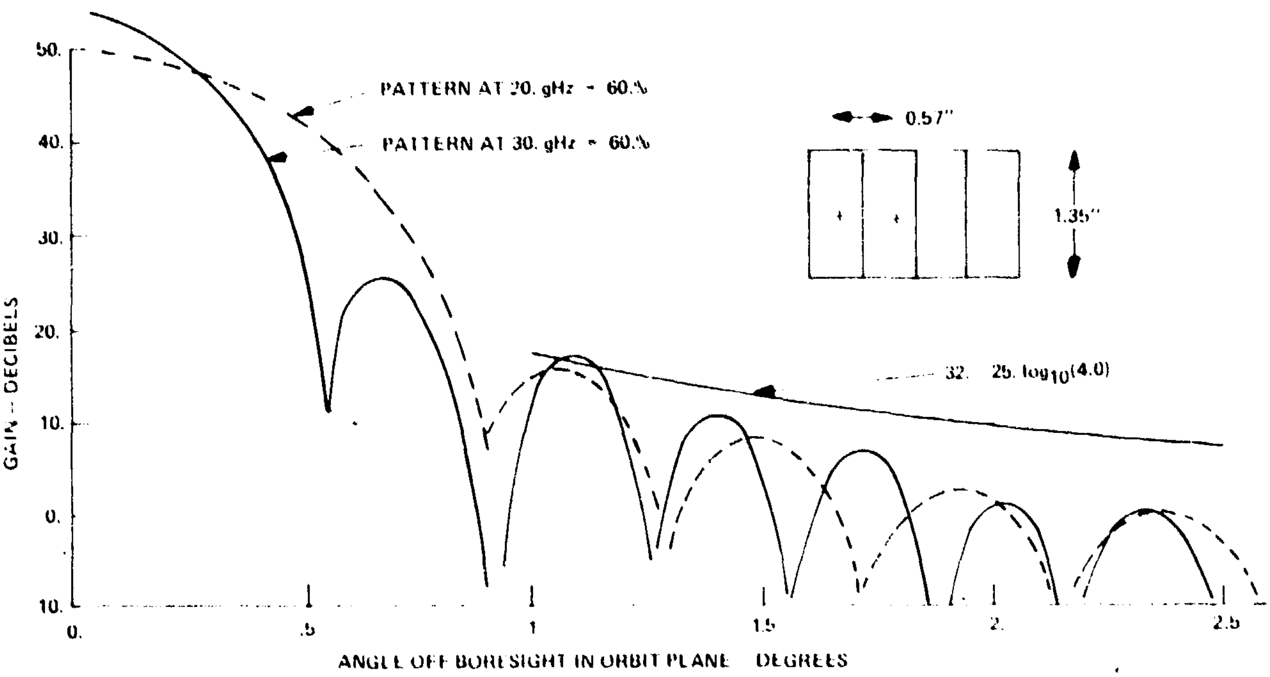


Figure 3-66. Six Foot Aperture Open Cassegrain Suppressed Sidelobes Virtual Focus Calculation

SECTION 4

SS-FDMA SWITCH AND CHANNELIZATION CONCEPTS

SECTION 4

SS-FDMA SWITCH AND CHANNELIZATION CONCEPTS

4.1 INTRODUCTION

This section describes the channelization and switching requirements for SS-FDMA and indicates attractive hardware solutions. In essence, the available bandwidth in any beam is divided into several segments each consisting of a contiguous FDMA "stack" of RF bandwidths, as depicted in Figure 4-1. Each individual RF bandwidth is identified as a "path" in an N Beam antenna having N^2 possible routes. An RF path or paths can be used to complete desired routes within the satellite. There may be one or many RF modulated carriers within each path. The purpose of the channelization and switching system is to filter each path so each can be "identified" with respect to all the others, and then to apply the channelized RF paths to a crosspoint switch or similar device so they may be switched to the proper downlink beam. If the switching and channelization is done at the proper frequencies, the weight and power impact to the satellite is minimized despite the large number of filters and switches. There appears to be a need for several different path bandwidths - depending on the nature of the traffic and the number of antenna beams. Using the frequency plan of Figure 4-1, the general arrangement of Figure 4-2 is obtained, showing several switch matrices each

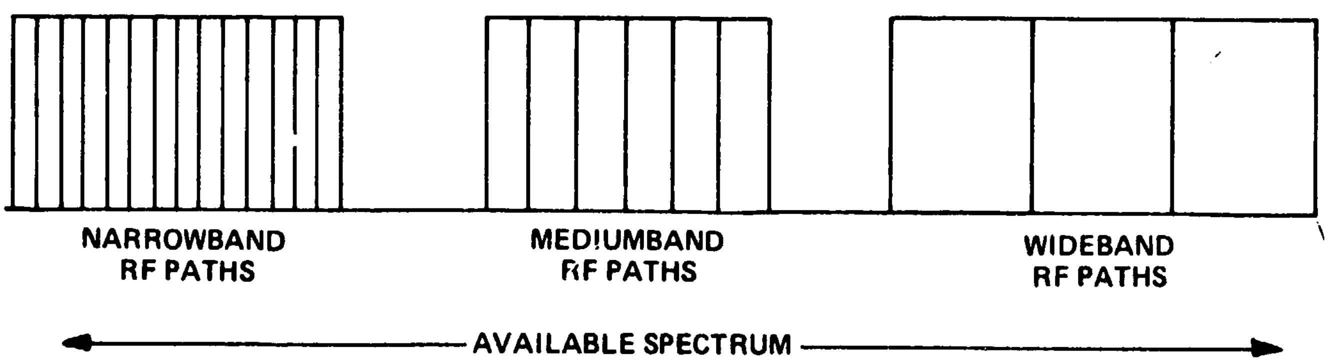


Figure 4-1. FDMA Spectral Allocation Example (Per Beam)

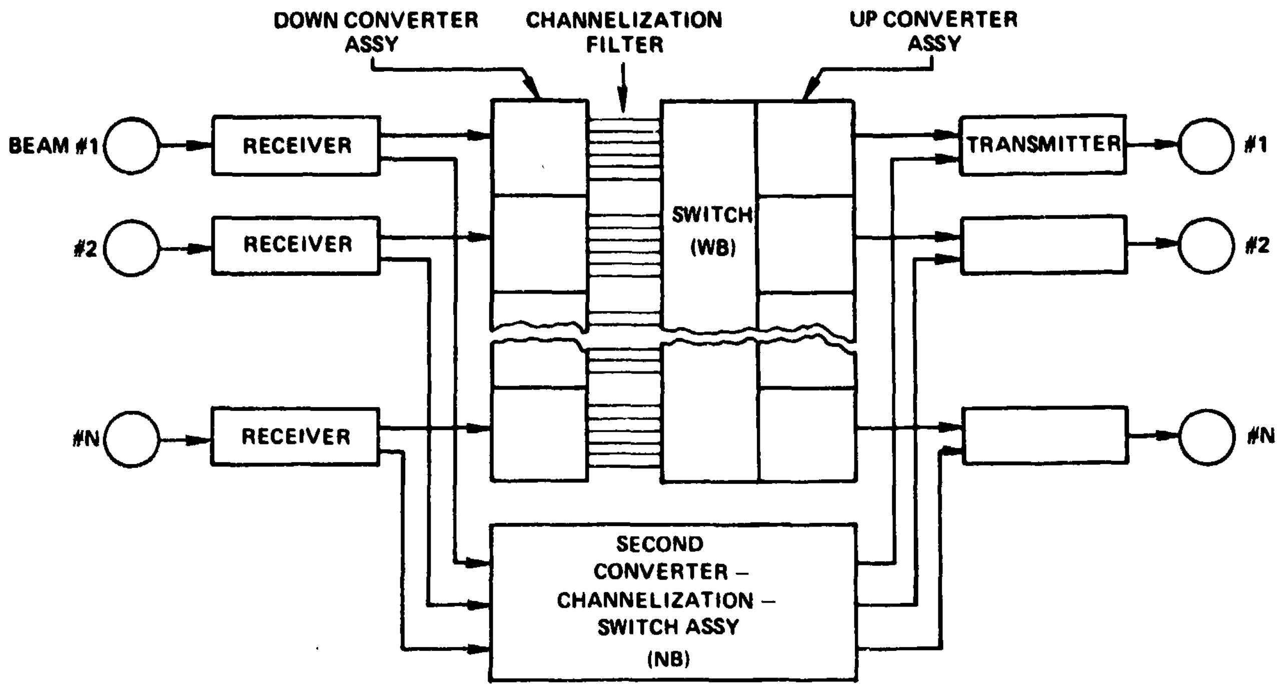


Figure 4-2. Representative Channelization Filtering and Switching Arrangement

switching different RF path bandwidths. The frequency converters convert the incoming FDMA bands to convenient frequencies (one or more conversions may be needed), each path is then filtered and the filter output serves as an input to the (crosspoint) switch which switches each path to the proper downlink - at this point up conversion and final amplification takes place. In terms of technology, the switch matrix and channelization filters have stringent performance, weight and power requirements. The frequency converters themselves also represent an engineering challenge in obtaining proper levels, frequency stability and acceptable spurious signals.

Typical RF path bandwidths are 1.0, 5.0 and 36.0 MHz. Satellite hardware is greatly simplified because of the replication of components such as mixers, filters and switches. The small numbers of basic building blocks result in economical design, development, fabrication and testing cycles and permit the use of LSI techniques.

It was recognized early in the study that the best way to identify technology requirements is to carry thru a number of trial system designs. Two system design extremes were identified for this purpose - a 100 beam system and a 10 beam system. The final system design will most likely fall between these two extremes and consist of some mix of 1.0, 5.0, and 36.0 MHz RF-paths depending on the traffic handled.

4.1.1 EXAMPLE DESIGN (100 BEAMS @ 1.0 MHz PER PATH)

Figure 4-3 depicts a 100 beam design and Table 4-1 is a summary of size, weight and power for this selected baseline implementation. A bandwidth allotment of 100 MHz per beam is assumed with 100, 1 MHz RF paths per beam. Therefore, a total of 10^4 paths are handled in a 10 GHz satellite bandwidth.

Double downconversion is used with the 100 MHz beam bandwidth filters to obtain ten, 10 MHz slots each containing ten, 1.0 MHz paths. These slots are downconverted to the 25 to 34 MHz range. Each 10 MHz segment is then down converted again and band pass filtered such that each 1.0 MHz segment is nominally centered at 10 MHz. A $10^4 \times 10^4$ crosspoint switch, controlled via the common signalling channel, establishes the desired paths. The up conversion method is essentially the mirror image of the downconversion - with common local oscillators. Note the appreciable number of repeated channel building blocks; many of which may be used in the up-converter section as well.

The following performance assumptions are made:

1. The overall system noise figure is determined by the satellite front end (antenna, receiver, etc.) The subsystem shown may be implemented to give only a nominal degradation in noise temperature.
2. The channel hardware (filters, amplifiers, etc.) can be designed to meet the overall specifications on signal distortion parameters (channel-to-channel crosstalk, AM/PM conversion, amplitude flatness, etc.)
3. The configuration reliability is acceptable.

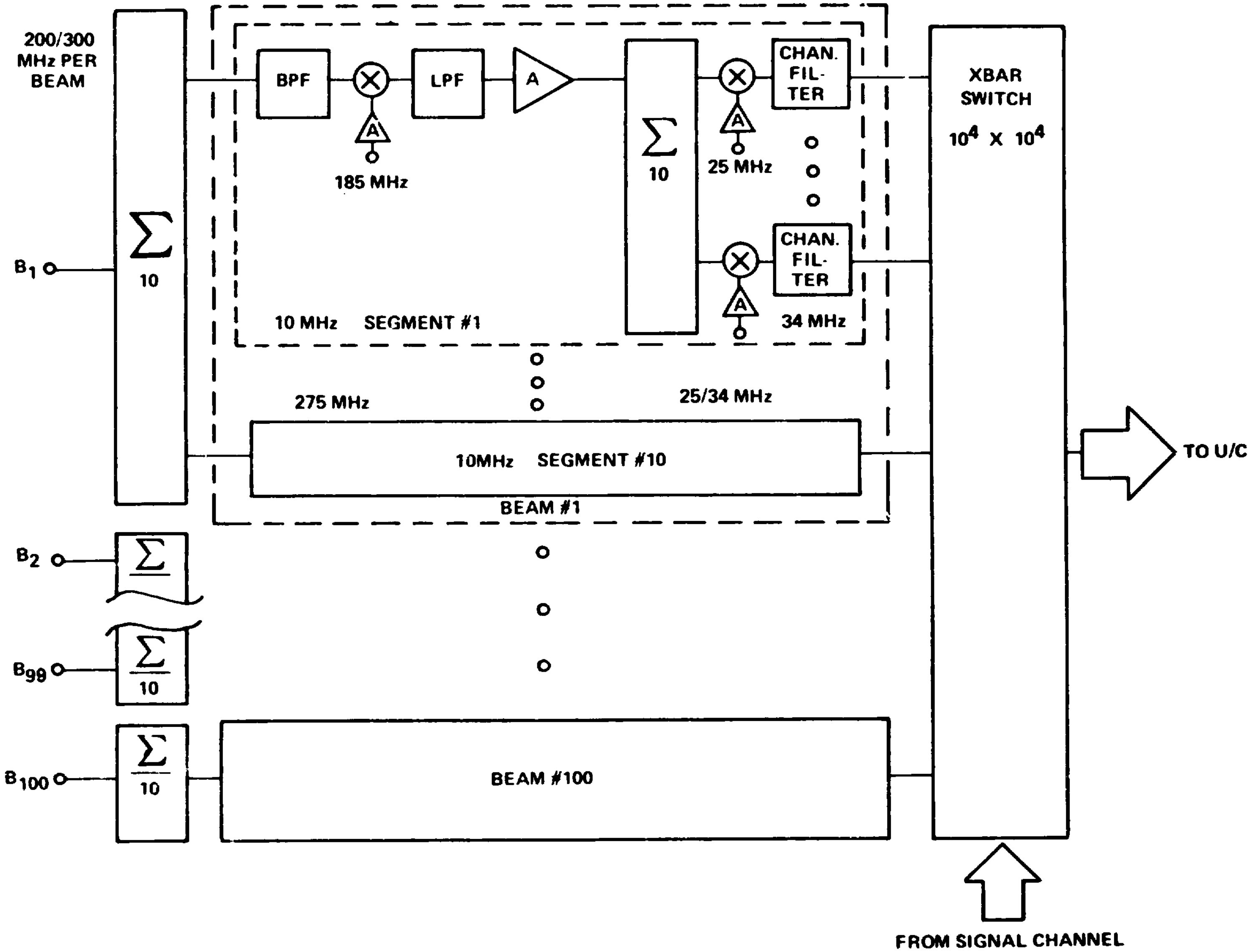


Figure 4-3. SS-FDMA System 100 Beams, 1 MHz Per Path

TABLE 4-1. HARDWARE SUMMARY 100 BEAM, 1 MHZ PER PATH

	Volume Per Element (Min)	Power Consumption Per Element (Min)	Implementation		Qty. Per Beam	Vol. Per Beam	Power Per Beam	Sat. Total 100 Beams	
			Baseline	Alternatives				(in^3)	Watts
Input Power Divider ÷ 10	1 x 1 x 0.1 = 0.1 in^3		Discrete		1	0.1		10/20	
BPF, 4-P CHEB .28 dB, BW=10 MHz	0.5 x 0.5 x .1 = .025 in^3		Saw	Combline Mic	20	0.5		50/100	
DBM + AMP	0.5 x 0.5 x 0.5 = 0.125 in^3	0.1 W	Mic		20	2.5	2.0	250/300	200
LPF 3-P CHEB BW=30 MHz, 0.28B	1 x 1 x 0.125 = 0.125 in^3		Discrete L/C		10	1.25		125/200	
RF AMP G=20 dB BW, 10-30 MHz	0.5 x 0.5 x 1 = 0.25 in^3	0.05 W	IC		20	5.0	1.0	500/600	100
Channel Power Divider ÷ 10	1 x 1 x 0.1 = 0.1 in^3		IC		10	1.0		100/200	
Channel DBM +	0.1 x 0.1 x 0.5 = 0.005 in^3	.005 W	LSI		200	1.0	1.0	100/200	100
Channel BPF 6-P CHEB, .28 dB, BW .73 MHz, f_o = 10 MHz	0.25 x 0.25 x 1.5 = 0.0937 0.1 in^3		Ceramic	Active Multimode Monolithic Xtal	100	10.		1000/1500	
Xbar Switch 100 x 100 = 10^4 Switches	0.3 x 0.3 x .1 = 0.01 in^3	3×10^{-2} Watts	CMOS-SOS		1	0.01	0.03	100/500	300
Freq. Synth. (185/275 MHz) Sources	0.2 in^3	0.5	LSI		20	4.0	10.0	400/500	1000
" (25/34 MHz)	0.02 in^3	0.1	LSI		200	4.0	20	400/500	2000

The following characteristics are evident from an inspection of the hardware summary of Table 4-1.

1. A $10^4 \times 10^4$ crosspoint switch is required. This requirement probably represents the most significant technology challenge.
2. The channel filter bank requires at least 10^4 individual high performance bandpass filters. This also represents a considerable technology challenge.
3. The frequency synthesizer apparently requires a significant amount of power. Methods for the generation of the 2200 local oscillators at reduced power levels can greatly facilitate system implementation. The use of low level mixing is another possible solution.

The modest power consumption of the crosspoint switch follows from the fact that a CMOS-SOS (complementary metal oxide semiconductor - silicon on saphire) implementation appears feasible.

4.1.2 EXAMPLE DESIGN (10 BEAMS @ 5.0 MHZ PER PATH)

Figure 4-4 depicts a trial ten beam design and Table 4-2 gives a summary of size, weight and power for this baseline implementation. A 100 MHz/beam allotment is also used with 20 MHz and 5 MHz paths per beam. A total of 200 paths is handled within a 1.0 GHz satellite bandwidth.

A single downconversion and up conversion is feasible using SAW band pass filters and higher frequency local oscillators. Each path in a beam is on 5.0 MHz centers ranging from 205 MHz to 300 MHz. The corresponding local oscillators range from 212.5 to 307.5 MHz. Each of the 200 paths is converted to a common I. F. of 7.5 ± 2.5 MHz and is routed to the crosspoint switch via separate lines. Up conversion again is essentially the mirror image of the down conversion process.

A comparison of the 10 and 100 beam systems shows the former to be less complex, more compact and to require less power. This is to be expected from the reduction in total bandwidth and in the number of channels. The crosspoint switch and channel

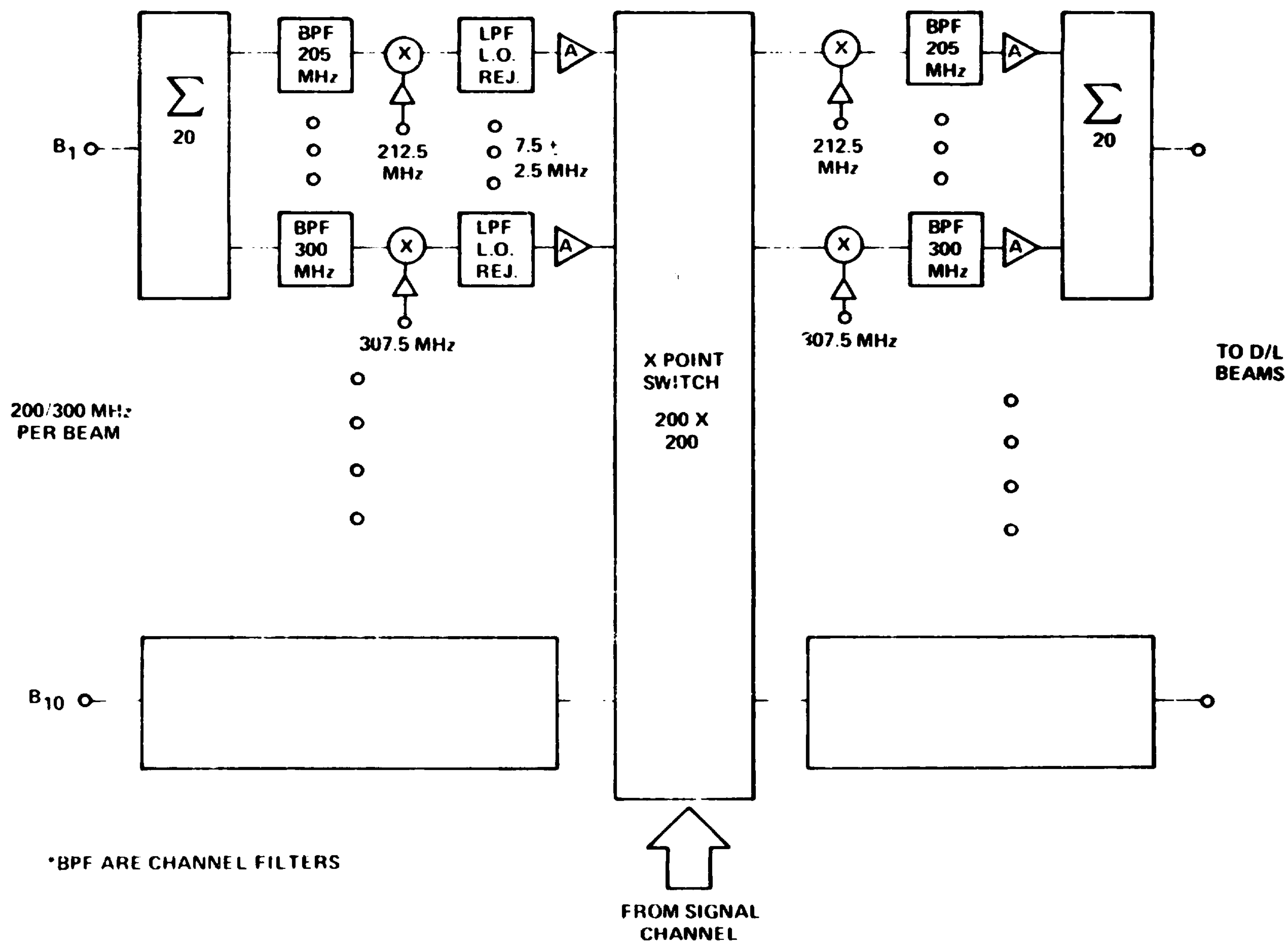


Figure 4-4. SS-FDMA System 10 Beams, 5 MHz Per Path

TABLE 4-2. HARDWARE SUMMARY 10 BEAMS, 5 MHZ PER PATH

	Size @	Power @ Consumption	Baseline Implementation	Qty	Total Size (in^3)	Total Power
Input Power Divider ÷20	1 x 1 x 0.2 = 0.2 in^3		Discrete	20	4/8	
Channel BPF 6-P CHEB	1 x 1 x 0.1 = 0.1 in^3		Saw	400	40/80	
Channel DBM · AMP	0.5 x 0.5 x 0.5 = 0.125 in^3	1.0 W	LSI	400	50/100	400/500
RF AMP	0.5 x 0.5 x 1 = 0.25 in^3	0.1 W	LSI	400	100/150	40/60
LPF · L.O. REJ	0.5 x 0.5 x 1 = 0.25 in^3		Discrete L/C	200	50/100	
XBAR Switch 200 x 200 = 40K	0.3 x 0.3 x 0.1 x 16 0.16 in^3	0.12 W	CMOS-SOS	1	0.16/10	0.12/5
Frequency Synthesizer	1.0 in^3	1.0 W	Discrete	400	400/600	400/600
					650/1000	800/1100

filter requirements are more reasonable, however, means for accomplishing a power reduction in the synthesizer and double balanced mixers still deserves investigation.

Consideration also was given to configurations for handling 36 MHz paths. Figure 4-5 shows a feasible configuration. While it is functionally similar to the 5 MHz path configuration, the local oscillators, filters, and rejection filters are somewhat higher in frequency. Also the channel amplifiers are necessarily broader band. The most significant difference is that a CMOS-SOS implementation is probably no longer possible because of the higher operating frequency. Instead PIN diodes or, where a large number of channels is needed, GaAs Fet's can be used. The 36 MHz system is larger (and heavier) and consumes more power than the 5.0 MHz system for an equivalent number of channels.

4.1.3 ALTERNATE CONFIGURATIONS

Variations of the configurations already discussed also were considered. For one reason or another these were not considered promising, however, they are mentioned for completeness.

4.1.3.1 TDM/FDMA

It is possible to perform an A/D conversion on a per path basis after front end down conversion and band pass filtering. The digitized samples enter a buffer storage and are read out at a higher data rate - and time sequenced in groups of - say 10 channels per group (TDM). The TDM digitized data is routed to the crosspoint switch which is programmed to sequence the data to the proper outputs, bit by bit, where the data is stored in high speed buffers and read out at a reduced rate. A D/A conversion is then performed on a per channel basis. This scheme illustrated in Figure 4-6 is similar to modern solid state terrestrial (electronic) switches such as the ATT #4 ESS now going into service. Applied to a satellite system the TDM/FDMA

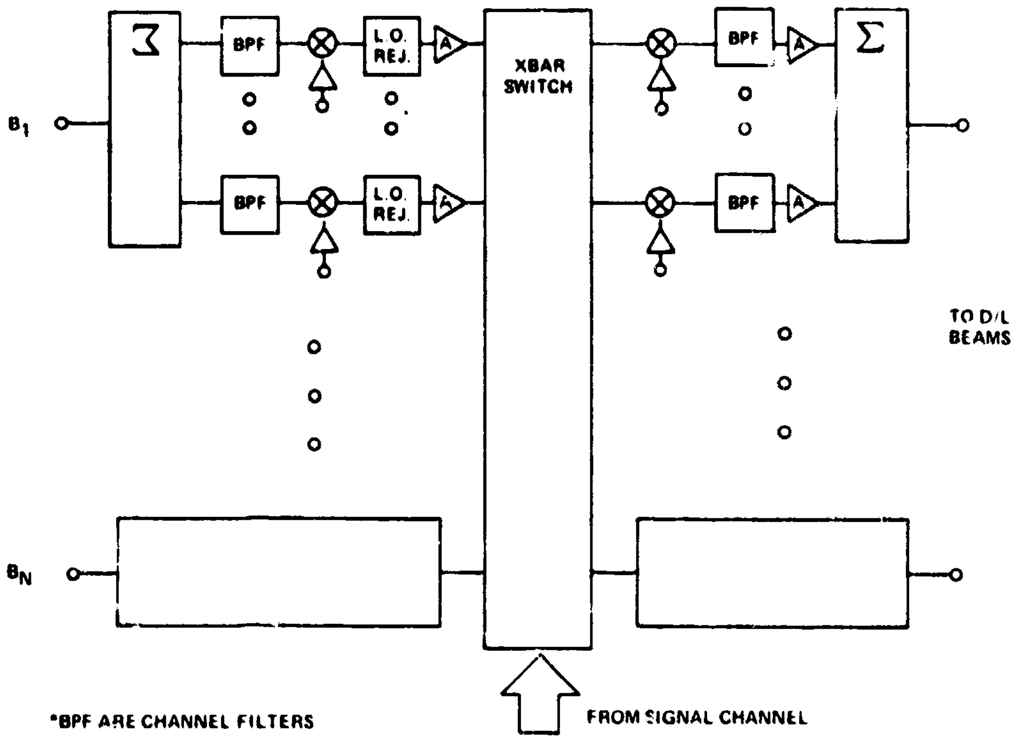


Figure 4-5. SS-FDMA System N Beams, 36 MHz Per Path

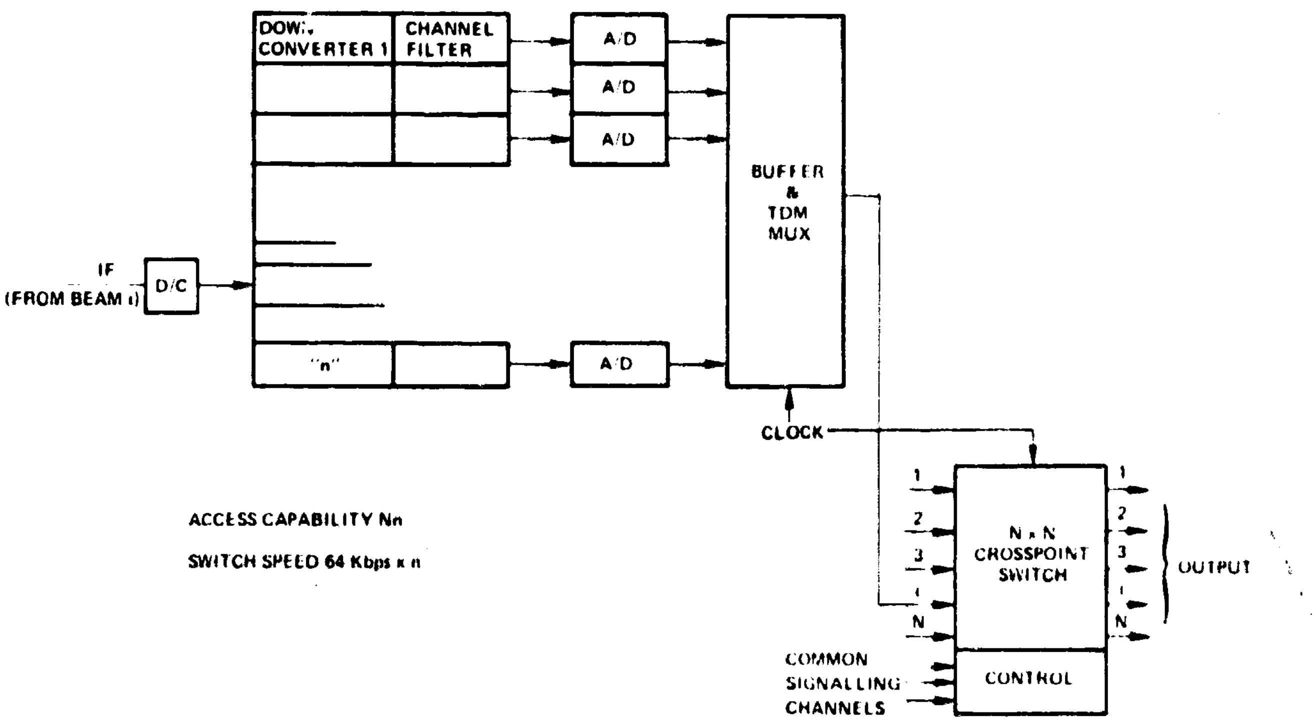


Figure 4-6. TDM Crossbar Switch Concept

arrangement reduces the number of switches required in the crosspoint switch but adds the complication of the A/D conversion and multiplexing. The satellite paths are still "transparent". Better utilization of the satellite switch is accomplished because it is switching, bit by bit (or perhaps block by block) in order to accomplish the routing instead of remaining quiescent. There are however several severe drawbacks in the arrangement. One is that the sampling rates when multiplexing become very high - ten 5 Mbps (5 MHz) paths when multiplexed together operate at 100 Mbps - precluding LSI implementation of the crosspoint switch and multiplexers. Also power consumption will be high at these speeds. In addition the TDM/FDMA arrangement now exhibits single points of catastrophic failures (points where failures result in substantial losses in traffic capability) - particularly in the multiplexer and crosspoint switches. For these reasons, the direct application of crosspoint switching is believed to be better for satellite switching, e.g., failures of a switch may reduce the number of paths providing a route or if the switch shorts it becomes a preassigned path, but other paths - so called "pooled" paths may be switched in. In general, it appears that degradation with switch failures will occur slowly. In the switching schemes envisioned in Section V, there are many ways to provide augmented capability in a given beam or path even after many switch failures so that flexible service still can be provided. Section V indicates some generic switch configurations (fully and partially implemented crosspoint switches for various bandwidth paths) in order to achieve flexibility of service. While much can be studied about switching, emphasis in the study has been placed on technology and arrangements compatible with satellite use where weight, power and reliability are primary considerations. Complex tandum stage switching configurations may be attractive to study once particular requirements are established. An important purpose of the Study is to identify practical, flexible switching schemes for satellites. Optimization for particular applications can come later.

4.2 CROSSPOINT SWITCH TECHNOLOGY

4.2.1 INTRODUCTION

Crosspoint switching as applied to direct access SS-FDMA sets up "paths" between the uplink users and downlink users. A multiplicity of independent modulated uplink carriers can exist within each path. Equal bandwidth paths in a given uplink are downconverted in the satellite to a convenient IF. Each path, after filtering, will then input the switch on a separate line. The switch is remotely commanded via a common signalling channel to allow any input line (or path) to be routed to any output line. Figure 4-7 gives a functional diagram of an M x M crossbar switch matrix· a switch closure at any line junction enables an input to be connected to an output. Equal numbers of input and output lines are chosen since this configuration appears most applicable to SS-FDMA operation. Furthermore, all inputs may be routed to all outputs simultaneously, but no input is routed to more than one output and no output is connected to more than one input. Figure 4-7 illustrates the following path routines: Note that a minimum of M^2 crosspoint switches is required. The SS-FDMA system trade-offs tentatively identify the range of M to be from 100 up to 10,000 consequently, the required number of crosspoint switching elements ranges from 10,000 to 100 million. Obviously weight, power, volume and reliability are important considerations.

