Cost Effective Options For The Thin Route And TV Broadcasting Network Of Northern Canada

VOLUME III

PREPARED BY MILLER COMMUNICATIONS SYSTEMS LTD.

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COST EFFECTIVE OPTIONS FOR THE THIN ROUTE AND TV BROADCASTING NETWORK OF NORTHERN CANADA

FINAL REPORT

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APPENDIX A

DESCRIPTION OF EXISTING THIN ROUTE SYSTEM

The work reported here was performed by Miller Communications Systems Limited for Research Sector, Department of Communications (contract officer Dr. Roger Kaye) under DSS Contract No. OST5-0051.

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APPENDIX A

DESCRIPTION OF EXISTING THIN ROUTE SYSTEM

1. INTRODUCTION

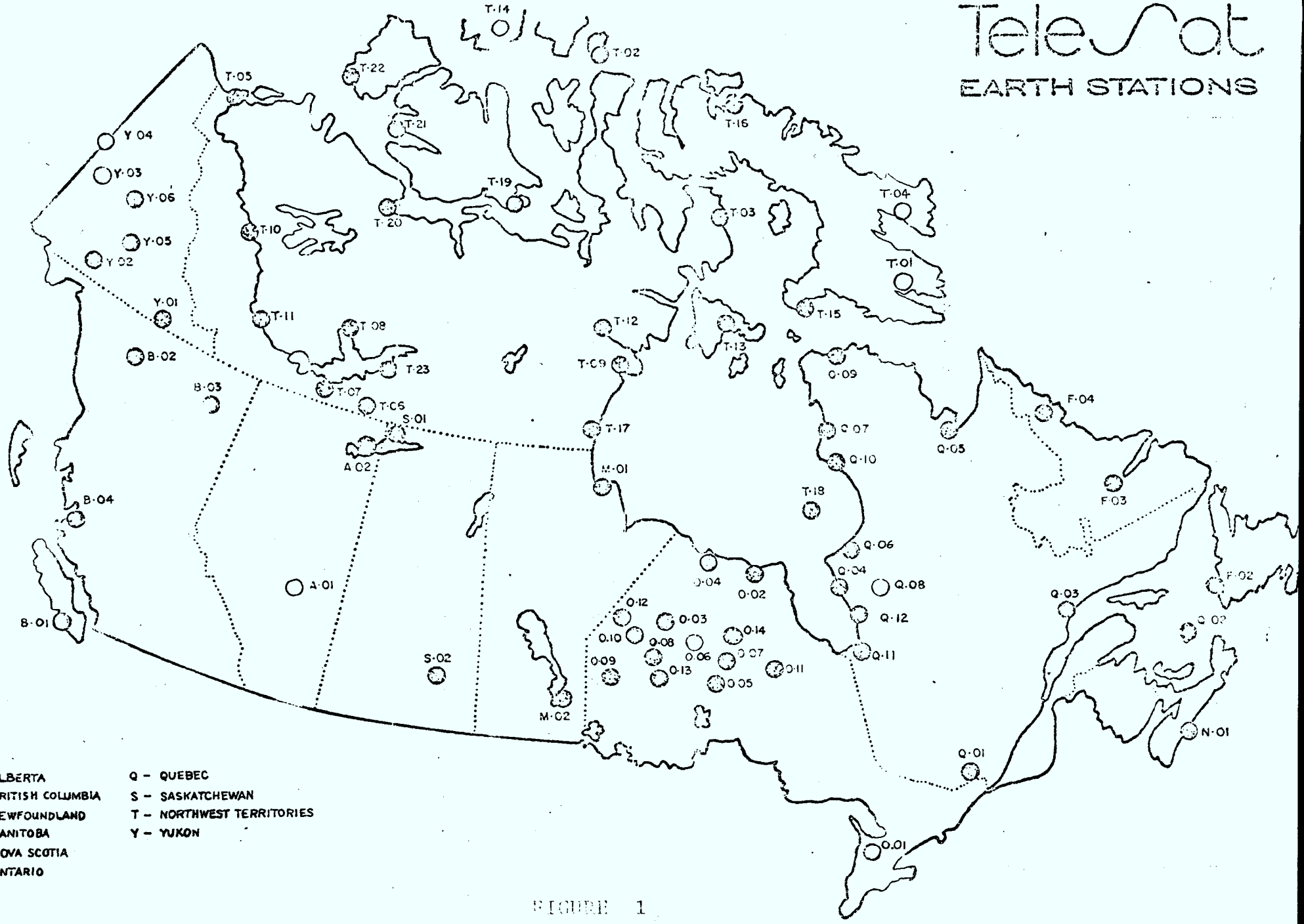
On February 1, 1973 the initial Thin Route System commenced commercial operation via the newly launched Canadian Domestic Satellite, Anik I. The service provided by Telesat to Bell Canada was designed to extend high quality reliable telephone communications to small remote communities in Northern Canada. Single Channel Per Carrier (SCPC) transmission techniques using delta modulation, 2-phase CPSK (DM-2ØCPSK) and frequency division multiple access to the satellite communications transponder were used.

To date and over the past three years the Thin Route System development has occurred in seven readily identifiable phases. Telesat's present network of earth stations are indicated on the map given in Figure 1 with associated index in Table 1.

This section outlines some of the most important features of the Thin Route network, its earth stations, and provides some perspectives on the constraints and problems attendant with growth of this flexible communications system.

Telesat

EARTH STATIONS



CODE

A – ALBERTA	Q – QUEBEC
B – BRITISH COLUMBIA	S – SASKATCHEWAN
F – NEWFOUNDLAND	T – NORTHWEST TERRITORIES
M – MANITOBA	Y – YUKON
N – NOVA SCOTIA	
O – ONTARIO	

FIGURE 1

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TABLE 1

EARTH STATION INDEX*

ALBERTA

A-01 Huggett
A-02 Ft. Chipewyan

BRITISH COLUMBIA

B-01 Lake Cowichan†
B-02 Cassiar
B-03 Ft. Nelson
B-04 Bella Bella

MANITOBA

M-01 Churchill
M-02 Belair

NEWFOUNDLAND

F-01 Bay Bulls
F-02 Port au Port
F-03 Goose Bay
F-04 Nain

NOVA SCOTIA

N-01 Harrietsfield

NORTHWEST TERRITORIES

T-01 Frobisher
T-02 Resolute
T-03 Iglookik†
T-04 Pangnirtung†
T-05 Inuvik
T-06 Ft. Smith
T-07 Pine Point
T-08 Yellowknife
T-09 Rankin Inlet†
T-10 Norman Wells
T-11 Ft. Simpson
T-12 Baker Lake†
T-13 Coral Harbour†
T-14 Rea Point†
T-15 Cape Dorset†
T-16 Pond Inlet†
T-17 Eskimo Point†
T-18 Sanikiluaq†
T-19 Cambridge Bay
T-20 Coppermine
T-21 Holman†
T-22 Sachs Harbour†
T-23 Snowdrift†

ONTARIO

O-01 Allan Park†
O-02 Winisk†
O-03 Big Trout Lake†
O-04 Ft. Severn†
O-05 Ft. Hope†
O-06 Kasabonika†
O-07 Landsdowne House†
O-08 Weagamow†
O-09 Cat Lake†
O-10 Muskrat Dam†
O-11 Ogoki†
O-12 Sachigo Lake†
O-13 Slate Falls†
O-14 Webique†

QUEBEC

Q-01 Riviere Rouge
Q-02 Magdalen Is.
Q-03 Sept Isles
Q-04 Ft. George
Q-05 Ft. Chimo
Q-06 Poste de la Baleine†
Q-07 Povungnituk†
Q-08 Radisson†
Q-09 Saglouc†
Q-10 Inoucdjouac†
Q-11 Ft. Rupert
Q-12 Noveau Comptoir

SASKATCHEWAN

S-01 Uranium City
S-02 Qu'Appelle

YUKON

Y-01 Watson Lake
Y-02 Whitehorse
Y-03 Dawson
Y-04 Clinton Creek
Y-05 Faro
Y-06 Elsa

* Stations are numbered by Province in order of start of service

† Stations possessing Thin Route Equipment

2. SYSTEM DEVELOPMENT

2.1 Background

The Thin Route System derives from an agreement between Bell Canada and Telesat in which Telesat was required to carry voice frequency signals (in contrast to FDM basebands) presented by Bell within a full Canada coverage service area. Bell in turn has selected the locations to be served and the means of integrating Thin Route facilities into the existing switched telephone network.

Initially the transmission techniques specified were delta modulation, PSK single voice channel per carrier, frequency division multiple access, although subsequent changes in modulation techniques were not precluded. This agreement is referred to as the Bell-Telesat Northern Service No. 2 agreement to distinguish it from the Northern Service No. 1 agreement which defines the FM-FDM-FDMA system linking Frobisher Bay and Resolute with the South.

The NS No. 2 Agreement required Telesat to provide a full 36 MHz RF transponder for the Thin Route system.

Thusfar pre-assigned North-South circuits have been implemented, although features such as end-to-end compatibility and frequency

agility have been retained where possible to permit the addition of demand assignment at minimum cost. In general, terrestrial radio links (possibly existing) can and have been used to extend the radius of access of a remote terminal, and in some cases allow the "sharing" of a terminal by two or more closely spaced northern communities.

Service was scheduled to commence in January 1973 with the communities of Igloolik, NWT and Pangnirtung, NWT each being linked to Allan Park, Ont. by two preassigned voice circuits. Service actually started 1 Feb. 1973 at which time Telesat began to receive the full rental revenue from the Thin Route transponder.

Even prior to commencement of service of the initial system, agreements were reached and contracts were awarded to extend Thin Route coverage to an additional fifteen northern communities. A summary of the ensuing system expansion is given in Table 2, and each phase is described in the following paragraphs.

PHASE	PROGRAM	START OF SERVICE
1	Thin Route I	Feb 1973
2	Thin Route II	Jan 1974
3	Rea Point Terminal	Aug 1974
4	Leased Message Transportable	Feb 1975
5	Purchased Message Transportable	Early 1976
6	CNT Program	Spring 1976
7	Northern Ontario Program	Summer 1976

TABLE 2. Thin Route System Expansion

2.2 Thin Route I Program

The Thin Route I program provided two preassigned circuits between Iglookik, NWT and Allan Park, Ont., and two preassigned circuits between Pangnirtung, NWT and Allan Park.

The SCPC equipment considered for this program included PCM-PSK, DM-PSK and FM versions. A contract award based on objective and subjective performance evaluation, price, and delivery was made to Canadian General Electric for an in-service date of Jan. 1, 1973. The chosen system was a 40 kbps DM-2ØCPSK system similar to one which was in an advanced state of development at G.E., Valley Forge, Pa.

Raytheon Canada Ltd. designed, manufactured and installed the two remote earth stations except for the SCPC equipment. In addition Raytheon provided the redundant downconverters for Allan Park.

The transmitters including high power klystron amplifiers for Allan Park were provided by RCA Ltd. as an addition to the primary contract for the Heavy Route Earth Station.

On Feb. 1, 1973 subsequent to the completion of acceptance testing the three stations entered commercial service.

Early problems encountered included:

i Erratic signalling (2600 Hz SF, in band) performance due to delta codec deficiencies

ii Reliability of SCPC equipment

iii Frequency stability of local oscillators in translation equipment.

The deficiencies were referred to the contractors for correction. Subsequent expansions of the Thin Route system have employed SCPC equipment from other manufacturers.

2.3 Thin Route II Program

The TRII program has been the major Thin Route implementation to date in terms of number of locations served and contract price.

A single contract award was made in Dec. 1972 to Raytheon Canada to provide:

1. Twelve new Thin Route earth stations

2. Three Thin Route add-on packages to existing RTV stations

3. Upgrading of Thin Route I stations to be compatible with Thin Route II

4. Channel equipment for Allan Park to permit one for one termination with remote earth stations.

The field implementation was handled in two phases as follows

Phase I - In Service 1974 (1st & 2nd qtr)	Phase II - In Service Winter 74/Spring 75
* Fort Chimo, P.Q	Cape Dorset,NWT
Povungnituk, P.Q.	Eskimo Point, NWT
Coral Harbour, NWT	Pond Inlet, NWT
Baker Lake, NWT	Sanikiluac, NWT
* Rankin Inlet, NWT	Saglouc, P.Q.
Winisk, Ont.	Inoucdjouac, P.Q.
Big Trout Lake, Ont.	* Great Whale R, P.Q.
	Fort Severn, Ont.

* denotes RTV add-on package.

Each of these stations is fully reduncant and uses a 26 ft. diameter antenna to provide a G/T of 22 dB/^{o}K. During construction each station was equipped with two voice channel units with provisions to accept up to 6 further units on a plug-in basis.

The SCPC equipment was developed by Raytheon Canada specifically for this program again using 40 kbps DM 2Ø CPSK techniques but with no attempt to ensure codec compatibility with the TRI SCPC equipment. Plans to replace Thin Route I SCPC equipment with Thin Route II type equipment were subsequently dropped.

Both Thin Route I and Thin Route II form part of the Bell Canada public network.

2.4 Rea Point Terminal

The Rea Point Terminal was the fore-runner of Telesat's air transportable earth stations. Conceived as a quickly erectable air transportable earth terminal it was built as a company funded venture. Since Aug. 1974 the terminal has been in service in Rea Pt. NWT and provides private line service for Panarctic Oils from the Rea Pt. base camp to their Calgary office via Allan Park.

This terminal is fully redundant, uses a 12 ft. diameter antenna and provides a G/T of 12 dB/^{o}K.

The terminal was assembled by Telesat Engineering using Raytheon type Thin Route II communications equipment and was airlifted to Rea Pt. by a Lockheed Electra.

2.5 Leased Message Transportable Program

Business forecasts early in 1974 prompted the study of low cost air transportable message terminals particularly for service on oil rigs, in mining camps, with mobile exploration crews etc. In order to meet potential market requirements in the shortest time frame, Telesat signed a two year lease agreement with Hughes Aircraft Systems International (HASI) in September 1974 for five air transportable message terminals and the corresponding channel equipment for Allan Park.

Each earth station was initially equipped for 1 voice and 1 Telegraph (110 baud) circuit on a "speech plus" basis. The stations were delivered to Telesat in the March/May 1975 period and several are in commercial service. These terminals use FM-SCPC together with threshold extension demodulators. Twelve foot diameter antennas are used in conjunction with GaAs FET transistor low noise amplifiers, resulting in a G/T of 14 dB/^{o}K.

The design of these earth stations is unusual in that the station electronics consists of an outdoor unit bolted to the antenna support structure and containing the transmitter and receiver, and an indoor unit containing channel unit, signalling equipment and frequency references. The indoor unit may be operated within the customers premises in such a way as to avoid the

necessity of constructing a special shelter.

The terminal is highly transportable in a light aircraft (e.g Twin Otter) and can be erected by two men within a few hours. Approximate physical characteristics are as follows:

Weight -	Antenna	720 lbs
	Electronics (inc. crates)	600 lbs
Power Requirement:		less than 2 KVA

If permanent operation at a site is not required, temporary antenna foundations can be prepared from locally available materials such as timbers, concrete blocks, sandbags, etc.

The major significance of this type of terminal is that communications is available anywhere in Canada within hours or days compared with previous earth station implementation or alternative techniques which required several months of preparation. In addition to augmenting Bell's existing Thin Route Network, transportables can be used to provide private circuits between two locations on a short or long term basis. *

* A transportable possesses sufficient EIRP to allow any of the following types of pre-assigned circuits: transportable to TRII fixed station; transportable to transportable; transportable to Allan Park.

2.6 Purchased Message Transportables

In September 1974 Telesat awarded a contract to Raytheon Canada to design and supply fifteen Air Transportable Thin Route earth terminals. These terminals will form part of the permanent inventory of Telesat and will eventually replace the leased terminals.

Each earth station will disassemble into small packages and will be capable of transportation in a Twin Otter type of aircraft including shelter and antenna in a single load. The station uses a 12 ft. antenna, and transistor LNA to yield a G/T of 14.2 dB/K. A single 20 W transmitter is used.

Delivery of the first terminal is expected early in 1976.

The channel unit in the SCPC equipment supplied under this contract merits special mention. The design is based on the Raytheon Thin Route II design (which uses DM-2Ø CPSK at 40 kbps) but has substantial additional data carrying capacity.

For example the channel unit can operate in the following modes:

Mode 1 - Voice only (40 kbps)

Mode 2 - Voice (~~3.75~~ 37.5 kbps) plus 4 X 50/110 baud TTY channels (2.5 kbps) *

Mode 3 - Voice (37.5 kbps) plus 1 X 2400 bps data channel (2.5 kbps) *

Voice activation of the transmitted carrier is used to energise the carrier only at those times when voice or data is actually present. This technique is capable of reducing transponder loading by 3 dB or more, or conversely, of increasing the system full load capacity.

A further design improvement is the use of digital bit stream signalling, a technique which is expected to result in high integrity dialling at channel bit error rates (BER) of 10^{-2} or more.

As all other stations operating in the Thin Route system, the transportables operate on a preassigned channel basis.

Home base for the transportable terminals will be in the Telesat Headquarters area. While primarily designed to provide a service to short term users, the transportables are equally capable of providing a rapid reaction to requirements for extended service.

---

* The voice, data and teletype signals are carried on separate voice circuit backhaul facilities and may be independently switched.

Interfaces with the customer/user are as follows:

Voice - 4 wire VF interface 600 ohms balanced

TTY - current loop

Data - EIA RS-232-C

Capital costs for this type of terminal are in the region of $80,000 in small quantities.

2.7 CNT Program

The CNT (Canadian National Telecommunications) program is important for two major reasons:

a. It will be the first operational hybrid Thin Route/Heavy Route communications link.

b. It marks the start of Thin Route service using the Lake Cowichan Heavy Route earth station.

Three communities, Sachs Harbour, Holman and Snowdrift (all NWT) will be provided with earth station facilities and telephone service in Spring 1976. These communities are in an area served by Canadian National Telecommunications and to provide service to them, CNT has leased a fraction of the Thin Route Transponder from Bell.

In order to minimise the utilisation of the Thin Route Transponder the system is configured to employ SCPC DM-2Ø CPSK in the remote station to L. Cowichan links only. For the L. Cowichan to remote station links, the voice circuits are multiplexed into the CN portion of the 960 channel West-East message baseband (leased jointly by TCTS and CN-CP) and transmitted using single carrier per transponder FM.* This hybrid configuration has three disadvantages:

- A remote terminal G/T in excess of 22 dB/^{o}K is required to ensure the 960 carrier is received above threshold

- The two-way channels provided are not compatible with the rest of the SCPC network and therefore could not be integrated into a demand assignment system.

- The capacity of the West-East carrier is reduced, leaving an "excess" capacity on the East-West carrier. Since the 960 service operates well below capacity this is not a practical consideration at the moment.

The remote terminals will use a 15 ft. diameter antenna and 100^{o}K LNA to yield a G/T of 22.5 dB/^{o}K which with a conventional FM discriminator will provide a message channel noise performance

* If specifically designed for the purpose, an FDM/FM-SCPC hybrid scheme probably offers the most cost effective solution to a North-South only traffic requirement.

of 44 dBrnCO. A 35 watt transmitter is used to support the transmission of two voice circuits per site into L. Cowichan.

CNT interfaces with Telesat are at:

a. VF in the Thin Route portion.
b. FDM Baseband in the Heavy Route portion - CNT arranges mux/demux.

Although the system is designed to make most economical use of the space segment, flexibility is limited and single hop North to North communications is precluded by this hybrid approach.

2.8 Northern Ontario Program

The Northern Ontario program is the most recent major Thin Route implementation program and will provide service to nine remote communities in N. Ontario. Telesat's customer is Bell Canada.

Earth stations will be installed in two phases

Phase I - 1976	Phase II - 1977
Fort Hope	Cat Lake
Kesabonika	Slate Falls
Lansdowne House	Muskrat Dam
Weagamow	Sachigo Lake
	Ogoki

1976 sites will be fully redundant; 1977 sites will be non-redundant.

The SCPC equipment will be functionally similar to that used in the Raytheon transportables but in this case will be provided by Canadian Digital Communications Ltd. RCA Ltd. are contractors for the radio equipment portion.

The remote stations will use 15 ft. diameter antennas, transistor LNA's and provide a G/T of 19 dB/K.

All remote traffic will home on Allan Park earth station on a preassigned frequency plan. From Allan Park all traffic will be back-hauled to a Bell switching centre in Thunder Bay, Ont.

3. NETWORK CONSIDERATIONS

3.1 Communications Patterns

As explained in section 2 the Thin Route System is composed of many small remote earth stations communicating with two large earth stations in southern Canada. If the CNT program is excluded all traffic is between the remote stations and Allan Park on fixed, pre-assigned channels. This reflects Bell Canada's requirements and represents the simplest network configuration. With this type of network, North-South connections require one satellite hop while North-North connections require two hops.

The main disadvantages of the double hop technique from the subscribers point of view are:

a. degraded noise performance

b. double delay.

From an economic point of view, the double hop connection uses two voice channels instead of one. This inefficient use of transponder capacity may not be significant in a lightly loaded transponder but will become so when the RF transponder approaches channel fill. At this stage some form of demand assignment

will become necessary to permit direct remote to remote calling rather than expanding into another RF channel.

3.2 Backhaul - Voice Circuits

3.2.1 General

The common carrier selects the appropriate technique for backhauling voice circuits from the earth station to the switching facilities.

3.2.2 Remote Stations

Typically the remote station is located within a mile or two of the community it serves. Most such communities already have small central offices and the circuits between the central office and the earth station are carried via 4 wire VF land lines. All Thin Route II stations except Winisk use this approach.

In the case of Winisk a UHF radio link to the central office is used.

3.2.3 Transportable Stations

The leased transportables are flexible in terms of interconnection. They can either be connected to the central office as described in 3.2.2 or operated with a modified

4 wire telephone connected directly to the terminal.

3.2.4 Southern Termini

At Allan Park/Lake Cowichan the 4 wire VF circuits are multiplexed into the FDM baseband of a conventional high capacity radio relay system and backhauled to Toronto/Vancouver for further distribution as required. Some examples of the central offices handling Thin Route circuits are:

i Ottawa - Thin Route I and II except Winisk

ii Cochrane - Winisk

iii Thunder Bay - Northern Ontario network

iv Hay River - CNT

3.3 Signalling

At present, Thin Route connections generally require operator assistance, the remote locations not being part of the Direct Distance Dialling network (DDD). However, in special circumstances, such as when the leased transportables are used, a "Foreign Exchange" service may be arranged, whereby the trans-

portable is assigned a southern exchange number and appears to users virtually identical to any other telephone on that exchange e.g. a transportable in Sandy Lake, Ont. may be assigned a Toronto exchange number.

The signalling techniques employed on the Thin Route Network are In-Band Single Frequency techniques. The signalling frequency used is 2600 Hz. Dial pulses are bursts of 2600 Hz at a rate of 10 pps. Signalling equipment is normally provided by the telephone company and located in its own central office. This is true, too, for echo suppressors, 2/4 wire termination sets and other VF termination equipment.

All supervisory and billing functions are the responsibility of the Telephone Company; Telesat is not involved in these aspects of the business. More detailed discussion of signalling and terrestrial/satellite network interfacing is given in Appendix C.

3.4 Operations and Maintenance

All remote Thin Route earth stations are designed to operate unmanned, and maintenance is performed on an as required basis. Operations personnel are dispatched from Ottawa HQ or from a small staff permanently located in Frobisher Bay, NWT, to perform maintenance. Allan Park is the main spares depot and is also the main repair and overhaul centre.

The time required to reinstate service at a remote site is a strong function of travel time rather than direct repair time since repairs are effected on a module replacement basis. Travel time may vary from several hours to several days depending on location, weather, and availability of scheduled and chartered air services.

Typical costs for a repair trip to a remote site are $500 - $1500 for 1 technician excluding salary, e.g. Pond Inlet earth station:

Ott - Mtl Airfare	36.00
Mtl - Resolute Airfare	474.00
Resolute - Pond I. Airfare	300.00
Accommodation - 3 days	150.00
Food	75.00
TOTAL	$1035.00

The use of local people as agents can prove quite effective in diagnosing equipment problems. Pre-requisites for this type of arrangement are:

a. more than one voice circuit
b. well instrumented equipment i.e., meters, lights, etc.

A more detailed description of Operations and Maintenance for the present system is given in Appendix D.

3.5 Circuit Availability

Circuit availability is here defined as the % of time any two-way circuit between a thin route station and Allan Park is available. One of n circuit availability is the availability of at least one two-way circuit from a thin route station to Allan Park. The following table compares in-service availabilities from the beginning of January 1974 till the end of March 1975.

SERVICE AVAILABILITY

	1974 (12 Months)	1975 (3 Months)
Individual Circuit	97.37%	98.42%
1 of 2 Circuit	98.10%	99.65%

In general the availability performance has steadily improved as subsystem bugs and operational difficulties have been ironed out.

4. EARTH STATION DESIGN AND PERFORMANCE

4.1 General

This section describes the design and performance of two representative Thin Route earth station configurations used by Telesat viz

i The Thin Route II Earth station

ii The Air Transportable Earth station

Raytheon Canada Ltd. supplies both types to Telesat.

4.2 Thin Route II Earth Stations

4.2.1 Earth Station Configuration

Table 3 summarizes the primary parameters of both the Thin Route II earth stations and the Allan Park earth station.

TABLE 3

EARTH STATION PRIMARY PARAMETERS

PARAMETER	NORTHERN TRII STATIONS	ALLAN PARK
Transmit Frequency	5985±18 MHz	5985±18 MHz
Receive Frequency	3760±18 MHz	3760±18 MHz
ANTENNA	26 ft. Dia. (non-tracking)	97 ft. Dia. (Tracking)
Tx Gain (dB)	50.5	62
Rx Gain (dB)	48.5	59
G/T (dB/^{o}K)	22/19	37
LNA Noise Temp. (^{o}K)	300/700	100
TRANSMITTER	2x35 W TWT Hybrid Combined at O/P	1.5 KW KLYSTRON (Redundant Multicarrier)
POWER - Main - Back-Up	Commercial AC 4 hr-lead acid Battery	Commercial AC Diesel/Battery
STAFFING	Unmanned	Manned-24 hrs.

4.2.1.1 Communications Equipment

Figure 2 shows the basic communications equipment configuration at a typical northern earth station.

4.2.1.2 Transmitters

The transmitters consist of two identical units coupled at the output in a 3 dB hybrid. Each transmitter uses a high level upconverter and a 35 W output TWT. Single upconversion is employed.

4.2.1.3 Receivers

The main LNA is a single stage parametric amplifier followed by a 3.7 - 4.2 GHz transistor amplifier; the main LNA noise temp is 300^{o}K. The standby LNA is a multistage transistor amplifier with a noise figure of approximately 5 dB.

The two redundant receivers accept TR carriers at the LNA (4 GHz) input at a level of approximately - 110 dBm per carrier and deliver - 21 dBm per carrier at the 7- MHz IF output. Double downconversion is employed with a first IF of 1112.5 MHz ± 18 MHz.

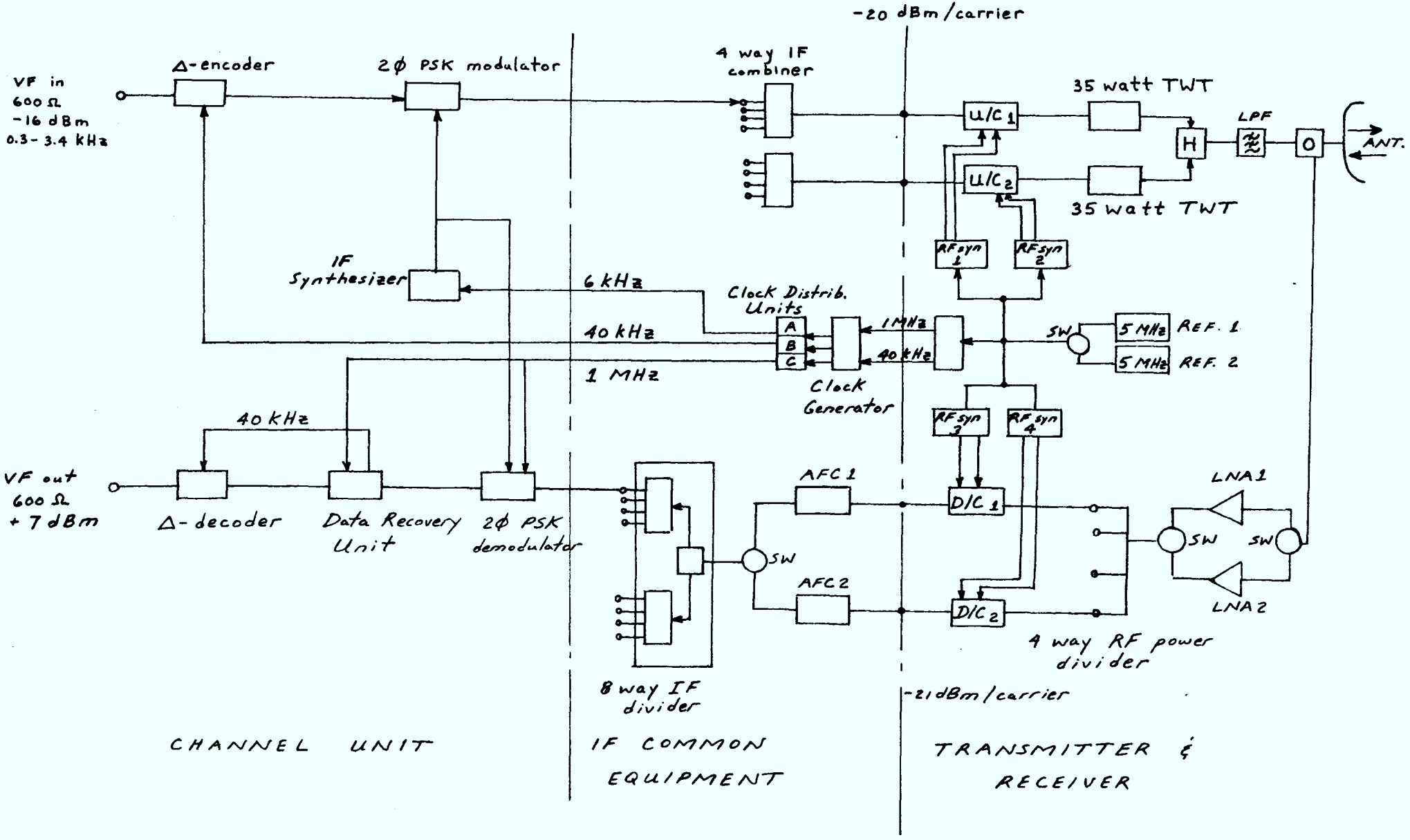

FIGURE 2 THIN ROUTE II COMMUNICATIONS EQUIPMENT

Both main and standby LNA's have 40 dB gain.