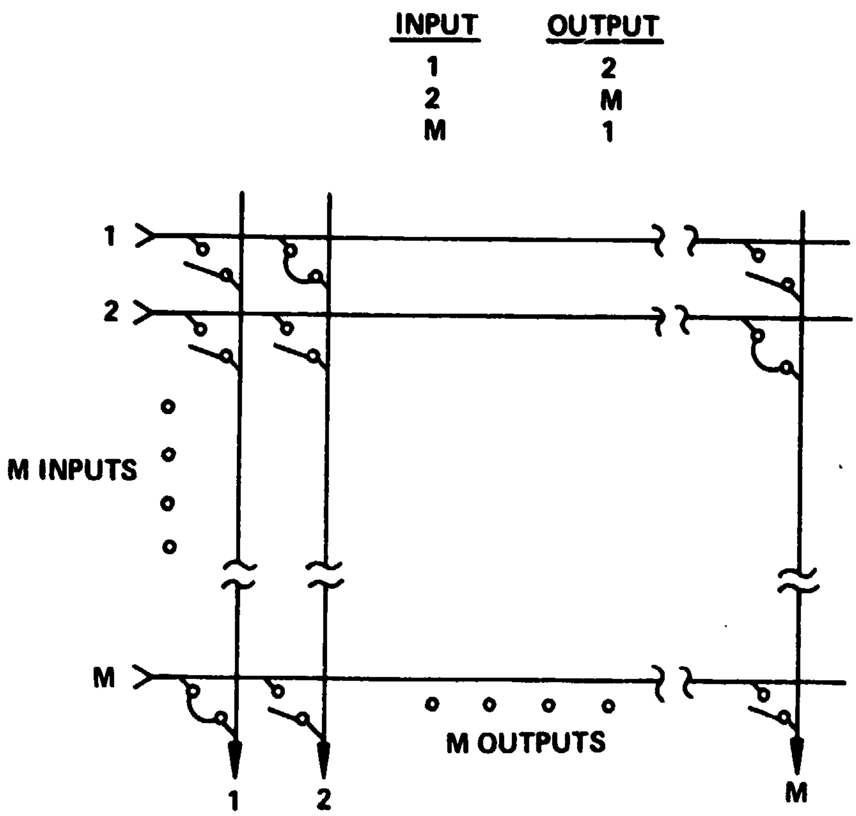


Figure 4-7. Functional Diagram of a Crossbar Switch

4.2.2 SWITCH IMPLEMENTATION CONSIDERATIONS

Table 4-3 lists important considerations and parameters considered in the evaluation of candidate switch implementations. This section discusses the inter-relationships and considerations among these parameters leading to a selected baseline switch concept.

TABLE 4-3. CROSSBAR SWITCH MATRIX
TRADE OFF PARAMETERS

- PERFORMANCE
  - 3 dB BANDWIDTH
  - FREQUENCY OF OPERATION
  - INTERCHANNEL INTERFERENCE (ISOLATION)
  - AM/PM CONVERSION
  - SIGNAL LEVEL
  - MATRIX SETTLING TIME
  - IMPEDANCE, VSWR
  - AMPLITUDE FLATNESS
  - GAIN SLOPE
  - GROUP DELAY/PHASE LINEARITY
  - CONTROL TO SIGNAL INTERFERENCE (CONTROL SIGNAL ISOLATION)
  - INSERTION LOSS
  - DIFFERENTIAL CHANNEL DELAY
  - NOISE FIGURE
- NUMBER OF INPUTS AND OUTPUTS
- RELIABILITY/REDUNDANCY/OPERATING LIFETIME, FAIL "SOFT" FEATURES, FAILURE MODES, RECONFIGURABILITY (FLEXIBILITY)
- MATRIX ORGANIZATION (CONFIGURATION), POWER DIVIDER MATRIX, N-WAY MATRIX
- SWITCH/DRIVER CONFIGURATION--SERIES, SERIES/PARALLEL, PARALLEL TRANSFER
- IMPLEMENTATION ALTERNATES--CMOS-SOS, PIN, SINGLE GATE FET, MECHANICAL RELAY, BI-POLAR TRANSISTOR
- SIZE, WEIGHT, POWER CONSUMPTION
- GROWTH CAPABILITY
- MATRIX CONTROL--LATCH, S R.
- MECHANICAL-PACKAGE
- ENVIRONMENT--TEMPERATURE, SHORT, VIBRATION, ACCELERATION

4.2.2.1 Operating Frequency

The maximum operating frequency requirement for each path through the crosspoint switch should be low to facilitate LSI implementation, typically 2-15 MHz. The maximum frequency handling capability of each switch path also is:

1. A strong function of the dimensions of the switch matrix (e.g., number of inputs and outputs), and
2. A strong function of the parasitic resistance and capacitance of the individual switches and associated input/output lines.

All crosspoint switch implementations considered (FET's, relays, bi-polar transistors, etc.) may be approximated by the equivalent circuit shown in Figure 4-8.

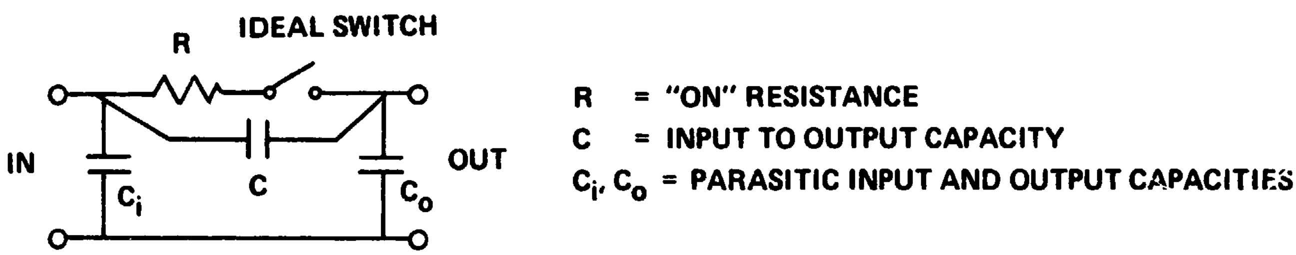


Figure 4-8. Typical Solid State Switch Representation (equivalent circuit)

Consider the case illustrated in Figure 4-7 where only one input is connected to only one output via a closed crosspoint switch. It is clear that the M-1 open switch networks which are common to an output line load that one closed crosspoint switch circuit. In addition, the M-1 open switch networks common to the input line load the input line driver. The net effect is to introduce a low pass network between every input and output line having a frequency response determined by the values of R, C, C_i, Co, M, and the equivalent driving point impedance of the driving source and the termination impedance. Parasitic elements are inherent in the technology selected to implement the switch, such as CMOS-SOS. The control of the effects of these parasitic impedances is achieved by limiting M by partitioning the switches and by controlling source and load impedances.

4.2.3 ISOLATION

An inspection of Figure 4-7 shows that every input line is coupled to every output line via the crosspoint switch capacity C. The net coupled interference signal (V_I) coupled to a given output line is undesirable and should be reduced to an acceptably low value. It is convenient to define the following ratio:

$$\frac{V_{DESIRABLE}}{V_I} = \bar{R}$$

$\bar{R}$ is a function of the same switch parameters which determine the frequency response of a switch path. In a practical case where a single crosspoint switch is used, the following approximations may be made:

$$f_{3db} \approx \frac{1}{2\pi MR\bar{C}} \quad \text{where } \bar{C} = C + C_o$$

$$\text{and } C_o = 3C$$

and

$$\bar{R} \approx \frac{1}{2\pi R f_M M^{1/2} C}$$

where f_{3b} is the 3 db bandwidth of the equivalent low pass response to the RF path and f_m = maximum frequency of operation.

Therefore, $R \times fm \approx f_{3db} \times M^{1/2}$

To a first order approximation the isolation-operating frequency product is a constant determined by the crosspoint switch parasitic impedance and the matrix dimensions, and is a useful design formula, since an estimate of allowable matrix size follows directly from a knowledge of switch parasitics and the desired isolation and operating frequency.

The following assumptions were made in the foregoing analysis:

1. The input driving impedance is small relative to crosspoint switch resistance R.
2. The interfering signals are non-coherent (in total).
3. The switch packaging concept introduces low parasitic capacity between adjacent input and output lines.

4.2.4 MATRIX PARTITIONING

The relationships derived in the previous paragraph show the futility of implementing a crosspoint switch directly along the lines indicated in Figure 4-7. Reasonable requirements for a crosspoint switch are as follows:

$$R \geq 40 \text{ (32 dB)}$$

$$fm \geq 10 \text{ MHz}$$

Estimated parasitics for a first cut CMOS-SOS crosspoint switch implementation are as follows:

$C \simeq 0.026$ pf and

$R \simeq 2500$ ohms, from which $M \simeq 40$, therefore

$$f_{3db} = \frac{Rfm}{M^{1/2}} = 64 \text{ MHz}$$

A typical 10-beam satellite requires an M=200 and a typical 100-beam satellite an $M=10^4$ for complete connectivity. It is evident then that a method for integrating smaller submatrices in order to obtain a larger effective M is useful. Another method for obtaining a larger matrix dimension is to employ a more complex crosspoint switch and this will be discussed later.

Figure 4-9 presents one approach to partitioning. An M x M matrix is partitioned into m x m submatrices. Output impedance buffers are included to preserve bandwidth while driving the parasitic switch capacities. A total of M, $\frac{M}{m}$ x 1 matrices are used to interrogate the m x m matrix columns. Therefore, $\frac{M^2}{m}$ additional crosspoint switches are added. The $\frac{M}{m}$ x 1 matrices preserve inter-line interference levels by providing additional isolation of the m x m matrix outputs. Low impedance drivers are included for the m x m input lines to conserve bandwidth. Although square matrices are shown, this may not be the optimum partitioning scheme. In fact, if very low impedance drivers are available ($<$ 10 ohms), then an m x n matrix (where $n > m$) is more optimum.

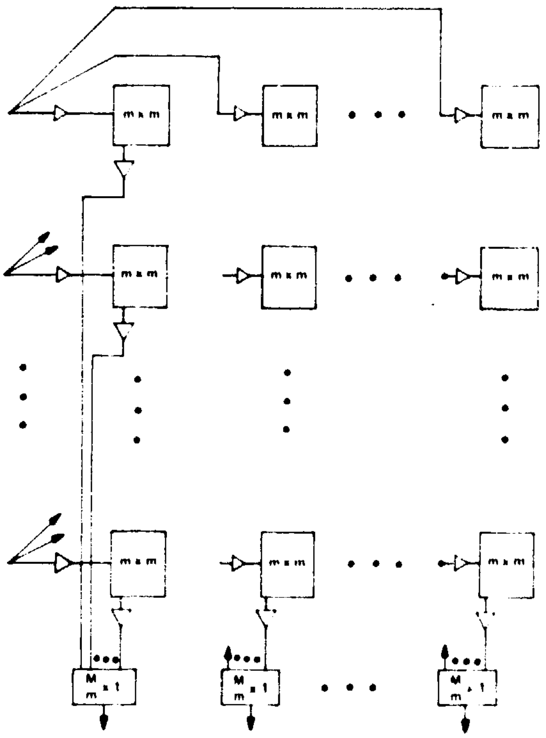


Figure 4-9. Partitioned M x M Switch Matrix

Using the M = 40 matrix, it should be possible with this scheme to construct a much larger crosspoint switch matrix and maintain acceptable bandwidth and isolation levels. Care must be taken, of course, to limit parasitic capacity between the matrix interconnections. The price paid is:

1. higher power consumption due to the impedance convertors, and
2. a modest increase in the number of crosspoint switches.

It is estimated that a 400 x 400 crosspoint switch is feasible employing this method. This will require 100, 40 x 40 submatrices. The implementation of a $10^4 \times 10^4$ matrix is a totally different matter and will require further detailed analysis.

4.2.5 CROSSPOINT SWITCH IMPLEMENTATION

Switches may be operated in series, shunt, or in combination as shown in Figure 4-10.

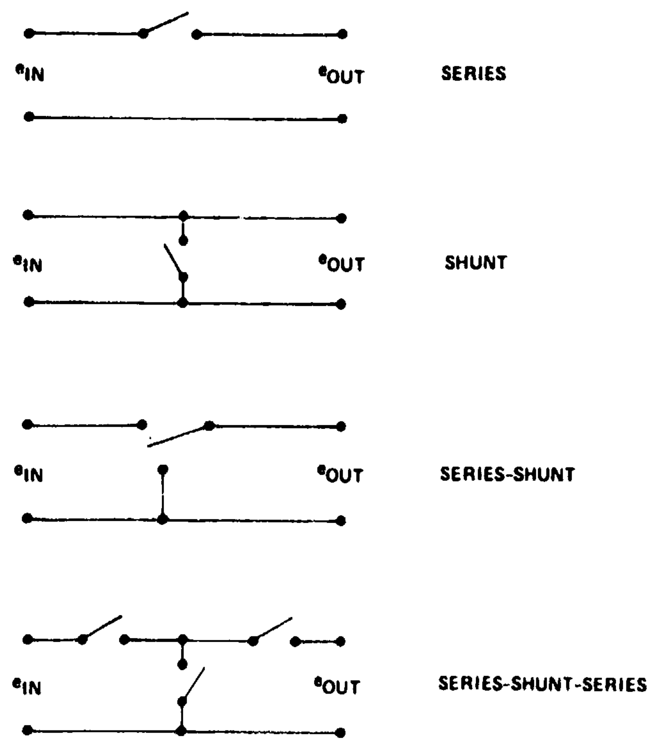


Figure 4-10. Representative Switch Configurations

Each configuration has its own special properties, but it appears that the series and the series-shunt-series configurations are most useful for these applications.

A series switch is simple to emplement and also disconnects the source from the load; a parallel switch in conjunction with a series switch can improve the switch isolation over that of a simple series switch.

A variety of electronic switching devices are available for crosspoint switch implementation. The best device (or devices) permit the most favorable combination of the matrix trade-off parameters listed in Table 4-3. Some likely candidates with a listing of characteristics is given in Table 4-4. Trial 32 x 32 switch matrix designs were analyzed for each implementation and conclusions drawn. CMOS-SOS (Complimentary Metal Oxide Semiconductor - Silicon On Sapphire) was tentatively selected since acceptable performance requirements can be achieved with negligible power consumption; it is potentially reliable; it is capable of compact LSI circuit implementation; and a substantial effort is being expended by industry to apply CMOS-SOS LSI to space systems.

An example of a CMOS-SOS switch design, illustrated in Figure 4-11, was used to explore performance and packaging problems. A crosspoint switch logic and circuit is shown in Figure 4-12. Figure 4-12a, the logic representation, depicts a single series crosspoint switch operated by a latching circuit. Figure 4-12b, depicts the basic circuit elements. Transistors P1 and N1 are "long channel" (high impedance) devices so that the internal mode of the latch can be driven directly through switches H2 and H3. The switch is controlled by using the signal input and output lines, unique to each switch, to control the switch state. Simultaneous pulses on the input and output lines are necessary to change the switch state. If these pulses are coincident with a "data" pulse indicating polarity (logic "0" or "1") the switch state will revert to that of the "data" pulse polarity. Controlling the switch through the signal lines is an acceptable convenience because signals are not expected to be present anyway during switching. Figure 4-13 depicts the parasitic impedances for the tvpical circuit which values were used previously to explore bandwidth and isolation properties. The greatest uncertainty here is the value C_{DS}, the drain to source capacitance of the switch. It was arrived at with a previously calculated value for parallel metal runs and making allowances to account for the different dielectric and for the presence of the gate electrode. It is also possible, based on some recent work at GE on air gap crossovers, to reduce C_{cu}. A preliminary CMOS-SOS chip layout was performed to explore the problems of an LSI implementation and to get some ideas of packaging arrangements. (The packaging design is described subsequently). It appears feasible to place approximately 1000 switch circuits, (7 transistors each) on a chip 1/4" by 1/4". Ample room for transistors is available however provision of lead-wires for input/output lines etc is a substantial problem. Figure 4-14 is an example of such a layout showing the seven transistors and the wire pad connections. The arrangement is "pad" limited. The cell occupies 11.1 mil^2, is based on a 2 Kilohm switch (V_{OD} = 10V). Because of pod spacing limitation, input and output lines are alternated on opposite sides of the chip.

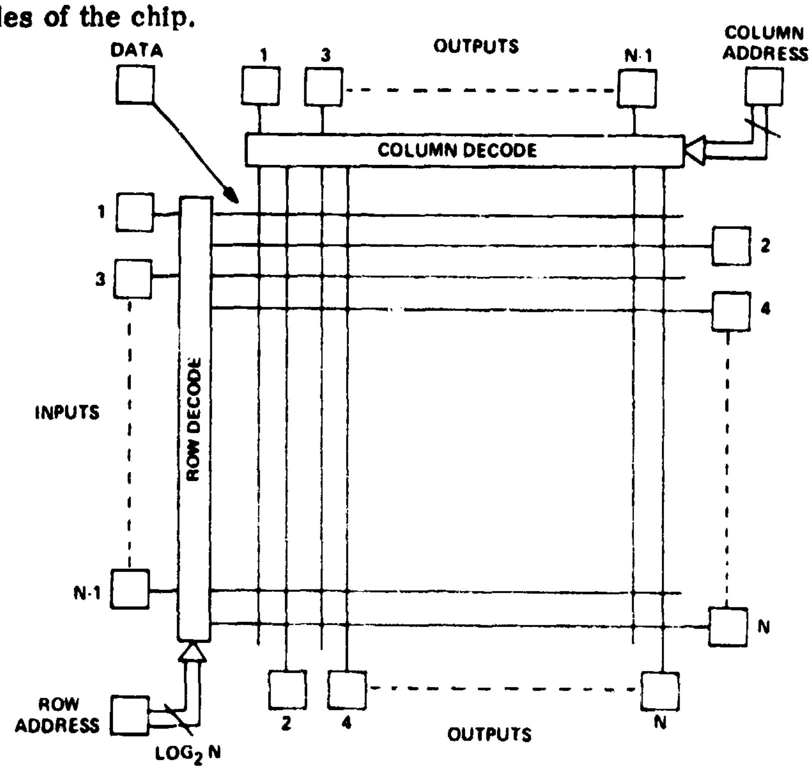


Figure 4-11. N x N CMOS-SOS Crosspoint Switch with Memory

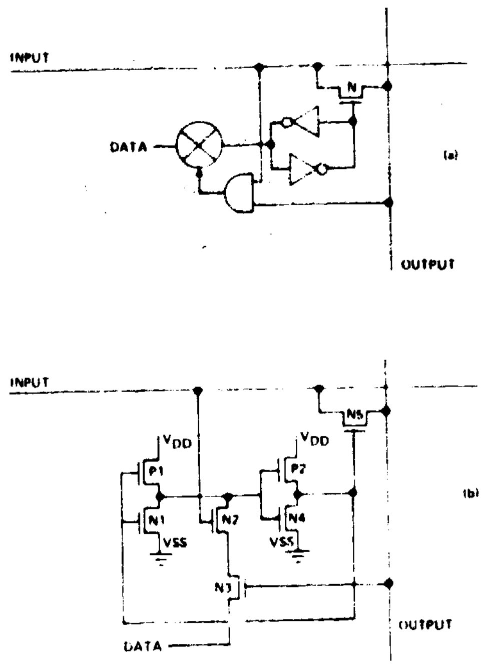


Figure 4-12. CMOS-SOS Crosspoint Switch and Latch

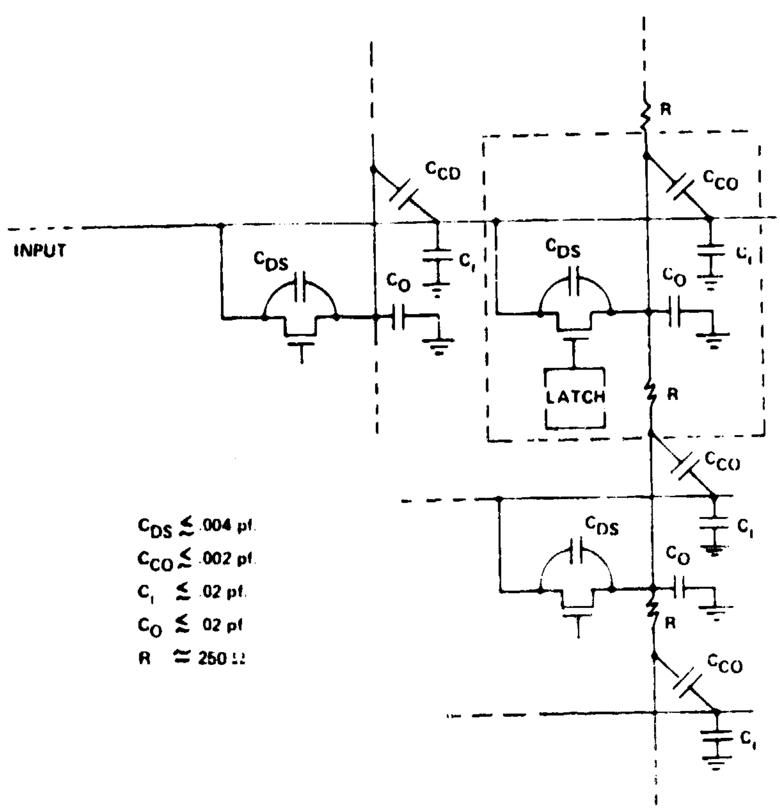


Figure 4-13. Parasitic Elements in the Preliminary Crosspoint Switch Layout, based on CMOS-SOS

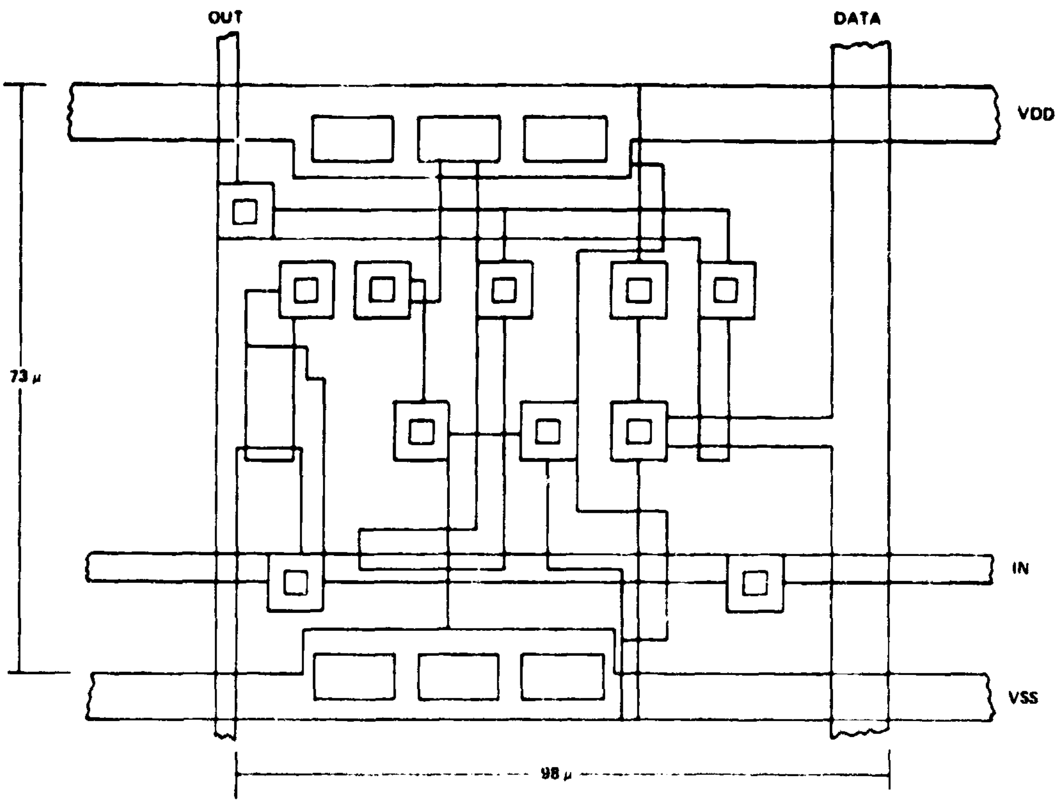


Figure 4-14. Preliminary Layout CMOS-SOS Crosspoint Switch and Latch

The other performance parameters of Table 4-3 are not considered to be of major importance. Most switch devices and configurations will meet acceptable performance levels. Amplitude flatness, differential channel delay, gain slope, phase linearity, are determined primarily by the "path" frequency response. Switch noise figure will not be significant (within limits) since the system noise figure is basically established by the microwave receiver "front end". Acceptable matrix settling times can probably be achieved even with mechanical relays. AM/PM conversion is a function of device non-linearity, and acceptable signal levels (-10 dBm to 0 dBm) can be used to control this effect. Control-to-signal interference may be made negligible by proper circuit design. Modest insertion losses may be insured by proper design practice or compensated by amplifiers.

TABLE 4-4. 32 x 32 MATRIX IMPLEMENTATION EXAMPLES

Crosspoint Switch Implementation	Crosspoint Switch Configuration	Isolation @10 MHz (dB)	f_{3dB} (Mhz)	Matrix Power Consumption (W)	Matrix Size (IN^3)	Remarks
RELAY (LATCHING)	S	>50	>10	NEGLIGIBLE	150	• SWITCHING TIME = 10 M SEC • LARGE VOLUME, WEIGHT • WIDE BAND
PIN DIODE	S-P-S	>60	1000	100	LSI	• HIGH POWER CONSUMPTION
BIPOLAR XISTOR	S-P-S	20	2	50	LSI	• POOR ISOLATION • SMALL BANDWIDTH
N-P CHANNEL FET	S-P-S	42	40	NEGLIGIBLE	LSI	• LOW IMPEDANCE
CMOS/SOS	S-P-S	42	40	NEGLIGIBLE	0.2	• HIGH IMPEDANCE
GaAs FET	S-P-S	>45	500	NEGLIGIBLE	LSI	• WIDE BAND • LOW POWER CONSUMPTION

4.2.6 RELIABILITY

The required useful operating lifetime of the crosspoint switch in a spacecraft application should approach 10 years. At this point in the study, it is not clear what design philosophy should be used to achieve this order of reliability. It should be noted, however, that a crosspoint switch per Figure 4-7 does degrade gracefully inasmuch as alternate multiple paths are available between ground stations in case of a crosspoint switch failure.

A number of methods have been traditionally used to enhance the reliability of spacecraft hardware. Block or piece-part redundancy may be employed. So-called "Re-arrangeable Multistage Networks" may also be useful (1).

4.2.7 SWITCH PACKAGING

A preliminary packaging design was performed in order to identify potential problems. A 500 x 500 element (250,000 switch elements) CMOS-SOS crosspoint switch with a 10 MHz bandwidth, was chosen for illustration. State-of-the-Art packaging techniques can provide the required density of circuits and interconnections in a modular configuration. The arrangement, illustrated in Figure 4-15, consists of a 5 x 5 array of thin film hybrid circuits on a multi-layer printed circuit board. The overall package is approximately 15" x 15" x 3" and weighs approximately 10 lbs.

The basic building block illustrated in Figure 4-16 is a CMOS-SOS integrated circuit 0.25 x 0.25 inches square, having 200 solder bump contacts arranged around the periphery of the active area. Each chip contains 2500 switch elements arranged in a 50 x 50 array. It is clear from Figure 4-15 that interconnection of the switch array will be a major problem.

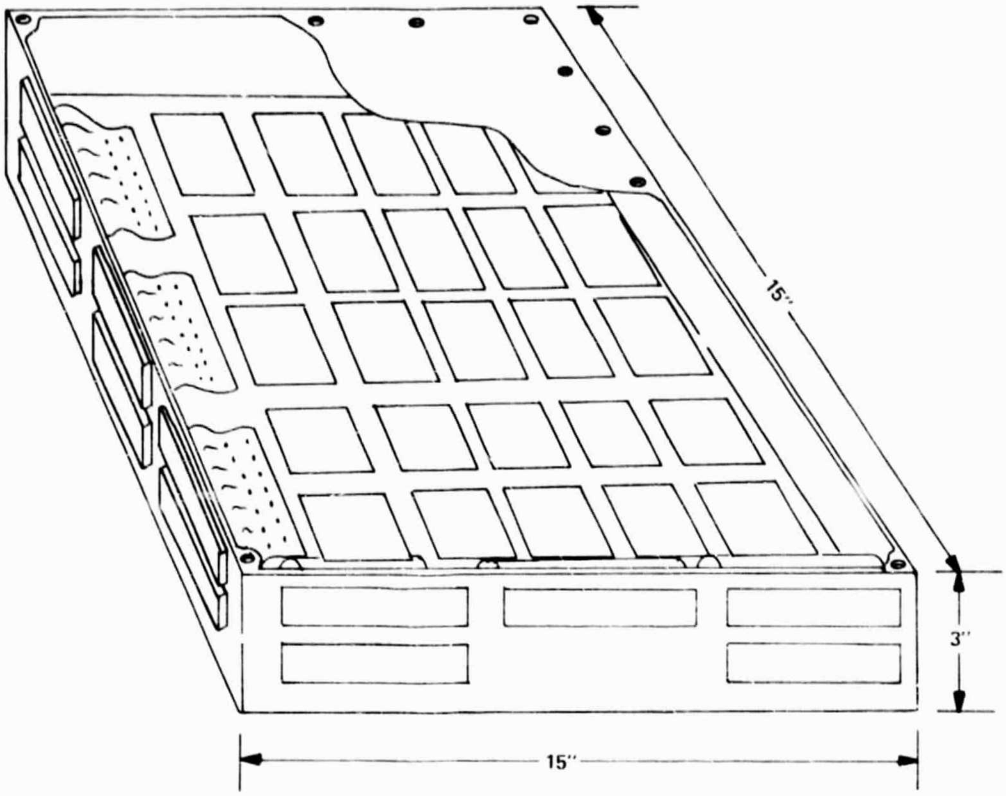


Figure 4-15. General Electric 500 x 500 Crossbar Switch Arrangement

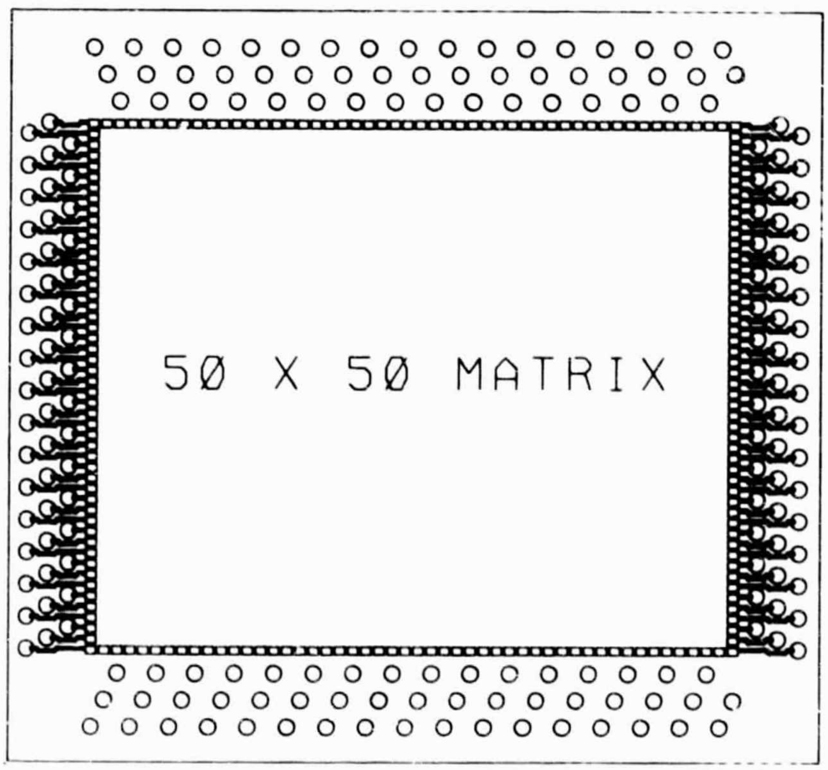


Figure 4-16. 50 x 50 Element CMOS-SOS Switch on a Single Chip

The chips are arranged in modules having four (4) IC's each mounted flip-chip style to a ceramic substrate. Because of the high interconnection ratio, it might be desirable to group more parts per module, but production yields on such a large device tend to be low, cancelling out any reduction in top assembly wiring complexity.

Each module (ceramic substrate) containing four (4) IC's has 408 pins on 0.1 inch grids exiting from the base. This provides for 400 inputs and outputs, leaving 8 pins for power, grounds, telemetry, and spares. The geometry of the substrate is shown in Figure 4-7. Thin film deposition techniques are used to provide the interconnection pattern between the solder bump pads and module pins which consists of 1 mil lines and 1 mil spaces. The area of each module is about 2.4 x 2.4 inches. Provision will be made to hermetically seal the active circuits from the environment.

An alternate substrate design can be evaluated using thick film technology. A multi-layer co-fired thick-film circuit can accomplish the interconnection function, using 5 mil lines and 5 mil spaces on an eight-layer substrate. The yields for both thin and thick-film approaches should be considered before choosing a substrate design.

The interconnecting method for the 25 modules and 1000 input/output connections is a multi-layer printed circuit board, approximately 15 inches per side, having up to 12 layers and 11,300 plated through holes. A possible arrangement is shown in Figure 4-18. Boards of this complexity are routinely made for the computer industry.

The P/C board is mounted in an aluminum housing. The small amount of heat developed by each switch element is easily dissipated by conduction through the chassis mounting points. Inputs and outputs are brought off the board by 3-layer Kapon/copper flex cables which are soldered into the plated-through-holes in the terminal areas of the board. The cables are terminated in high density "D" style connectors (104 pins each) mounted in the chassis. Five connectors are required for inputs, five for outputs and one for power and grounds.

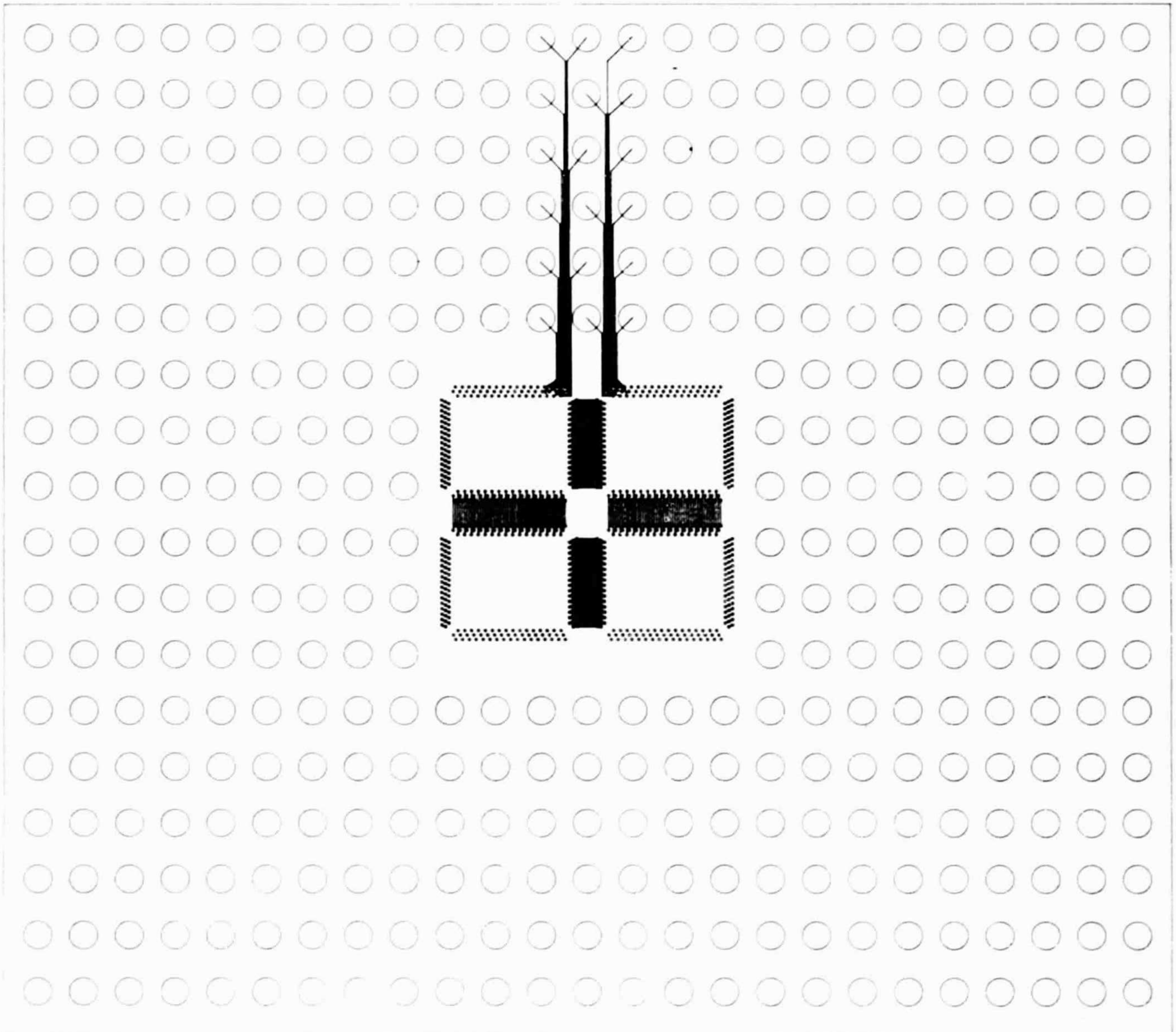


Figure 4-17. Geometry of Module Containing Four IC's

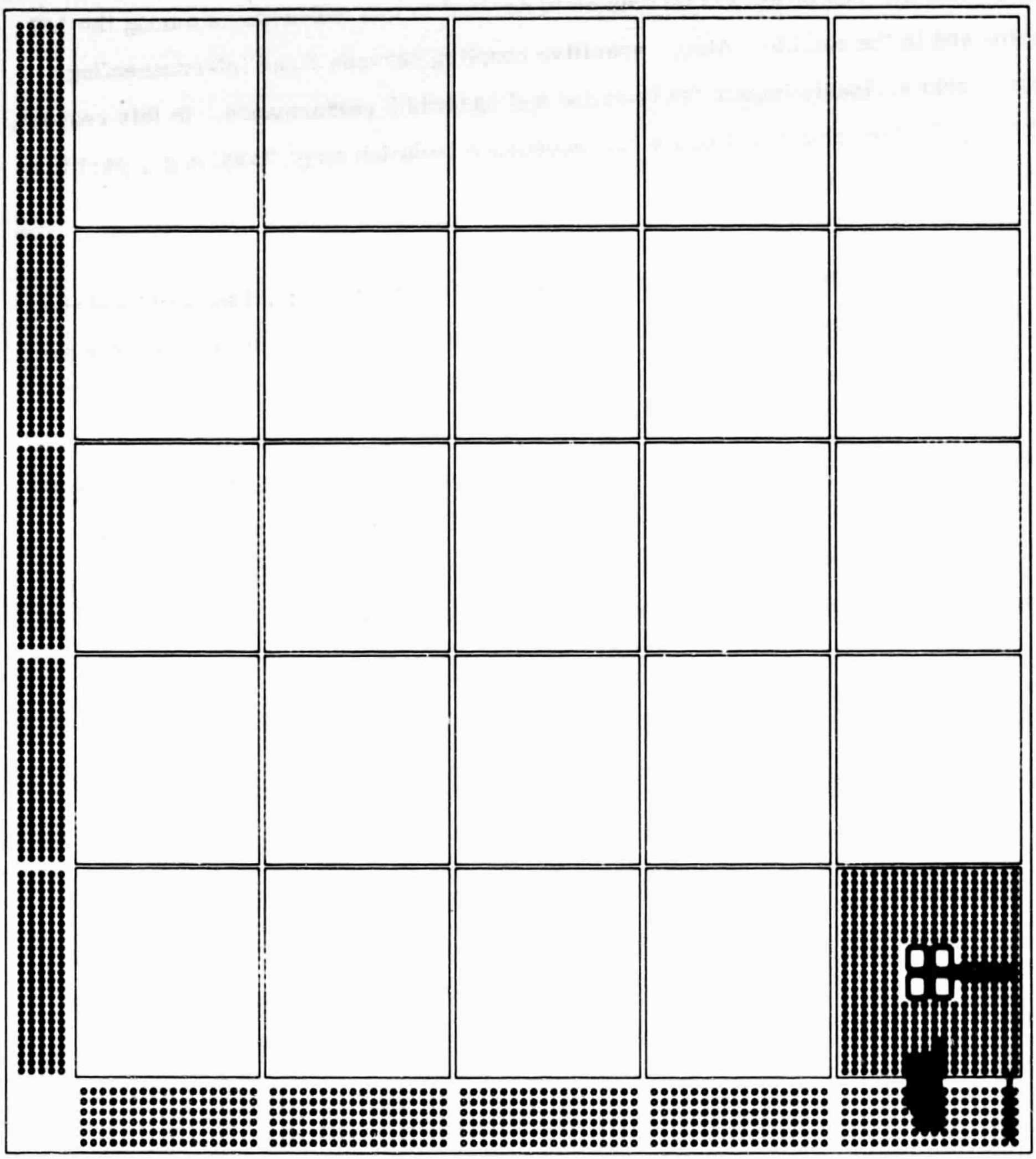


Figure 4-18. Multi-Layer Printed Board Pattern

All of the packaging techniques described in this section have been developed and do not constitute a significant technical risk on an R&D basis. The question of production yields and long-life reliability remain the important questions to be evaluated.

A trial design yields reasonable size, weight and power parameters. It should be noted that the bulk of the switch volume is devoted to interconnections among the LSI chips and to the outside. Also, capacitive coupling between these interconnecting runs could seriously impair the isolation and bandwidth performance. In this respect, it is highly desirable to introduce low impedance isolation amplifiers; e. g., partitioning, on the LSI chips.

The achievement of adequate communications performance, e. g., bandwidth and isolation, and achievement of a producible, light weight package requires an iterative design effort between the packaging/LSI engineer and the communications engineer. Partitioning, switch control circuits, T&C and reliability assessment, particularly from the point of view of avoiding catastrophic failure and achieving a design which fails "gracefully" are important considerations. The design described herein is believed to be representative of the needed SS-FDMA switch technology. The steps to accomplish technology development are described in Section 6.

4.3 FILTERING TECHNOLOGY

4.3.1 INTRODUCTION

A number of different filtering requirements are identified as part of the up/down conversion and frequency multiplexing/demultiplexing operations for SS-FDMA. The "Path" band pass filters present a severe challenge in terms of size, weight, power and performance because at least one channelization filter (and perhaps two) is required per path so that a large number of filters are needed. Other filters also are needed to reject local oscillator signals, avoid images in mixing etc. In addition, large percentage bandwdths are required. Good bandwidth utilization and minimal "path" performance degradation requires good "out of band" rejection and good "in-band" performance.

4.3.2 REQUIREMENTS

Two classes of filters appear useful in the SS-FDMA systems considered; one centered in the 200 to 300 MHz band and one centered at approximately 2 MHz. The requirements are shown in Table 4-5. The class I filter is used as a channel filter in the 10 beam system or in the 100 beam system. The class II (narrowband) filter is used as a channel filter in the 100 beam system. The skirt selectivity for these filters is roughly equivalent to a 6 pole, 0.3 dB ripple Chebychev design. The channel spacing is 5 MHz and 10 MHz for class I and 1 MHz for Class II, thus achieving an 80% bandwidth utilization. A 3 dB insertion loss is desired to reduce requirements for analogue amplification in the path (which consume power) however, larger losses are otherwise tolerable (indeed SAW devices have insertion losses exceeding 20 dB). Specifications on ultimate rejection, amplitude and phase ripple, and temperature stability are typical, and in most cases, insure suitable "in band" channel transparency. Since the channel filters are used in large quantity, small size and weight is important. Passive implementations are highly desirable in order to conserve power.

4.3.3 FILTER IMPLEMENTATION

A technology survey, including a leterative survey was conducted in order to identify filter technologies (present and future) applicable to SS FDMA systems. These surveys showed that SAW (Surface Acoustic Wave) devices are a candidates for the compact, high performance filters needed for SS-FDMA. Existing SAW band pass filters use Quartz and Lithium Niobate, however other base materials are being

TABLE 4-5. BAND PASS FILTER PERFORMANCE REQUIREMENTS

	CLASS I	CLASS II
CENTER FREQUENCY	200/300 MHz	2 MHz
3 dB BANDWIDTH	4 MHz & 8 MHz	0.8 MHz
40 dB BANDWIDTH	6 MHz & 12 MHz	1.2 MHz
INSERTION LOSS	3 dB	3 dB
ULTIMATE REJECTION	40 dB	40 dB
AMPLITUDE RIPPLE	0.5 dB P-P	0.5 dB P-P
PHASE RIPPLE	3° P-P	3° P-P
TEMPERATURE STABILITY	100 KHz/50°C	20 KHz/50°C
SIZE	0.5 x 0.5 x 0.1 in	0.25 x 0.25 x 1.5 in

pursued - Lithium Tantalite is an example. Low pass filters are presently impractical. The useful frequency range is from 20 to 400 MHz but considerable effort is being expended at present to extend operation to higher frequencies. Discrete devices are available from a number of vendors, notably the General Electric Company and Anderson Labs. The underlying physics is well understood and filter synthesis procedures are refined. The fabrication techniques are amenable to LSI implementation since multiple filters can share a common substrate. Although the SAW filter is passive its insertion loss is high ($\approx$ 20 dB) requiring complementary analog amplifiers. Ultimate rejection exceeding 35 dB is difficult to achieve and "out of band" spurious responses are common. This can be a particularly severe problem because the FDMA percented bandwidth can be large.

Composite quartz crystal also are candidates. These devices are relatively large and are useful only at the lower frequencies. They represent a relatively mature technology and very sharp skirt selectivity is obtainable. Although a passive device, a composite filter meeting SS-FDMA requirements has an insertion loss $\approx$ 10 dB. Its application to a high performance filter bank does not appear warranted in view of the SAW filter potential.

Ceramic filters are potentially compact and require no standby power. They appear most attractive at the lower frequency range (1 to 10 MHz) as band pass and/or L. O. rejection filters. Unfortunately, they have not been extensively applied in high performance band pass applications and no extensive filter synthesis techniques are reported in the literature. The published papers indicate that Japanese firms presently dominate this field. Ceramic technology probably has a place in future SS-FDMA systems, however, a substantial development effort will doubtless be required.

Discrete L/C and Mechanical filters are relatively bulky. The former suffers from low resonant Q's and the latter are suitable only to operation at low operating frequencies ($\approx$ 100 KHz).

Cavity type filters (comb line, interdigital, coaxial, etc,) deserve consideration in areas requiring high performance at relatively high frequencies, but in limited quantities. These filters are large and heavy and are not particularly attractive.

Active operational amplifier filters can be made relatively compact if LSI techniques are employed. They are presently restricted in operating frequency (primarily by the operational amplifier response) and require relatively large amounts of power. This power increases with operating frequency. The development of a micro power high frequency operational amplifier (voltage gain greater than 20 at 10 MHz) would change this picture drastically, especially in subsystem implementations using low pass channel filters.

A possible implementation is shown in Figure 4-19 containing low pass filters, mixers and quadraure L. O.'s. It may be shown that this configuration is equivalent to a band pass filter with a band pass equivalent to that of the low pass filter. Furthermore, the configuration may be used to filter and translate a segment of bandwidth centered at f_1 to one centered at f_2. This is accomplished by using f_1 for the first L. O. mixer pair and f_2 for the second L. O. mixer pair.

The Switched Capacitor and Charge Coupled Device (CCD) implementations are potentially compact using LSI techniques, but again are restricted to the lower frequencies ($\approx$ 500 KHz). Standby power may be acceptable, but a breakthrough toward higher frequency operation is required.

A Digital Filter implementation is probably not practicable even at the narrowist path bandwidths considered. A sampling A/D conversion is necessary, preceeded by a noise limiting analog filter. In order to achieve a good approximation to a band pass response, a high ($\approx$ 10 times the Nyquist rate) sampling rate is mandatory.

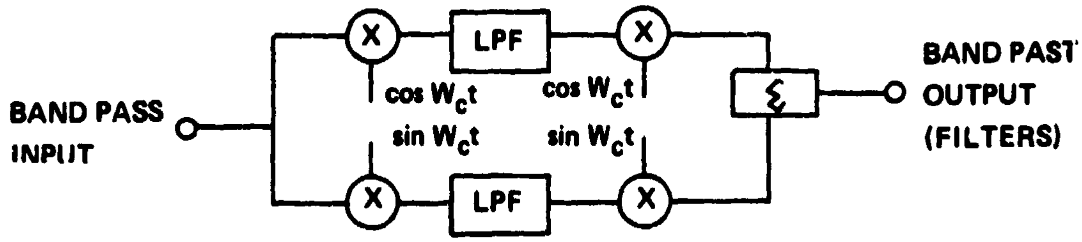


Figure 4-19. Band Pass Filter Using Low Pass Filter

A/D conversion using a 6 BIT resolution is probably adequate. A bit rate approaching 200 megabits is not unusual for a single 1 MHz bandwidth path. Power and equipment complexity appear prohibitive even with a modest number of channels. A list of candidate filter technologies with appropriate comments on characteristics and SS-FDMA applicability is given in Table 4-6.

4.3.4 SAW DEVICES

SAW devices are now being researched and developed which have demanding specifications and can therfore have a significant impact on both the performance and system design philosophy. SAW devices have the following advantages:

1. Planar structures which use photolithographic techniques for fabrication.
2. The slow velocity of the acoustic wave allows miniature devices to be made.
3. The signal can be tapped during propagation and therefore allows for sophisticated signal processing.
4. The design is predominantly built into the mask used in the fabrication of the structure.
5. The surface wave is stable and on some materials is very insensitive to temperature change.

TABLE 4-6. FILTER IMPLEMENTATION CHARACTERISTICS

FILTER TECHNOLOGY	SALIENT CHARACTERISTICS	APPLICABLE TO SS-FDMA
SAW QUARTZ LITHIUM	• COMPACT • HIGH PERFORMANCE • NO STANDBY POWER • RELATIVELY MATURE TECHNOLOGY	YES
QUARTZ BULK MONOLITHIC	• RELATIVELY LARGE SIZE • NARROW BAND WIDTH • NO STANDBY POWER • RELATIVELY MATURE TECHNOLOGY	PROBABLY NOT
CERAMIC - BULK	• POTENTIALLY COMPACT • NO STANDBY POWER • RELATIVELY IMMATURE TECHNOLOGY	PROBABLY YES
DISCRETE L/C	• LARGE VOLUME • LOW "Q"	NO
MECHANICAL	• LOW FREQUENCY • LARGE VOLUME	NO
CAVITY	• HIGH FREQUENCY • LARGE VOLUME	PROBABLY YES
ACTIVE OP. AMP	• HIGH STANDBY POWER • LOW FREQUENCY • COMPACT	PROBABLY NOT
SWITCHED CAPACITOR & CCD	• COMPACT • LOW FREQUENCY RESPONSE • LOW PASS POTENTIAL • MODEST STANDBY POWER	PROBABLY
DIGITAL FILTER	• REQUIRES SAMPLING AND NO CONVERSION • LOW FREQUENCY RESPONSE • HIGH STANDBY POWER	PROBABLY NOT

The attractive properties of Surface Wave bandpass filters are especially applicable to radar and communications systems, SAW's have the ability to meet a precisely defined amplitude and phase characteristic simultaneously in single filters, especially since wide-band surface wave filters may be small enough for hybrid integration. They can be designed to interface with standard, wideband IC amplifiers and therefore offer rugged amplifiers with plug-in exchangeable filters.

SAW band pass filters have typical insertion losses of 20 dB. An important objective of a technology study would be to study techniques for reducing the insertion loss of SAW filters. The size of a passive SAW device is determined by its center frequency and bandwidth; narrowband filters being somewhat larger than widerband filters.

SAW bandpass filters can be designed with complex phase and amplitude characteristics. The SAW filter will replace the bulky lumped constant filter, especially where stability, ruggedness, reliability, weight and size are of prime importance, SAW's are especially attractive for achieving low out-of-band spurious response. The type of filter should be especially useful for low distortion requirements, SAW filters have many advantages and applications in modern electronic systems where an accurate pass and stop band can be well defined.

The Electronics Laboratory of the General Electric Company has been pursuing the development of SAW devices for the past seven years. As the center of Research for SAW technology of the G. E. Company, the Laboratory has been investigating various types of SAW devices such as bandpass filters, dispersive delay lines, single pass delay lines, resonators and oscillators.

A recent in house program involving the study of high performance bandpass filters proved very successfull. Reproducible SAW filters were required for a sidelobe canceller receiver loop in a radar system which had a center frequency near 300

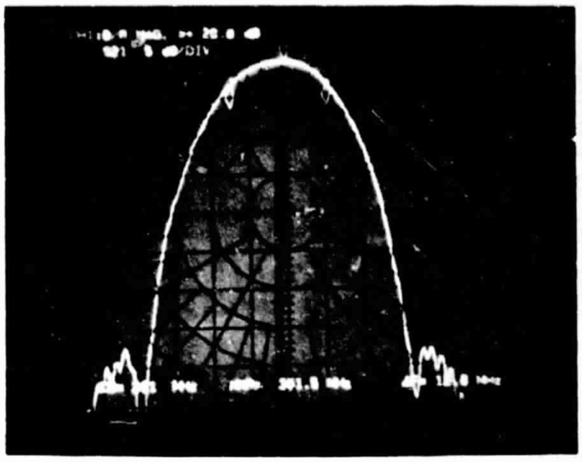
MHz. Figures 4-20 through 4-22 show the excellent transfer response of three representative filters. The filter requirements were:

f_o	=	$\simeq$ 300 MHz
BW	=	5 MHz
Sidelobe	=	$>$30 dB
IL	=	$<$30 dB
Amplitude	=	$\pm$ .25 dB
Phase Linearity	=	$\pm 1.25^{\circ}$
Reproducible	=	All filters must be identical in the passband 30 dB down.

These specifications were all met and were exceeded in the actual filter in some parameters. As shown in the first three figures, of special significance is the extremely low amplitude ripple $\pm$.03 dB and purity of bandpass characteristic. Figure 4-23 shows the excellent linear phase characteristic for the three filters. It would indeed be difficult if not impossible to achieve such linearity in a lumped constant type filter.

Figures 4-24 and 4-25 show the group delay response over a $\pm \Delta f$ scale of 1.1 MHz. Such linear phase and extremely low distortion depicts the advantages of using a high performance SAW filter in a radar as a communication system for processing of various types of waveforms or signals. Figure 4-26 shows the highly reproducible response of 6 matched filters superimposed from a network analyzer test response. Note that the 5 filters lie right on top of each other in the passband response. Figures 4-27 and 4-28 are close up views of 6 filters and a single filter respectively. The results of this program have clearly shown the advantages of SAW filters for a specific application where flat magnitude and group delay responses are of prime importance.

Surface wave filters are becoming increasingly attractive in communication systems at UHF and VHF frequencies. Their major advantages are small size, low cost, good reproducibility and easy integratability with other microelectronic circuitry.



MSR 6A-1

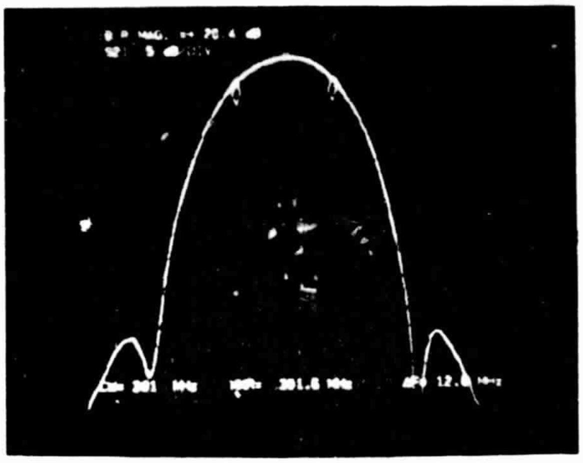
f_o = 301.5 MHz

$\pm \Delta f$ = 12.0 MHz

IL = 22.2 dB

Vertical Scale - 5 dB/Div.

TRANSFER



MSR 6A-2

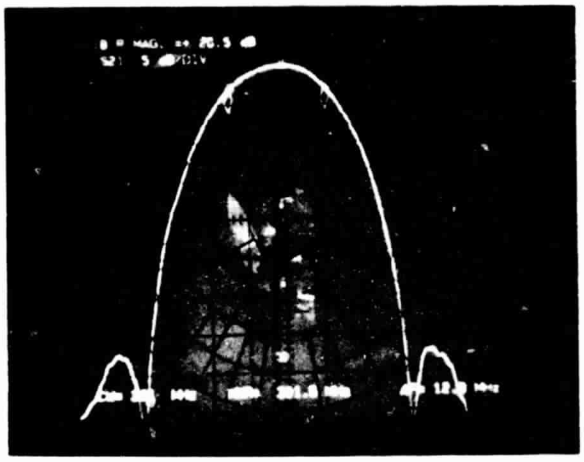
f_o = 301.6 MHz

$\pm \Delta f$ = 12.0 MHz

IL = 21.8 dB

Vertical Scale - 5 dB/Div.

TRANSFER



MSR 6A-3

f_o = 301.5 MHz

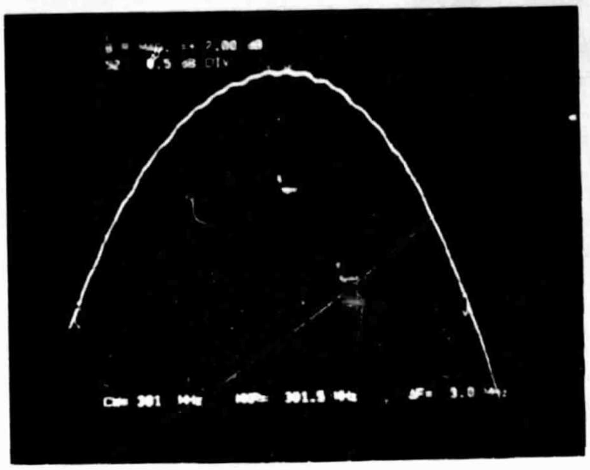
$\pm \Delta f$ = 12.0 MHz

IL = 21.7 dB

Vertical Scale - 5 dB/Div.

TRANSFER

Figure 4-20. Transfer Response of SAW Filters



MSR 6A-1

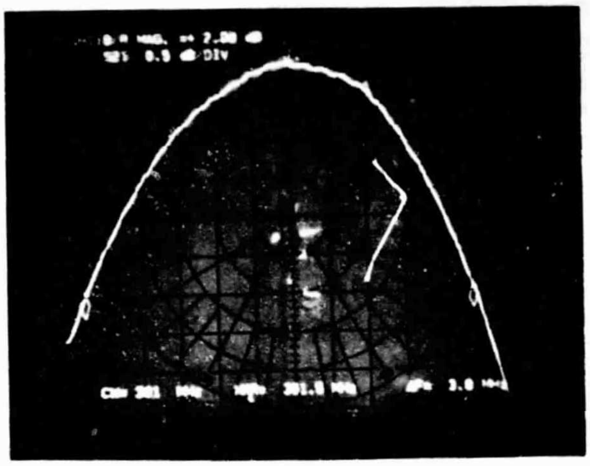
f_o = 301.5 MHz

$\pm \Delta f$ = 3.0 MHz

Vertical Scale - 0.5 dB/Div.

f_1 = 298.63 MHz

f_2 = 303.77 MHz



MSR 6A-2

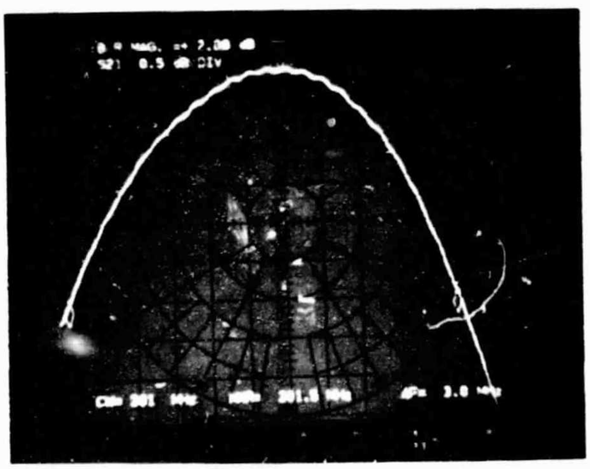
f_o = 301.5 MHz

$\pm \Delta f$ = 3.0 MHz

Vertical Scale - 0.5 dB/Div.

f_1 = 298.78 MHz

f_2 = 303.93 MHz



MSR 6A-3

f_o = 301.5 MHz

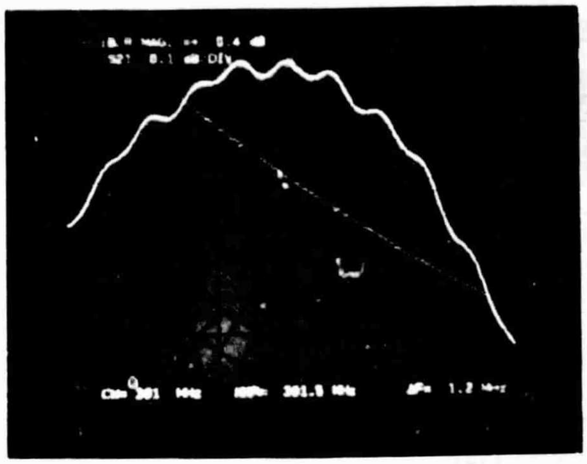
$\pm \Delta f$ = 3.0 MHz

Vertical Sca[illegible] - 0.5 dB/Div.

f_1 = [illegible]5 MHz

f_2 = 303.84 MHz

Figure 4- Bandwidth Response of SAW Filters



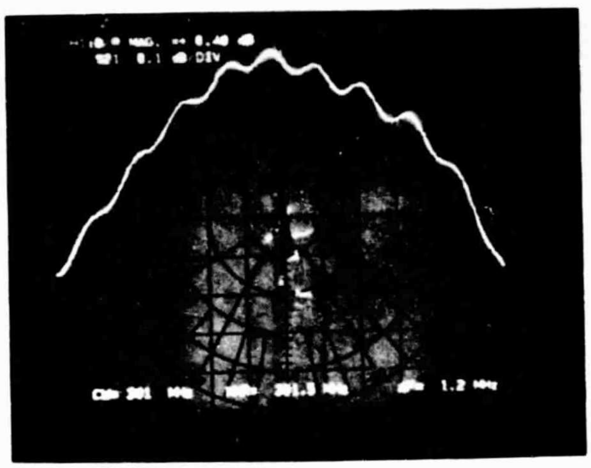
MSR 6A-1

f_o = 301.5 MHz

$\pm \Delta f$ = 1.2 MHz

Vertical Scale - 0.1 dB/Div.

RIPPLE



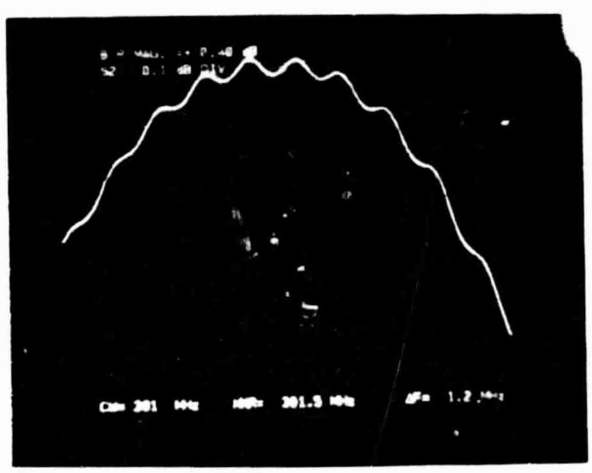
MSR 6A-2

f_o = 301.5 MHz

$\pm \Delta f$ = 1.2 MHz

Vertical Scale - 0.1 dB/Div.

RIPPLE



MSR 6A-3

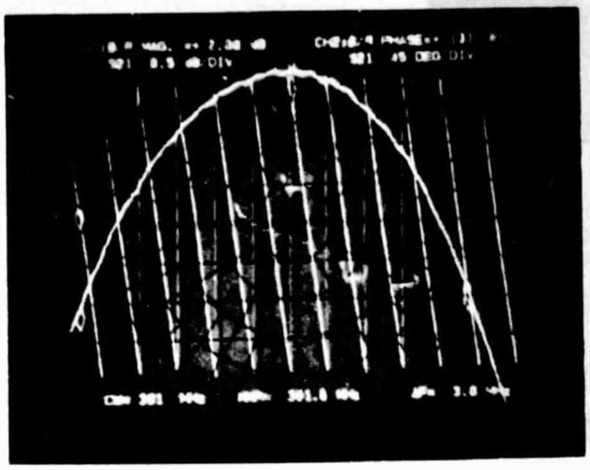
f_o = 301.5 MHz

$\pm \Delta f$ = 1.2 MHz

Vertical Scale - 0.1 dB/Div.

RIPPLE

Figure 4-22. Amplitude Response of SAW Filters



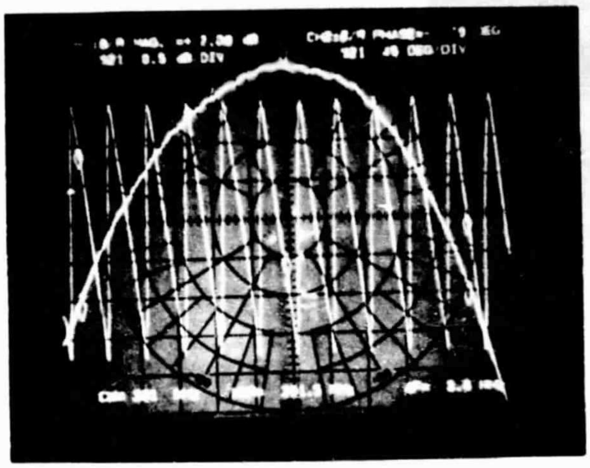
MSR 6A-1

f_o = 301.5 MHz

$\pm \Delta f$ = 3.0 MHz

Vertical Scales - 0.5 dB/Div.
45^o/Div.

PHASE



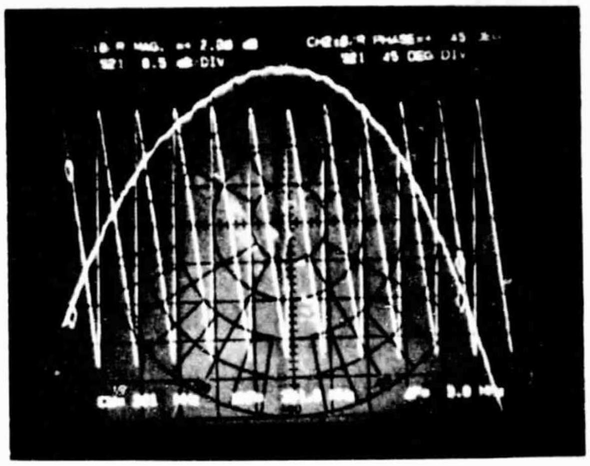
MSR 6A-2

f_o = 301.5 MHz

$\pm \Delta f$ = 3.0 MHz

Vertical Scales - 0.5 dB/Div.
45^o/Div.

PHASE



MSR 6A-3

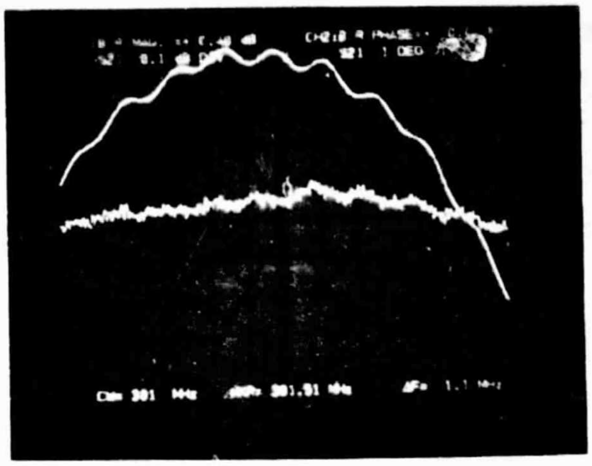
f_o = 301.6 MHz

$\pm \Delta f$ = 3.0 MHz

Vertical Scales - 0.5 dB/Div.
45^o/Div.