4.2.1.4 Frequency References

Ultra stable frequencies are essential to the Thin Route type of system particularly in the transmitters. The design approach adopted was to derive all station frequency references from a master oscillator with a long term stability of a few parts in 10^{10} per day. Synthesizer techniques are used to derive the RF local oscillator signals for the transmitters and receivers and the 6 kHz/40 kHz/1 MHz referneces for the baseband equipment. By this means all the derived frequencies retain the basic stability of the 5 MHz master oscillator.

4.2.1.5 Common IF Equipment

The common IF equipment consists of:

- clock generator and monitor unit (6 kHz)

- clock distribution units 6 kHz, 40 kHz, 1 MHz

- automatic frequency control unit (AFC)

- 8 way power combiner (passive)

- 8 way power divider (passive)

- metering and control module

- power supplies

All active common IF equipment is protected by full redundancy and automatic switch to standby upon failure of an online unit.

Of the common IF equipment listed above only the AFC unit requires further discussion. The function of the AFC unit is twofold:

a. To centre the full received IF spectrum by phase locking the incoming 69.96 MHz pilot tone to a locally generated stable pilot tone. The purpose here is to remove the effect of frequency offsets which are caused by drifts in the frequencies of both the satellite transponder L.O. and the earth station downconverter. The 69.96 MHz pilot tone is originated at Allan Park

b. To frequency translate the received IF spectrum to a new centre frequency of

93.28 MHz and perform spectrum inversion about the pilot frequency.

c. This function permits the same IF synthesizer to be used to select both transmit and receive frequencies.

(i.e., Transmit and receive frequencies are paired and are mirror images in frequency about the pilot frequency.)

4.2.1.6 Channel Units

The function of the channel unit is to digitally encode at 40 kbps a single analog voice channel, and phase modulate this signal onto an IF carrier generated by the IF synthesizer. On the receive side the same channel unit selects the required receive IF carrier, demodulates, re-generates and decodes the signal into its analog form.

One channel unit is thus required for each duplex voice circuit. Initially the remote stations will be equipped for two voice circuits per site with provision for plug in expansion for up to eight circuits.

A channel unit consists of

- 40 kbps Delta Codec (Encoder/Decoder)
- 2Ø PSK Modulator
- 2Ø CPSK Demodulator
- Data Recovery Unit
- IF Synthesizer
- Metering Unit and power supplies

CODEC

The codec employs a continuous variable slope delta modulation technique. The operation is such that the analog input is sampled at 25 usec. intervals (40 kilobits/sec) and encoded to produce a binary (1,0) output according to whether the present analog sample is greater or less than the previous sample.

The continuously variable slope feature is controlled by the digital output of the encoder and results in a substantially improved signal to noise ratio over a wider dynamic range of input signals than for a linear delta encoder.

The decoder at the receive end performs the inverse process to re-construct the analog wave form.

The codec also contains the 0.3 - 3.4 kHz base-band filtering.

The major objective parameters of the delta codec are summarized in Table 4.

PARAMETERS	SPECIFICATION VALUE	
Sampling Rate	40	kbps
Idle Circuit Noise	<37.5	dBrnC0
Dynamic Range (for S/Q > 25 dB measured at 1 kHz)	>30	dB
Maximum S/Q measured at 1 kHz	30	dB
Linearity	± 1	dB
Frequency response (measured at -10 dBm0, referred to 1 kHz		
300 - 3400 Hz	+1,-10	dB
300 - 3000 Hz	+1,- 6	dB
600 - 2400 Hz	+1,- 3	dB

TABLE 4

MAJOR PARAMETERS OF THIN ROUTE DELTA CODEC (LOW BER)

MODULATOR

The 2Ø PSK modulator accepts the digital TTL output of the encoder and the input frequency from the IF synthesizer and produces a modulated band limited (35 kHz wide) spectrum at the desired IF frequency.

AGC is employed to keep the modulator output constant for all channel frequencies.

IF SYNTHESIZER

The IF synthesizer is used to select, by means of Thumbwheel switches, any transmit/receive channel pair within the 36 MHz IF bandwidth. Channel frequencies are spaced at 60 kHz intervals. Thus there are 600 possible frequencies in the band.

DEMODULATOR

The input to the demodulator is a replica of the 93 ±18 MHz multicarrier spectrum from the IF power divider. The required channel is selected by the IF synthesizer, double downconverted, AGC'd, and preselected at 1.0 MHz. Coherent

demodulation is then performed using the local 1.0 MHz local reference signal. A data-aided carrier tracking loop (or decision feed back) is used to establish phase coherence of the PSK spectrum and the local 1 MHz reference.

The output of the demodulator is thus a noisy analog version of the transmitted data.

DATA RECOVERY UNIT

The function of the data recovery unit is to generate a 40 kHz reference clock from a 1 MHz local reference clock and phase lock it to the basic incoming 40 kbps data. When this is done the recovered 40 kHz is used to provide timing information to mid-bit-sample and re-generate the output of the demodulator which is then fed to the decoder for recovery of the analog signal.

VF INTERFACE

(Not strictly part of the channel unit) The VF interface is a jackfield (type 310 jacks) from which all VF testing may be accomplished.

Input Impedance - 600 ohm balanced to ground

Output Inpedance - 600 ohm balanced to ground

Input test tone - -16 dBm 1 kHz

Output test tone - + 7 dBm 1 kHz

4.2.2. Shelter Design

In the northern stations the electronics equipment is housed in a 20' x 8' x 8' fiberglass shelter. Outside ambient temperatures may range from -50°F to 80°F over the year so that active thermal control is necessary to maintain the inside ambient temperatures at 70°F ± 20°F. A thermostatically controlled heater/chiller unit is provided together with intake fans to accomplish the climate control.

Automatic de-icing is provided for the antenna feed horn window and sub-reflector by means of infra-red lamps. No de-icing provisions are made for the main reflector.

4.2.3 Power Subsystem

The station power system is comprised as follows:

Primary Power: - 208/120 V 3-phase

OR - 240/120 V 1-phase

Communications Equipment Supply	- 24 Vdc from the station battery charger
Standby power	- 4 hours provided by lead-acid batteries for communication equipment only
Utility power	- 120 V 1-phase

4.2.4 Voice Channel Acceptance Tests

The following tests are performed on each channel unit pair over the satellite as part of the acceptance test demonstration.

- Input/Output Test Tone Level Setting

- Signal to quantization noise as a function of input test tone level (1 kHz)

- Dynamic range (1 kHz)

- Input output linearity

- Frequency Response at -10 dBmO

- Idle circuit noise

- Audio crosstalk

- Gain stability

4.3 The Air Transportable Earth Station

4.3.1 General

The Raytheon Air Transportable Earth Station consists of a 12 ft. diameter antenna, an equipment shelter with environmental control and a communications equipment rack. All of these items can be disassembled and packaged into units weighing not more than 150 lbs. - each easily handled by two men. The station can be erected and operational in under eight (8) hours. When in operation the station will provide voice circuit capability with four channels of teletype or one channel of 2400 B/s data.

4.3.2 Transmitter

The transmitter (see Fig. 3) accepts a carrier from the channel unit in the band 52-88 MHz. This carrier is translated to a 6 GHz satellite transponder channel by the upconverter. The upconverter local oscillator is locked to a high stability 5 MHz reference signal which also provides clock frequencies for the channel unit(s).

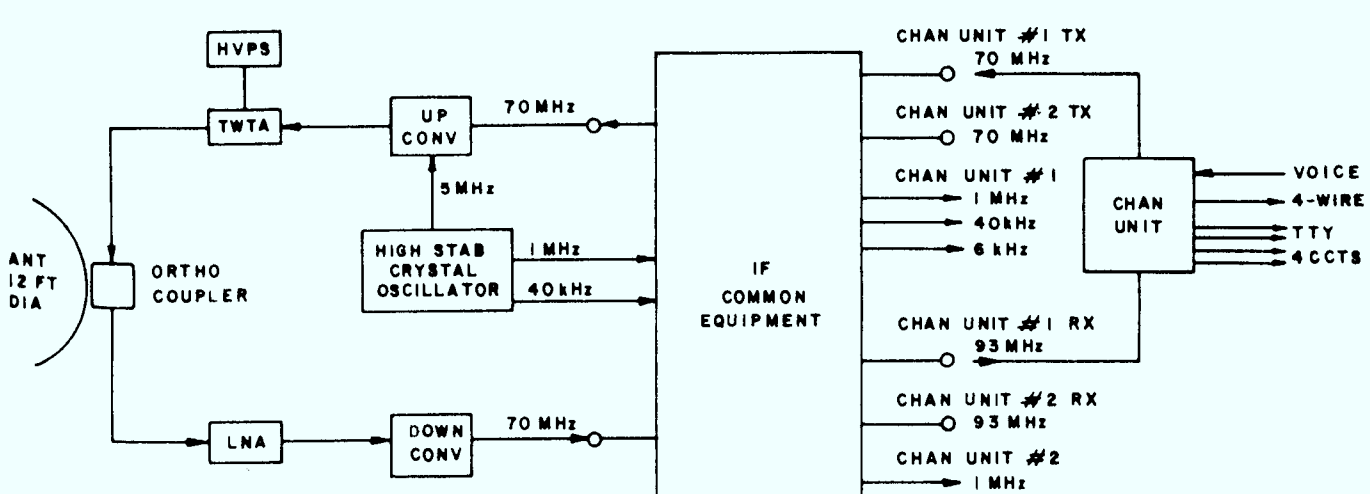

FIGURE 3 - EARTH STATION CONFIGURATION

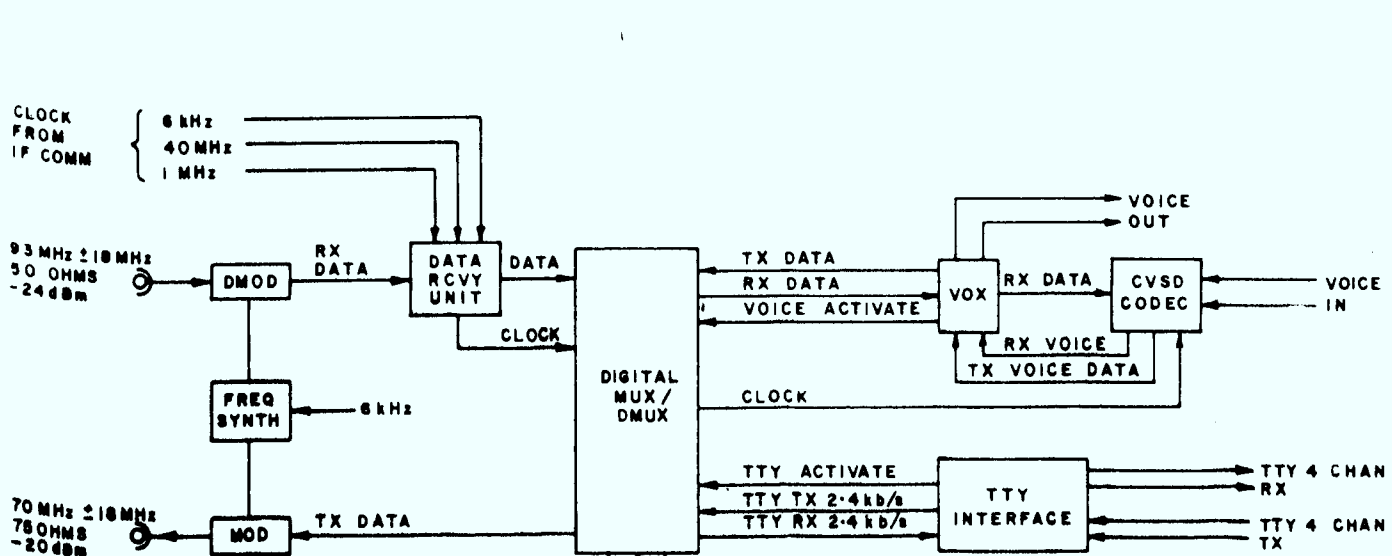

FIGURE 4 CHANNEL UNIT

The upconverter output is amplified in a travelling wave tube amplifier capable of providing more than 10W output to the antenna transmit port. The 6 GHz carrier is then radiated towards the spacecraft by the antenna.

4.3.3 Receiver

The down-link carrier from the spacecraft is in a channel 36 MHz wide in the 4 GHz band. This carrier enters the antenna which provides more than 41 dB of gain. The signal from the antenna receive port is amplified a further 40 dB by a low noise transistor amplifier having a noise figure of 3.5 dB. The low noise amplifier (LNA) then feeds the downconverter where all of the carriers in the down-link transponder channel are converted to the 70 MHz band. i.e.52 - 88 MHz.

4.3.4 IF Common Equipment

The IF Common Equipment provides for combination and separation of the transmit and receive carriers to and from each channel unit. In addition, clock frequencies at 1 MHz and 40 kHz derived from the high stability crystal oscillator are distributed via the IF Common

Equipment. The most important function is the centering of the down-link spectrum. In this process the IF Common locks on to a pilot at 69.96 MHz keeping the output link spectrum centered at 93 MHz and removing any satellite and receiver local oscillator offset.

4.3.5 Channel Unit

In the transmit path, (see Figure 4) a voice signal between 300 and 3400 Hz is digitally encoded by means of a continuously variable slope detector (CVSD) encoder. The voice data together with up to four 110 baud teletype signals are then digitally synchronized and multiplexed to produce a composite 40 kbps signal. This modulates the carrier in a 2Ø CPSK modulator to give a 70 MHz transmit carrier. Voice activation (VOX) is also provided to improve transponder channel efficiency.

In the receive path the channel unit selects the required carrier from the 93 MHz output of the IF common equipment. The selected 2Ø CPSK carrier is demodulated and the 40 kbps data is recovered and demultiplexed thus providing the voice data and teletype data streams. The voice data is decoded by the CVSD decoder and the four channels of teletype are demultiplexed and distributed to their

respective interface points. For voice and data operation, the TTY module is replaced with a data module. This permits a 2400 bit data stream to be combined with the voice data and permits operation with a 201B modem.

4.3.6 Technical Specifications

4.3.6.1 General

Environmental	- -50°C to +40°C outside temperature
Power Requirements	- 115 Vac 60 Hz 2.5 kVA
Weight	- total station, shelter antenna and all equipment with one channel unit and all transit cases for aircraft transportation .. 2300 lbs.
Shipping Volume	- complete station as defined above will fit inside a deHavilland "Twin Otter" or other similar light aircraft.

4.3.6.2 Transmitter

Frequency Band	- 5925 - 6425 MHz
Output Power 6 GHz	- 10 W
Gain Frequency Response	- ±0.5 dB over ±18 MHz
Local Oscillator Stability	- 5 parts/10^{10}/day
Input Frequency	- 52 MHz - 88 MHz
Input Level	- 20 dBm per carrier
Input Impedance	- 75 ohms

4.3.6.3 Receiver

Frequency Band	- 3700 MHz - 4200 MHz
Receiver Noise Temp	- 400^{o}K

Figure of Merit (G/T) - 14.25 dB/^{o}K

Output Frequency - 52 MHz - 88 MHz

AFC Pull In Range - ±40 kHz

4.3.6.4 Channel Unit

Frequency Agility - 600 frequencies every 60 kHz in the band 70 MHz ± 18 MHz. Can be selected remotely.

Teletype Capacity - 4 TTY channels 110/50 baud

Data Capacity - 2400 bps

Modem Bit Error Rate - BER = 10^{-4} at E_b/N_o. = 10.9dB

- BER = 10^{-2} at E_b/No. = 7.0dB

TTY Character Error Rate - 1 in 10^{6} for modem BER = 10^{-4}

Idle Circuit Noise - 33 dBrnc0 at BER-10^{-6}

- 38 dBrnc0 at BER-10^{-3}

Signal to Quantizing - S/Q>24 dB; 0 dBmO to
Noise for 1 kHz -30 dBmo
sinusoid

Audio Linearity - ± 1 dB: 0 dBmO to
-30 dBmO

Audio Gain Stability - ±1 dB in 24 hours.

COST EFFECTIVE OPTIONS FOR THE THIN ROUTE AND TV BROADCASTING NETWORK OF NORTHERN CANADA

FINAL REPORT

APPENDIX B

SINGLE CHANNEL PER CARRIER (SCPC) PERFORMANCE ANALYSIS

The work reported here was performed by
Miller Communications Systems Limited
for Dr. Roger Kaye, Research Branch, D.O.C.
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April, 1976

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APPENDIX B

SCPC PERFORMANCE ANALYSIS

1. INTRODUCTION

This Appendix gives a detailed evaluation of viable thin route satellite system techniques for telephony communications. Various single channel per carrier access arrangements and two modulation/processing techniques, namely pre-emphasized companded FM and companded Δ-modulation/coherent phase shift keying, are considered in relation to system capacity obtainable and earth station performance requirements. This information is combined with earth segment cost data in the body of the report to arrive at the optimum method of providing thin route telephony communications to and between sparsely populated northern communities.

Only Anik I type 4/6 GHz satellite technology is considered here. This is necessary to restrict the size of the study and also it is felt that this technology is most applicable for the time frame considered (next 5-7 years) and the type of service and region being considered. More specifically:

1. There already exists an extensive thin route service to the North on Anik I serving approximately 25 communities with plans for expansion (see Appendix A for description of existing system). Thus, to derive maximum benefit from any new system, compati-

bility or inter-connectability with the existing system should be of major consideration.

2. There is normally no requirement for backhaul to communities in the North as there is no interference problem to restrict the location of the earth stations near or in the community. Thus there is no economic penalty due to backhaul employing 4/6 GHz as there is in the South. Also to provide the same space link availability at 12/14 GHz as 4/6 GHz will require larger link margins which results in less efficient utilization of power.

3. There has been considerable technology development in the area of small ground terminals operating in the 4/6 GHz band especially in the area of solid state devices which aid in significantly reducing earth station costs and improving their reliability. This same level of development does not as yet exist in the 12/14 GHz band.

4. The thin route service coverage area and inter-connectivity requirements are such that little advantage can be derived from narrow satellite spot beams more easily available in the 12/14 GHz band.

This analysis considers the design and optimization of non-homogeneous networks (i.e. networks possessing unequal earth station G/T's and

carrier powers); however, capacity vs G/T results are presented only for homogeneous configurations.

A minimum cost algorithm for selecting successive earth station G/T's to meet the known requirements of a growing network is presented. This algorithm can include constraints on the range of G/T's considered and refurbishments of existing stations.

No allowance is made in link budgets for interference from adjacent satellites, as the CCIR has not yet finalized recommendations on this subject. Nor are constraints on up-link off-axis EIRP considered explicitly.

An interference model is presented which will permit the addition of the effects of interference to all system calculations presented. The problem of adjacent satellite interference and associated regulatory aspects are considered briefly in the body of the report.

2. DESIGN CONSIDERATIONS

2.1 Access Techniques

There are many available methods of modulating the carrier and accessing the satellite in communication systems. The ones that are relevant to a network application such as envisioned for the remote areas of the north are those which efficiently utilize the satellite power/bandwidth but still provide for many earth stations with a few circuits each, are easy to implement and expand, and do not require sophisticated and complex equipment.

Time Division Multiple Access (TDMA) does not satisfy the requirement of relatively inexpensive and simple terminal equipment, and, in fact, has not yet seen commercial service. Furthermore, unless a very large frame length is chosen (with its associated penalty in buffer storage requirements), TDMA efficiency drops substantially when the number of accesses becomes large. Finally, the implementation of digital speech interpolation (equivalent to voice activation) and demand assignment add significantly to terminal complexity, and are as yet untried.

Frequency division multiplexed (FDM) systems employing FM modulation and frequency division multiple access (FDMA) meet

the requirement for simple terminal equipment with a capability to expand the number of circuits. However, they do not make efficient use of bandwidth and power, especially when the number of circuits per carrier is small. Moreover, FDM/FM systems are virtually limited to pre-assigned operation due to the nature of the baseband multiplex arrangement.

The method that suits the stated criteria best is the use of a single voice channel to modulate an RF carrier with the carriers accessing the satellite transponder on a frequency division basis. This type of FDMA system if referred to as single channel per carrier (SCPC). SCPC is a flexible system as it can be designed to grow as the system capacity requirements increase. Thus it is possible to gradually increase the capacity at a node or by implementing voice activation and/or demand assignment techniques substantially increase the overall system capacity to meet the network growth requirements in a graceful fashion.

There are two access techniques most commonly employed in SCPC systems at the present time. These are 'fixed' assignment access and demand assignment access systems.

The fixed assignment access, as the name implies, employs pre-assigned frequencies for transmitting and receiving RF carriers at each earth station. Thus each voice circuit be-

tween any two nodes requires a dedicated pair of frequencies and transmitters and receivers.

The advantages of a pre-assigned system are:

1. It requires a minimum of earth station sophistication (e.g. non-agile receivers and transmitters, little or no software or switching equipment)

2. It requires a minimum of system control and supervision.

The major disadvantage of a fixed assignment system is its poor utilization of space segment bandwidth and its inherent inflexibility to short term traffic load fluctuations between network nodes.

It is possible to substantially increase the capacity of an SCPC system by utilizing a voice activation scheme. Voice activation is a method by which advantage is taken of the idle periods that exist in speech. These idle periods are detected and the earth station transmitter is switched off during the pauses in speech. This has two effects that improve the utilization of the space link namely (1):

- by conserving the transponder power during pauses in speech

it is possible to increase the overall number of voice channels simultaneously accessing the transponder.

- with the increased randomness of location of active carriers at any instant in time and the fact they always appear modulated, the average intermodulation noise power falling within the noise bandwidth of any carrier in the transponder decreases, resulting in an intermodulation advantage which allows an increase in the number of active voice channels accessing the transponder. However, this advantage is offset by a lessened ability to choose a set of staggered carrier frequencies which reduces the number of IM products falling on carriers (2).

Employing voice activation it is possible to increase the power-limited capacity of an SCPC system by a factor of 2.5 - 3.5 times.

An example of a thin route fixed assignment system is Bell Canada's system serving Northern Quebec and Ontario and the NWT. The evolution and present configuration of the system are described more fully in Appendix A. Pre-assigned duplex channels between remote northern earth stations and the Heavy Route earth station at Allan Park are provided. There are presently no pre-assigned channels allocated between earth stations in the north. Thus a north-south call employs

one satellite hop and occupies two RF channel slots in the transponder whereas a north-north call will employ two satellite hops and utilize four RF channel slots. Figure 1 graphically illustrates how the transponder power and bandwidth utilization efficiencies of the presently configured system decrease as the percentage of north to north calls increases. Due to the higher G/T at Allan Park the power limited capacity with 100% N - N calling is only reduced by about 24%.

There is also the disadvantage of extremely long round trip circuit delays (1 second) and additional performance degradation due to the two hops.

The advantages of this arrangement are:

1. The transmit power requirement of the small remote stations are minimized since all the small stations are transmitting only to the large station at Allan Park. This results in a substantial reduction in the cost of a remote station, and also means that off-axis EIRP restrictions are less stringent.

2. The scheme allows for centralized billing and control of calls from the north to be accomplished in the south.

As traffic loading increases a pre-assigned SCPC system can evolve to a fully or partially variable demand assigned system.

In this arrangement, nodes or earth stations are not assigned fixed frequency RF channels in the transponder but rather any one or one of several of the available transmit/receive pair of RF channels can be assigned to any two earth stations on demand. Thus the pool of RF channels in the transponder are shared by earth stations and are available on request. This arrangement not only increases the utilization of the space system significantly over a pre-assigned scheme but also provides increased routing flexibility between nodes in the network.

The disadvantages of demand assignment are:

1. The earth station complexity and cost increase since frequency agile transmitters/receivers are required.

2. A method of control for frequency assignment, monitoring of system status and billing must be provided.

Various methods of implementing demand assignment and interfacing it with the existing telephone network and Telesat SCPC System are identified and compared in Appendix C of the report.

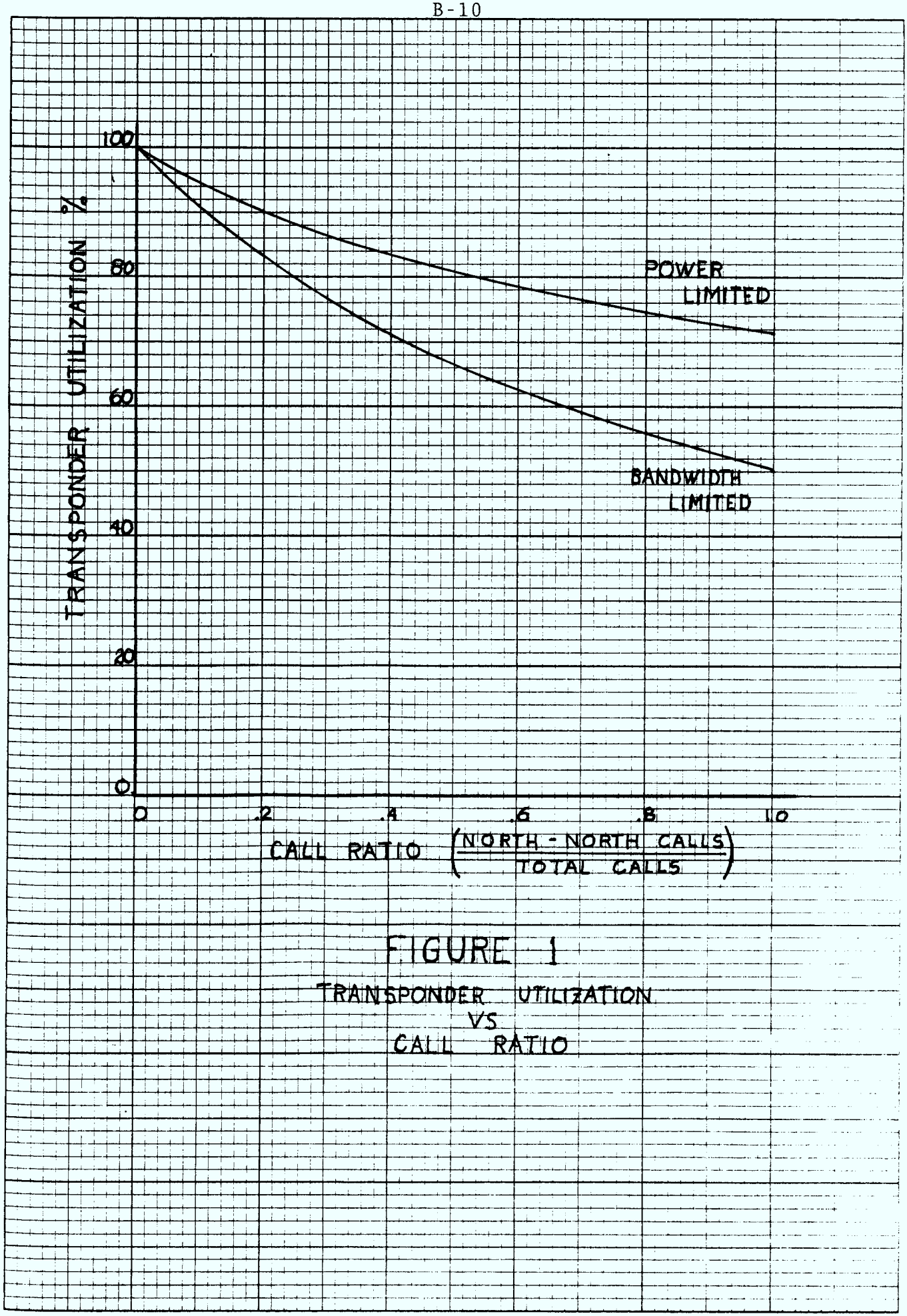


FIGURE 1

TRANSPONDER UTILIZATION VS CALL RATIO

46 0782

K+E 10 X 10 TO THE INCH • 7 X 10 INCHES KEUFFEL & ESSER CO. MADE IN U.S.A.

2.2 Performance Objectives

High transmission performance objectives lead to higher transmission costs. Thus high standards are usually associated with a service with high utilization and generally in areas of high population densities. In remote areas of northern Canada, it cannot be expected that usage would support the costs of implementing the higher performance systems found in the south.

As this analysis is one of several inputs supporting a choice of the most cost-effective modulation and access techniques for an SCPC satellite system, no specific level of performance constraint will be used; rather a range of performance levels that would be adequate for serving remote areas will be considered. In this manner, it will be possible to relate the effect of various performance levels on system capacity and hence ultimately on costs. Suggested standards and associated system costs are discussed in the body of the report.

Many parameters are associated with specifying the performance of a telephony system, however this analysis only considers two of the important ones, namely:

1. Desired circuit quality, usually expressed in terms of picowatts of idle noise or its subjective equivalent,

and, for the digital case, a transmission error probability.

2. Circuit availability, expressed as the percentage of the time a given minimum level of circuit quality is exceeded.

The availability aspect will be examined in detail in Appendix C of the report, and is considered here only insofar as allowing adequate link margins. Note that at 4/6 GHz the major portion of the availability budget selected for a satellite system is normally allocated to earth station electronics and associated outages; i.e. the availability budget is relatively independent of the earth station to satellite and satellite to earth station links.