PHASE

Figure 4-23. Phase response of SAW Filters



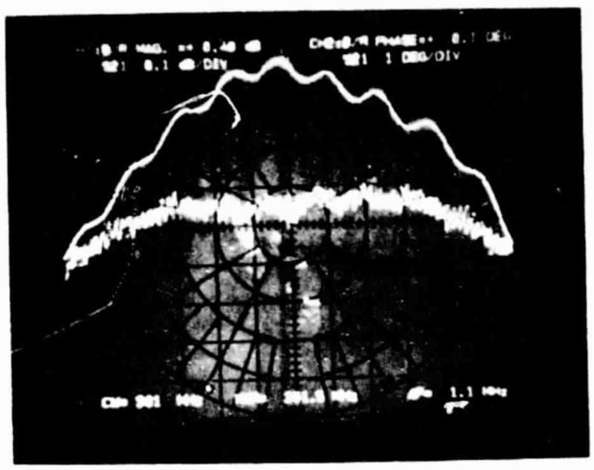
MSR 6A-1

f_o = 301.51 MHz

$\pm \Delta f$ = 1.1 MHz

Vertical Scales - 0.1 dB/Div.

1^o/Div.



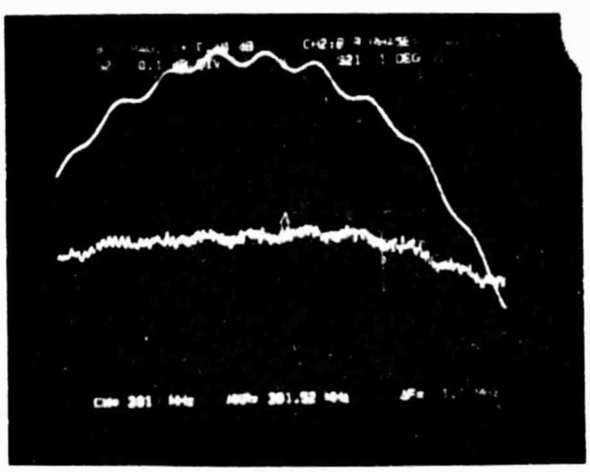
MSR 6A-2

f_o = 301.5 MHz

$\pm \Delta f$ = 1.1 MHz

Vertical Scales - 0.1 dB/Div.

1^o/Div.



MSR 6A-3

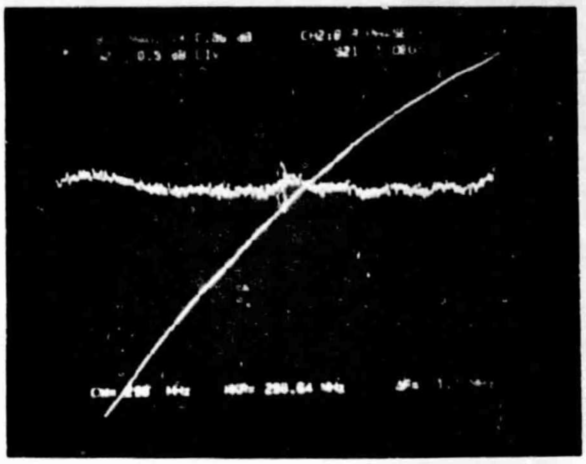
f_o = 301.52 MHz

$\pm \Delta f$ = 1.1 MHz

Vertical Scales - 0.1 dB/Div.

1^o/Div.

Figure 4-24. Phase and Group Delay Responses of SAW Filters



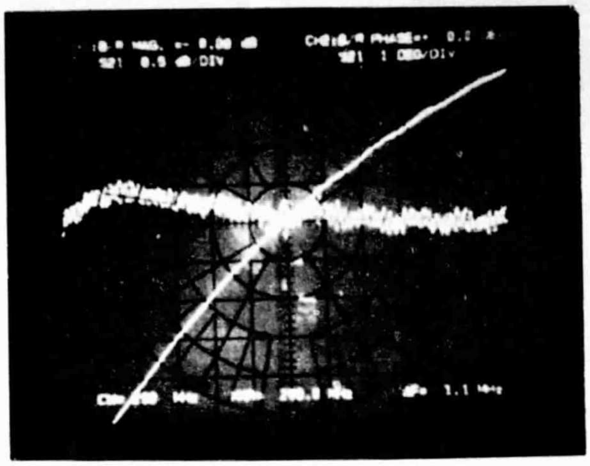
MSR 6A-1

f_o = 298.64 MHz

$\pm \Delta f$ = 1.1 MHz

Vertical Scales - 0.5 dB/Div.

1^o/Div.



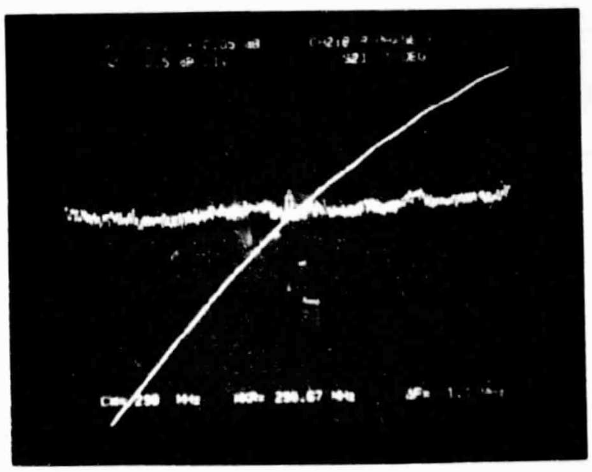
MSR 6A-2

f_o = 298.7 MHz

$\pm \Delta f$ = 1.1 MHz

Vertical Scales - 0.5 dB/Div.

1^o/Div.



MSR 6A-3

f_o = 298.67 MHz

$\pm \Delta f$ = 1.1 MHz

Vertical Scales - 0.5 dB/Div.

1^o/Div.

Figure 4-25. Amplitude and Group Delay Responses of SAW Filters

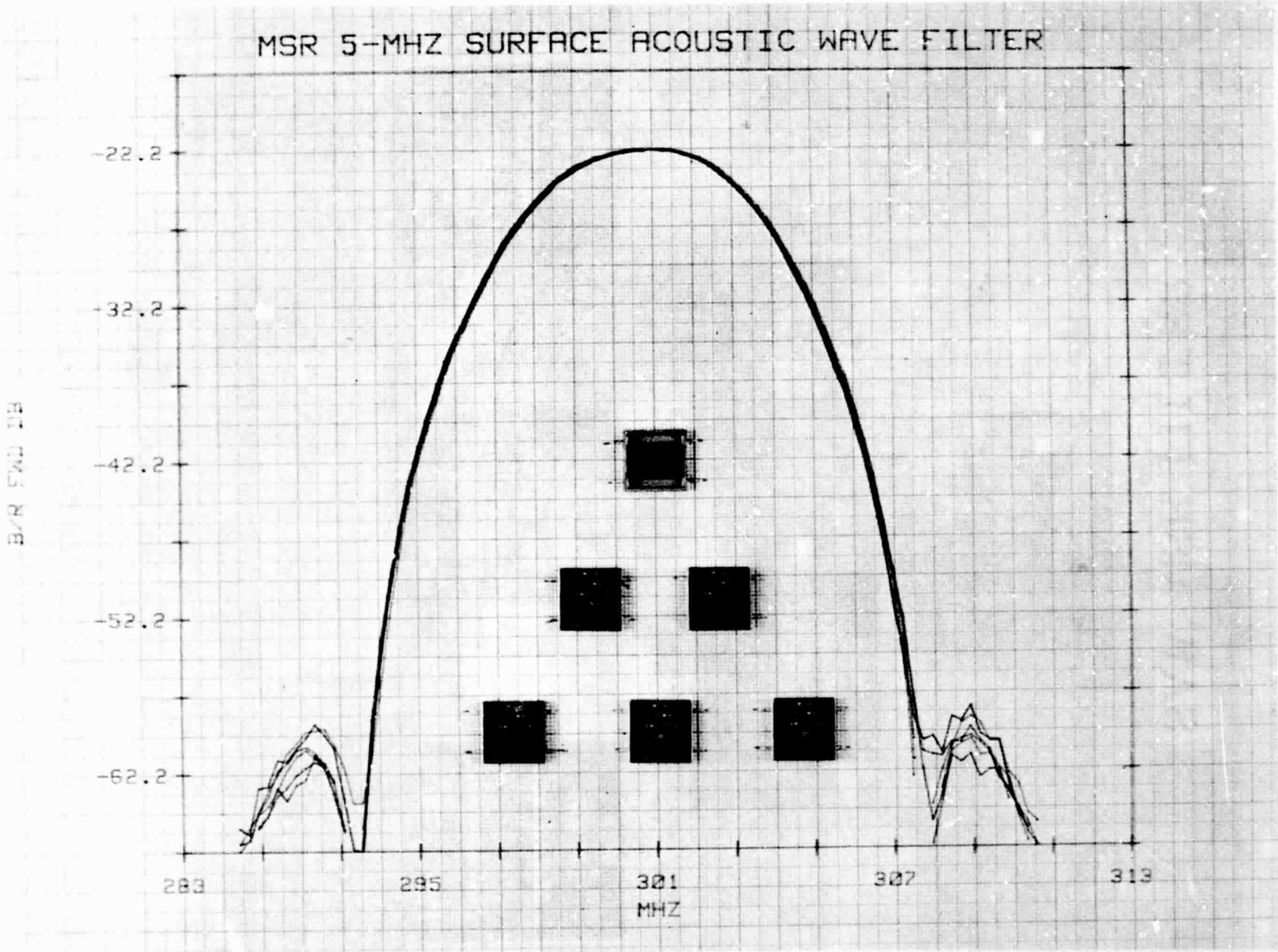


Figure 4-26. Filter Responses of Five "Matched" Filters

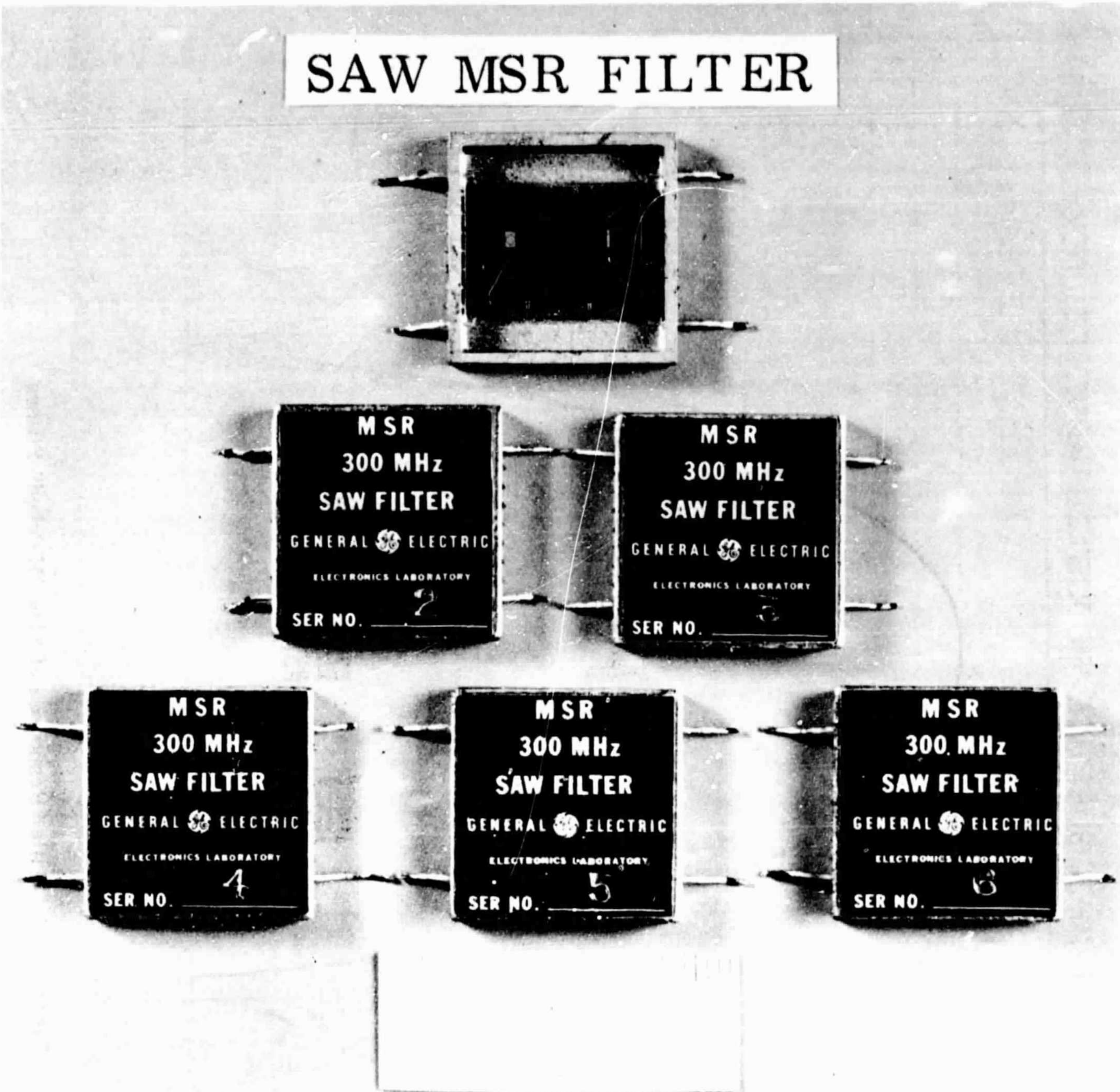


Figure 4-27. View of Six SAW Filters

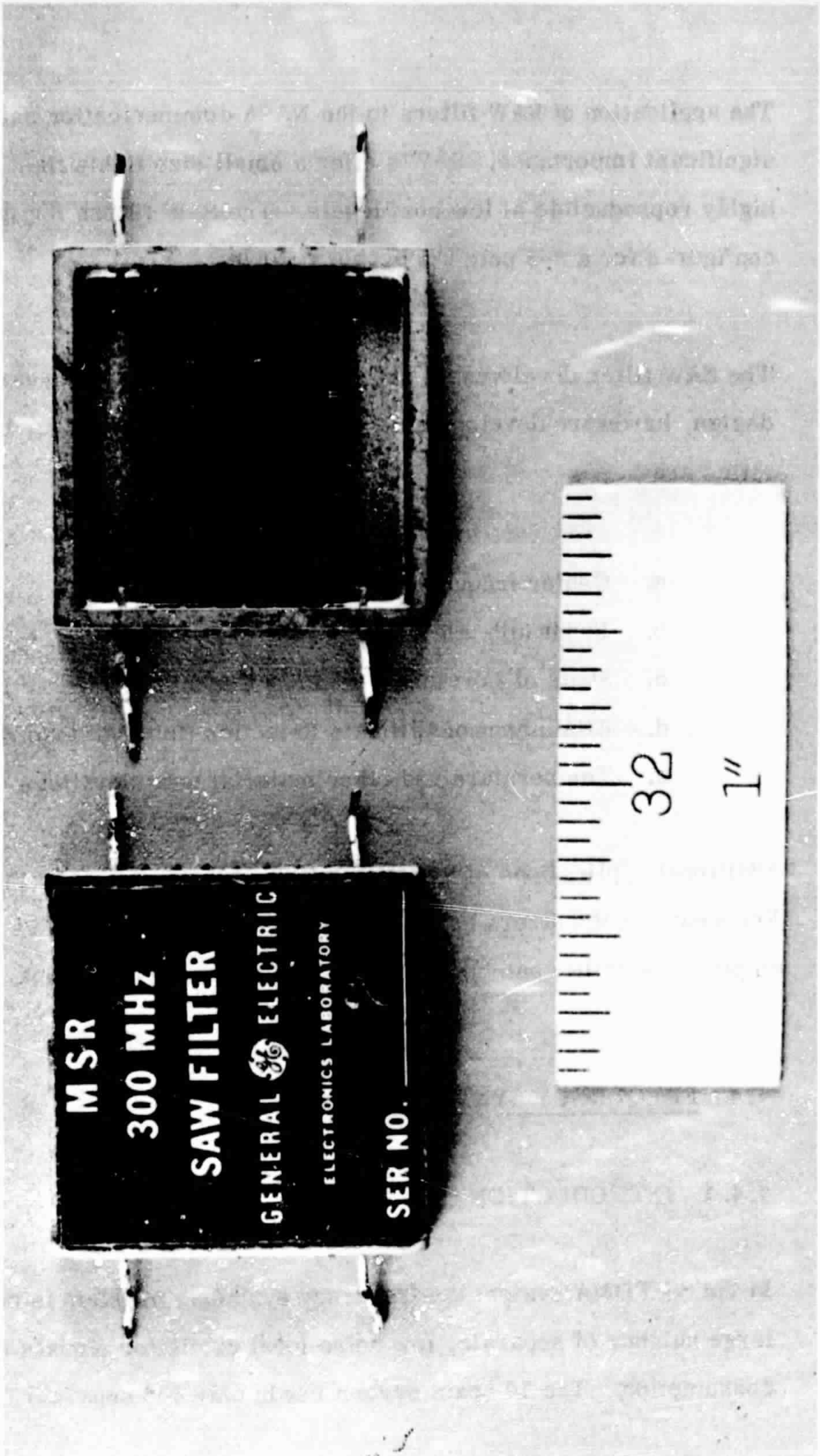


Figure 4-28. Close up of Single SAW Filter

The application of SAW filters to the NASA communication satellite program is of significant importance. SAW's offer a small size lightweight filter that could be highly reproducible at low cost levels. The SAW filters for this program would be configured for a 6-8 pole Chebychef response.

The SAW filter development program (approximately one year in duration) requires design, hardware development (Proof of Concept) and test. Representative characteristics are:

a. Center frequency $\approx$ 200-300 MHz
b. Bandwidth $\approx$ 5 MHz
c. Study of Low Insertion Loss Techniques.
d. Attainment of Ultimate Rejection (for side lobe region)
e. Temperature and other material considerations.

Additional applications of SAW devices can play an important part in the program. Frequency synthesizers using SAW DDL's or SAW oscillators with good phase and amplitude stability can be utilized in the communications link.

4.4 FREQUENCY SYNTHESIS

4.4.1 INTRODUCTION

In the SS-FDMA system the frequency synthesis problem is one of generating a very large number of separate, low noise local oscillator signals with an acceptable power consumption. The 10 beam system needs only 400 separate L. O.'s whereas the 100 beam system needs 2.2×10^4 L. O. 's. Fortunately, many of the frequencies are duplicated and this aids in reducing complexity and in conserving power. However, the ancillary problems associated with phase noise, crosstalk, spurious signal control and the large number of entry connections still presents a challenging engineering design problem. Some effort was devoted to the identification of these major design problems which are discussed in the following sections.

4.4.2 SYNTHESIZER CONFIGURATIONS

The frequency synthesizer may be organized in a variety of ways; a typical arrangement is shown in Figure 4-29. A 1 MHz stable oscillator is used as a reference (its frequencies may be periodically updated via ground command). The reference frequency is multiplied up to the L. O. frequency via a phase locked loop (PLL). One such loop is used for each output frequency which is then routed to the individual mixer L. O. drive ports requiring that frequency. The multiplying factor is determined by a programmable "divide by N" counter in the loop feedback path. In this case, the number N is hardwired, but in some cases it is useful to remotely program the counter by means of a digital number set into the counter. The range of N selected is for the second downconversion stage of the 100 beam system.

This approach has some useful features. The same basic design is used for all L. O. 's; and PLL's are hard wired to the proper output frequency. VCO's are efficient sources of R. F. power. The PLL may be designed to have a narrow loop bandwidth which reduce phase noise and spurious signals. LSI techniques also may be employed - a monolithic unit is presently available covering this frequency range in DIP form (0.29 x 0.75 x 0.2 inches).

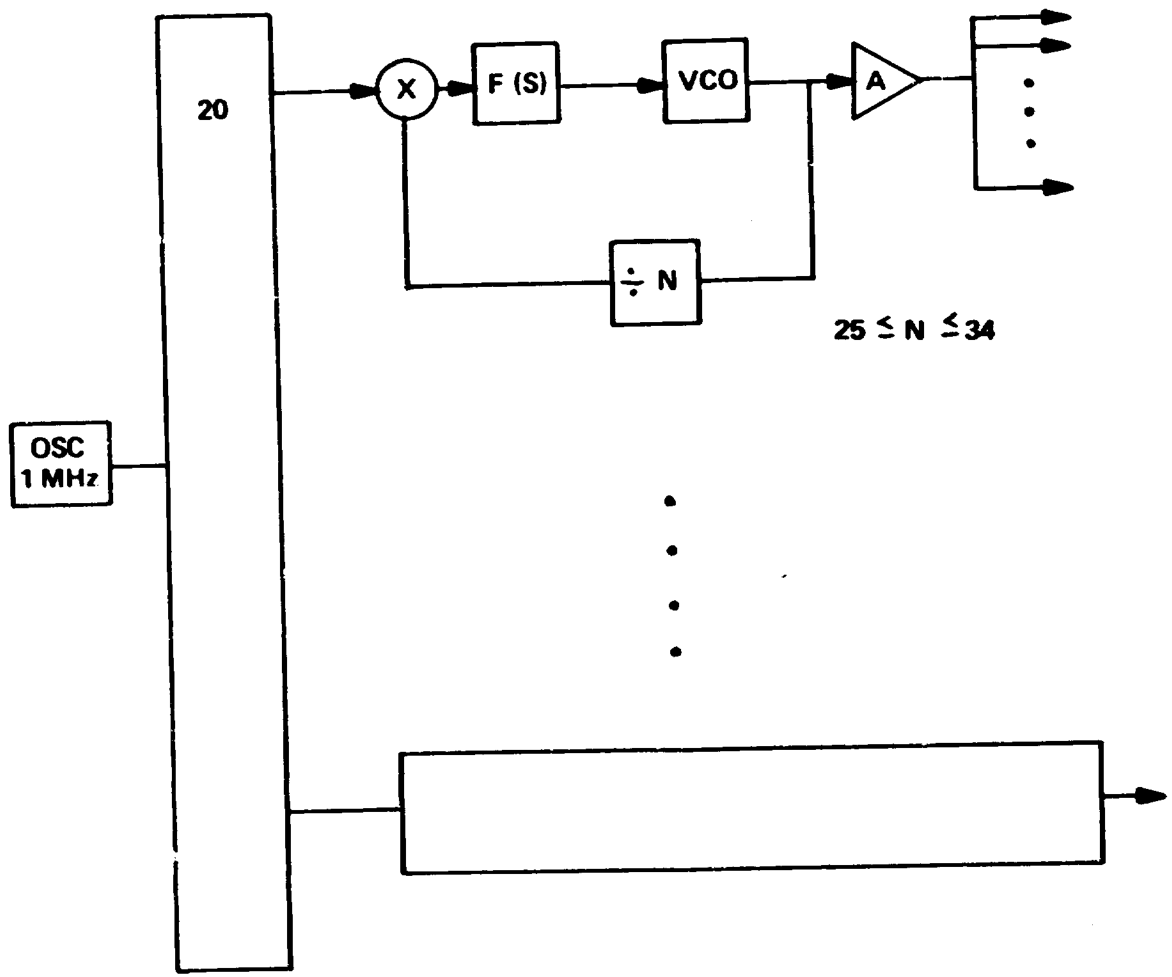


Figure 4-29. Programmed Frequency Synthesizer

An alternative synthesizer configuration routes the reference source to a comb generator. The composite comb is then mixed up in frequency to the range desired. The desired frequencies are selected via narrow band filters.

The above two organizations should be compared on the basis of performance, reliability, size, weight, power, etc. Other approaches are possible - indeed a hybrid of the two may have advantages.

The estimated synthesizer power consumption appears to be significant. This is primarily due to mixer power requirements. Conventional diode double balanced mixers are assumed and these require from 5 to 10 dBm of L. O. power per mixer. A mixer need not necessarily require such a high power level. A low level mixer operating at signal levels of -30 dBm and local oscillator drive levels of -10 dBm may indeed be feasible using FET's as the multiplying element. Such an approach may well reduce the synthesizer power by an order of magnitude.

Harmonic crystal oscillators with stabilities in excess of 1 part in 10^6 per year are presently available. Periodic updating via the telemetry link should enable one to exceed this order of stability over a ten year period.

Oscillator phase noise is most likely not a problem for the satellite source. Careful design of the synthesizer will minimize signal contamination by the L. O. source. It should be noted that the down conversion/up conversion scheme used in the SS-FDMA system is tolerant of coherent phase noise introduced at both mixers. The phase noise at the first and second mixers tend to cancel.

The above discussion is not intended to minimize the engineering problems associated with so many frequency conversions, which are substantial indeed, but our conclusions are that the frequency conversion system required for SS-FDMA does not require technology development. It is believed that the basic engineering problems (of levels, spurious signals, L.O. rejection, stability, phase noise etc.) are soluable once a specific application is identified which permits detailed design and analysis.

SECTION 5

OPERATIONAL CHARACTERISTICS

SECTION 5

OPERATIONAL CHARACTERISTICS

5.1 INTRODUCTION

The purpose of this section is to describe the salient features of hypothetical high capacity operational Ka-Band satellites that might be operational in the 1990's time frame. The design descriptions, while conceptual in nature illustrate the important aspects of such a system, e.g.:

- Frequency reuse via a multiple beam antenna
- Flexible operation with simple, low cost earth stations
- Operation with single antenna earth stations having high availability
- Ka-Band equipment performance characteristics (satellite and earth stations)
- Satellite mass, power, and capacity
- Ramifications of DAMA
- Example switch configuration based on hypothetical traffic matrices

This information provides guidance for the technology recommendations discussed in Section 6, identifies important economic "drivers" and also provides assurance that the SS-FDMA concept is economically sound when applied to a representative situation. In this regard satellite and earth station costs are developed so that representative user costs can be obtained. 1980 dollars are used throughout.

5.2 EARTH STATION CHARACTERISTICS

USER DEFINITION

An important goal in the design of a direct user system is an earth station and satellite cost that is favorable at relatively low data rates. Almost all data communications in the U.S. are handled by telephone lines capable of data rates up to 4.8 Kbps unequalized, and up to 9.6 Kbps with equalization. As described in Section 2, compared with existing terrestrial facilities a direct access system capability can be characterized by its "break even" data rate and (route) mileage, e.g., at higher data rates, or longer (route) mileage than the breakeven the direct access satellite system will be economically favored. The lower the breakeven data rate and mileage the broader the customer base will be. It will be seen that this breakeven characteristic is dominated by earth station cost for the direct user system. Having good breakeven characteristics does not automatically assure that the satellite system will capture that market, other factors also are important such as satellite delay (echo suppressors for satellite use have proven a troublesome maintenance problem and not all data transmission protocols are compatible with the long satellite delays), the mix of routes (long, short and of different bandwidths), and historical prejudices - some customers will just prefer terrestrial facilities. The lowest earth station cost can be achieved by the simplest earth station, for example one containing a single MODEM and a low power solid state HPA. It is expected that such an earth station can have wide applicability. However, some customers require a variety of services - several voice lines and several data lines - and these customers can be satisfied by a more sophisticated earth station containing several MODEMS, and a high power tube type HPA. This customer can afford a more sophisticated terminal because he has higher capacity requirements. Viewed another way the cost per MODEM in this case is less because the pro rata cost of the earth station common equipment, antennas, receiver, HPA, signalling and switching, power supplies, etc., are shared among the many MODEM channels, e.g., his cost per channel is less. To illustrate the range of representative services two earth station concepts are used. These are identified in the following paragraphs.

5.2.1 SINGLE MODEM EARTH STATION

A single MODEM earth station is depicted in Figure 5-1, consisting of a single, small antenna, a single-thread receiver chain consisting of a bandpass filter, low noise receiver or LNR(possibly a GaAsFet type), and a down converter; a single thread transmitter chain consisting of an upconverter, HPA (possibly a GaAsFet or IMPATT type), an output attenuator to control the radiated power and a filter to control harmonics and out of band spurious radiation. An orthomode transducer connects the transmit and receive chains to the antenna. The MODEM is DAMA operated, e.g., its carrier frequency is provided by a frequency synthesizer and the MODEM is used either for communication or for signalling. When the earth station is on-hook (not in service) the MODEM is operated in a TDMA mode on the common signalling channel frequency - dedicated to this purpose. In this mode earth station equipment status is periodically transmitted back to Network Control and earth station commands

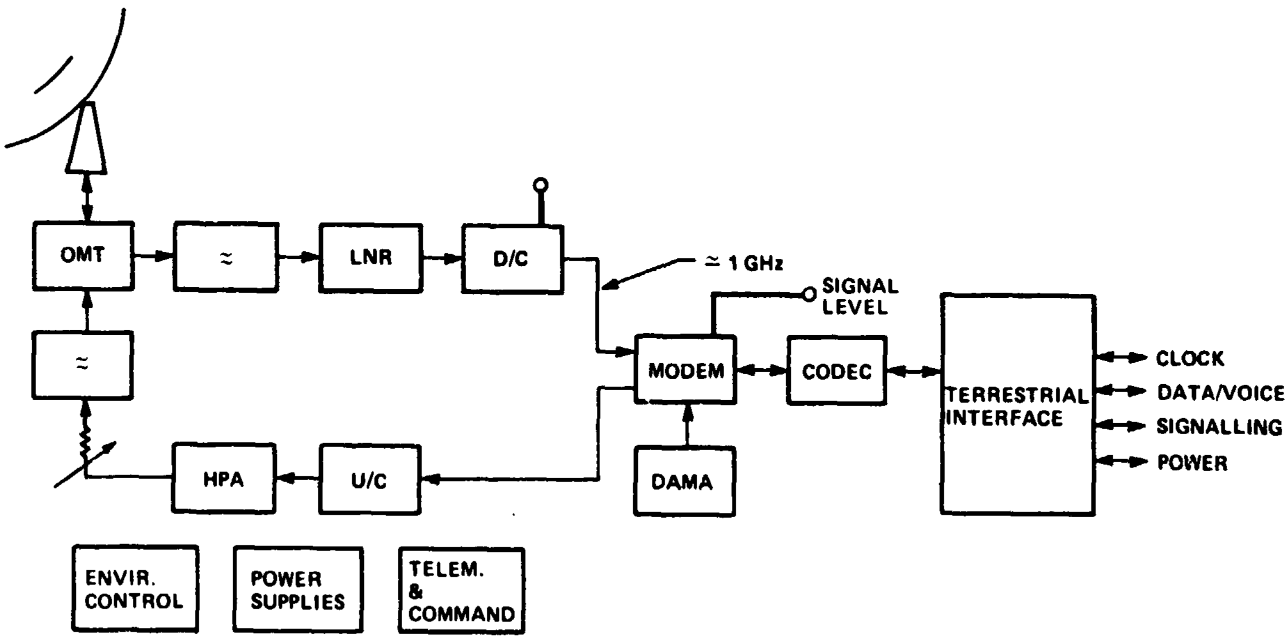


Figure 5-1. Single Modem Earth Station 32 kbPS 4Ø CPSK AVD

(including commands to change radiation power to overcome precipitation attenuation affects) and signalling commands can be received. In addition the earth station monitors the CSC received carrier level as an indication of precipitation attenuation. A reduction in level is telemetered back to Network Control as an indication of precipitation attenuation. This implies that if severe precipitation attenuation occurs during a conversational mode, (when the single MODEM is in the off-hook condition), the earth station cannot take any action to compensate for the fade. If this should happen the station can return to on-hook re-establishing the CSC and enabling the proper actions to take place. The call can then be re-established after a few seconds and can proceed normally from this point.

The MODEM operates at 32 Kbps and in the off hook condition can transmit either voice or data, the latter can be multiplexed data from a variety of terminal equipments and can involve multiple destinations. A delta-modulated CODEC is used for voice encoding and a convolutional coder can be used to reduce errors in data. Therefore, while only a single MODEM is available, several different services can be provided, some on an alternate voice or data, or AVD mode. Low costs are not achievable unless the component parts are simple and there can exist a standardized design capable of low cost mass production and test, which requires only a simple installation procedure. With regard to simplicity the FDMA type station results in the simplest components. The HPA is low powered, possibly solid state, the antenna is small and the MODEM/CODEC and signalling system relatively unsophisticated and low speed. Even so, a maximum of microwave integrated circuitry (MIC) must be used to reduce the cost of these high frequency items, and a maximum of large scale integration (LSI) must be used to reduce the cost of the low frequency items. This requires a standard earth station that has wide applicability. The small antenna and factory tested subsystems should facilitate rapid installation and service inauguration and simple repair procedures. These items are summarized below:

- Relatively simple FDMA Design
- Standard Configuration with Wide Applicability
- Mass production and Test
- Maximum use of MIC, LSI
- Centralized DAMA (e.g., minimal processing at earth station)
- EIRP, Frequency are network controlled
- Simple small modules (and antenna) for rapid installation, service inauguration and maintenance)

While this study did not consider the technical design problem of the low cost earth station there is ample evidence that an earth station can be developed which can achieve a cost in the range stiplulated.

5.2.2 MULTI MODEM EARTH STATION

Many users will have applications for earth stations which are more sophisticated, consisting of serveral or many MODEMs with carriers having different destinations. An example of such an earth station is given in Figure 5-2, which for brevity omits the microwave portion, containing ten 32 Kbps (AVD) MODEMS and two 64 Kbps data MODEMS. In this situation, it is likely that a 32 Kbps MODEM will be dedicated to signalling. In addition to the MODEM cost the earth station might have a larger antenna and HPA (including a power factor for linearization), or both and may employ redundancy particularly in the microwave portion. There are a myriad of combinations of MODEMS, redundancy and other equipment arrangements to suit more sophisticated users; however, it is likely that the number of such terminals is less, that cost is not as critical (this will be proved subsequently) and that efficient use of the satellite is not terribly important, particularly with DAMA type services.