Other circuit performance measures such as frequency response, dynamic range, voice-band data handling capability and subjective response are treated briefly in a non-analytical manner.

2.2.1 Circuit Noise Objectives

Three levels of circuit quality have been used in satellite systems and are discussed here briefly.

a. Major Traffic Toll Quality: This quality of telephone service would be equivalent to existing toll quality objectives for a 2000 mile long haul analog system in the south. Allowing for multiplex and rearward facilities, this represents a satellite link performance objective (from modulator input to demodulator output) of 5623 pWCo or 37.5 dBrnCo.

This performance objective is the same as that of Telesat Canada for its East-West 960 channel message service and FDM/FM/FDMA Northern Channel service into Allan Park (i.e. north-to-south circuits only).

b. CCIR International Satellite System Objective: Present international satellite systems have a mean circuit noise power objective at 10000 pWpO in a 3.1 kHz telephone channel for 80% of the worst month. This objective is based on a 2500 km hypothetical reference circuit (3) connection. This is approximately equivalent to a 40 dBrnCo noise level objective.

In addition the level of noise should not exceed 50,000 pWpO (47 dBrnCo) for more than

0.3% of any month. Assuming operation above threshold for an FM system this is equivalent to a 7 dB degradation in carrier-to-noise ratio , considerably more than expected for satellite systems operating at 4/6 GHz.

c. Tentative Northern Service Quality: As mentioned earlier the cost of a telephone transmission system is directly related to the constraint placed on circuit quality. At present, systems serving many of the remote communities such as HF and tropo-scatter systems do not provide the high quality offered by the major southern toll systems. HF radio is substandard in the aspects of service reliability and signal/noise quality; also medium length tropo-scatter systems operating in remote areas provide a performance of the order of 44 dBrnCo under median signal conditions which has been acceptable performance up to the present time. The difference in acceptable circuit qualities to North and South is reflected in the specification of the Telesat FDM/FM/FDMA northern message system which is designed to provide 37.5 dBrnCo into Allan Park and 44 dBrnCo into Frobisher Bay and Resolute Bay.

Thus a third level of circuit quality of 44 dBrnCo is considered which could provide an acceptable level of telephony service for intra-north communications.

A fourth level of circuit quality described as "mobile" quality in the 40,000 to 50,000 pWpO (46 to 47 dBrnCo) range is not considered for fixed service applications to remote locations.*

In summary, the three levels of quality service considered in the study are:

37.5 dBrnCo - equivalent to Southern Toll system standards

40 dBrnCo - CCIR International Toll quality objective

44 dBrnCo - possible acceptable quality level for intro-north toll circuits.

The idle circuit noise power relative to a 0 dBmO test tone level specification used above to define

---

* FM/SCPC equipment will be used in the Marisat system to provide this circuit quality to shipboard terminals.

circuit quality is derived from and is easily related to standard FDM/FM transmission performance. However, in the case of companded PCM, Δ-mod, or FM systems, the received noise power depends on signal level and therefore neither idle noise power not test-tone-to-noise ratio measurements directly define an FDM equivalent circuit quality. Therefore using either further analysis or the results of subjective tests we must relate the subjective noise performance of the SCPC systems considered to the above stated objective performance specifications.

This turns out in fact to be a contentious and as yet unresolved problem that introduces difficulties in comparing the performances of companded FM and Δ-mod circuits with each other and with standard reference circuits.

2.3 SCPC Modulation and Processing Techniques

The two types of systems examined here are pre-emphasized companded FM and variable slope Δ-modulation with phase shift keying modulation (4). Other alternatives have been considered and implemented (5), but these are clearly less appropriate to the requirement of providing communications to remote locations in Canada. In particular, companded Δ-modulation

was chosen over companded PCM because:

- it is simple and cheaper (e.g. Δ-mod does not require frame synchronization; recovers from phase reversals; and does not generate repetitive '0' or '1' patterns that can result in loss of bit timing)

- it degrades gracefully in the presence of noise

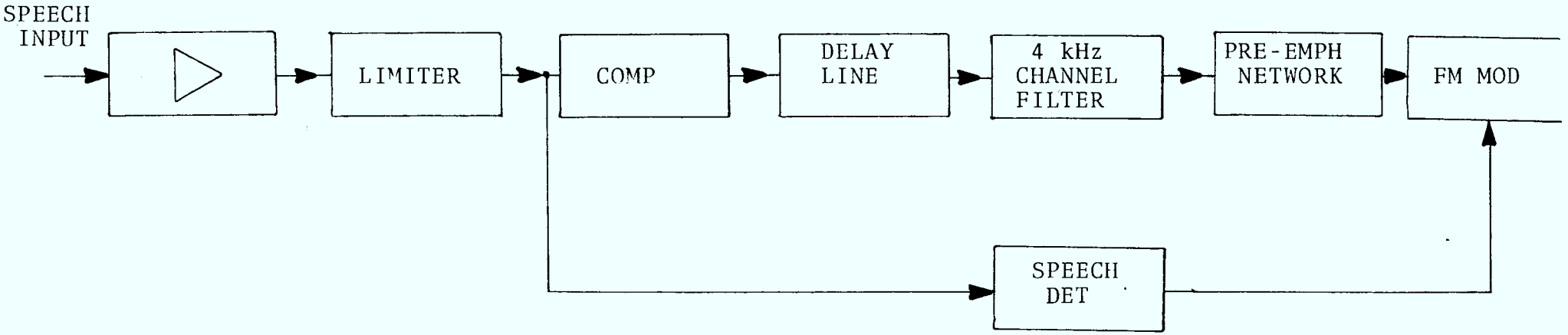
- it is more efficient in terms of transponder required.

2.3.1 Pre-emphasized and Companded FM (6)

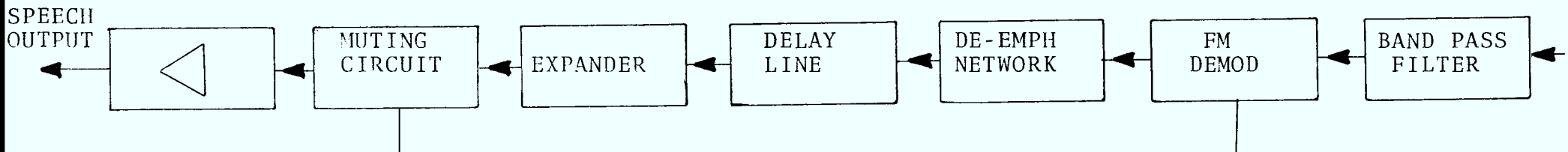
A functional block diagram of an FM SCPC channel unit with voice activation is presented in Figure 2. The following is a discussion of some of the discrete signal processing functions.

2.3.1.1 Emphasis

An FM system is characterized at the demodulator output by a noise spectral characteristic which rises at the rate of 6 dB per octave. Without the use of de-emphasis following the demodulator (and the complementary pre-emphasis preceeding the modulator) the noise output of the demodulator



A) TRANSMITTER



B) RECEIVER

FIGURE 2

would be almost entirely controlled by the noise power present at the high-frequency end of the FM demodulator output spectrum. This thermal noise can be made to appear flat with frequency if an emphasis characteristic is used which is complementary to the noise itself, i.e., a pre-emphasis which rises at 6 dB per octave across the communications band and a de-emphasis which decreases at the rate of 6 dB per octave. In actual practice, 6 dB per octave may not be desirable for one of several reasons. When a system is operating well above threshold, 6 dB per octave will minimize the total noise present as measured at the output of the system. However, this much pre-emphasis will substantially increase the high-frequency components of the transmitted signal passing through the FM system and can cause distortion if those high-frequency components are relatively large. This becomes a potential problem with the use of a full voiceband of voice-frequency carrier telegraph signals. For voice signals, the high-frequency components of the voice of loud talkers may reach levels which could exceed the modulation characteristics of the FM system. If the system is designed to include a hard limiting over-

deviation limiter preceeding the modulator but following the pre-emphasis, the high-frequency components of the loud talker will be peak clipped and there will be resulting distortion. However, this type of limiting has virtually no subjective impairment of the voice signal (7).

While the spectral characteristics of thermal noise rise at the rate of 6 dB per octave, impulse noise encountered in the threshold region of the FM demodulator is essentially flat in its spectral characteristics. The use of 6 dB-per-octave pre-emphasis to offset thermal noise will substantially increase the low-noise components of impulse noise, and if operation is anticipated at or near threshold, the 6 dB per-octave de-emphasis will severely degrade both the objective and subjective operation of the system. A compromise emphasis characteristic which is of the RC type with a knee at 1000 Hz, i.e., with a time constant of 159 microseconds will still give substantial improvement relative to thermal noise without such severe degradation in the impulse noise region. Across a voice band of 300 to 3400 Hz, 6 dB-per-octave emphasis will give an emphasis improvement of 6.3 dB, an

emphasis network of 159 microsecond time constant will give a measured emphasis improvement of 5.8 dB. The subjective improvement in the threshold region, however, with the 159 microsecond emphasis is several decibels better than 6 dB-per-octave emphasis and the 159 microsecond emphasis is a reasonable compromise permitting occasional operation in the threshold region without severely degrading the performance of the system due to impulse noise.

If operation at low carrier-to-noise ratios is anticipated as a normal operating point, then the predominant noise to be concerned with is impulsive, rather than thermal, and emphasis of any sort should be avoided. The Marisat system does not use emphasis because it does operate in the threshold region of its demodulator. Use of emphasis with that system would severely degrade its performance at the operational C/N_o.

2.3.1.2 Threshold Extension Demodulators (T.E.D.'s)

TED's used with FM SCPC systems all have the objective of increasing the operational range of the FM demodulator to lower C/N_o's than would otherwise be possible, thereby allowing more efficient use of satellite transponder power and bandwidth or permitting an earth station possessing a lower G/T to receive SCPC carriers without encountering an excess amount of impulse noise.

Two types of threshold extension demodulators are presently in use, phase-lock demodulators and tracking filter demodulators. Both demodulators have a characteristic of lowering the threshold of the system several decibels below that obtainable without threshold extension. In these applications, threshold is usually defined as the C/N_o or C/N at which the dB-for-dB relationship between input carrier-to-noise ratio and output signal-to-noise ratio is degraded by 1 decibel. At this point, there will be substantial amounts of impulse noise present in the output of the demodulator. It is important to

note that the apparent threshold of a demodulator is impacted both by the amount and shape of the emphasis response and by the demodulated bandwidth in which the noise measurement is made. For example, a TED demodulator measured without de-emphasis and with noise bandwidth limited only by the basic capability of the demodulator (about 8 kHz) can show a carrier-to-noise ratio of 5.5 dB at threshold but when the same demodulator is used with 159 microsecond de-emphasis and with its noise bandwidth limited to the upper end of the transmitted voice band, i.e., 3.4 kHz the measured threshold is 7.5 dB.

In a voice activated system the PLL (i.e. Phase Locked Loop) receiver must operate on bursts of received carrier with randomly varying durations and arrival times. Thus the PLL receiver must have a wide enough loop bandwidth to provide rapid acquisition to prevent speech clipping. This constraint may reduce the threshold advantage obtained with a PLL.

2.3.1.3 Voice-Operated Carrier (VOX)

The circuitry used for voice-operated carrier operation in FM systems is somewhat different than that which is appropriate for digital systems. The problem with transmit control of any voice-operated carrier system is associated with the detection of the voice signal and how to differentiate the presence of the voice signal from a momentary noise burst or impulse. In both systems, a delay line is used between the voice detector and the carrier-control switch. This delay is necessary to prevent clipping of the first syllable of a voice-operated transmission. In digital systems, this is done in a digital delay line, in FM systems it is done using "bucket brigade" chargecoupled MOS devices. These bucket-brigade delay lines will produce delays of several milliseconds in a small integrated circuit package. When properly designed they introduce essentially no distortion and little noise into the voice path.

At the demodulator end of the transmission path, it is necessary to have additional delay lines to permit control of the system squelch operation

without clipping of the voice and without permitting noise bursts to reach the output at the end of each voice-operated transmission.

2.3.1.4 Compandors

Compandors used in systems which are presently in operation are designed in conformance with CCITT Recommendation G162. These are 2:1 syllabic compandors with unaffected levels of 0 dBmO. Their purpose is to substantially reduce the amount of noise which is heard subjectively on the circuits. With systems whose subjective noise objective is 10,000 pWpO the syllabic compandor will give 15 to 17 dB noise improvement over a non-companded circuit, i.e., the weighted test-tone signal-to-noise ratio measured on a system before compandors are added, will be between 33 and 35 dB while achieving a subjective noise performance of 10,000 pWpO. The systems operating with lower non-companded test-tone-to noise ratios will have less compandor improvement because of the subjective nature of the compandor action. Further the compandor improvement is a function of talker levels, the 15 to 17 dB value being appropriate for the average talker; softer

talkers will have more compandor improvement, louder talkers will have less (8). Compandors can be used in channels transmitting alternate voice/data because the constant amplitude of a data circuit will "lock" the compandor gain at a fixed value and will not disturb data transmission. Note that the nature of the compressor is to raise the transmission level of a data signal going through the system. Since data signals are not usually transmitted at high levels which approach the modulation capabilities of the FM channel, this increase in transmission level through the FM system actually improves the performance of the data signal above that which it would be if compandors were not used.

A number of telephone administrations have expressed concern about the use of compandors with FM SCPC systems because of unsatisfactory past experiences. Syllabic compandors have two characteristics which could be disruptive when applied to telephone networks; they amplify any gain changes in the transmission path by two times, and if not properly designed, they can distort voice signals when certain types of

noise signals are present at the input to the compressor/expandor. In FM SCPC systems using compandors it is essential that the transmission path between compressor and expandor have excellent stability so that the gain variation doubling effects of the compandor will still yield a system with good gain stability. This is, of course, a manufacturer's design criterion. When the signal applied to a compressor is quite noisy, an inadequate compandor design will result in modulation of the voice signal by components of the noise. The resulting impairment is undesirable and it is unnecessary if the compandor designer has done his task properly. The concern of administrations about the use of compandors stems largely from these characteristics of compandors which can make their use undesirable if not properly controlled. Companded FM SCPC systems should not and need not suffer from these problems.

2.3.1.5 Deviation Limiters

With the effects of voice processing such as pre-emphasis and companding, which may be present in the FM terminal, it is important in FM systems

that the peak deviation be kept within the nominal IF passband of the demodulator. This is particularly true when operating close to threshold (which most FM SCPC systems do) because excessive modulation will increase the amount of impulse noise present, as well as have the potential for disturbing adjacent channels. Deviation limiters used with these FM systems are usually hard limiters rather than the relatively soft limiters used with frequency-division-multiplex (FDM) systems. Since limiters can generate harmonics and higher-order intermodulation products that fall above the normal voice band, limiters are followed by low-pass filters before the processed voice signal is applied to the input of the FM modulator itself.

2.3.1.6 Click Suppressors

For those systems which are designed to operate in the threshold region of the demodulator, the impulses or clicks represent a substantial amount of the noise which appears at the demodulator output. It is possible to detect the presence of these clicks and to suppress them.