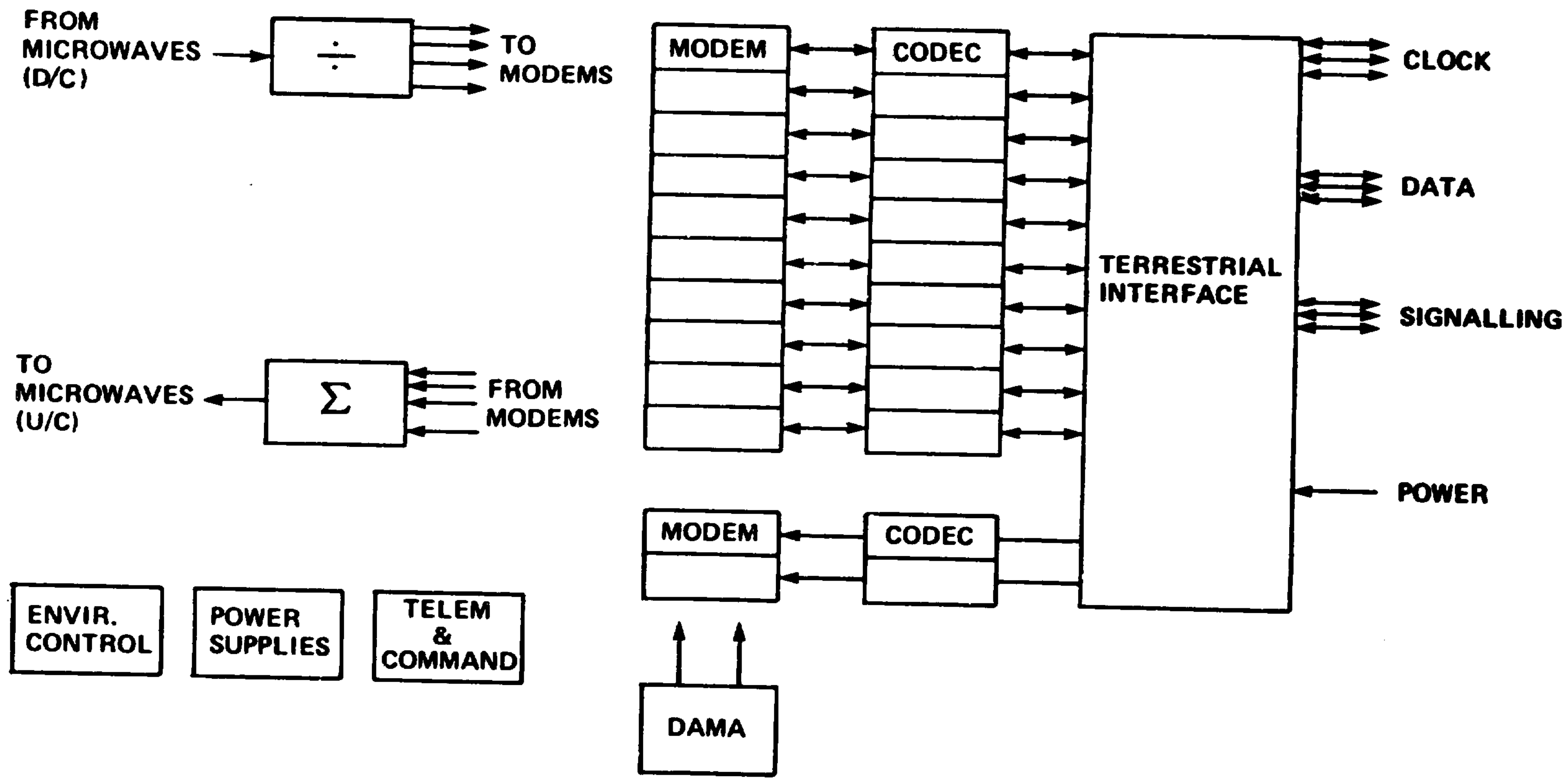


Figure 5-2. Multi-Modem Earth Station

This can still result in substantial amounts of standardization, perhaps several baseline terminals with fixed options - so that mass production techniques discussed previously still can apply.

5.2.3 EARTH STATION PERFORMANCE AND COST

Table 5-1 is a representative link budget based on the use of a 1 meter earth station antenna, a 32 Kbps 4Ø PSK carrier and a satellite antenna beamwidth (in a multiple beam system) of nominally 1.5° e.g., a 10 beam system. The modest output backoff for a nominal 25 db intermodulation signal to noise ratio applies more readily to linear GaAsFet amplifiers and this value is not acceptable for conventional TWT's unless they are linearized. Antenna gain averaging assumes that the SCPC/DAMA system adjusts the earth station HPA levels such that the satellite radiated eirp per given bandwidth is the same regardless of the earth station location in the satellite antenna beam. The earth station system noise temperature is assumed to be a timely state of the art for the hypothetical operational system. Thermal CNR's are adjusted to 25 db because these are the values so indicated in Section 2 for the dynamic carrier power sharing technique, for availabilities approaching 0.9999 in the Mid Atlantic region. Even so, this satellite power is very large, approximately 0.19 watts per 32 Kbps carrier.* If the satellite beam were bandwidth limited and 833 MHz were allocated per beam, then 33,333 preassigned carriers or 13,333 VOX carriers result. Under these circumstances, the transponder saturated power per beam is 0.19 watts/carrier x 13,333 = 2500 watts. This high power is a combination of several factors:

- Operation at Ka-Band with poor performance (e.g., high earth station noise temperature
- Impact of fading
- Impact of interference limited operation (intermodulation) and antenna up-down interference of the order of 25 to 30 db
- Small earth station antennas

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* For 1 meter ES antenna

TABLE 5-1. KA BAND LINK BUDGET SINGLE SERVICE, 32kbPS, 1 METER ES ANTENNA

SINGLE SERVICES, 32KBPS, 1 METER ES ANTENNA			
FREQUENCY GHz	20	30	
POWER, dbw	0	+8.5	
LOSSES, db	-2	-2	
OUTPUT BACKOFF, db	-3.5	0	
ANT GAIN, MINIMUM, dbi	37.5 (1.5° cell)	42.2 (GS)	
ANTENNA GAIN AVERAGING, db	+2	–	
EIRP, dbw	32	40.8	46.2
MISC. LOSSES	-.7	-.7	
ANTENNA GAIN, dbi	43.5 (ES)	39	
NOISE TEMP db > 1°K	24.8 (300)	26.9 (490)← PARAMP	
CNR, CLEAR SKY, (THERM), db	25	25 + 15	
SAT POWER	~.19 WATTS/CH	–	

The dominant affect is due to the small earth station antenna; however, all factors are important. Even if the earth station antenna is 3 meters, the transponder power is still 278 watts. It is clear that a direct access operational Ka-Band satellite system will be high powered.

The uplink also requires good power performance. Even with a good 30 GHz paramp the requirement to provide a thermal noise CNR in excess of 40 db requires an HPA power in excess of seven watts. From both uplink and downlink considerations it appears desirable that the earth station antenna diameter be in the range of two to three meters.

Using the above link as a rough guide the cost of the earth station can be estimated. An example of what can be expected by the standardization and productization described above can be obtained from reference (1) which refers to extensive earth station equipment suppliers surveys and which has compiled the costs of earth station equipment as a function of frequency band (where applicable), production lots and performance. These results are for the period up to 1985 and are based on trends or projections which could be foreseen in the 1975-1976 time frame. While reference (1) does not include Ka-Band equipment it is believed that only small errors are introduced by assuming these costs are similar to comparable Ku-Band equipment (for instance a fixed Ku-Band or Ka-Band antenna of the same size, or similarly, both with tracking). Of course, oscillator frequency stability and phase noise (particularly critical for narrow band systems such as direct access systems) will be worse at Ka-Band, but solutions at Ka-Band appear to be applicable.

A summary of results are tabulated in Table 5-2 in production lots of 100,000* for the telephone/data terminals and in production lots of 1000 for video teleconferencing, including installation. These numbers should be viewed as costs to a second party such as a carrier. These numbers are certainly optimistic by present experience where simple single MODEM earth stations (in small lots), cost between \$70,000 to \$140,000 depending on application. On the other hand the General Electric Company was engaged in a contract to construct similar prototype earth stations for a follow-on production of ten thousand terminals (based on SCPC DAMA and TV receive only capability at Ku-Band). If this design is extrapolated to the single MODEM design case its cost in 1980 dollars falls within the postulated range of \$10,000 to \$30,000. Nevertheless, the achievement of such a low cost earth station will be uncertain unless a specific design can be proven in actual production lots. Consequently, lacking such detailed design and production information a range of earth station costs is used for the purposes of this study. It will be seen

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(1) Communication System Technology Assessment Study, NAS 3-20364

* Learning curve experience was confirmed by reference (1). A typical value used is 95%. This can be used to compute production costs at lower production rates.

TABLE 5-2. COST IN 1980 DOLLARS FOR REPRESENTATIVE EARTH STATION

	Earth Station #1	Earth Station #2	Earth Station #3
	Qty $\simeq$100,000 Single Service 32 Kbps Duplex (AVD) $.9999 < A < .999$	Qty $\simeq$100,000 Multi-service 10 32 Kbps Duplex (AVD) 2 64 Kbps Duplex (AVD)	Qty $\simeq$1000 Single Service TV Teleconferencing 40 MBPS Duplex $.9999 < A < .999$
Antenna Diam, M	2 Meters	2 Meters	6 Meters
HPA Power	$\simeq$1 Watt	$\simeq$60 Watts	$\simeq$400 Watts
Costs (1980 Dollars)*			
HPA	$ 1,500	$ 4,500	$40,000
Antenna	1,100	1,100	15,500
LNR & D/C (300°K)	2,000	2,000	2,000
Up Converter	1,500	1,500	1,500
Modem	1,000	13,000	5,000
DAMA/Signalling	1,000	2,000	1,000
Miscellaneous and Installation	2,100	3,900	25,000
	$10,000	$29,000	$90,000
*Quantity 100,000			
Assumed Cost Range	$10,000-$30,000	$30,000-$50,000	$90,000

subsequently that the direct access satellite system is very attractive if earth station costs at the low end of the scale are achievable and less so if these costs exceed $30,000. In fact, since earth station costs are the prime economic drivers in direct access system (for low data rate short route-mile distances) the problem can be turned around by setting this cost range as an objective in a technology development program to achieve the cost range, considering acquisition costs, as well as O&M costs.

As discussed previously, the cost of the multi MODEM earth station is less important and the subsequent analyses will show this can be substantially above the $30,000 to $50,000 range and still be very attractive. Of course, cost becomes more significant as the number of MODEMS decreases.

5.3 SATELLITE DEFINITION AND CHARACTERISTICS

5.3.1 INTRODUCTION

10 beam and 100 beam, Ka-Band, direct access satellites are defined in this section in order to illustrate broad characteristics applicable over a ten to one range in capacity and frequency reuse. Satellite block diagrams, including antenna pattern topology, switch matrices, channelization, hypothetical traffic matrices and other information previously generated will be used to illustrate the basic design characteristics including in-orbit weight and prime power, as a function of earth station antenna diameter. While not optimized, and while some of the technology (particularly higher powered satellite amplifier and switching network characteristics) lacks the clarity of hardware development, the basic satellite arrangements do shed considerable light on directions to take for at least initial technology development, and do show some striking contrasts in basic configuration when compared with existing satellites. Later, using a satellite cost model, the satellite information will be used to derive user costs.

5.3.2 SATELLITE BLOCK DIAGRAMS

A generalized block diagram of a multiple contiguous beam direct access satellite is depicted in Figure 5-3 for N contiguous beams. Each up link beam feed contains a receiver consisting basically of a band pass filter, low noise receiver, down converter and a divider to route FDMA signals to the two switching networks. These networks contain additional frequency converters, channelization filters and switching networks to provide the proper interconnection of the RF paths or routes. The down link paths destined for a specific beam are summed in that beam, up converted to 20 GHz and amplified by the final amplifier so that only traffic destined for that beam is amplified by that amplifier. The amplifier is operated in its linear region because many FDMA signals (digital or analogue carriers) are present

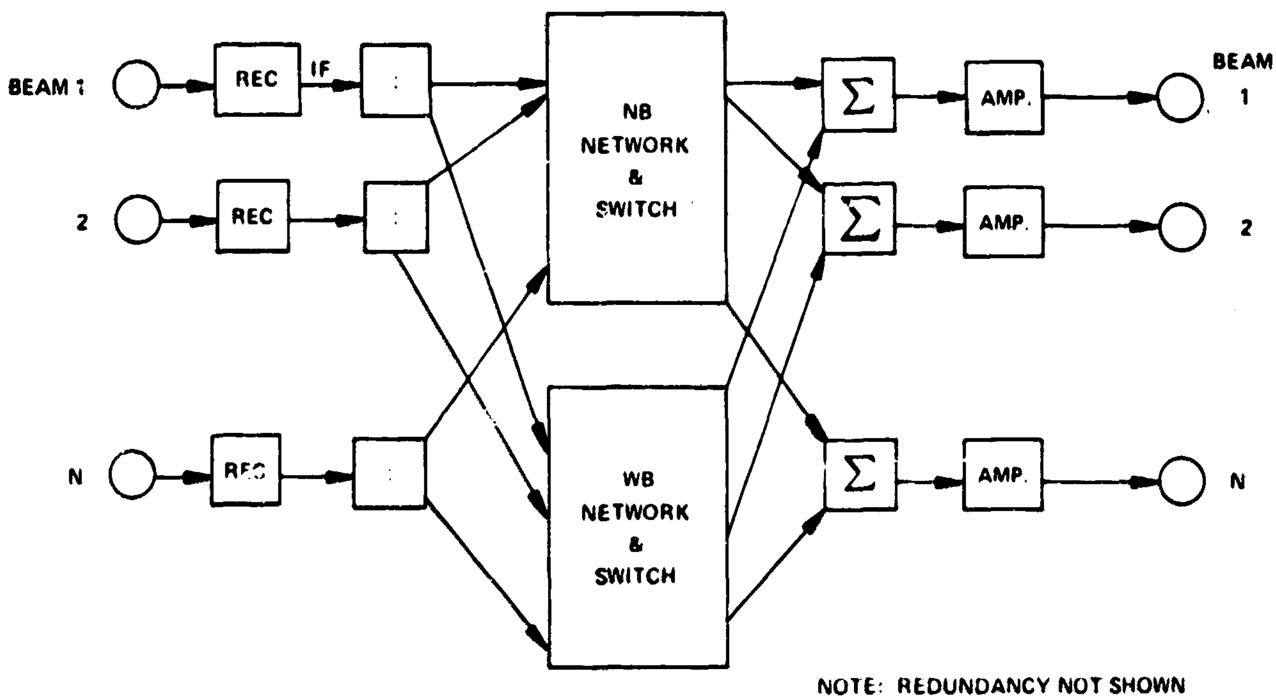

Figure 5-3. Generalized Block Diagram of Multiple Contiguous Beam Direct Access SS-FDMA Satellite.

simultaneously. Redundant elements with concommitant switching are not shown, but are included in the weight estimates. The actual number of switch networks may be more than the two depicted in Figure 5-3, in fact, this is true for both the 10 beam and 100 beam configurations, as will be seen shortly.

5.3.3 TRAFFIC MATRIX AND SWITCH DESIGN

In a multiple beam satellite system it is unlikely that the traffic demand will be the same in each of the beams. For example, at the end of a satellite lifetime, during the busy hour it is a reasonable expectation that a satellite beam pointed at the vicinity New York City will be saturated while one pointing at the Rocky Mountains region, exclusive of Denver will not. Such a disparity must be taken into account in the satellite design. In addition, experience shows that it will not be possible to accurately predict the growth of traffic in each beam and for the interconnecting routes. Particularly in lightly loaded beams, unexpected changes can be severe. For example, a customer might be acquired requiring 40 MbPs for interactive television teleconferencing; this customer might appear in any beam. Consequently flexibility is important and flexibility seriously impacts the satellite switching arrangement.

Lacking a specific customer model, a first order typical traffic matrix can be derived by estimating the population density within each of the multiple beams since population density generally emulates commercial, industrial and government activity. A typical contiguous beam plan is given in Figure 5-4 and the approximate percentage population distribution is given in Table 5-3.

For simplicity the Plan 4 topology based on triads is assumed; each of the ten beams is composed of 7 component beams to achieve increased antenna resolution, necessary to achieve minimum antenna isolation of approximately 25 db.* In this scheme the frequency reuse factor is one third e.g., one third of the total bandwidth allocation

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* See Table 3-6 for results of antenna analyses.

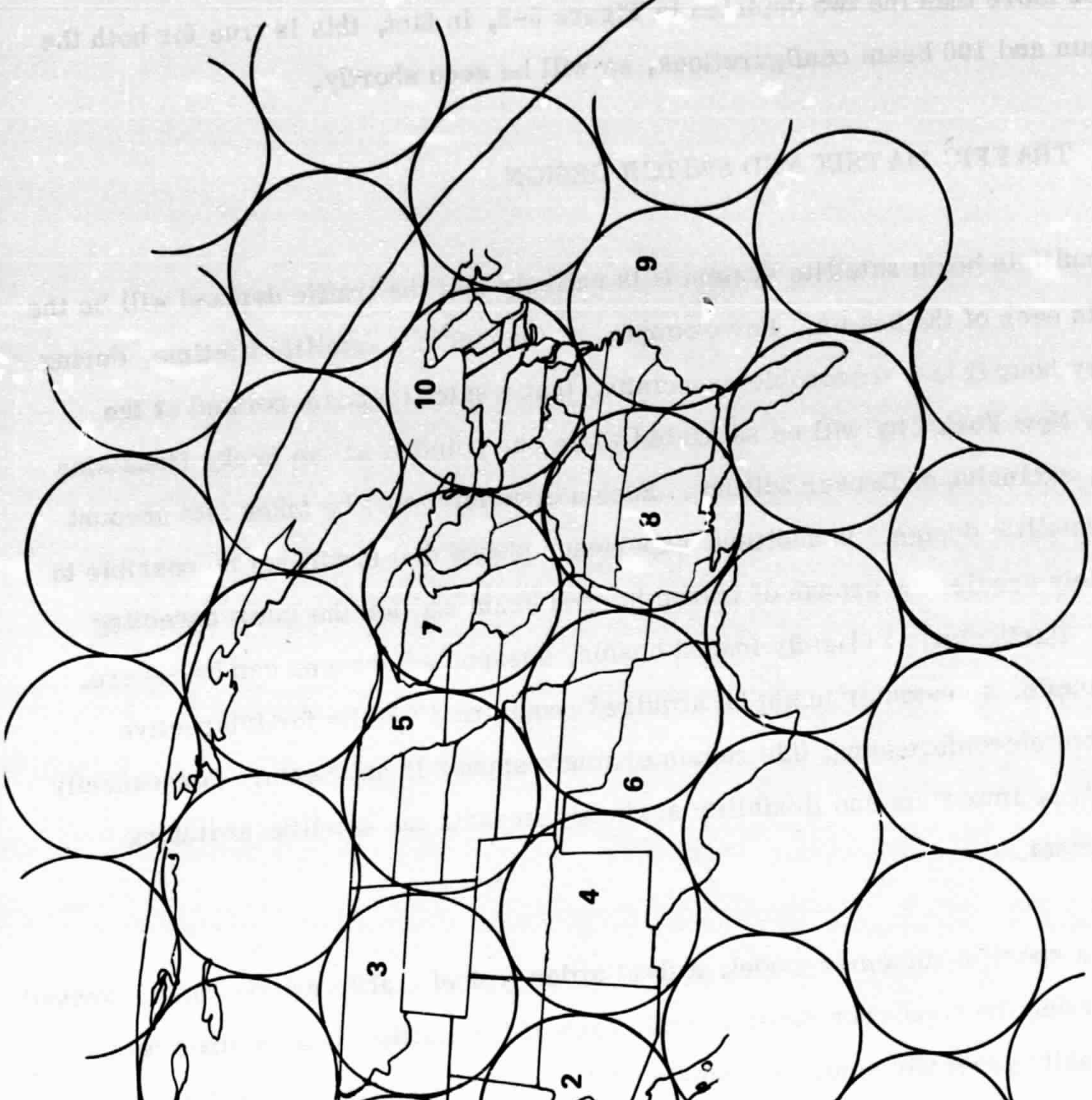

Figure 5-4. Example of Typical 10 Beam Coverage Plan

TABLE 5-3. APPROXIMATE POPULATION DISTRIBUTION IN THE BEAM PLAN OF FIGURE 5-4.

Beam	Description	Population	%
1	Northwest (Seattle)	5,084M	2.70
2	Southwest (Calif)	21,407	11.20
3	Mountain Area	2,021	1.06
4	West Central (Denver)	8,295	4.34
5	North Mid-West (Minnesota)	7,567	3.96
6	Mid Southwest (Texas)	11,340	5.91
7	Midwest	43,797	22.92
8	South	34,486	18.05
9	Southeast	12,370	6.47
10	Northeast	44,700	23.39
		1,067,000	100%

is available in any one beam. For an assumed Ka-Band allocation of 2500 MHz then 833 MHz is available in any one beam. The distribution of bandwidth need not be uniform and the coverage beams themselves need not all be the same size. For example some of the lighter capacity beams can be coupled together to enlarge the coverage area and thus simplify the switching. The population distribution can be converted to MHz, recognizing that one beam (say the one covering New York City and its vicinity) requires 833 MHz at end of life during the busy hour.

Thus a traffic matrix can be constructed for the 10 beam satellite system, illustrated in Table 5-4. Note that some of the beams are heavily loaded; while others are lightly loaded. The relative capacity covers a spread of 21:1. Also traffic returning to the same beam is assumed, e.g., break even distance are assumed to be short enough to encourage this use. The table of numbers also

TABLE 5-4. TEN BEAM TRAFFIC MATRIX IN MHz

BEAM	1	2	3	4	5	6	7	8	9	10	
1	196	192	150	92.5	53.4	49.8	35.6	32.0	21.4	10.7	833.4
2	192	185	146	89.0	53.4	49.8	35.6	32.0	21.4	10.7	814.9
3	150	146	113.9	74.8	42.7	39.2	28.5	24.9	14.2	7.1	641.3
4	93.6	89.0	74.8	46.3	24.9	24.9	17.8	14.2	10.7	3.6	398.8
5	53.4	53.4	42.7	24.9	14.2	10.7	10.7	10.7	7.1	3.6	231.4
6	49.8	49.8	39.2	24.9	10.7	10.7	7.1	7.1	7.1	3.6	210.0
7	35.6	35.6	28.5	17.8	10.7	7.1	7.1	7.1	3.6	0	153.1
8	32	32	24.9	14.2	10.7	7.1	7.1	7.1	3.6	0	138.7
9	21.4	21.4	14.2	10.7	7.1	7.1	3.6	3.6	3.6	0	92.7
10	10.7	10.7	7.1	3.6	3.6	3.6	0	0	0	0	39.3
											3553.6 MHz or 1.42 reuses

shows the total bandwidth of each of the N^2 routes - in this case 100 routes all told Many of these at the upper left of the Figure are quite large, and many others, at the lower right of the Figure are quite light, (in fact, some are zero). In a multiple beam system routes with no demand are quite common. Note also that the total bandwidth generated by the system is 3553.6 MHz or 1.42 frequency reuses, caused by the non-uniform loading instead of 8333 MHz achievable if the satellite beams were uniformly loaded. The satellite switch/routing scheme must accommodate this matrix but also must accommodate substantial changes from it. This can be accomplished by considering three types of arrangements.

1. Preassignment of portions of the heavy routes.
2. Providing fully switched paths among the beams, e.g., N X N switching or "crosspoint" switching.
3. Providing extra capability from a common "pool": in this case the designation "pool" is misleading - flexibility is being provided but at reduced switch complexity.

It is convenient in this traffic matrix to consider two route (or path) sizes, one 36 MHz and the other 10 MHz. What results is a number of switch networks similar to those shown in Table 5-5.

In this scheme, which is by no means optimized neither for maximum flexibility nor minimum numbers of switches, the routing and flexibility requirements are satisfied by five networks. For beam 1, 10 preassigned, non switched 36 MHz paths, and 8 fully switched 36 MHz paths connect the heavier routes of the system. In addition, 8 fully switched 10 MHz paths also are provided. Also an additional 4 10 MHz paths can be added from a "pool." However, no more than 20 10 MHz paths and no more than 10 36 MHz paths are available in total; this reduces the size of the switch matrices. Beam 10 on the other hand has no preassigned or fully switched 36 MHz paths, but only 6 switched 10 MHz paths and the possibility for as many as 8 more 10 MHz paths and 4 more 36 MHz paths from the pool. Thus while the traffic matrix for beam 10 identifies a need for only 39.3 MHz of total bandwidth, as many as 18 additional paths with a total

TABLE 5-5. EXAMPLE OF A SWITCH MATRICES CAPABLE OF SATISFYING THE REQUIREMENT OF THE TRAFFIC MATRIX OF TABLE 5-4.

		36 MHz		10 MHz		10 MHz	36 MHz
		PREASS	SWITCHED	PREASS	SWITCHED	POOLED	POOLED
BEAM	1	10	8		8	4	0
	2	10	8		8	4	0
	3	9	7		10	4	2
	4	4	4		14	4	2
	5	2	2		8	6	2
	6	1	2		11	6	2
	7	1	1		10	6	4
	8	1	1		10	6	4
	9	0	0		10	8	4
	10	0	0		6	8	4
SWITCH MATRIX			33x33		95x95	56x20x56	24x10x24
NUMBER OF SWITCHES		1089		9025		2240	480

bandwidth of 284 MHz could be available should changing conditions warrant. While significantly more bandwidth cannot be added to Beam 1 the 21 36 MHz and 10 MHz switched paths permit a reasonably good distribution of bandwidth among its various routes. The switch matrices and total numbers of switches are listed at the bottom of Table 5-5. It is evident that switching must be accomplished at a low enough RF frequency to facilitate such large networks with adequate performance and acceptable sizes, weights and powers. Figure 5-5 depicts the resulting spectrum allocation by route bandwidth for several beams.

Beam 1, by definition is filled. Beam 10 on the other hand is not filled. Its required traffic is satisfied by the 6 to 10 MHz routes indicated with additional routes available from the pool. Typical arrangements for Beam 4, 6 and 8 also are shown. The arrangements shown are only unoptimized examples for illustration; the final frequency/path arrangements must be chosen with great care. For example, the Figure 5-5 frequency allocation can be rearranged to avoid or minimize a frequency interference between light traffic beams. The resulting 10 beam switch concept is shown in Figure 5-6.

Alternately a 100 beam system also is examined having ten times more frequency reuse than the ten beam design, and hence ten times the capacity. Of course the antenna and switching systems become very complex. A new traffic matrix can be developed, as before, from population density data. However, it is more convenient to extrapolate the hundred beam matrix from the ten beam matrix by allowing the first ten highest capacity beams in the hundred beam matrix to have the same capacity as the first beam in the ten beam matrix. Then the next ten in the hundred beam matrix (beams 11 to 20) will have the same capacity as the second beam in the ten beam matrix, etc. The result is depicted in Figure 5-7 in which the first entry may be interpreted "any beam from 1 to 10 has 19.6 MHz of traffic to each of the first 10 beams." The 100 beam matrix shows that the individual routes are now substantially

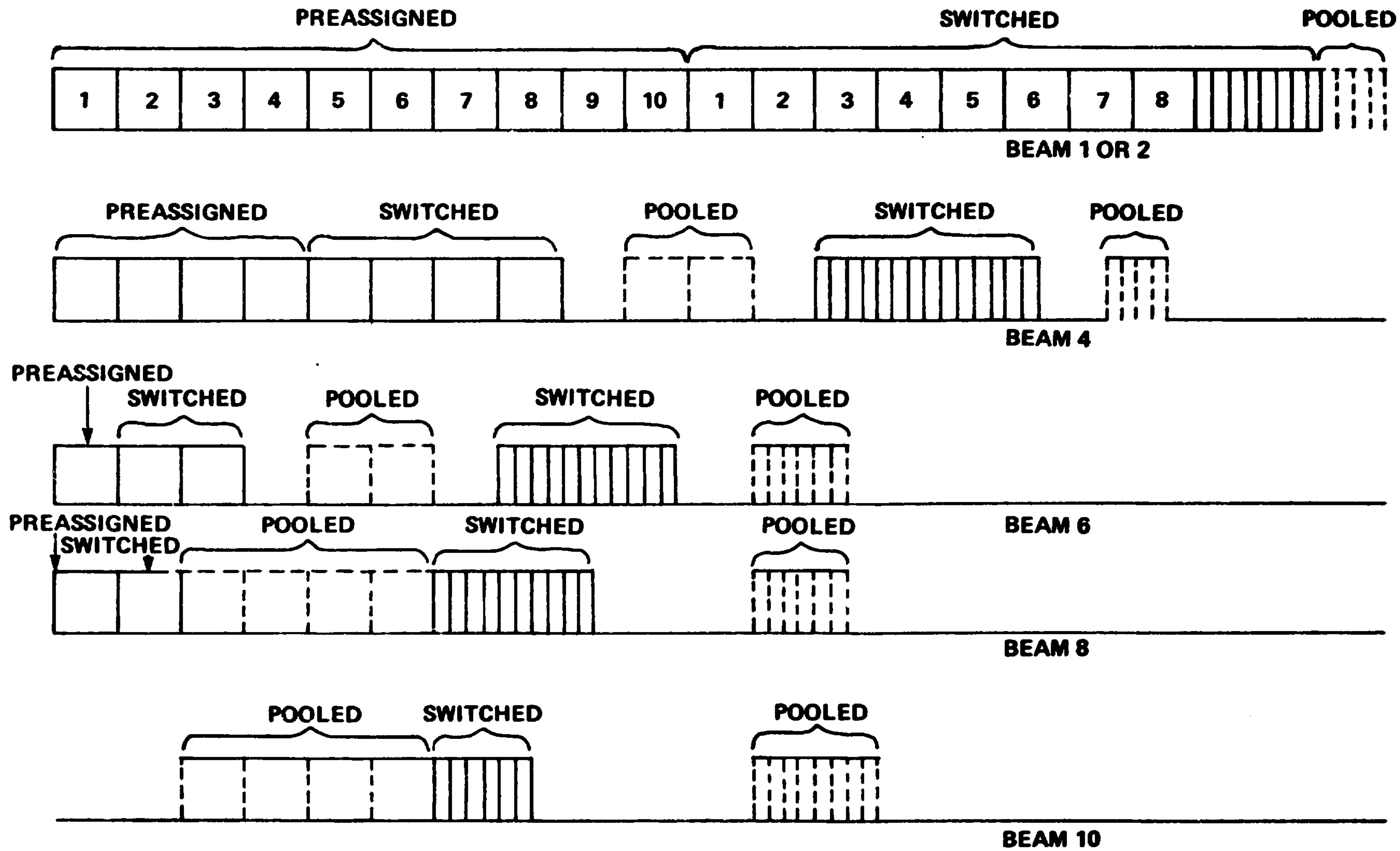

Figure 5-5. 10 Beam Satellite Channel Allocation

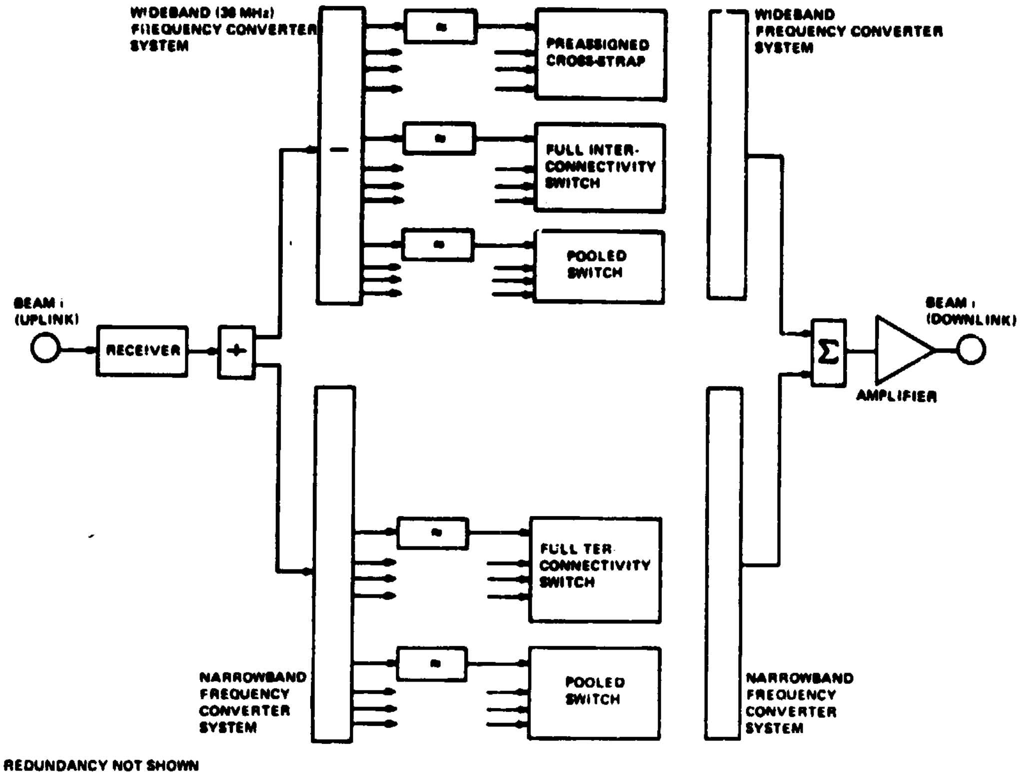

Figure 5-6. 10 Beam Switch Network - Transponder Configuration

reduced in bandwidth - the total bandwidth is increased 10 times but the number of routes have increased from 10^2 to 100^2. Even the heaviest route is only 19.6 MHz, e.g., the 833.3 MHz available for beam 1 must be distributed over 100 beams, and the thinner routes in the lower right of Figure 5-6 are of the order of 1 MHz. Consequently there appears to be a need for three path or route bandwidths, 36 MHz routes (occasional service), 10 MHz for "heavy routes" and 2 MHz for light routes. The different bandwidths are selected in order to simplify the switch matrices, it is important that each route have adequate bandwidth but it matters little whether there is an excess of bandwidth. The total usable bandwidth is 35536 MHz or 14.2 reuses of the 2500 MHz Ka-Band, which is a prodigious system capacity.

Resulting switch matrices (five) are listed in Table 5-6. While the traffic matrix identified no need for 36 MHz paths there is a possibility that such routes may be

TABLE 5-6. SWITCH PLAN, 100 BEAM SATELLITE

SWITCH PLAN, 100 BEAM SATELLITE

	WIDE BAND 36 MHz		10 MHz		2 MHz	
	FULLY SWITCHED	POOLED	FULLY SWITCHED	POOLED	FULLY SWITCHED	POOLED
1-10		4*	65**	30	60	20
11-20	N	4	65	30	60	20
21-30	O	4	60	30	70	20
31-40	N	4	30	30	50	20
41-50	E	4		30	50	20
51-60		4		30	30	40
61-70		4		30	20	50
71-80		4		20	30	50
81-90		4		20	40	60
91-100		4 (160)		20	20	80
TOTAL POOLED CHANNELS	N/A	400(1)	N/A	2700	N/A	7600(2)
NUMBER SWITCHES IN MATRIX	0	32,000	2200^2 = 4.84M	2.7M	4300^2 18.5M	7.6M

* EACH OF TEN BEAMS HAS ACCESS TO FOUR 36 MHz ROUTES
** EACH OF TEN BEAMS HAS ACCESS TO 65 10 MHz ROUTES

(1) 40 TOTAL
(2) 1000 TOTAL

BEAM	1-10	11-20	21-30	31-40	41-50	51-60	61-70	71-80	81-90	91-100	TOTAL
1-10	19.6	19.2	15	9.3	5.3	5.0	3.6	3.2	2.1	1.1	8340
11-20	19.2	18.5	14.6	8.9	5.3	5.0	3.6	3.2	2.1	1.1	
21-30	15.0	14.6	11.4	7.5	4.3	3.9	2.9	2.5	1.4	.7	
31-40	9.3	8.9	7.5	4.6	2.5	2.5	1.8	1.4	1.1	.4	
41-50	5.3	5.3	4.3	2.5	1.4	1.1	1.1	1.1	.7	.4	
51-60	5.0	5.0	3.9	2.5	1.1	1.1	.7	.7	.7	.4	
61-70	3.6	3.6	2.9	1.8	1.1	.7	.7	.7	3.6	0	
71-80	3.2	3.2	2.5	1.4	1.1	.7	.7	.7	3.6	0	
81-90	2.1	2.1	1.4	1.1	.7	.7	.4	.4	.4	0	
91-100	1.1	1.1	.7	.4	.4	.4	0	0	0	0	410
											35536 MHz or 14.2 reuses

Figure 5-7. 100 Beam Satellite Matrix (MHz)

required for interactive TV teleconferencing. As many as 40 simultaneous 36 MHz routes can be provided but no more than 4 per beam. 10 MHz fully switched paths amongest the 40 heavy route beams take care of the heavy route traffic. Additional 10 MHz routes are available from a pool of 500, with no more than 30 permitted per beam. The light route traffic is provided by a fully switched 2 MHz route network plus an additional 1000 routes from a pool. The 2 MHz route switch matrices are imposing challenges and perhaps a more judicious choice of route bandwidths, or use of more sophisticated switching techniques can simplify these requirements. On the other hand, these low frequency switches are amendable to large scale integration. With approximately 1000 switches per 1/4" x 1/4" chip, a million switches can be provided in approximately six cubic inches. The switch weight or volume is not the problem. Section 4 described the problem of packaging and power dissipation; these are more fundamental problems than the switch density on substrates.

The concepts for the 10 beam and 100 beam cases are deliberately chosen to be straight forward applications of the use of multiple beam antennas and on-board FDMA switching. Thus, each beam is uniform in size and each uplink beam contains a receiver and each down link beam is driven by an amplifier. The number of receivers and transmitters can be reduced in multiple beam systems with frequency reuse factors of 1/n. If the factor is 3 for example three uplink beams covering three different frequency bands can be multiplexed together and the composite signal can drive a single receiver. Correspondingly, a single amplifier can amplify signals covering the entire allocated spectrum and after amplification a multiplexer can direct the proper band to the proper beam. This is illustrated in Figure 5-8.

Of course, if the transmitters are already high powered it may not be desirable to combine them into one. Multiplexer losses at Ka-Band also may not be acceptable. It should be noted that the form of beam combination does not reduce the access problem from the individual geographic areas defined by the contiguous beams and hence does not simplify the on-board switch design.

Figure 5-9 is an example of a 10 beam plan and required bandwidth per beam e.g., a 10 beam traffic matrix, with the beam frequency plan shown circled, e.g., 1, 2 and 3. That is, beams 2, 5, 6 and 10 are co-frequency beams. Since the required bandwidth or traffic per beam is different, beams of different frequency bands can be combined in order to balance the bandwidth requirements for the composite.

For example beams 1, 4 and 10, beams 5, 6, 7 and 9, and beams 2, 3 and 8 can be multiplexed together to form three composite, non-interfering frequency bands.

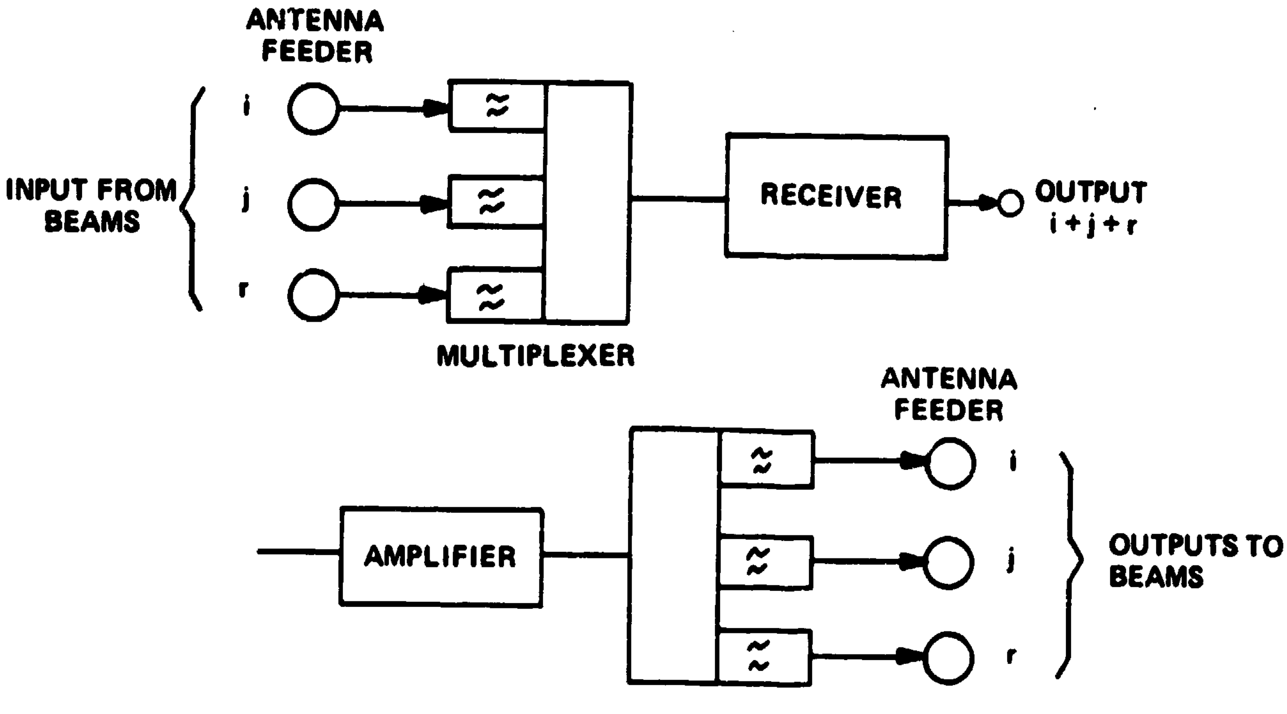

Figure 5-8. Example of Multiplexing to Reduce the Number of Receivers and Transmitters

Figure 5-10a depicts the results, and shows a reformed "traffic matrix" indicating that a considerable balance has been achieved by multiplexing. Figure 5-10b shows an example of a frequency/beam plan. Not only can the bandwidths be balanced but the frequency bands can be located such that frequency beam interference is reduced taking advantage of the fact that the bandwidths of the individual beams are not filled.

This example illustrated the complexities of frequency planning and what can be accomplished by:

- beam combining to balance traffic
- minimization of interbeam interference
- distribution of effects of precipitation attenuation*
- multiplexing - frequency bands can be separated to ease multiplexing problems

In order ot reduce switch complexity the number of geographical areas requiring interconnection must be reduced. Telecommunication activity is very intense in the northeast (Boston - Chicage - Atlanta triangle), and somewhat less so for the far west (LA - San Francisco). Spot beams covering other areas have less traffic demand so that advantage can be taken to combine the lighter traffic beams into "super" beams or shaped beams of wider area coverage. Figure 5-11 is an example; spot beams have been retained for the north east and mid west (Chicago, St. Louis, Detroit area), and the far west, however, the U.S. south including Atlanta is now served by a "super" beam composed of three spot beams and the west and north west are served by a super beam composed of four spot beams. In terms of access, a 10 beam system with 100 potential routes has been reduced to a 5 beam system with only 25 routes. The power required in each of the beams is proportional to both the traffic covered and the coverage area and must be adjusted accordingly.

The arrangement of beams, frequency plans, routing plans and switching plans are intimately tied to specific network characteristics and are complex in their myriad forms.

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* In the example, beams 9 and 10 contain traffic in Florida and the Gulf of Mexico region which experience severe precipitation attenuation fading: these two beams are arranged so that the extra power needs for this region is provided in different amplifiers.

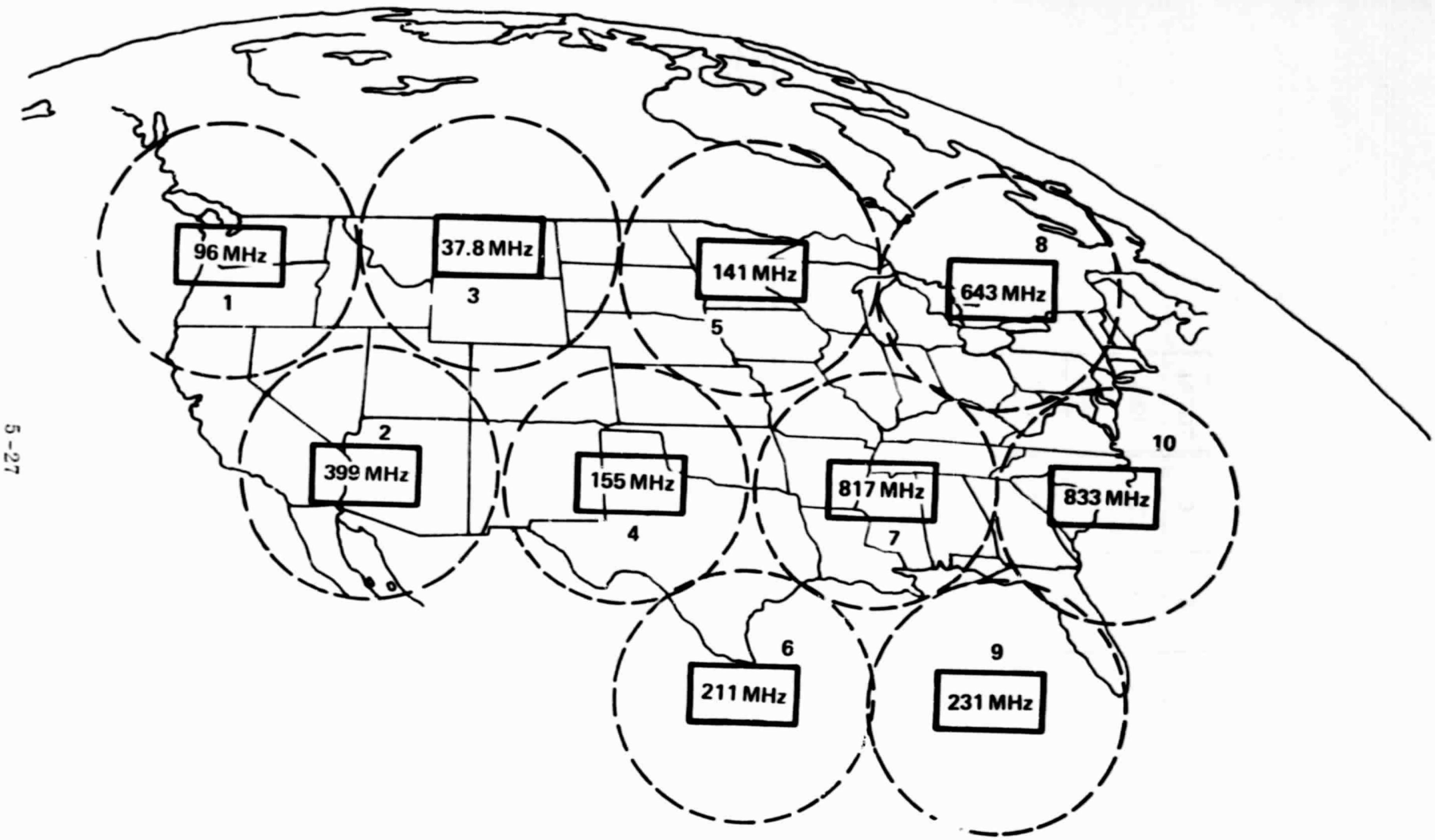
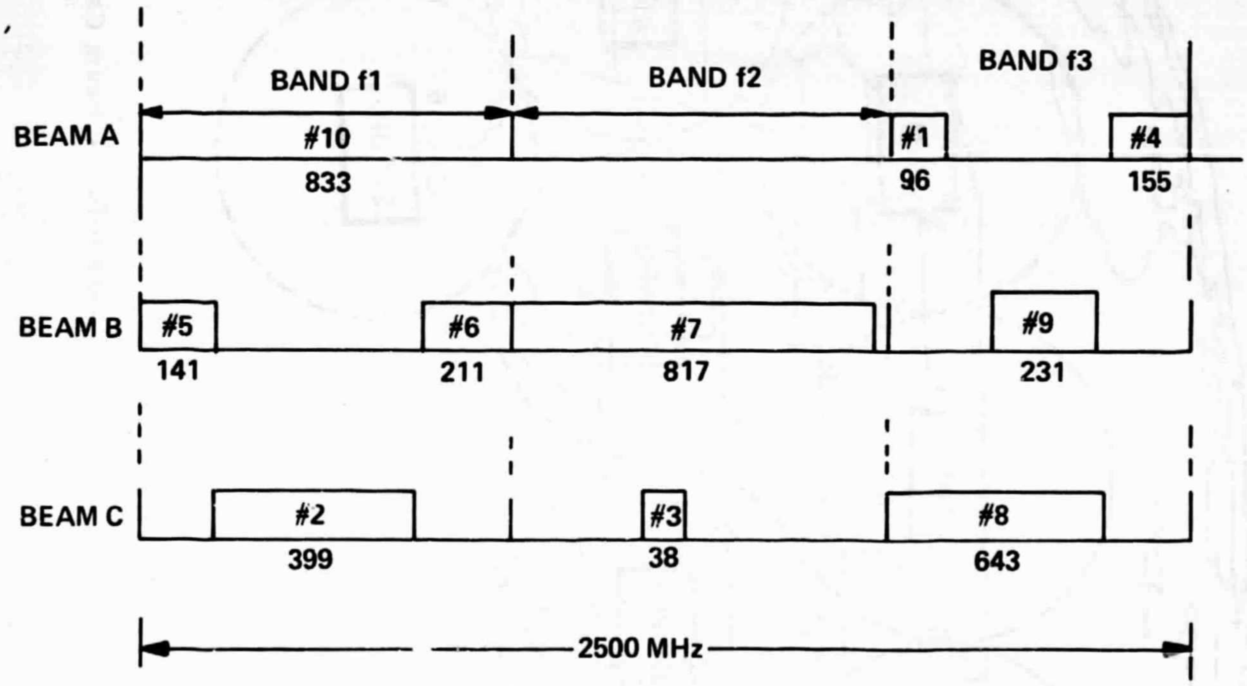


Figure 5-9. 10 Beam Case

BEAM	A	B	C	TOTAL
A	330	426	328	1084
B	426	550	424	1400
C	328	424	327	1080
TOTAL	1084	1400	1080	3564

A.



B.

Figure 5-10. Example of Multiplexer to Incorporate A/O Beam System Into a Three Receiver, Three Transmitter System, A) is Multi-plexed Arrangement, B) is Resulting, Frequency Plan

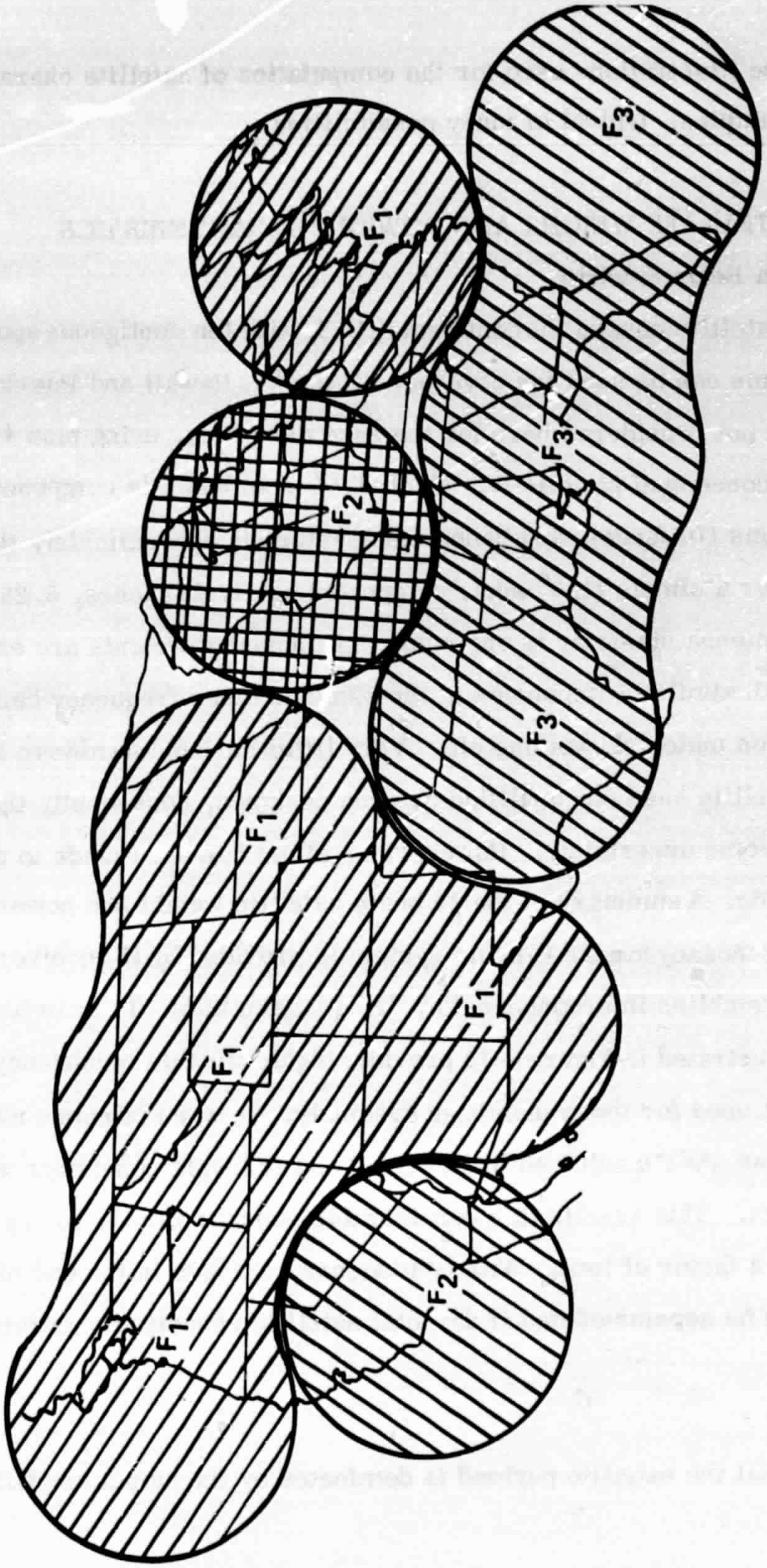


Figure 5-11. Example Beam Pattern

Conseqently the illustrations used for the computation of satellite characteristics are simple examples, typical of many possibilities.

5.3.4. SATELLITE WEIGHT AND POWER CHARACTERISTICS

Ten Beam Satellite

The ten beam satellite covers the continental U.S. with ten contiguous spot beams, additional spot beams can be used for coverage of Alaska, Hawaii and Puerto Rico, but this requirement is not considered here for the sake of brevity, using plan #4 of Section 3 because of its conceptual simplicity. Each 1.5° spot beam is composed of seven component beams (70 feeds), and hence the aperture is approximately three times that required for a simple (1.5° spot, or approximately 99 inches, 5.25 feet). The earth station antenna diameter is one meter. Component weights are estimated by comparison with similar components at the same or other frequency bands, by computation based on material, volume etc. Very little Ka Band hardware applicable to this type of satellite has been qualified or even designed, conseqently the estimates are subject to some uncertainty. However, an effort has been made to err on the conservative side. A summary of the 10 beam satellite weight and power is given in Table 5-7. Redundancy for the 10 beam system is provided by 15 receiver and 15 transmitter assemblies interconnected by "ring" circuits of "T" switches. This arrangement illustrated in Figure 5-12 provides highly flexible redundancy. A similar configuration is used for the transmitter assembly. Power estimates assume the use of variable power TWT's adjusted to saturated power levels consistant with the traffic level in its beam. This results in a considerable reduction in the prime power needed (approximately a factor of two). With contingency, changes in the end of life, busy hour traffic can be accommodated if the total satellite capacity is approximately constant.

It is apparent that the satellite payload is dominated by the output amplifier weight.

TABLE 5-7. 10 BEAM SATELLITE WEIGHT AND POWER ESTIMATE, 2M ES ANTENNA

Item	Unit Wtg. Lbs.	Qty	Total Lbs.	Unit Power Watts	Qty	Total
Antenna Reflector (1)	54		78			
Support	5		7			
Feeds (20 GHz)	.03	70	2.1			
Feeds (30 GHz)	.02	70	1.4			
Pwr. Dividers (20 GHz)	.05	70	3.5			
Pwr. Dividers (30 GHz)	.05	70	3.5			
Wavequide (20 GHz) 10 ft.	.86	10	8.6			
Waveguide (30 GHz) 10 ft.	.57	10	5.7			
			109.8			
Receiver (30 GHz -Uhf)						
BPF (3 Pole) invar	.03	10	.3			
LNR (st)	.10	15	1.5	.06	10	.6
Switches (T) WG	1.0	15	15			
Switches (T)coax	.3	15	4.5			
BPF (4 GHz) 2P	.3	15	4.5			
BPF (Uhf) 2P	.3	15	4.5			
Uhf Amp 3 Stage	.3	15	4.5	.06	10	.6
Mixer Assy (30 GHz)	.1	15				
Mixer Assy (4 GHz)	.1	15	1.5			
LO (Ka)	2.5	3	7.5	25	1	25
LO (C)	1.5	3	4.5	3	1	3
SWITCH	.1	2	.2			
Divider	.05	2	.1			
Shields, Boxes	lbs set	10	16			
Contingency(10%)			6			2.9
Total			66.1 lbs			32.1 watts
Control			10			10
38 Preassigned (36 GHz filters)	4.1	1	4.1	67	1	67
36 GHz						
33 x 33 (36 GHz) cross point SW	6.0	1	6.0	167	1	167
24 x 10 x 24 (36 GHz) pool	3.9	1	3.9	100	1	100
95 x 95 (10 MHz) x point	6.5	1	6.5	39	1	39
56 x 20 x 56 (10M) pooled	3.7	1	3.7	23	1	23
			24.2			396 watts
Shields, Boxes			5			
Contingency			2.4			40 watts
Sub Total			31.6			436 watts
Transmitter Section						
Outpit BPF (20GHz)	.10	10	1.0			
Switch (WG)	1.0	15	15			
Switch Coax	.3	15	4.5			
BPF, 2nd converter	.05	15	.75			
BPF, C Band	.3	15	4.5			
4 GHz SS amp (3 stage)	.6	15	9.0	.6	10	6
U/C Mixer (C)	.1	15	1.5			
U/C Mixer (KA)	.1	15	1.5			
LO #1	1.5	3	4.5	25	1	25
LO #2	2.5	3	7.5	25	1	25
SW	.1	2	.2			
Dividers	.05	2	.1			
Amplifiers* (TWTA)*	220	15	3300	700/.4	10	17.5KW (variable
Shields, Boxes			10			
Contingency			6			1.7
Sub Total			3366			19.2 KW +
Total			3584 Lbs.			19678 watts

TABLE 5-7. 10 BEAM SATELLITE WEIGHT AND POWER ESTIMATE, 2M ES ANTENNA

Item	Unit Wgt. Lbs.	Qty	Total Lbs.	Unit Power Watts	Qty	Total
Antenna Reflector (1)	54		78			
Support	5		7			
Feeds (20 GHz)	.03	70	2.1			
Feeds (30 GHz)	.02	70	1.4			
Pwr. Dividers (20 GHz)	.05	70	3.5			
Pwr. Dividers (30 GHz)	.05	70	3.5			
Wavequide (20 GHz) 10 ft.	.86	10	8.6			
Waveguide (30 GHz) 10 ft.	.57	10	5.7			
			109.8			
Receiver (30 GHz -Uhf)						
BPF (3 Poie) invar	.03	10	.3			
LNR (st)	.10	15	1.5	.06	10	.6
Switches (T) WG	1.0	15	15			
Switches (T)coax	.3	15	4.5			
BPF (4 GHz) 2P	.3	15	4.5			
BPF (Uhf) 2P	.3	15	4.5			
Uhf Amp 3 Stage	.3	15	4.5	.06	10	.6
Mixer Assy (30 GHz)	.1	15				
Mixer Assy (4 GHz)	.1	15	1.5			
LO (Ka)	2.5	3	7.5	25	1	25
LO (C)	1.5	3	4.5	3	1	3
SWITCH	.1	2	.2			
Divider	.05	2	.1			
Shields, Boxes	lbs set	10	16			
Contingency(10%)			6			2.9
Total			66.1 lbs			32.1 watts
Control			10			10
3* Preassigned (36 GHz filters)	4.1	1	4.1	67	1	67
36 GHz						
33 x 33 (36 GHz) cross point SW	6.0	1	6.0	167	1	167
24 x 10 x 24 (36 GHz) pool	3.9	1	3.9	100	1	100
95 x 95 (10 MHz) xpoint	6.5	1	6.5	39	1	39
56 x 20 x 56 (10M) pooled	3.7	1	3.7	23	1	23
			24.2			396 watts
Shields, Boxes			5			
Contingency			2.4			40 watts
Sub Total			31.6			436 watts
Transmitter Section						
Outpit BPF (20GHz)	.10	10	1.0			
Switch (WG)	1.0	15	15			
Switch Coax	.3	15	4.5			
BPF, 2nd converter	.05	15	.75			
BPF, C Band	.3	15	4.5			
4 GHz SS amp (3 stage)	.6	15	9.0	.6	10	6
U C Mixer (C)	.1	15	1.5			
U C Mixer (KA)	.1	15	1.5			
LO #1	1.5	3	4.5	25	1	25
LO #2	2.5	3	7.5	25	1	25
SW	.1	2	.2			
Dividers	.05	2	.1			
Amplifiers* (TWTA)*	220	15	3300	700 .4	10	17.5KW (variable
Shields, Boxes			10			
Contingency			6			1.7
Sub Total			3366			19.2 KW
Total			3584 Lbs.			19678 watts

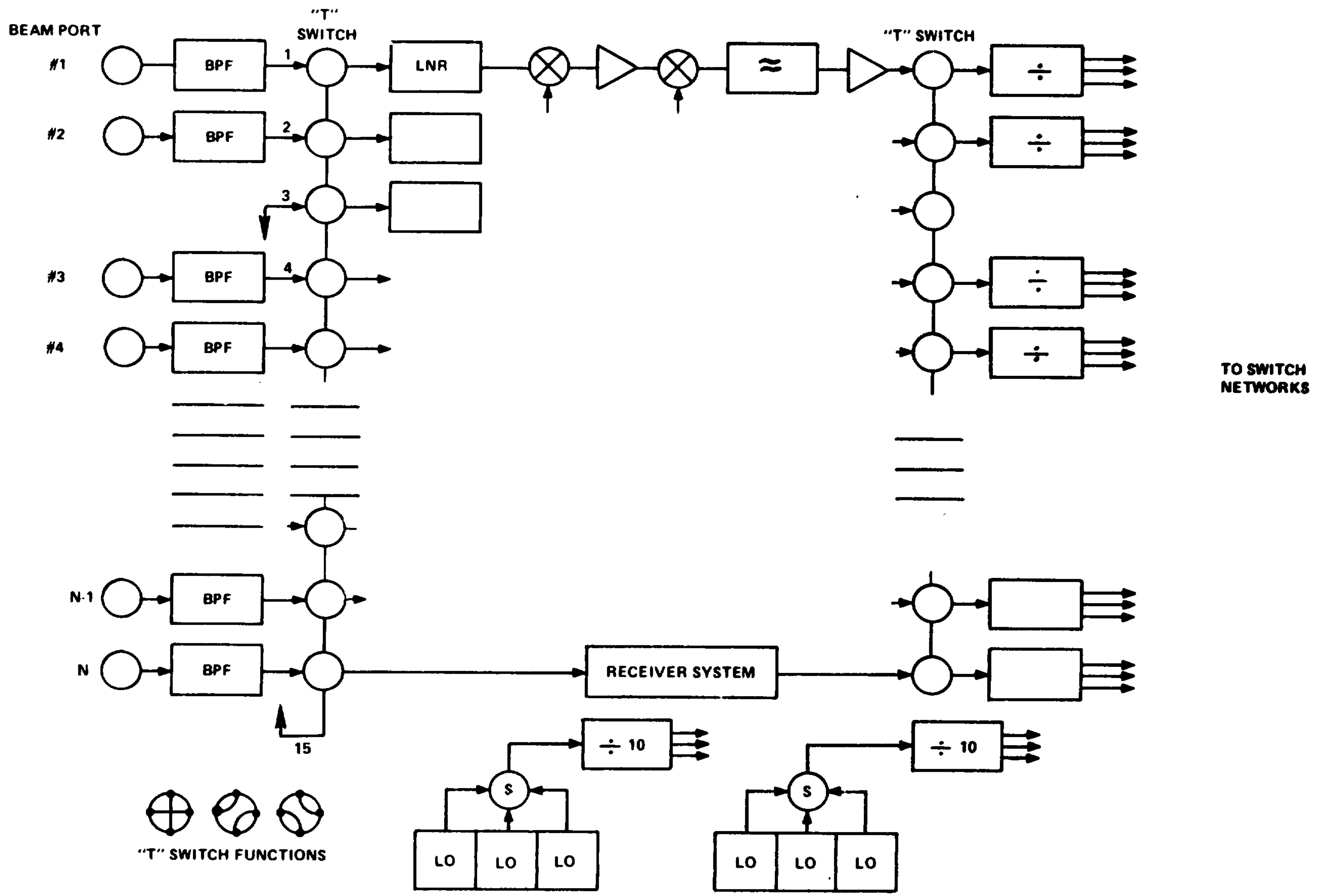


Figure 5-12. Example of Receiver Section with 15:10 Redunding

Table 5-8 summarizes the weight and power for the 100 beam satellite. Again a substantial portion of the satellite payload is consumed by the output amplifiers. However switch network weight and power also has increased significantly because of the complex routing demands of the 100 neam satellite. The reader will excuse us if we do not provide a block diagram for the 100 beam satellite.

A summary of weight and power characteristics based on the 1 meter earth station link previously given is depicted in Table 5-9. Total spacecraft weight can be computed by converting the daytime prime power requirements into satellite power system weight at the rate of 15 watts per pound. This is beyond the performance of existing satellites but is believed achievable for large amounts of power generation in the future. Power systems of this efficiency have been tested in space. The battery energy storage system is designed to take advantage of FDMA operation. During eclipse period the satellite system traffic will be considerably reduced (off peak hour data dumping can be scheduled to occur outside of the eclipse period), and consequently the transponder output power is reduced - either through the use of variable power tubes or by switching to low power tubes. A reduction in output power to 10% of the daytime load is assumed. Even though the eclipse period traffic will be considerably less then that value the battery system weight savings is not significant beyond this point. Energy during eclispe is generated at the rate of 6 watts/lb during eclipse, (eg, nickel cadmium batteries). A summary of these weight characteristics for the 10 beam satellite is given in Table 5-10. Even with only 10 beams a Ka - Band direct access satellite is predicted to be quite large and require a large amount of power, particulary for s mall earth station antennas. Even if the earth station antenna is 3 meters the satellite weight is comparable to that of the largest satellites today but with considerably more power, e.g., 2.6 KW. On the other hand the satellite capacity is large, approximately 74,000 32 KbPs, preassigned trunks which is considerably higher than that of existing satellites. It is apparent from Table 5-10 that the earth station antenna diameter has a substantial weight and cost.