While click suppressors may not suppress all clicks, particularly when the click rate is high, the suppressor can substantially improve both the measured and the subjective quality of a noisy channel.

For systems which are designed to operate above threshold in order to control the thermal noise output of the demodulator, the click suppressor is useful only for the small percentage of the time when transmission degradations will have reduced the C/N_o to where operation in the threshold region occurs. The complexity of the click suppressor and the distortion which it causes to modulation peaks suggest that it should only be used in systems in which impulse noise occurs under normal operating conditions.

2.3.1.7 Crosstalk and Intermodulation

A matter of considerable concern with analog systems such as FM SCPC is the potential of intelligible crosstalk due to non-linearities in the spacecraft transponder amplifiers. These are almost always operated as close to saturation

as possible in order to increase the radiated power from the satellite. But operation close to saturation will result in substantial amounts of intermodulation among the multiplicity of SCPC channels which may be present in the same transponder. Theory and measurement indicate that for the large number of carriers normally associated with an SCPC system intelligible crosstalk will not occur.

2.3.1.8 FM Noise

Proper SCPC design must keep other sources of noise besides thermal under control, such as intermodulation noise, VF circuit noise, power supply coupling, and FM noise. Good circuit design prevents power supply coupling and reduces VF circuit noise to a negligible level. FM noise usually originates in the frequency translation devices between the modulator and the demodulator. The translation stages are the transmit station frequency select synthesizer and upconverter, the satellite and the receive station downconverter and synthesizer. This noise can be drastically increased at low frequencies by the

receive de-emphasis network, whose response increases at 6 dB/octave with decreasing frequency. If the frequency converters and synthesizers are not adequately designed, low frequency distortion will be introduced and circuit quality impaired. This becomes more noticeable with lower deviations and puts a lower bound on acceptable test-tone deviations. Figure 3 shows typical FM noise contributions of various translation equipment. For comparison, thermal and satellite FM noise densities are also presented. Designers of the system must put significant amount of time and effort into optimizing crystal coupling, phase-lock bandwidths, and noise generating components to obtain the results shown in Figure 3.

2.3.1.9 Program Channels

FM SCPC is well suited for use as program channels for transmission of voice and music with frequency up to 15 kHz. The trade-offs among C/N_o, bandwidth, deviation, and performance make the design of such a system simple. Available compandors have noise improvements of about 10 dB for

K&E SEMI-LOGARITHMIC 46 5493
3 CYCLES X 70 DIVISIONS MADE IN U.S.A.
KEUFFEL & ESSER CO.

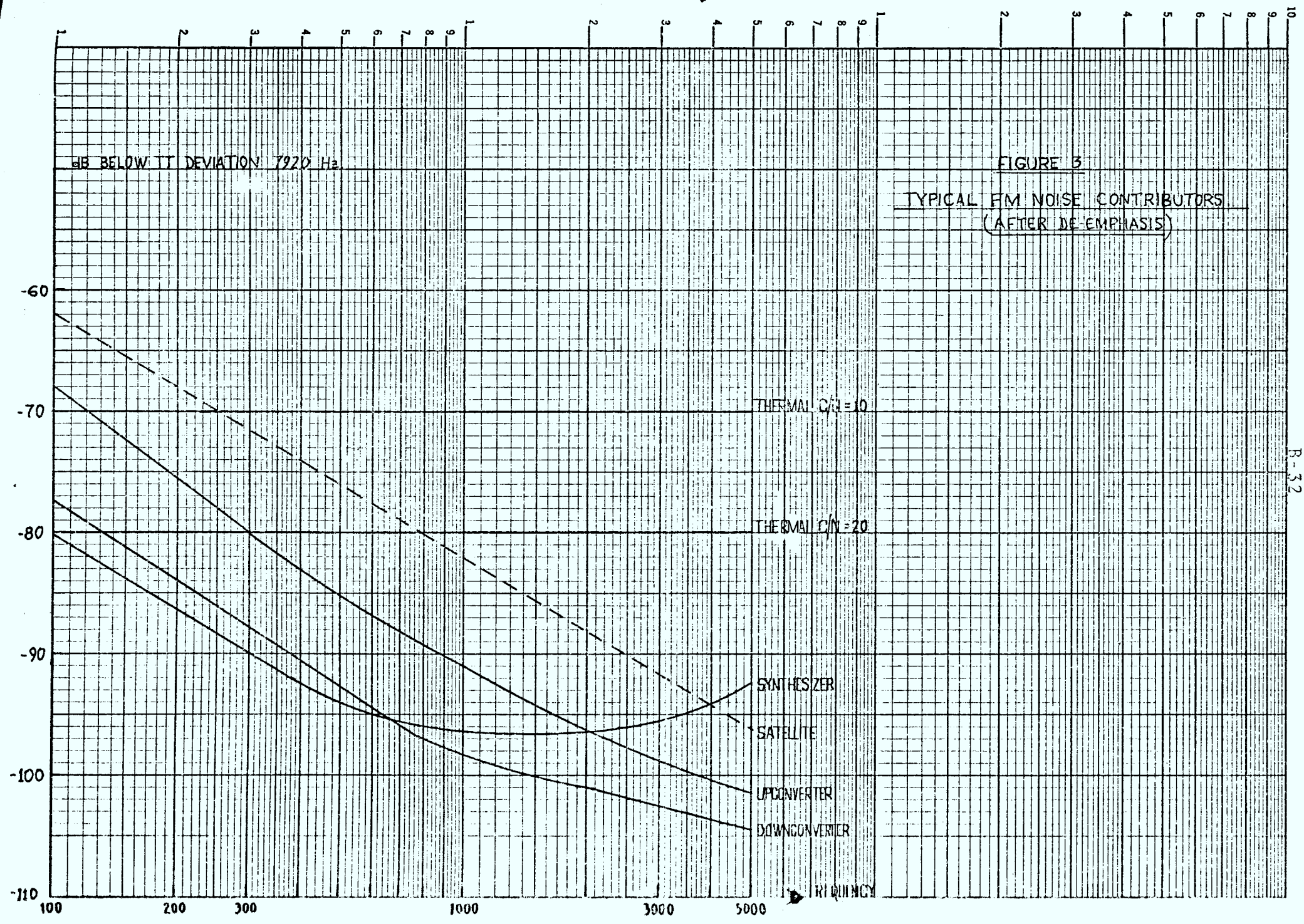


FIGURE 3

TYPICAL FM NOISE CONTRIBUTORS
(AFTER DE-EMPHASIS)

this type of application, and an emphasis network time constant of 75 μsec has proven to be suitable.

2.3.2 Variable Slope Δ-mod CPSK (4)

Figure 4 is a functional block diagram of a Δ-mod SCPC system employing 2 or 4 phase PSK modulation with differential encoding and coherent detection. The following is a description of the key elements of the system.

2.3.2.1 Variable Slope Δ-Mod Codec

This type of analog-to-digital conversion, discussed in detail in (9), involves predicting the next sample value from previous values and only encoding and transmitting the difference signal. If successive samples are highly correlated and the predictions are reasonably accurate, the difference signal will have a reduced dynamic range compared to the original signal, and therefore fewer bits are required to encode it.

With Δ-modulation, the signal is encoded to only

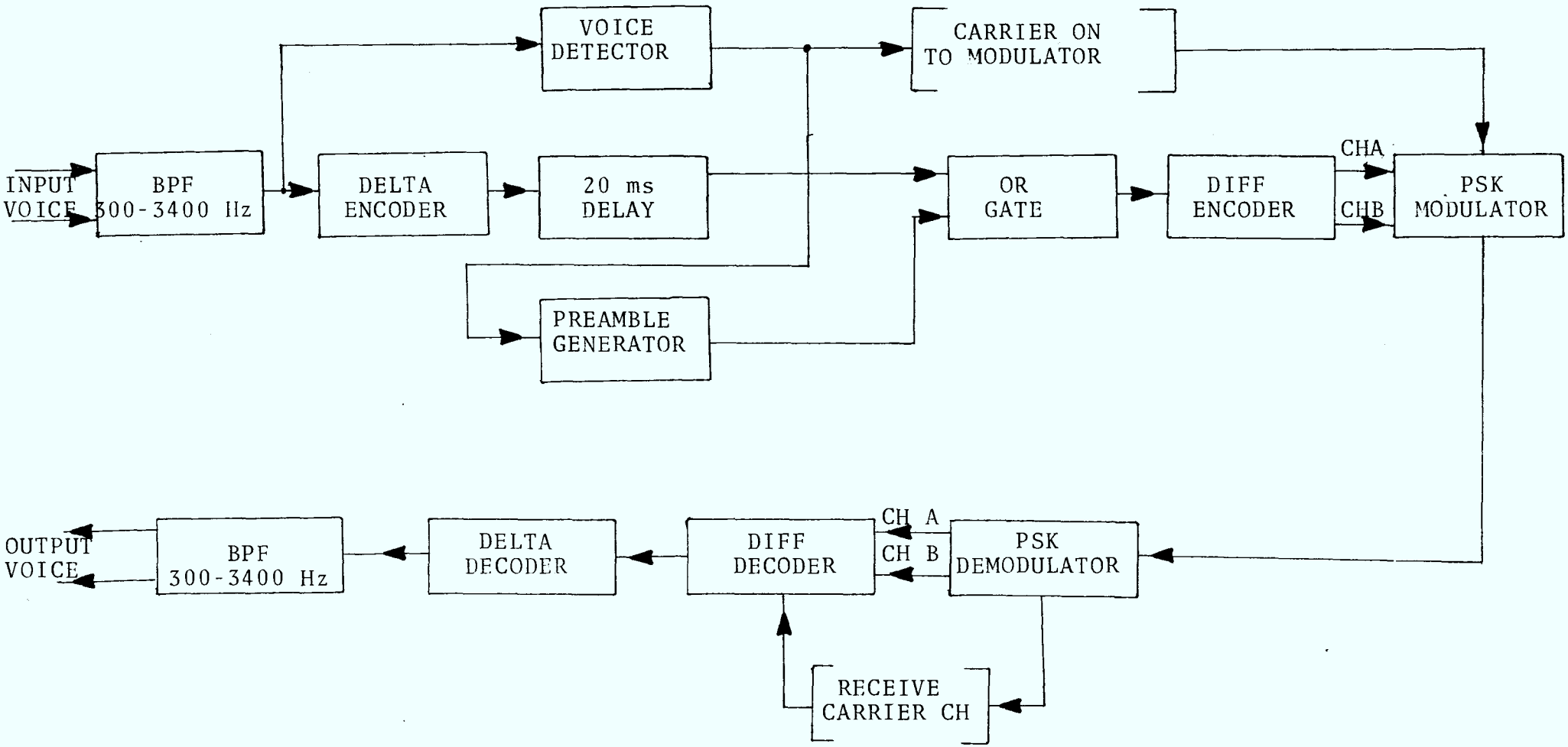

VOICE DETECTOR
CARRIER ON TO MODULATOR
INPUT VOICE
BPF 300-3400 Hz
DELTA ENCODER
20 ms DELAY
OR GATE
DIFF ENCODER
CHA
CHB
PSK MODULATOR
PREAMBLE GENERATOR
OUTPUT VOICE
BPF 300-3400 Hz
DELTA DECODER
DIFF DECODER
CH A
CH B
PSK DEMODULATOR
RECEIVE CARRIER CH

FIGURE 4

one bit per sample. With variable slope Δ-modulation, the levels are allowed to vary with time, dependent on past difference signal values to better accommodate both rapid signal fluctuations and nearly constant successive sample values.

Three encoding rates are considered in this study:

1. The 40 kbps rate presently employed in Telesat's baseline thin route system.

2. A 32 kbps rate which has been demonstrated in limited subjective tests (4) to match the performance of a 10,000 pWpO CCIR reference circuit.*

---

* In addition to providing more efficient space segment utilization, the 32 kbps rate offers the advantage of compatability with existing satellite and terrestrial 64 kbps facilities. For example, a 64 kbps modem can accommodate 1 PCM or 2 Δ-mod channel inputs (10).

3. A 20 kbps rate, which provides a circuit quality close to the 44 dBrnCo Northern service objective previously stated.

As Δ-modulation demonstrates a more graceful bit error threshold effect than PCM, a threshhold bit error rate (BER) of 10^{-2} is assumed for system analysis purposes. This error rate in fact corresponds to a noise level somewhat better than the 50,000 pWpO threshold level normally assumed in the design of telephony systems. Voice intelligibility through a Δ-mod system has in fact been maintained up to error rates near 10^{-1}.

In the Telesat baseline system a 10^{-3} BER threshold was assumed, and this is also considered here for comparison purposes in the case of the 40 kbps rate only.

2.3.2.2 CPSK Modem

The modem may be considered to be the heart of the SCPC channel unit because of the impact of its performance on that of the system and because of the design challenges posed by its specifications. The modulator portion of the

modem is fairly simple, and the same basic technique is used by most manufacturers. In the case of 4-phase PSK, two synchronous bit streams are accepted at half the bit rate and used to modulate quadrature components of a carrier. The two components are then combined in a passive summing device.

The demodulator portion of the modem poses a difficult problem since the designer is faced with specifications which call for the ability to track large frequency drifts relative to the symbol rate, lock up sufficiently rapidly to avoid impairing the speech signal at the beginning of each voice activated burst, and operate at relatively high bit error rates. The single most difficult design challenge is coherent demodulation in the face of conflicting requirements for low cycle skipping probability, fast acquisition, and the ability to track a large frequency offset. The second major function of the demodulator is coherent clock recovery. Several carrier recovery and bit timing techniques are available and have been reported extensively (11). When applied in a voice activated SCPC system these schemes all rely on the presence of

a pre-amble at the beginning of each burst. This means a pre-amble generator/combiner and additional delay (i.e., buffer storage) must be provided in the transmit portion of the channel unit (see Figure 4.)

2.3.2.3 Voice Operated Carrier

A digital voice detector is used to control voice activation of the carrier in a digital SCPC system. The digital voice detector can detect voice with a high degree of assurance within several bits providing the 300 Hz to 3400 Hz voice channel interfacing the SCPC system is sufficiently quiet. Implementation can consist simply of detecting a repetition of '1' 's or '0' 's (i.e., slope overload) from a rough linear Δ-mod encoder. The detection threshold or integration time can be altered to match the particular circuit noise and stability characteristics. In general, digital processing permits more rapid detection of voice than analog techniques.

2.3.2.4 Error Correcting Coding and Digital Data Transmission

Forward error correction (FEC) can be used to trade off transmission power and bandwidth in a digital communications system. Existing high rate FEC codes, however, offer little advantage to systems that are designed to operate up to a BER of 10^{-2}. Furthermore, modems may not operate well much above this error rate. Therefore error correcting coding is not considered further as a practical means of enhancing the performance of the SCPC Δ-mod system.

However, if the system is also used for the transmission of data in digital form, coding is an efficient and practical means of achieving a desired threshold error rate typically of 1 part in 10^{7}. Simple rate ½ convolutional codes which lower a demodulated error rate of 10^{-2} to less than 10^{-7} can be implemented quite cheaply. Rate 3/4 convolutional codes that employ viterbi (soft decision) decoding increase the data throughput at the expense of increased FEC codec complexity.