TABLE 5-8. 100 BEAM SATELLITE WEIGHT AND POWER ESTIMATE

Item	Unit Wgt. Lbs.	Qty.	Total Lbs.	Unit Pwr. Watts	Qty	Total Watts
Antenna Rfl. (2)			780			
Support			78			
Feeds (20 Ghz)	.03	700	21			
Feeds (30 Ghz)	.02	700	14			
Pwr. Dividers (30 Ghz)	.01	700	7			
Pwr. Dividers (20 Ghz)	.01	700	7			
Waveguides	.86	100	86			
Waveguides	.57	100	57			
			1050 Lbs.			
Receivers (30 Ghz - Uhf)						
Bpf (3P) invar	.03	100	3			
LNR (3 stage)	0.1	150	15	.06	100	6
Switches (T) WG	1	150	150			
Switches (T) Coax	.3	150	45			
BPF (4 Ghz) 2P	.3	150	45			
BPF (Uhf) 2P	.3	150	45			
Uhf amp, 3 stage	.3	150	45	.06	100	6
Mixer Ass'y (30 GHz)	.1	150	15			
Mixer Ass'y (4 GHz)	.1	150	15			
LO (Ka)	25	3	75	50	1	50
LO (C)	15	3	45	10	1	10
Switch	1	2	.2			
Divider	.05	2	.1			
Shield, Boxes		30	30			
Contingency (10%)		53				7
		582				79
Controle		30				30
Switch Network						
100 X 10 (36 MHz)	82	1	82	390	1	390 (Fet Switch)
220 X 220 xbar (10 MHz)	161	1	161	895	1	895
2700 X 500 X pool (10 MHz)	179	1	179	1085	1	1085
1300 X 1300 xpoint (2 MHz)	315	1	315	1775	1	1775
3800 X 1000, pool (2 MHz)	260	1	260	1543	1	1543
Shields, Boxes			15			569
Contingency (10%)			1113			6257
Transmitter Section						
Output BPF (20 Ghz)	.1	100	10			
Switches, WG	1.0	150	150			
Switches, Coax	.3	150	45			
BPF, 2nd converter	.05	150	7.5			
BPF, C Band	.3	150	45			
1 GHz SS Amp (3 stage)	.6	150	90	.6	100	60
U C (C)	.1	150	15			
U C (Ka)	.1	150	15			
LO #1	15	3	45	100	1	100
LO #2	25	3	75	100	1	100
SW	.1	2	.2			
Dividers	.5	2	1			
Amplifiers	44	150	6600	70 .4	100	17.5 KW
Shields, Boxes			30			
Contingency (10%)			50			1.75 KW
			7179 lbs			19510 Watts
Total			9954 lbs.			25876 Watts (Variable)

TABLE 5-8. 100 BEAM SATELLITE WEIGHT AND POWER ESTIMATE

Item	Unit Wgt. Lbs.	Qty.	Total Lbs.	Unit Pwr. Watts	Qty	Total Watts
Antenna Rfl. (2)			750			
Support			78			
Feeds (20 Ghz)	.03	700	21			
Feeds (30 Ghz)	.02	700	14			
Pwr. Dividers (30 Ghz)	.01	700	7			
Pwr. Dividers (20 Ghz)	.01	700	7			
Waveguides	.86	100	86			
Waveguides	.57	100	57			
			1050 Lbs.			
Receivers (30 Ghz - Uhf)						
Bpf (3P) invar	.03	100	3			
LNR (3 stage)	0.1	150	15	.06	100	6
Switches (T) WG	1	150	150			
Switches (T) Coax	.3	150	45			
BPF (4 Ghz) 2P	.3	150	45			
BPF (Uhf) 2P	.3	150	45			
Uhf amp, 3 stage	.3	150	45	.06	100	6
Mixer Ass'y (30 GHz)	.1	150	15			
Mixer Ass'y (4 GHz)	.1	150	15			
LO (Ka)	25	3	75	50	1	50
LO (C)	15	3	45	10	1	10
Switch	1	2	.2			
Divider	.05	2	.1			
Shield, Boxes		30	30			
Contingency (10%)		53				7
		582				79
Controls		30				30
Switch Network						
400 X 40 (36 MHz)	82	1	82	390	1	390 (Fet Switch)
220 X 220 xbar (10 MHz)	161	1	161	895	1	895
2700 X 500 X pool (10 MHz)	179	1	179	1085	1	1085
1300 X 1300 xpoint (2 MHz)	315	1	315	1775	1	1775
3800 X 1000, pool (2 MHz)	260	1	260	1543	1	1543
Shields, Boxes			15			569
Contingency (10%)			1113			6257
Transmitter Section						
Output BPF (20 Ghz)	.1	100	10			
Switches, WG	1.0	150	150			
Switches, Coax	.3	150	45			
BPF, 2nd converter	.05	150	7.5			
BPF, C Band	.3	150	45			
4 GHz SS Amp (3 stage)	.6	150	90	.6	100	60
U C (C)	.1	150	15			
U C (Ka)	.1	150	15			
LO #1	15	3	45	100	1	100
LO #2	25	3	75	100	1	100
SW	.1	2	.2			
Dividers	.5	2	1			
Amplifiers	44	150	6600	70 .4	100	17.5 KW
Shields, Boxes			30			
Contingency (10%)			50			1.75 KW
			7179 lbs			19510 Watts
Total			9951 lbs.			25876 Watts
						(Variable)

TABLE 5-9. SATELLITE WEIGHT AND POWER (1 METER EARTH STATION ANTENNA)

	10 BEAM SATELLITE		100 BEAM SATELLITE	
	WT	POWER	WT	POWER
ANTENNA SYSTEM	110 LBS	0 WATTS	1050 LBS	0 WATTS
RECEIVER	66	32	582	79
CONTROL	10	10	30	30
FDMA SWITCH NETWORK	31	436	1113	6257
TRANSMITTER	3366*	19,200 (DAYTIME)	7179**	19,510 (DAYTIME)

*1600 WATTS PER SATELLITE TRANSPONDER
*160 WATTS PER SATELLITE TRANSPONDER

TABLE 5-9. SATELLITE WEIGHT AND POWER (1 METER EARTH STATION ANTENNA)

	10 BEAM SATELLITE		100 BEAM SATELLITE	
	WT	POWER	WT	POWER
ANTENNA SYSTEM	110 LBS	0 WATTS	1050 LBS	0 WATTS
RECEIVER	66	32	582	79
CONTROL	10	10	30	30
FDMA SWITCH NETWORK	31	436	1113	6257
TRANSMITTER	3366*	19,200 (DAYTIME)	7179**	19,510 (DAYTIME)

*1600 WATTS PER SATELLITE TRANSPONDER
*160 WATTS PER SATELLITE TRANSPONDER

TABLE 5-10. SPACECRAFT CHARACTERISTICS FOR 10 BEAM SATELLITE AS A FUNCTION OF THE EARTH STATION ANTENNA DIAMETER

	1M	2M	3M
COMMUNICATIONS PAYLOAD WEIGHT, LBS	3584	1702	937
ARRAY WEIGHT, LBS	1315	356	178
BATTERY WEIGHT LBS	407	159	123
SUBTOTAL	5299	2217	1238
SPACECRAFT WEIGHT, LBS*	11776	4927	2751
PRIME POWER, KW	19.7	5.3	2.6

*BEGINNING OF LIFE IN SYNCHRONOUS ORBIT

5.4 SATELLITE COST MODEL

5.4.1 INTRODUCTION

A satellite cost model is needed so that user cost characteristics can be computed. For the direct user system parametric cost analyses are necessary in order to show the relationships between user cost and important system characteristics such as satellite capacity, influence of earth station antenna size, satellite weight and power, impact of DAMA, number of antenna beams, fading, etc. Since future operational system characteristics are of most interest (e.g., the cost effectiveness of a Ka-Band direct user system compared to alternatives) a basic assumption is made that enabling technology is already developed and demonstrated and that the satellite is state-of-the-art when designed and that only non-recurring design costs and recurring costs are important. Correspondingly, the range of earth station costs do not include the basic enabling technology development costs, but do include imbedded non-recurring costs. Section 6 discusses recommendations for developing the enabling technology for a Ka-Band direct access system.

5.4.2 COST MODEL DESCRIPTION

The costs model considered is the SAMSO satellite cost model developed by the Aerospace Company (1) satellite modeling techniques by the Communications Satellite Corporation (2) and cost estimates for use of Shuttle STS (3). 1980 dollars are used throughout. The procedure for computing costs is as follows:
Cost is computed from satellite weight distribution and power. To encompass all the situations likely to arise, satellite costs are computed for a range of spacecraft

(1) Unmanned Spacecraft Cost Model, 4th Edition, February 1978 (SAMSO)
(2) Communication Stalite Modeling Technique; Comsat Technical Review, Volume 2, No. 1, Spring, 1972
(3) Space Transportation System Reimbursement Guide TCS-11802

weights extending from 1250 lbs. (beginning of life in synchronous orbit) to 10,000 lbs. The SAMSO cost model is used because it enables a computation of the communication and power systems non-recurring and recurring costs which can be analyzed later to see if the SS-FDMA subsystem costs follow reasonable trends in such a cost model. It should be noted that the SAMSO cost model is derived from a large number of historic spacecraft programs all of which are dissimilar to the multi beam SS-FDMA satellite, and the SAMSO model is used beyond its experience range. However, it is believed that the extrapolation for an operational system is still reasonably accurate. The procedure follows:

1. Develop a typical weight distribution for a satellite
2. Compute the satellite subsystem, non-recurring and recurring costs over the range of interest. Two different power levels are computed to cover the detailed design to be performed later.
3. Compute corresponding Shuttle/PAM costs.
4. Convert costs to annual costs over an assumed 10-year life, depreciation occurs only during the 10-year period. Add other ancillary costs required to design, launch and operate a satellite system.

It is assumed for the purpose of the study that the cost vs. weight is a continuous function. This will enable the cost of the 10 beam and 100 beam satellite listed in Section 5.3 to be identified and enables these costs to be parametrically manipulated. By relating costs, weight and capacity the user space segment costs can then be derived.

Tabel 5-11 shows a breakdown of satellite weight computed several different ways; by reference (2) which is a composit of many typical synchronous orbit communications satellites, an Intelsat IV breakdown (spinner) and a typical breakdown for a 3-axis satellite, the latter two for illustration. The Comsat model is used in this analysis.

The SAMSO model using the "Normalized Cost Estimation Relationships" (which applies a learning curve to the cost data to allow for industry maturation over the

TABLE 5-11. WEIGHT DISTRIBUTION FOR TYPICAL COMMUNICATIONS SATELLITES

	Comsat	Intelsat IV	GE
Positioning & Orientation Subsystem (POS) (Dry)	8%	15 %	5.4%
Hydrazine (10 years)	22	18.4	20.8
AKM case	8	28.1	7.9
Structure/Thermal/Balance	17		16.4
TT&C	3	3	1.7
Sub Total	58%	64.5%	52.2%
Available for Communication and Power System	42%	35.4%	47.8%

years) can allow for refined estimates taking into account "degree of complexity" and "technology carry over". However, since we are computing the cost of an operational system, these factors may not apply - it is assumed that the technology is well in hand during the satellite design phase. The SAMSO cost estimating relationships are given in Table 5-12 and the weight breakdown and power are given in Table 5-13 for four specific satellite designs covering the range of interest. Because the cost estimating relationship for power uses power as the independent variable two power levels are considered, given in Table 5-12 as Case I and Case II.

A 13% fee is applied. Since these are manufacturer's prices, fee includes cost of money prior to launch. Results are given in Table 5-14 for the four satellites. Note that the cost of a higher power satellite is a slow function of primary power which suggests that on Shuttle, extra power to avoid power limited operation with small antenna earth stations should not be an important cost consideration. This is particularly true if battery weight can be minimized - as in intended in the satellite design.

TABLE 5-11. WEIGHT DISTRIBUTION FOR TYPICAL COMMUNICATIONS SATELLITES

	Comsat	Intelsat IV	GE
Positioning & Orientation Subsystem (POS) (Dry)	8%	15 %	5.4%
Hydrazine (10 years)	22	18.4	20.8
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TABLE 5-12. COST ESTIMATING RELATIONSHIPS (SAMSO COST MODEL)

	NRC	RC
Structure TC & Interstage	$y = 477.45 + 128.27\,x^{0.55}$	$24.11 + 8.78\,x^{.75}$
TT&C	$315.95 + 29.46\,x$	$99.11 + 13.87\,x$
Communications	$-97.07 + 461.34\,x^{.51}$	$94.05 + 33.89\,x^{.83}$
ACS	$139.8 + 138.89\,x^{.8}$	$-68.24 + 41.39\,x^{.82}$
Power (x = BOL Power)	$126.19\,x^{.5219}$	$121.57\,x^{.1727}$
Program Level Costs (vs. total)	$.4490\,x$	$.4123\,x$
Convert to 1980 dollars by	152.5/109.693 = 1.39	152.655/109.725 = 1.39
Fee	13%	13%

TABLE 5-13. WEIGHT AND POWER BREAKDOWN

	Satellite Weight		1250 lbs	2500 lbs	5000 lbs	10,000 lbs
	Weight Distribution					
	POS	8%	100 lbs	200 lbs	400 lbs	800 lbs
	Hydrazine	22%	275	550	1100	2200
	AKM case	8%	100	200	400	800
	Struct/Th/Bal	17%	212.5	425	850	1700
	TT&C	3%	37.5	75	150	300
	S/C Bus	58%	725	1450	2900	5800
	Payload	42%	525	1050	2100	4200
I	Comm PL	36.6%	420	840	1680	3360
I	Power Wt (1/5 of PL)	8.4%	105	210	420	840
I	Power (5W/lb)	--	525W	1050W	2100W	4200W
II	Comm PL	21%	262.5	525	1050	2100
II	Power Wt (1/2 of PL)	21%	262.5	525	1050	2100
II	Power (5W/lb)		1312.5W	2625W	5250	10,500W

Conversion to annual costs is accomplished by the methodology indicated in Table 5-15 which gives spacecraft and launch vehicle expenditures during the satellite development period, and includes insurance, interest at 10%, cost of a satellite control center, launch costs, O&M, parts and spacecraft engineering. All acquisition costs are accumulated until launch date (t = 0, launch is instantaneous) at which time

TABLE 5-14. SUMMARY SATELLITE WEIGHT NON RECURRING, RECURRING COSTS (1980 DOLLARS)

Satellite Wt, Lbs	1250	2500	5000	10,000
NRC (000's $)	55,796	86,027	135,258	216,860
RC (000's $)				
Case I (000's $)	14,746	25,683**	45,192	80,072
Case II (000's $)	11,970	20,679**	36,217	64,028
Program Cost, 3 sat., Case I	100,034	163,076*	270,834	457,076
Program Cost, per sat., Case I	33,345	54,358	90,278	152,359
Program Cost, 3 sat., Case II	91,706	148,064	243,909	408,944
Program Cost, per sat., Case II	30,569	49,354	81,303	136,315
Shuttle Costs ($M)	13.0	21.2	57.6	115
Space Segment				
Annual Cost Case I	28.1	45.0	84.7	150.4
($M) Case II	26.4	42.0	79.2	140.7

* 7 satellite = $265,808
** Comsat Model RC = 20.2M @ 2500 lbs.
NRC = 96M @ 2500 lbs.

acquisition (capitol) costs are depreciated in 10 years at a return-on-investment of 15%. The control center is depreciated in 20 years, however. Annual expenses are added to the depreciation. At the end of 10 years the salvage is zero, (except for the control center). The resulting annual costs are listed in Table 5-14, and are displayed in Figure 5-13 as a function of spacecraft orbit weight. Over the range of interest neither non-recurring, recurring or annual costs are proportional to weight but increase much more slowly - at a power of 0.5 to 0.7.

TABLE 5-15. SATELLITE COST MODEL (SAMSO, 1978) 1980 DOLLARS

SUBSYSTEM COST — BY FORMULA (NORMALIZED)
PM 45%
FEE 13%
3 SATS, 2 LV
SHUTTLE-OPTIMIZED
3 YEAR CONSTRUCTION

COSTS

SAT WT. (BOL, LBS	1250	2500	5000	10000
COMM NRC, $M	13.8	19.7	28.2	40.2
COMM RC, $M	7.2	12.8	22.5	39.9
TOTAL NRC, $M	55.8	86.0	13.5	21.7
RC, $M	14.8	25.7	45.2	80.0
AVER. SAT COST $M	[illegible]	54.4	90.3	152
LV COST $M	13.0	21.2	57.6	~115

ANNUALIZING FACTORS

10% INTEREST (TO LAUNCH)
LAUNCH COSTS
GROUND SPARE & STORAGE
INSURANCE
O&M (TT&C) & SAT. CONTROL (20 YRS)
10 YEARS DEPRECIATION
ROI = 15%
FILL FACTOR 58% (15% GROWTH)

WEIGHT MODEL

POS (DAY)	8%
HYDROGINAL (10 YRS)	22
AKM CASE	8
STRUCTURE/THERM	17
TT&C	3
COMM./POWER	42
	100%

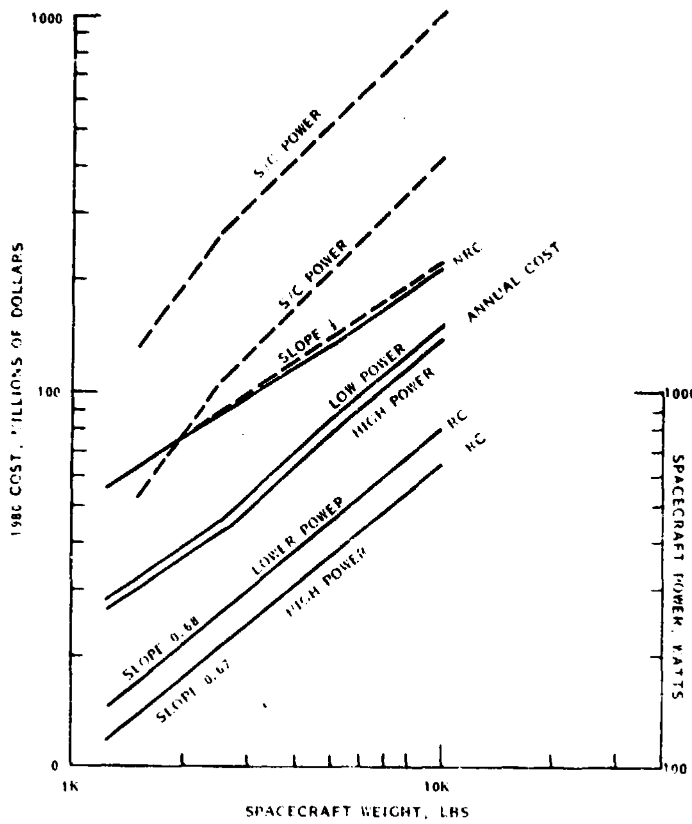

Figure 5-13. Space Segment NRC, RC and Annual Costs (1980 Dollars)

For example, consider the implications of SS-FDMA switching. Its weight impact on a satellite is obviously more than it is for SS-TDMA. In Section 4, this impact was 132 lbs. out of 11776 lbs. or 1% for the 10 beam satellite and 2568 lbs. out of 24448 lbs. or 10.5% for the 100 beam satellite. The impact on spacecraft cost is less than 7%. Note also, that the higher power satellite costs less. This is believed to be a logical result of the SAMSO modeling because the lower power satellites tend to have more transponders and transponders are costly. The principle reason for the high power is due to operating with small earth station antenna for the direct access system. Operation in the FDMA mode actually saves power because of the advantage in a fading environment and advantages in sizing satellite power and energy systems.

The complete user annual space segment costs per circuit require one further step. The space segment annual cost is multiplied by a factor F, the satellite "fill" factor.

The satellite "fill" factor increases the satellite cost to take into account the fact that traffic grows exponentially (for the foreseeable future) and consequently the satellite is not fully utilized until its last year. The satellite fill factor f is the average traffic load presented to it over the 10-year operational life, and is a function of the traffic average growth rate. If traffic increases at the rate of 15% per annum for 10 years the annual fill factor is 58%. This value is used in the Study.

5.5 USER COST AND SERVICE

5.5.1 SATELLITE COST

The previous sections can be used to compute the annual cost of the satellite as a function of the earth station diameter in meters. This is accomplished by first calculating the satellite weight using the methodology developed in Section 5.3 and then calculating the annual cost associated with that weight as developed in Section 5.4. If the annual cost of the satellite is divided by the number of 32 kbps circuits (two channels per circuit) or trunks then we have the annual cost per pre-assigned 32 kbps trunk, in 1980 dollars: these and other key relationships are displayed in Figure 5-14 as a function of earth station antenna diameter.

The 10 beam satellite annual cost declines from $2700 per trunk for a 1 meter antenna to about $1200 per trunk for a 3 meter earth station antenna. Over the same range, the satellite weight and power vary from 12,000 lbs and 20 kw to 2700 lbs and 2400 watts respectively.

The 100 beam satellite annual cost declines from $830 per trunk for a 1 meter earth station antenna to about $430 per trunk for a 3 meter earth station antenna. Over the

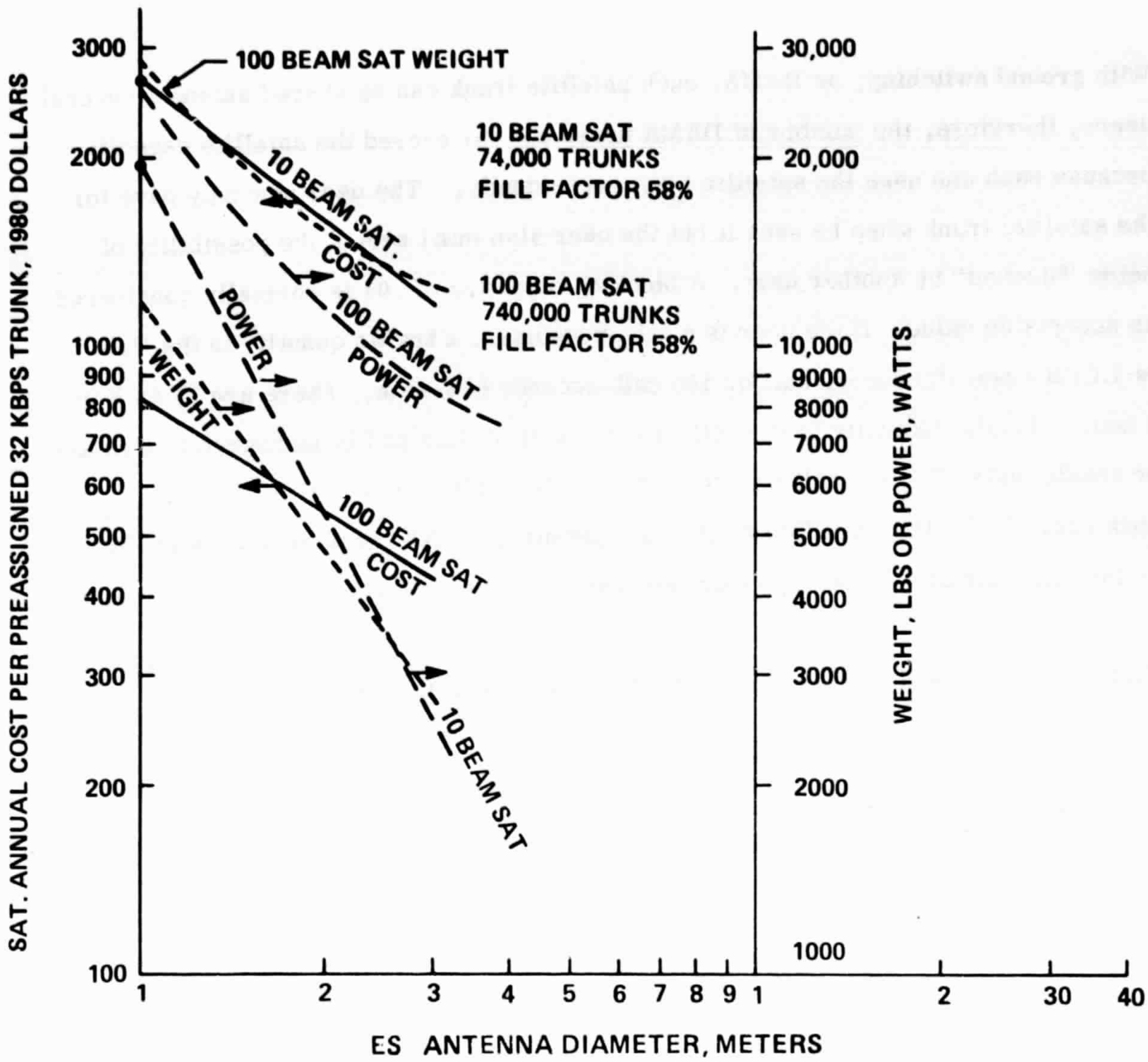

Figure 5-14. Satellite Annual Cost, Weight and Power vs. Earth Station Antenna Diameter

same range the satellite weight and power vary from 28,000 lbs and 25,000 watts to 1300 lbs and 8500 watts, respectively. Note that the capacity of the 10 beam satellite and 100 beam satellite are fixed at 74,000 trunks and 740,000 trunks, respectively. As the earth station antenna diameter changes the satellite power is changed to maintain constant link performance.

With ground switching, or DAMA, each satellite trunk can be shared amongst several users, therefore, the number of DAMA users can far exceed the satellite capacity because each one uses the satellite only occasionally. The user now only pays for the satellite trunk when he uses it but the user also must accept the possibility of being "blocked" by another user. A blocking objective of .01 is normally considered an acceptable value. If the user is a telephone user, a traffic quantity in the U.S. is 1 CCS = one 100 second call or 100 call-seconds of traffic. There are 36 CCS in 1 hour. Traffic intensity is the call rate per unit of time and is measured in Erlangs or traffic units (TU) - 1 call minute/minute. One Erlang on one trunk implies continuous use. In the U.S. traffic intensity is measured in CCS/hour or unit calls (UC) so that one erland = 36 UC = 36 CCS per hour.

Traffic intensity and blocking are related by the Erlang B equation*:

$$P = \frac{y^n}{n!} \quad \frac{1}{\sum_{o}^{n} \frac{y^n}{n!}}$$

where

P = probability of blocking

n = number of trunks

y = Erlangs

This formula can be used to construct Table 5-16 based on a blocking probability of .01. If a trunk is utilized 100% of the time, then there are 36 UC/trunk. This is the maximum number of 100 second calls on any trunk. Such high utilization requires many available trunks and high traffic intensity. Note that with only 150 trunks

* This assumes blocked calls will try again

TABLE 5-16. ERLANG B CHARACTERISTICS, P = .01

Trunks	UC	UC/Trunk	Efficiency
1	.4	.4	1.1%
10	161	16.1	44.7
100	3026	30.26	84.0
150	4737	31.58	87.7

available the efficiency (number of UC/36 x 100) is 87.7%. For 32 Kbps trunks using delta modulation and 4Ø PSK, there are 20 trunks per megahertz. If the path is 10 MHz there are 200 trunks available if the path is 100 MHz there are 2000 trunks available. Consequently the 10 beam and 100 beam satellites handle the bulk of the traffic efficiently in terms of calls per hour. A detailed study is required to actually configure the satellite for the busy hours with its mixed demand for service and to calculate therefrom its DAMA capacity. If the bulk of the subscribers are telephone users, the analysis predicts that the satellite could handle approximately 30 subscribers per hour per trunk during the busy hours. In this case the satellite's apparent capacity is very large indeed. On the other hand there will be some pre-assigned service, and there may be demands for other classes of service, for example, teleconferencing which has different service characteristics Teleconferencing conversations, assuming the use of 32 Kbps for either voice or facsimile graphics can be expected to have longer durations than telephone conversations. Teleconferencing may even use faster rates, 64 Kbps or 128 Kbps for interactive facsimile or TV "still" pictures. Interactive full motion video teleconferencing can occur at 40 Mbps. Lower usage per trunk, and fewer trunks per path result in lower trunk utilization. On the other hand teleconferencing requires scheduling the attendance at both ends of the circuit. Advantage can be taken of this by scheduling the circuit as well as the conference - this will improve the efficiency; however, the call duration for the service will be considerably longer than for a telephone call - perhaps of the order of an hour. Considering the above, and noting the absence of real statistics or mixes of service, the narrowband services are assumed to have a DAMA factor of 20:1 and wideband services (video) are assumed to have a DAMA factor of 5:1. These factors reduce the space segment cost directly.

5.5.2 EARTH STATION COSTS

Earth Station costs are based on three general classes of users. One will use the single MODEM earth station described in Section 5.2 which operates at 32 Kbps AVD, and which could include a time division multiplexer to multiplex several low-speed data streams. The second earth station uses a multiplicity of MODEMS, 10 32 Kbps MODEMS for voice or data and 64 Kbps MODEMS for data, providing a broad range of communication services including voice, computer interaction, teleconferencing, telemail, etc. These earth stations do not include costs for user equipment such as facsimile equipment, computers, etc., but only the transmission equipment. The third earth station is for interactive 4.2 MHz video teleconferencing and its earth station costs include the camera and TV monitoring equipment. Antenna costs including installation is assumed to be a variable user cost and is given in Figure 5-15, based on Reference 1. HPA costs also will be variable with antenna diameter; however, these are not taken into account because previous studies have indicated that the HPA costs are a slowly varying function with power and do not affect the antenna trade-off substantially. Earth station installation costs are converted to annual costs by the following factors:

Capital Recovery Factor	0.23852
O&M	0.09481
	0.33333

which are based on 20% return on investment in 10 years. Capital recovery factor identifies the annual payment of 10 equal payments which depreciates the investment in 10 years paying a return on investment of 20%. The following analysis computes

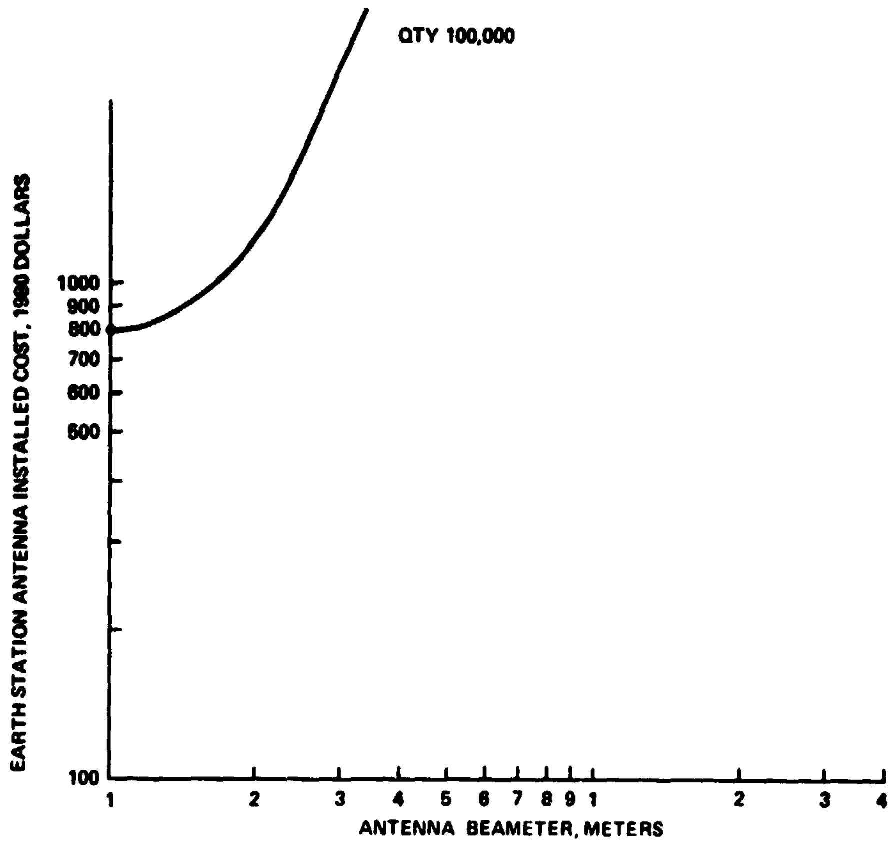

Figure 5-15. Antenna Costs Including Installation as a Function of Antenna Diameter

user end to end costs, including cost of two earth stations and a prorated satellite trunk cost - with and without DAMA so that comparisons can be made with end to end terrestrial trunks. Subscriber costs, since satellite trunk costs are generally small, will be approximately half the end to end costs. The range of costs are as follows:

- single MODEM earth station $10,000-$30,000
- multi MODEM earth station $30,000-$50,000
- interactive TV $79,000

Again it is emphasized that these costs are production costs of standardized earth stations produced in large lots with maximum incorporation of MIC and LSI technology. While these costs are in 1980 dollars the operational implementation is anticipated to be in the 1990's.