The design and performance of the SCPC system postulated in this study is assumed determined by voice communications requirements. The extent to which the SCPC links can accommodate voice band data must also be considered, since this capability is considered by many to be an essential feature of the switched telephone network. Consideration of SCPC interfaces other than standard voice band, however, is clearly beyond the scope of the study. Suffice it to say that the digital SCPC system offers obvious advantages in the transmission of digital data.

2.3.3 Echo Suppression

The same detectors used to control voice operation of a carrier can also be used in conjunction with the delay lines to provide a very effective echo suppressor. When designed with appropriate comparator circuits, time constants, and break-in circuitry, the VOX and its associated detectors will provide an echo suppressor function which has all the characteristics of the standard CCITT echo suppressor (Rec. G161). It is to be noted that the use of the delay lines would permit substantial reduction in the break-in hangover time given in the CCITT recommendation. A reduction of the hangover time

improves the subjective performance of the echo suppressor since by reducing the time in which the circuit is open in both directions it reduces the annoyance of echoes during this period.

2.3.4 Frequency Stability and Compensation

A commonality among SCPC systems is the necessity for providing a means of correction for drifts in the translation oscillators used in the spacecraft. SCPC systems are by their nature narrow band (as low as 27.5 kHz allocated bandwidth) and the need for high frequency stability is obvious with transmission frequencies in the 6 and 4 GHz regions. To accomplish this stability, most SCPC system designs include the use of a pilot signal transmitted from one station to correct common frequency errors, namely those occurring in the satellite and earth station downconverters. Errors in the transmitting station upconverters cannot be corrected through the use of a pilot, and their frequency stability must be excellent. Further system designs require that the frequency of the individual channels be readily changed, either manually or by DAMA (i.e. Demand Assignment Multiple Access) control. To achieve stability objectives and the frequency agility requirement, SCPC systems employ frequency synthesizers in the channel

equipment. Some systems have common synthesizers for transmit and receive directions which require that the two directions of transmission be "paired" in the system frequency plan (SPADE and Telesat's baseline systems are this way); other systems have separate synthesizers for the two directions of transmission giving considerably greater freedom in the design of a DAMA system.

The modulator and synthesizer circuits are typically designed within the following constraints (70 MHz IF assumed):

1. The modulation index must be constant for output frequencies between 52 and 97 MHz (FM case only)

2. The output frequency stability must be better than ±1 part in 10^8 per month.

3. Changing frequency must be accomplished rapidly (typically 20 ms) and accurately to any incremental frequency in the 52-97 MHz band.

4. Output RF power must be constant for each frequency.

2.3.5 Transmitter Backoff Requirements

Unlike an FM carrier, a band-limited digitally phase modulated (PSK) carrier exhibits substantial amplitude modulation, the degree of AM increasing as the symbol rate-to-channel bandwidth approaches the Nyquist limit. When such a signal is passed through the earth station power amplifier (e.g. TWT), which typically exhibits a non-linear input - output power characteristic and AM to PM transfer, the signal may be distorted and energy spread beyond the original frequency band (12). The in band distortion, resulting primarily from AM/PM conversion, degrades the bit error rate vs E_b/N_o performance of the link (13). The spread energy, resulting mostly from amplitude clipping, does not normally represent a substantial loss of carrier power, but can constitute an important source of interference to the adjacent SCPC channels. For either 2 phase of 4 phase CPSK signals, at least 1 dB of TWT output backoff is required to avoid severe non-linear distortion and an associated penalty in performance.

For single carrier per transmitter operation this backoff constraint results in a 1 dB increase in required power amplifier (PA) size. An FM carrier, on

the other hand can operate at saturation without undergoing significant distortion, and has an advantage in this regard.

For multiple SCPC carriers per PA, net envelope variations are dominated by the addition of randomly phased signals, and intermodulation is the dominant effect. Therefore there is little difference between the multi-carrier backoff requirements for FM and PSK carriers.*

2.3.6 Voice-Band Data Applications

For private-line data both the modulation level of the data carrier and the RF level of the SCPC carrier can be separately adjusted from the rest of the system to achieve whatever bit error rate may be required for such dedicated service. For alternate voice/data operation, an optional data transmission detector can be made available which, when used, will automatically disable the echo suppressor function during the period of data transmission. In addition, this same detector may be used in marginal applications to increase the level of the RF carrier when data is to be transmitted.

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* If anything an FM/SCPC is more sensitive due to the possibility of intelligible crosstalk when a small number of carriers share a transmitter.

Measurement of voice-band data transmission performance for FM and Δ-mod SCPC systems are presented and discussed in section 4 of the report. It appears that FM is superior in this regard, operating at rates up to 9.6 kbps, where 32 kbps Δ-mod is limited to 2.4 kbps or at best 4.8 kbps.

2.4 Earth Station Performance Considerations

One of the key parameters that determines earth station performance (and to a large extent the number of circuits the satellite transponder can support) is the receive gain to noise temperature ratio (G/T) of the station. For example, in the capacity analysis section of the report, the capacity of the various modulation and access arrangements is derived in terms of the earth station G/T.

The earth station equipment that essentially determines the G/T are the antenna gain and the first stage of the receiver low noise amplifier. Thus a trade-off exists for obtaining the desired earth station G/T. However, in most system applications, both technical and economic constraints exist on the antenna size and low noise amplifier.

For applications in remote areas of Canada where capacity requirements are typically low, access is limited to air trans-

port in many cases, and the cost of civil engineering and construction of antenna foundations is high, the maximum size of the antenna is economically limited to the order of 26'-33' diameter. The minimum size of antenna is constrained by three factors namely:

1. Systems employing small antennas with corresponding low values of G/T result in inefficient utilization of the satellite transponder.

2. Systems employing small antennas with corresponding low values of G/T require larger HPA's (high power amplifiers) both because transmit antenna gain is lower and because the required EIRP per carrier is higher. For the existing north-to-south SCPC system HPA costs are an important component of the total cost. When implementing direct north-to-north links HPA requirements become even greater. (For this reason future Anik's offering thin route service should include transponders possessing saturating flux densities less than -81 dBW/m^2).

3. The minimum size antenna is also constrained by interference considerations into adjacent satellites. This trade-off is complex depending on assumptions made on orbital arrangements, co-ordination of co-channel traffic,

side-lobe characteristics of the antennas used, polarization discrimination, etc.

For all of the above reasons the smallest size transmit/receive antenna to be employed in the Anik system is Telesat's 12' diameter transportable earth station.

Low noise amplifiers most commonly used for receiver front ends are:

- Parametric amplifiers whose noise temperatures currently range from 20°K for the cryogenically cooled units to approximately 300°K for uncooled paramps. Cryogenic paramps although providing a low noise temperature are expensive, relatively unreliable, and require more maintenance which renders them impractical for consideration in unattended stations in remote areas.

  Thermal electric cooled paramps with noise temperatures in the range of 30°K - 150°K are most reliable and less costly than cryogenic paramps and have found applications in the domestic satellite earth station market.

- Tunnel diode amplifiers although highly reliable and relatively inexpensive are limited in noise temperature

performance due to basic limitations of the diode. TDA's are available at noise temperatures in the range of 600°K to 900°K.

- The latest development in LNA technology is the transistorized amplifier. Commercial units are available at operating temperatures of 200°K - 350°K with predicitions in the near future of reducing the noise temperature down to 160°K using thermo electric cooling. These units are highly suitable for unattended stations in remote areas being relatively inexpensive, small in size and offering high reliability.

Figure 5 provides the G/T of an earth station versus antenna diameters at 4 GHz operational frequency considering:

- Thermal electric cooled and uncooled paramps

- Transistorized amplifiers

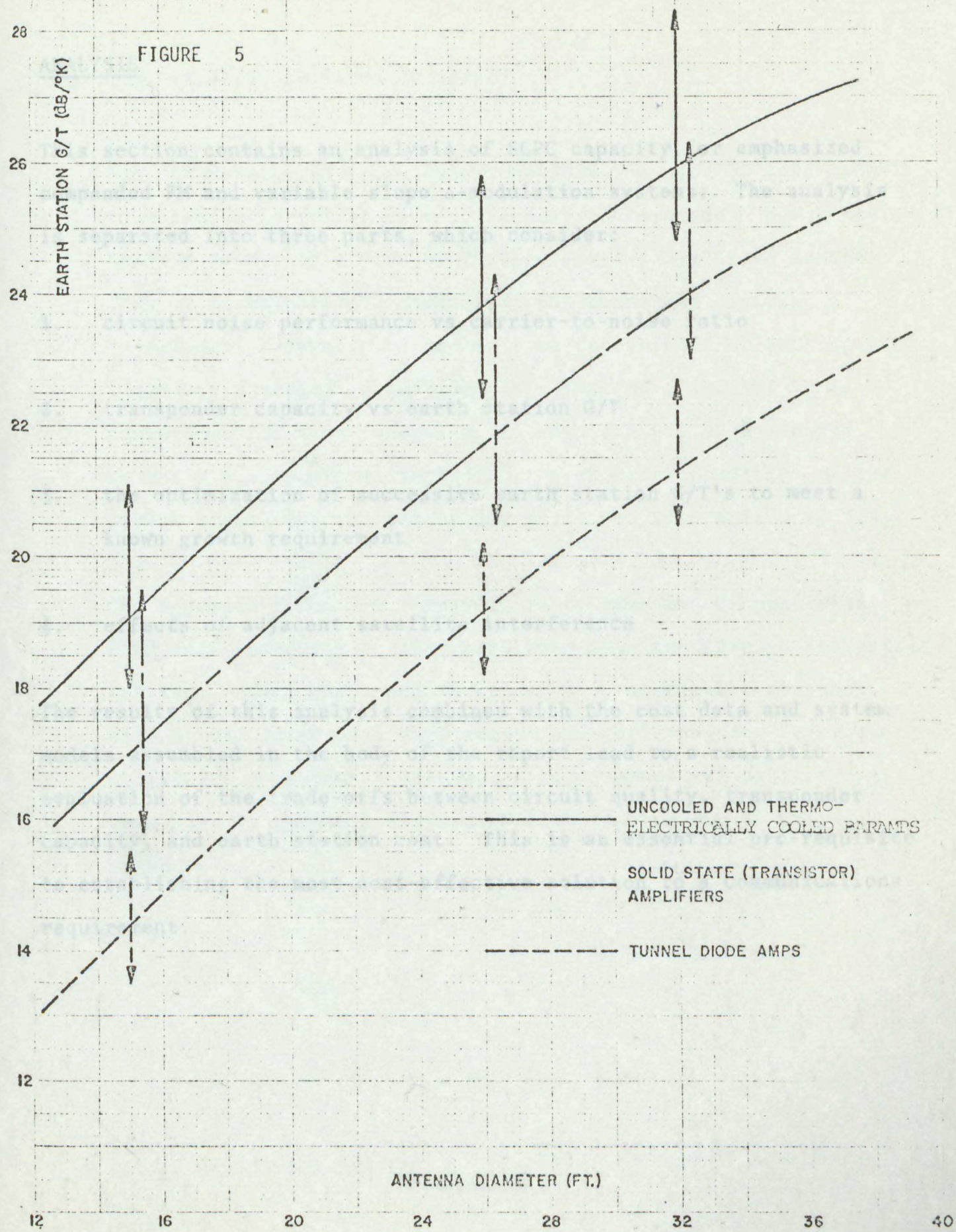
- Tunnel Diode Amplifiers.

The curves show the average earth station G/T using the above devices with the vertical lines indicating the range in G/T due to the range in LNA noise temperatures as discussed above.

ANTENNA & LOW NOISE RECEIVER CAPABILITY

(4 GHz Receive Frequency)
4°-5° Antenna Elevation Angle

FIGURE 5



3. ANALYSIS

This section contains an analysis of SCPC capacity for emphasized companded FM and variable slope Δ-modulation systems. The analysis is separated into three parts, which consider:

1. circuit noise performance vs carrier-to-noise ratio

2. transponder capacity vs earth station G/T

3. the optimization of successive earth station G/T's to meet a known growth requirement

4. effects of adjacent satellite interference

The results of this analysis combined with the cost data and system models assembled in the body of the report lead to a realistic evaluation of the trade-offs between circuit quality, transponder capacity, and earth station cost. This is an essential pre-requisite to establishing the most cost-effective solution to a communications requirement.

3.1 Circuit Noise Performance vs C/N_o

For the modulation techniques considered, a plot of test tone-to-noise ratio (TT/N) vs C/N_o is first determined. Since the desired result is a plot of "articulation index" (14) or the TT/N of a subjectively equivalent non-companded circuit, a relationship between the two must be determined for each case. A graphical method is outlined in reference (15), but it is cumbersome and time consuming. A computational method based on certain assumptions regarding the spectral densities of the speech signal and the perturbing noise has also been used (14).

The approach taken here however, is to make use of published compandor improvements that have been obtained through subjective testing. Since companded FM or Δ-mod circuits have not seen wide-spread commercial application and corresponding rigorous subjective performance testing by telecommunications carriers, this part of the analysis necessarily remains one of possible weakness.

3.1.1 Test Tone-To-Noise Ratio

3.1.1.1 Emphasized Compandored FM

The FM link equation in the region well above threshold is given approximately by:*

$$\frac{TT}{N} = 3 \left[\frac{C}{N_o}\right]_T \frac{(\Delta tt_{RMS})^2}{f_m^3} \quad P \qquad (1)$$

where:

$\left(\frac{C}{N_o}\right)_T$ = carrier-to-noise density ratio into the demodulator including effects of up-link and down-link thermal noise and inter-modulation noise generated in the satellite transponder.

Δtt_{RMS} = rms test tone deviation

f_m = top baseband frequency (3.4 kHz)

and

P = pre-emphasis advantage

---

* An exact expression for the TT/N at the output of an FM discriminator is given by Shimbo (14).

The transmission bandwidth is given by Carson's Rule:

$$B_{CR} = 2\ (\Delta tt_{RMS}\ \ell.g + f_m) \qquad (2)$$

where

ℓ = rms speech to-test-tone ratio

g = peak speech-to-rms speech ratio

The value used for the peak-to-rms ratio for speech depends on the amount of speech clipping that is to be permitted. If speech clipping is to occur no more than 0.1% of the time, then it should take place at a level of 14.2 dB greater than the rms level (17). This is a typical clipping level for high quality systems.