5.5.3 RESULTS

Figure 5-16 depicts the results for the single MODEM earth station user and the 10 beam satellite showing total annual end to end costs as a function of earth station antenna diameter. The single MODEM user service can be voice, computer interaction, point of sales, electronic fund transfer, narrowband teleconferencing, telemail, etc. The shaded-area curves show the range of earth station acquisition costs from $10,000 to $30,000 and considers both pre-assigned satellite trunks and DAMA satellite trunks with a 20:1 average use factor. User costs straddle the range from about $7000 per annum to about $22,000 per annum. Since the individual curves are nearly horizontal, earth station antenna diameters at the range of 1M to 3M are nearly optimum - size selection does not seriously impact user costs. Reduction in costs due to DAMA are not substantial in this case because the satellite charge is small compared with the annual earth station cost. An understanding of the significance

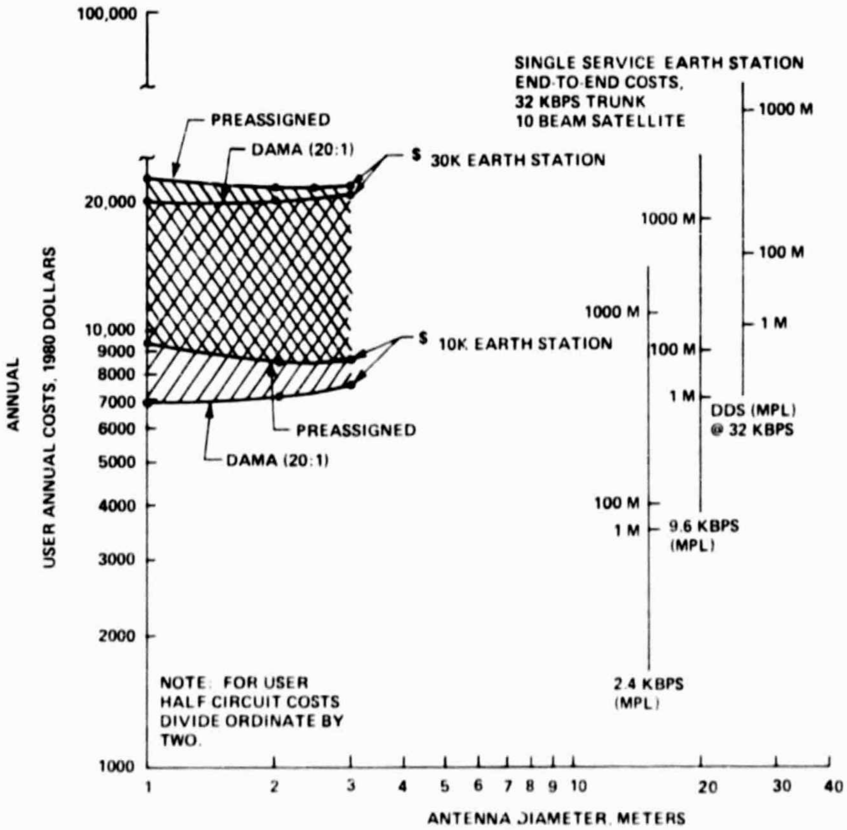

Figure 5-16. Single Modem Earth Station

of these costs can be obtained by comparing the terrestrial system rates shown at the right of Figure 5-16. For the direct access single MODEM system to compete with the 2.4 Kbps MPL rates an earth station cost approaching $10,000 is needed; a somewhat higher cost earth station is permissible in order to compete with the 9.6 Kbps MPL - say $20,000: and a $30,000 earth station can confortably compete with the DDS (this charge is prorated for 32 Kbps instead of 56 Kbps). The comparison at 2.4 Kbps and 9.6 Kbps may be viewed as a 2.4 Kbps stream buffered for transmission through the satellite at 32 Kbps in order to reduce the system requirement for frequency and phase stability. The terrestrial charge emphasizes the importance of low cost earth stations and sets appropriate cost goals for system operators if they wish to compete for this market. As described previously such low earth station costs are not achievable with certainty; however, under the circumstances defined in Section 5.2 they may be achievable and there is evidence to support this contention. The market available for the direct access satellite system could be immense under these circumstances.

Figure 5-17 shows the similar results for a multi MODEM earth station user. The multi-MODEM user embraces more of the same kinds of services identified for the single MODEM user. The earth station cost, however, is prorated by data rate amongst the various services so that the earth state cost per service is much reduced. User costs now range from $1600 per annum to $5000 per annum. Since the earth station antenna cost is now shared, the satellite costs are more important particularly for pre-assigned satellite trunks and these curves show a propensity for larger earth station diameters (large negative slope) in order to reduce the satellite charge. The curves are truncated at 4 meters since direct access terminals beyond this size are not practical. On the other hand, DAMA reduces the satellite charges so that 2 meter earth station antennas are attractive. DAMA now has a great influence on user costs.

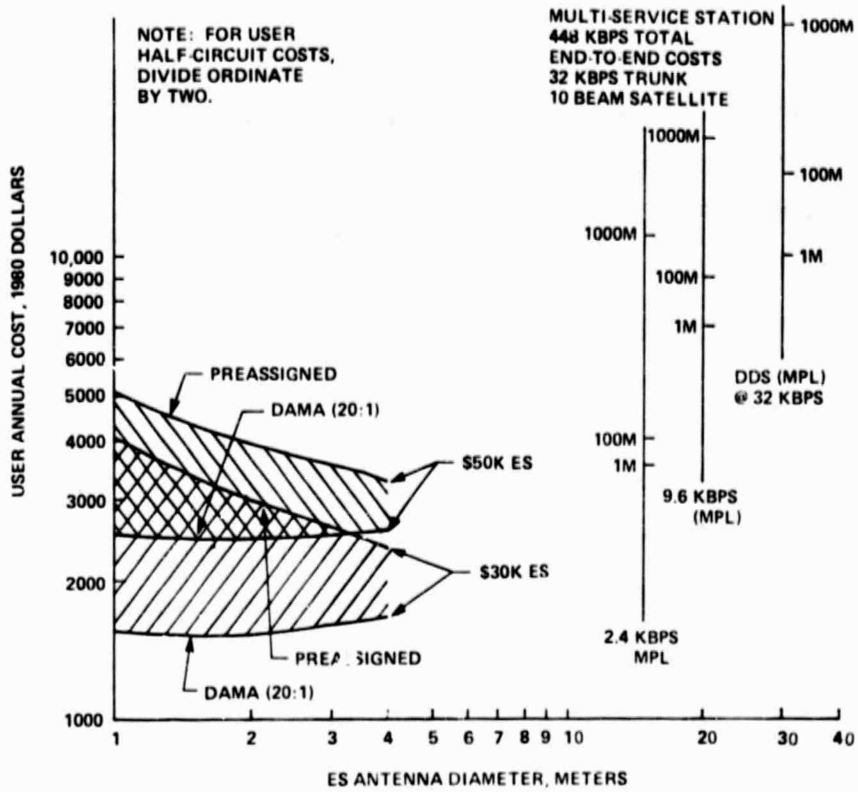

Figure 5-17. Multi-Modem Station

The terrestrial charges at the right indicate that the direct access satellite services compare favorably with terrestrial systems with break-even distance less than 100 miles even for 2.4 Kbps MPL. The earth station allowable cost could be as high as $70,000 and still compete for the 2.4 Kbps market. 9.6 Kbps and 32 Kbps (or 56 Kbps services) are very attractive by satellite even for earth station costs higher than $70,000.

Figure 5-18 shows similar results for the 100 beam satellite and a single MODEM earth station user. In this case the range in costs are very similar to that of the 10 beam satellite because satellite costs are not significant, (and consequentially DAMA also is not important), earth station antenna diameters are close to optimum over the range of 1 to 3 meters. Again, a $10,000 earth station is needed to compete in the 2.4 Kbps market whereas $30,000 is probably adequate at 32 Kbps and 56 Kbps.

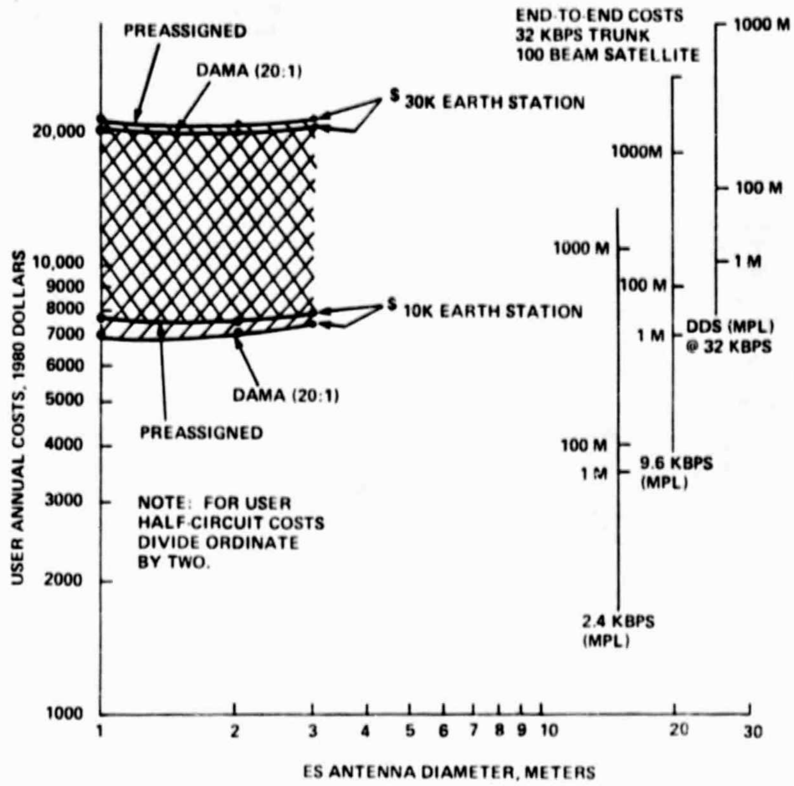

Figure 5-18. Single Modem Station

Figure 5-19 shows corresponding results for the multi MODEM earth station and 100 beam satellite. Again, earth station antenna diameters around 2 meters are optimum, DAMA has some impact for the multi MODEM service and earth station costs higher than \$50,000 - perhaps \$70,000 or higher are still attractive.

User costs for interactive television teleconferencing are given in Figure 5-20 for the 100 beam satellite. Despite the large satellite capacity and use of DAMA the satellite charge is still substantial and Figure 5-20 indicates that minimum user cost will occur for an earth station antenna diameter of around 10 meters. Subscriber costs of around \$40,000 to \$50,000 will be attractive to some users but the present guess is TV teleconferencing will not be a large market at these prices particularly since graphic displays will have relatively poor resolution.

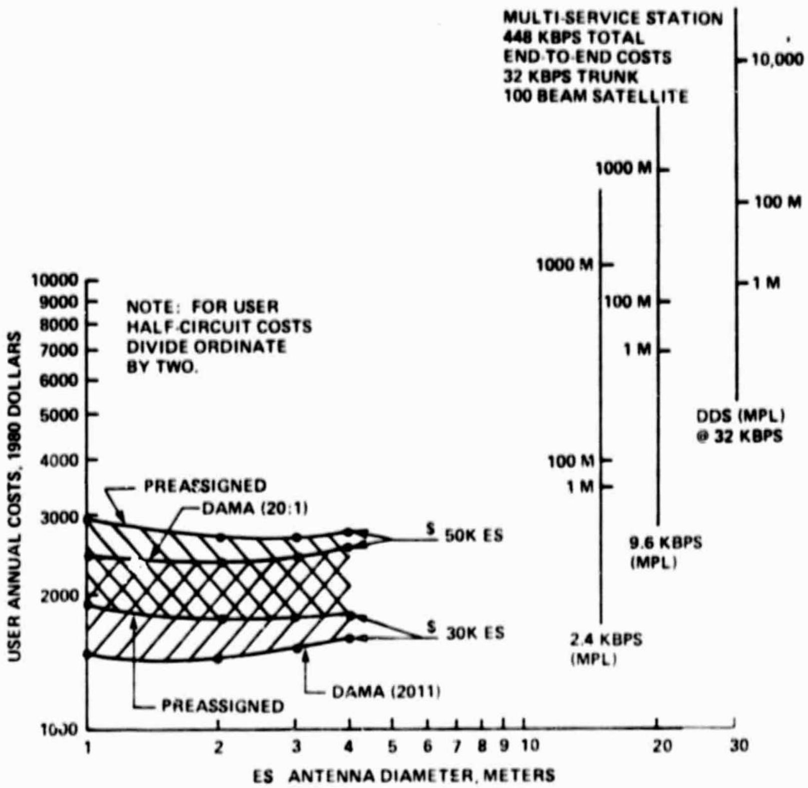

Figure 5-19. Multi-Modem Station

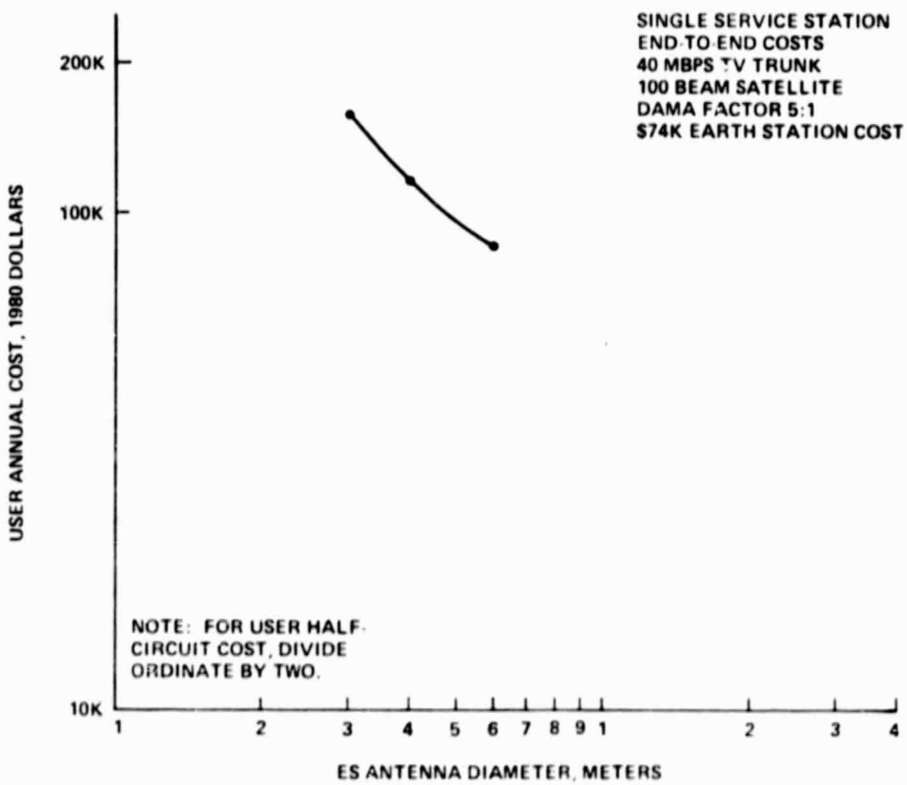

Figure 5-20. Single Modem Station

5.5.4 SUMMARY RESULTS

Results of the technology and economic analyses performed in the Study show that a direct access system based on SS-FDMA can provide flexible service, at user costs which are very attractive compared to existing terrestrial tariffs, and at lower costs than conventional trunking satellite systems. Operation at Ka-Band, using carrier power sharing, permissible with FDMA, permits single earth station antenna operation (no antenna diversity) with high user availability. Thus, flexibility, including complete switching capability (including a TELCO interface if necessary) low cost and high reliability is achievable. The system design, in essence reverses current trends in satellite communications by placing more tehnology in the satellite in order to simplify the earth station configuration and thus reduce overall cost.

Access directly to satellite avoids performance and cost limitations of terrestrial loops and offers new versatility - bandwidth on demand - for system users.

Technical feasibility requires development and demonstration of the low cost earth station - in a system context where standardization, large lot production and a common signalling/switching system (SCPC/DAMA) is possible. Satellite technology involving high power (with diurnal power variation), high power (adaptive amplifiers), on board channelization and switching is capable of being demonstrated by an extensive technology program. The nature of this program is the subject of Section 6.

SECTION 6

RECOMMENDATIONS FOR SS-FDMA DEVELOPMENT

SECTION 6

RECOMMENDATIONS FOR SS-FDMA DEVELOPMENT

6.1 INTRODUCTION

The SS-FDMA System Concept requires a new approach to the design of both satellites and earth stations in order to realize the system characteristics described herein. The present NASA program involving both technology development and experimental satellite test and demonstrations can result in confirmation of these characteristics, including proof of concept and confirmation of technology readiness. This approach is not new in the respect that present carrier systems include direct access SCPC networks, but because of the considerations described herein these systems are not attractive economically. FDMA also is not new and in fact continues to be the work horse access method of today's satellites despite great strides forward in the design and production of TDMA. Intelsat V is an example of an FDMA satellite with on-board switching - rudimentary compared with the techniques advocated in the study - but sophisticated by present state of the art considerations. Since the SS-FDMA technology is relatively simple in application if the satellite has just a few beams and rapidly become more complex as the number of beams increases it is possible to foresee an evolutionary approach to SS-FDMA design, becoming increasingly more complex as the satellite resort to more and more antenna beams. A special contribution can be made in the development of low cost earth stations because special technology (consistent with low-cost technology) and high non-recurring costs inhibit an evolutionary approach. The existence of NASA developed hardware, system tests, and experimentation can provide the inspiration and confidence for later commercial exploitation. In addition the demonstration of special service (like teleconferencing, and general access to a wideband transmission medium) and techniques such as signalling and switching and use of carrier power sharing to overcome fading effects can add to the utility of the tests and demonstrations.

This section identifies key technologies and special engineering problems important to the realization of a futuristic Ka-Band operational system and then suggest a program for, initially, a technology development and later for proof of concept (e.g., a satellite experiment and demonstration).

6.2 SATELLITE TECHNOLOGY (SS-FDMA)

a. SS-FDMA Switch

This consists of multiple switch elements and control elements built on LSI chips. The LSI implementation should explore both a wideband switch, at 36 MHz or higher and a narrowband switch at 5-10 MHz. Representative chips and chip control circuits should be fabricated and tested for compatibility with communication objectives and in a packaging arrangement suitable for satellite use.

b. Channel Filter

This can consist of either (or both) SAW filters and ceramic piezo-electric filters, depending on bandwidth and center frequency to demonstrate achievement of communication parameters. A 6 pole or 8 pole Tchebychev filter with control of out of band roll-off is needed to demonstrate compliance with requirements for channelization in an FDMA system. Some consideration should be given to packaging and the interplay of the filter with the down conversion/up conversion plan.

c. Ka-Band Components

NASA already has plans for developing Ka-Band amplifiers, both solid state and TWT types and low noise receivers. This study identifies a need for substantial amounts of amplifier power, practical only with TWT's for the foreseeable future - perhaps several hundred watts or perhaps more depending on the ultimate operational traffic capacity and the G/T realizable with the low-cost earth stations. FDMA also has additional advantages with a variable power TWT with relatively constant efficiency to enable adjustment of power according to traffic flow through a given TWT: this minimizes spacecraft power which will be high in any case for a direct user system. In addition, a power range of at least a factor of 10 between daytime and eclipse operation also can minimize the battery system. Linearization of the TWT to minimize phase shift with amplitude can increase the useful power output. These requirements, all attractive for implementation of an FDMA satellite represent a significant advance in the TWT state of the art particularly at Ka-Band. While it would be desirable to achieve this performance by use of solid state devices the high power levels appear to be prohibitive.

The need to use a solid-state HPA in the earth station to minimize initial costs and O&M costs with adequate power margin for operation in the power sharing mode appears to require a good satellite G/T. Particularly, if the number of satellite beams are small it appears desirable to achieve system noise temperatures in the vicinity of 100°K, requiring the use of paramps. Perhaps with more antenna beams somewhat higher noise temperature can be permitted. Since earlier implementation and experiments will likely use less rather than more satellite antenna beams a paramp development for satellite use is attractive. Converters for use at Ka-Band are believed to be available from industry.

d. High Power Satellite

Solar arrays, slip rings, batteries are available to develop the required power, active louvers and heat pipes to control all expected diurnal variations in satellite dissipation also have been demonstrated in space (ATS-6, CTS, GE Broadcast Satellite are examples of satellites exhibiting these characteristics). There does not appear to be any necessity for technology development. In this regard attitude control technology also is believed to be adequate despite the narrow beams used.

e. Satellite Antenna

A contiguous multiple beam antenna has not been tested to date although several antenna applications, namely DSCS-III and Intelsat V in particular are examples of relevant technology. In both these cases shaped beams are formed from separate feeds consisting of arrays of horns. The antenna implementation suggested herein, based on offset fed parabolic antenna technology, is highly predictable by analysis. Nevertheless, the achievement of C/I ratios of approximately 30 db is indicated by system considerations and this should be demonstrated by test with a configuration which includes the optics, the effects of horn coupling and an evaluation of misalignments and mechanical error.

6.3 EARTH STATION TECHNOLOGY

Low cost is the most important objective for the earth station design, requiring special system design, hardware and O&M considerations. With regard to the NASA program it is recognized that final proof of low cost comes with the production of large numbers of earth stations. This is clearly beyond the scope and mandate of the NASA program. Nevertheless, it is believed that substantial advances can be made by demonstrating the performance of several prototype earth stations designed and constructed to meet the cost goals, and exhibiting a substantial number of MIC/LSI subsystem. Costs (and performance) of future operational Ka-Band systems will depend on cost of devices, particularly, Ka-Band devices and these can only be extimated during the present program; costs in general will depend on earth station production rates, standardization of design, and costs for devices. The NASA prototype earth station should contain examples of:

- Standardized designs
- Maximum use of MIC/LSI
- Factory test
- Simple installation
- Test and O&M instructions from a network control center

With regard to standardization of design it is believed that advantages can be taken of lower satellite charges to operate at standard rates (standard MODEMS/CODECS etc.) even though these may only roughly correspond to the user input/output rates. For the microwave circuits which represent the principle component parts cost, attention should be given to monolithic circuit construction and super component MIC (filter LNR-downconverter combination) etc to simplify assembly and testing. Of course the signalling and switching (DAMA) interface also must be standardized to permit central control and testing by a central control center. The single MODEM earth station design is most sensitive to cost. Its flexibility can be enhanced by an input multiplexer which can provide alternate voice or data, AVD at 32kbps or multiplexed data. The combination of MODEM, frequency synthesizer, signalling unit, CODEC/multiplexer and miscellaneous interface equipment are amenable to LSI construction, in fact CODECS and multiplexers of various types already exist in chip form. Once standardized these subsystems can be adapted to LSI form for low cost manufacture. The MODEM is perhaps the greatest challenge since it has a substantial number of high frequency circuits and complex analogue functions (filtering, tracking loops, etc) difficult to convert to standard low frequency LSI construction. However, the LSI state of the art continues to advance and it is believed that a substantial cost reduction can be achieved by proper facilitation. Probably the biggest unknowns with regard to low cost are the Ka-Band devices principally GaAsFets. At 20 GHz and

30 GHz these devices are barely out of the laboratory and production in any substantial quantity - necessary for low device prices - is years away. In fact, it may be difficult even to achieve adequate performance at the time of the NASA program so that even performance may change radically in the late eighties. Even Ku-Band devices are still expensive since the commercial market is still relatively small. However, it may be possible to extrapolate Ka-Band experience and possibly C-Band experience to show analogously what devices price experience might be expected to occur at Ka-Band.

6.4 TECHNOLOGY PROGRAM PLAN

The Technology Program recommended for the SS-FDMA development recommends that the present NASA Technology Program should be augmented to include the FDMA-peculiar items and that these technology developments should be pursued in a timely manner so that there will be high confidence on the part of NASA and the system contractor concerning the feasibility of an FDMA system experiment and demonstration.

This means technology is developed or progress made such that NASA and the System Contractors can take advantage of it during the Phase II and particularly during the Phase III. This does not necessarily mean that the output of the technology development is directly available during the Phase III design activities.

The plan to be described assumes that NASA develops a definitized system model from which system specifications and objectives can be developed. The present study emphasized antenna and switch technology with supportive system engineering performed to illustrate attractive direct access system features. This study stopped short of developing a detailed system concept including frequency plans for both the satellite and earth station and specific satellite and earth station designs. An alternative to the above is to require the Phase II system contractors to perform detailed designs to serve as guidelines for Technology Development. If this alternative is selected the Technology Program start dates shown in Figure 1 ought to be slipped at least 3 months and possibly by 6 months to allow the System Contractors sufficient time to develop these designs.

Figure 6-1 depicts a plan for technology accomplishment showing also the Phase II and III expected schedules. The FDMA switch, channelization filter and contiguous beam antenna each should start early in 1980 in order to obtain the maximum confidence with the System Contractors. With regard to the SS-FDMA switch it is recommended that the switch configuration be approximately 1000 switch units on a CMOS-SOS chip with bandwidth between 5MHz and 10MHz. It is desirable also to include a wider bandwidth switch unit, say 36MHz to 72MHz to also explore FET technology for a switch matrix of perhaps several hundred switch units. In design "A" period of Figure 6-1 there should be an interaction between a system designs establishing objectives, specifications and configuration, (including packaging) and an LSI design group investigating layouts etc. During the period designated B, the chip final design and testing jigs are completed and several chips are manufactured to investigate yield and achievement of basic design objectives. The chip design should include the design of basic control circuits considering methods to avoid single point failures. The period designated C is the time during which system integration (packaging), and test of chip and control circuit is accomplished. It is recommended that a reliability analysis and Failure Modes and Effects Analysis be included with this program. The final chip testing in this phase assures achievement of system parameters such as isolation, bandwidth and function.

Similarly a channelization filter based on SAW technology emulating a 6 or 8 pole Tchebychev filter particularly in the roll-off region should be designed and tested. Final testing should include typical filter parameter evaluation (flatness, noise bandwidth, group data, roll-off) with special emphasis on out of band roll off. Several examples of different bandwidths, 2 MHz to 72 MHz should be evaluated. The contiguous beam antenna ought to be evaluated on an antenna range. It is suggested that several active beams be constructed and that dummy (terminated) horns to used to simulate the remainder of the beam forming optics. Primary patterns can then be evaluated with

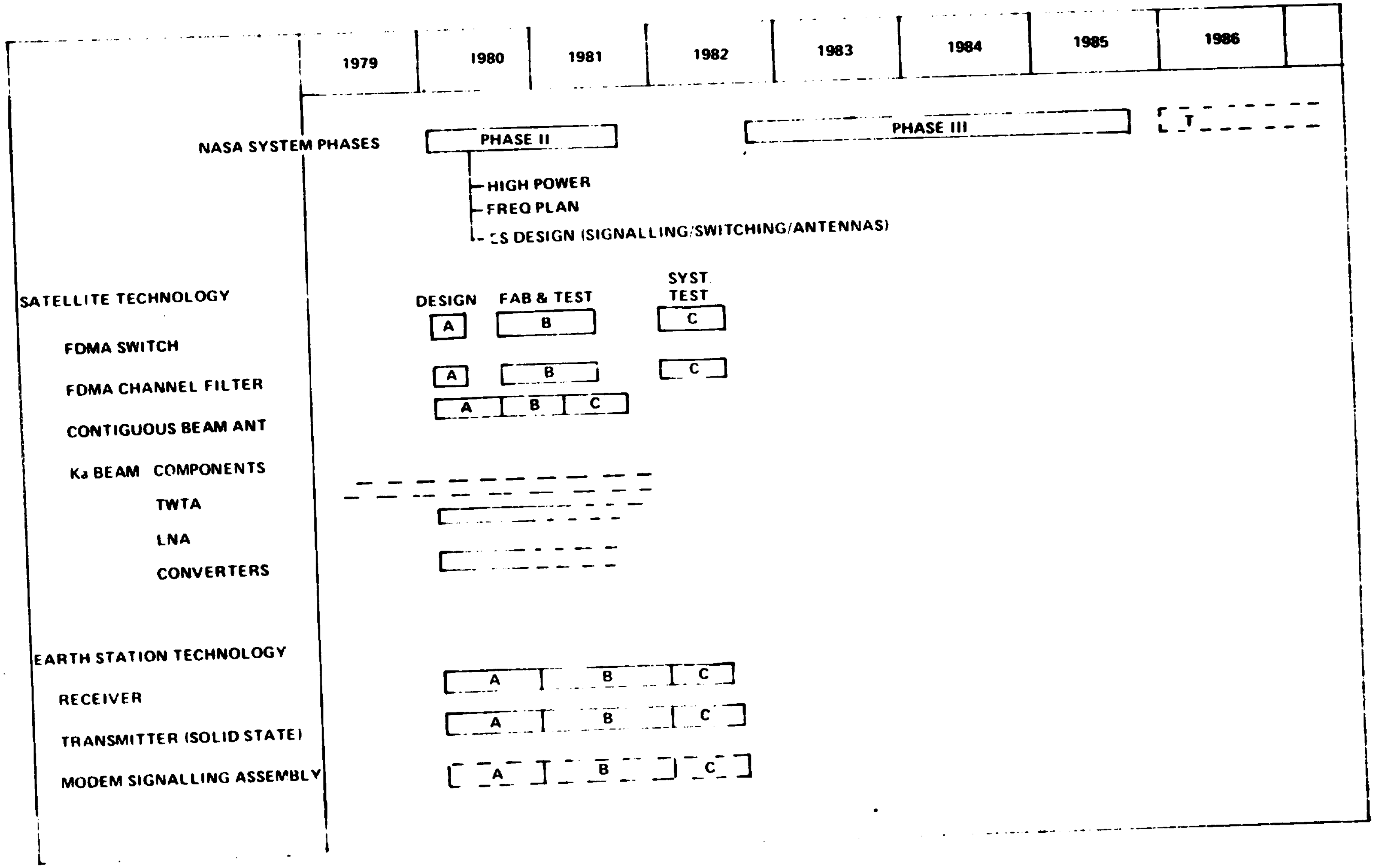

Figure 6-1. Technology Development Plan for SS-FDMA

regard to pattern bandwidth and coupling effects. Later, its optics (offset fed reflector) can be added to evaluate secondary pattern effects due to scanning. The entire evaluation will assure achievement of interference which is approximately 30 dB worst case. Since the beam topology plans result in various combination of frequency use and C/I a representative plan such as Plan 4 or 5 is recommended for evaluation.

The remaining suitable technology development requirements are satisfied by existing NASA plans. The TWTA development at the 75 watt level is ample for the anticipated experiment although operational Ka-Band direct access systems will require substantially more power. This high power level can be developed over the years by NASA technology programs to allow timely Ka-Band operational implementations in the 1990's. The conceptual designs described herein make use of multi level power amplifiers. This feature is important in the FDMA system because it minimizes satellite daytime power and night time energy storage requirements. Linearization also will help to reduce power.

The LNA program ought to investigate use of 30GHz paramps in order to reduce earth station HPA power.

The earth station technology development should include the earth station receiver (including converter) and solid state transmitter (including converter) and other important Ka-Band components. Since integration of several MIC or monolithic circuits is contemplated an interactive A/B design approach between system design and circuit design is suggested with system test accomplished in the C part. A program relating to the lower frequency circuits is recommended for MODEM, synthesizer, signalling unit etc in anticipation of an MIC/LSI earth station demonstration in the Phase III test and demonstration period. There already exists ample evidence that such circuits can be implemented in LSI form, given sufficient production demand. In this case this development can serve to identify sources of equipment, possibly with some modification, possibly with some development, to ready this lower frequency technology for later test and demonstration. In any case, demonstration of signalling and switching in a direct access system is believed to be important to future users.

APPENDIX A

SWITCH AND ANTENNA TECHNOLOGY IMPLICATIONS FOR LAND MOBILE SATELLITE

APPENDIX A

SWITCH AND ANTENNA TECHNOLOGY IMPLICATIONS FOR LAND MOBILE SATELLITE

A1. INTRODUCTION

A land mobile satellite system can also make use of a contiguous beam antenna and on-board switching. Therefore it was planned to example the relevancy of these technologies for application to either fixed or mobile services. This study however places great emphasis on the fixed service application, in terms of system and user application as well as technology relevant to a Ka-Band system. A similar concurrent Study also performed by GE for NASA was devoted exclusively to the general charateristics of a satellite - aided land mobile system. Consequently only the special antenna and switch requirements peculiar to a mobile system (but based on the work herein) will be described.

A2. SWITCHING

In the satellite aided mobile concept, contiguous multiple beams define the service area with at least one fixed station assigned in each beam. In general mobile traffic is local, requiring mobile to fixed communication within this service area beam. Long distance traffic or traffic destined outside a given beam can be handled in several ways.

1) The fixed station can transmit the long distance signal over land lines (or on fixed satellites) to the fixed station in the destination service area beam. In this case no satellite switching is required and reliance is placed on the capability of terrestial telephone networks.
2) Like the FDMA fixed service concept described herein, a portion of the available band can be set aside for the purpose of inter beam routing. Those mobile stations requiring this service are assigned frequencies in this special band. These are then switched in the conventional SS-FDMA format. Since only a small percentage of this traffic is inter beam, and the total traffic in a land mobile satellite is small compared to that of a fixed satellite the switch technology in terms of routes and paths is small compared to that of the fixed

services. The narrowband mobile traffic also would be switched at low RF frequencies required for LSI implementation. It is concluded therefore that this on board switch and filter technology is identical for the two systems and is described adequately in Section 4.

A3. CONTIGUOUS BEAM ANTENNA

In this case also, the operational characteristic of the mobile and fixed services are identical. Both will be constrained by the basic C/I that can be obtained by various beam topology plans and optical arrangements. Because of bandwidth requirements and a convenient mechanical arrangement suitable for spacecraft implementation the off-set fed parabolic antenna is the preferred choices. Since the Ka-Band antenna is physically small many different beam topologies can be considered depending on which of the system choices, such as weight, best C/I, best frequency reuse etc., prove optimum for the application, including plan 4 which provides good C/I only if the aperture is oversized. Therefore many choices are available and it was not possible (perhaps not desirable) to recommend a specific topology. Two factors are important to consider with regard to mobile services. First, the quality of the link, at least in terms of current usage is considerably less than that of the fixed services, in terms of voice intelligibility and background noise. Consequently it is believed that antenna C/I may not be quite as important as frequency reuse, particularly in a situation where the available bandwidth will cause a severe capacity restriction. Second, this lower frequency causes the physical aperture size to be a dominant engineering problem. Therefore engineering solutions will favor topologies like plan 1, which offer maximum frequency reuse and minimum aperture size. Antenna deployment itself has been the subject of much study and some testing by NASA and is certainly beyond the scope of this study except to point out that the assymetrical offset fed parabolic antenna geometry may be a problem for an orbit deployment. At least it is a somewhat different arrangement than the configurations previously studied and tested. ATS-6 antenna for example was a symmetrical prime focus fed parabola. A lens, the other candidate antenna configuration, while symmetrical, is very complex mechanically and heavy, and ought not to be considered for these reasons.

Consequently, the electrical performance calculated for the offset fed parabola is equally applicable to a land mobile antenna except for the ramification in physical size. Our conclusions therefore are that an offset fed parabolic antenna is the antenna configuration suitable for the land mobile service.