The average long term speech power of a talker for international circuits is -12VuO with a standard deviation of 5 dB (17). Therefore the average talker power is given by:

$$P_{avg} = Vo - 1.4 + .115\ \sigma^2$$

$$= -11.0\ \text{dBmO}$$

With g = 14 dB

and ℓ= -11 dB, (without companding)

$$B_{CR} = 2\ (1.4\ \Delta tt_{RMS} + f_m)$$

The noise bandwidth (Bn) of an FM system is typically 5% - 15% greater than the required distortionless transmission bandwidth B_{CR}. Assuming a value of 10% i.e.,

$$\frac{C}{N_o}_T = \frac{C}{N}_T B_N = \frac{C}{N_T}\ (1.1\ B_{CR})$$

and substituting into the FM equation (1)

$$\frac{TT}{N} = 3\left[\frac{C}{N}\right]_T \frac{(\Delta tt_{RMS})^2}{f_m^{\ 3}}\ (1.1\ B_{CR})\quad (P) \qquad (1a)$$

Now the theoretical C/N threshold of a conventional FM demodulator can be expressed by (16)

$$\left[\frac{C}{N}\right]_T = 6.64 + 2.9\ \log_{10} B_N - 2.9\ \log_{10} f_m \quad \text{(in dB)}$$

FM receivers may incorporate phase-locked loops (PLL) demodulators rather than discriminators. It has been demonstrated that at a modulation index above 3, FM-PLL receivers theoretically provide at least 2.5 dB of threshold extension (16). Assuming this value, the threshold C/N for a TED (threshold extension demodulator) is

$$\left(\frac{C}{N}\right)_T = 4.14 + 2.9 \log_{10} (B_N) - 2.9 \log_{10} f_m$$

Substituting the threshold equation into the FM equation (1a) and collecting terms, the final expression relating the theoretical test-tone-to-noise ratio and Carson's rule bandwidth at a threshold C/N is

$$\frac{TT}{N} = 10 \log_{10} \left[(B_{CR} - 2f_m)^2\right] + 12.9 \log_{10} B_{CR} \quad (3)$$
$$-X + P$$

where

$$X = \begin{cases} 113.15 \text{ for FM discrimination} \\ 115.65 \text{ for TED} \end{cases}$$

Equation (3) allows one to determine the maximum Carson's rule bandwidth (and hence maximum rms test tone deviation) to meet a given TT/N requirement. Assuming a power limited satellite transponder, the maximum (i.e., threshold limited) bandwidth is optimum since it minimizes the C/N_o required.

The pre-emphasis advantage is given by the reduction in noise power relative to a fixed signal power through the de-emphasis network. In order to compare a system that employs pre-emphasis to one without pre-emphasis it is important that the insertion of pre-emphasis does not alter the total speech spectrum power or equivalently, the rms frequency deviation of the modulation.

As pointed out in section 2.3.1.1 a pre-emphasis advantage of 6.3 dB can be obtained for a companded system; however, a pre-emphasis network giving a 5.8 dB advantage is more suitable in this case because it is less sensitive to operation in the threshold region. Assuming a 5.8 dB pre-emphasis advantege, and allowing 1

dB of implementation margin* the test tone-to-noise ratio at threshold becomes

$$\frac{TT}{N} = 10 \log_{10}\left[\left(B_{CR}-2f_m\right)^2\right] + 12.9 \log_{10} B_{CR} - X \qquad (4)$$

where:

$$X = \begin{cases} 108.35 \text{ for FM discriminator} \\ 110.85 \text{ for TED} \end{cases}$$

Figure 6a plots optimized Carson's Rule Bandwidth vs C/N_o and Figure 6b plots TT/N vs optimized Carson's Rule Bandwidth. In each case both normal and threshold extension type demodulation is considered. The foregoing FM system designs are based on the use of the Carson's Rule bandwidth formula for the required transmission bandwidth. It may be argued that Carson's Rule for distortionless transmission does not strictly apply to speech because the spectral characteristic of speech is unlike white noise. Analysis of the speech spectrum reveals that the energy at high frequencies diminishes rather rapidly. This

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* Includes margin for FM residual noise and departure from linearity in FM equation.

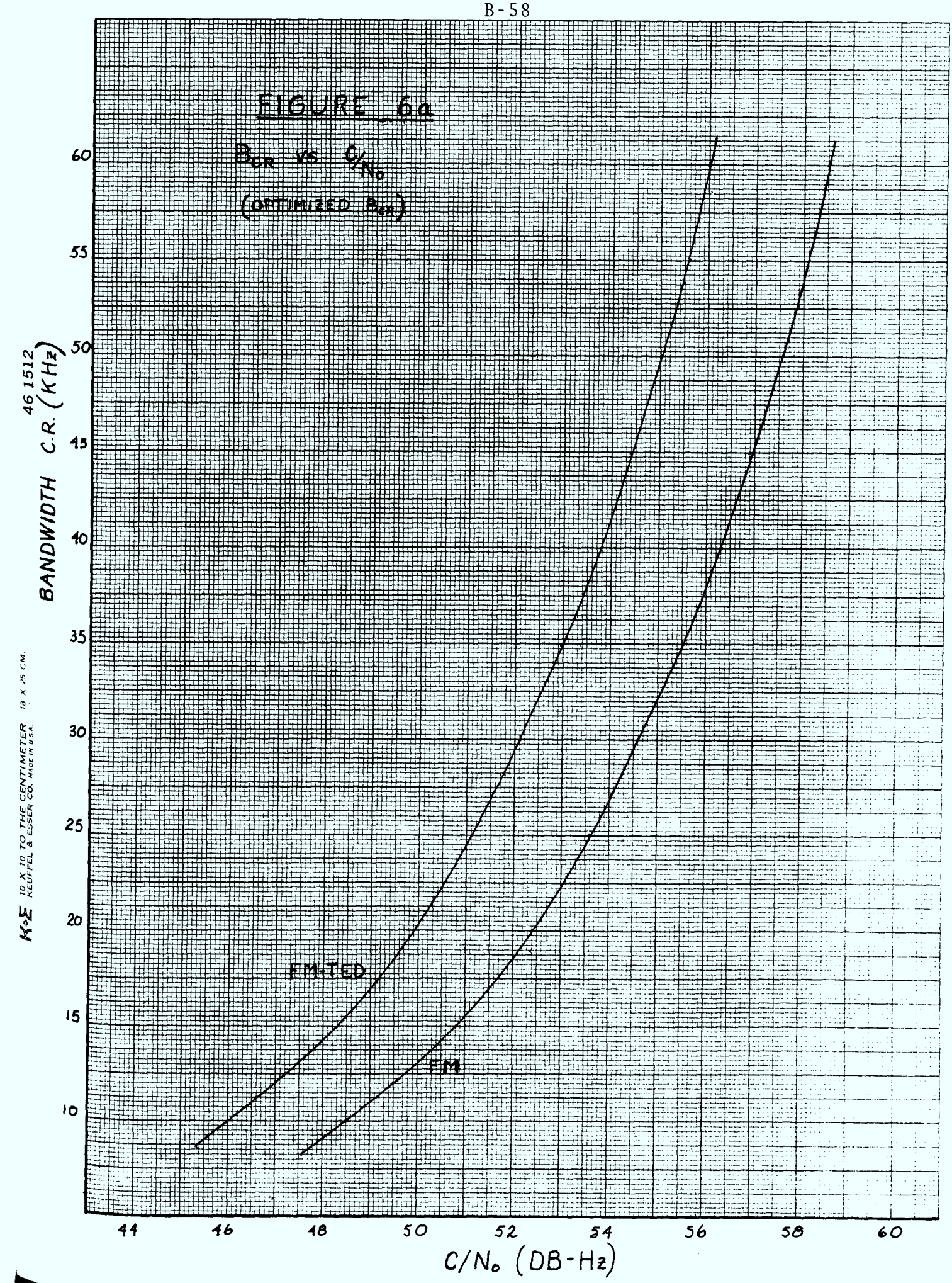


FIGURE 6a
B_{CR} VS C/N_0
(OPTIMIZED B_{CR})
BANDWIDTH C.R. (KHz)
60
55
50
45
40
35
30
25
20
15
10
FM-TED
FM
44
46
48
50
52
54
56
58
60
C/N_0 (DB-Hz)
46 1512
K&E 10 X 10 TO THE CENTIMETER KEUFFEL & ESSER CO. MADE IN U.S.A. 18 X 25 CM.

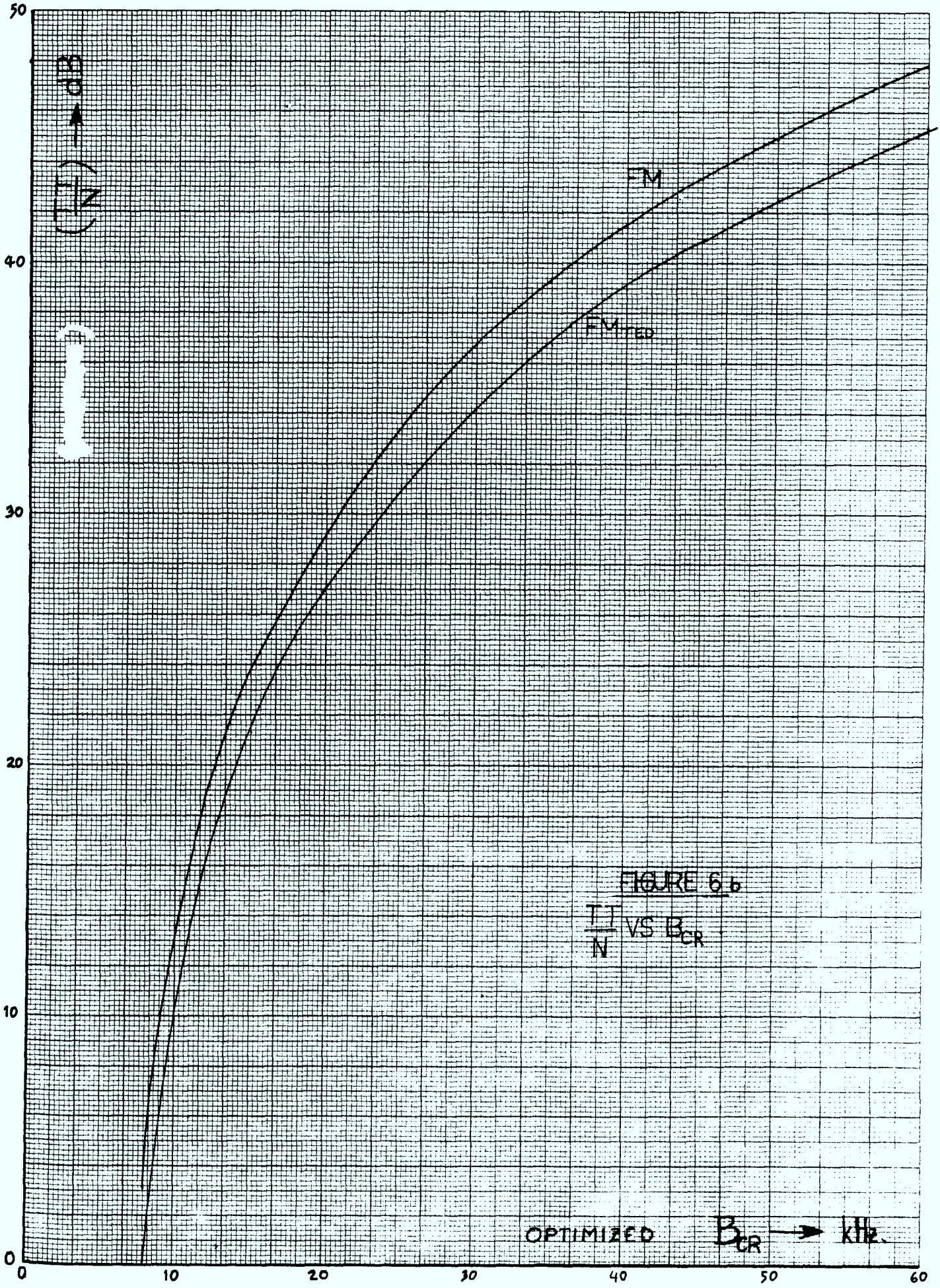


FIGURE 6.6

$\frac{TI}{N}$ VS B_{CR}

46 1512
K+E 10 X 10 TO THE CENTIMETER 18 X 25 CM.
KEUFFEL & ESSER CO. MADE IN U.S.A.

indicates that the transmission bandwidth necessary to contain the FM carrier components needed to reconstitute the speech signal is significantly less than that predicted by Carson's Rule. Reduction of the f_m term in the bandwidth equation leads to a substantial change for a relatively low deviation system.

Using this argument, narrow band FM (bandwidth of 10 kHz or less) has been analyzed and considered for aeronautical and marine applications (14). The circuit quality just above threshold does not meet our requirements, and since the narrow-band scheme is not particularly advantageous from a bandwidth conservation standpoint, no further consideration of narrow-band FM is given here.

3.1.1.2 Varible slope Δ-modulation CPSK

If slope overload is avoided, the signal-to-quantization noise ratio for Δ-mod can be expressed as (19)

$$\frac{TT}{N_q} = 3\pi\left(\frac{R}{2\pi f_t}\right)^3 \qquad (1)$$

where f_t is the test-tone frequency and R is the sampling rate. Slope overload can be avoided by assuming that the slope of correction introduced by the delta modulator at least equals the maximum slope expected of the signal. For a speech spectral density that is assumed to be flat to 1000 Hz and thereafter to descend at a rate of 6 dB/octave of more, the maximum slope would occur for a 1000-Hz test tone which occupies the full peak-to-peak range of the delta encoder. Newer digitally controlled variable slope Δ-mod codecs do in fact maintain a test tone-to-noise ratio close to that predicted in equation (1) over a dynamic range of 30 to 40 dB.

The signal-to-thermal noise ratio caused by random errors in the bit stream is (19)

$$\frac{TT}{N_E} = \frac{f_o R}{16 f_T^2 P_e} \qquad (2)$$

where f_o is the low frequency cutoff of the speech signal and P_e is the error probability. TT/N_E is dependent on f_o because the power spectral density of the noise in the regenerated

baseband signal is proportional to $1/f^2$. The voice band channel is assumed to start at 300 Hz; allowing for filter rolloff one can reasonably assume a value for f_o of 200 Hz. Combining (1) and (2), the following expression for test tone-to-noise ratio at 1000 Hz is obtained:

$$\frac{TT}{N} = \frac{R^3}{\frac{8}{3}\pi^2 \times 10^9 + 8 P_e R^2 \times 10^4} \qquad (3)$$

The theoretical relationship between error probability and C/N_o for 2-phase or 4-phase coherent PSK is given by:

$$P_e = \frac{1}{2} \operatorname{erfc} \left[\left(\frac{C}{N_o} \frac{1}{R} \right)^{\frac{1}{2}} \right]$$

where erfc (.) is the complementary error function. Allowing fixed 1.0 dB codec and 2.5 dB modem* margins in (3) gives a final expression relating the actual test tone-to-noise ratio of a Δ-mod CPSK system to (1) the C/N_o present

---

* Includes margins for filter degradation, phase and timing recovery jitter, and differential encoding/decoding (doubles the number of bit errors), but does not include satellite link fade margin.

at the demodulation input and (2) the encoding rate:

$$\frac{TT}{N} = \frac{R^3}{3.36\pi^2 \times 10^9 + 8R^2 \frac{\text{erfc}}{2}\left[(.562 \frac{C}{N_o} \frac{1}{R})^{\frac{1}{2}}\right]} \times 10^4 \qquad (4)$$

By differentiating this expression with respect to R, we can determine the best encoding rate for a given available C/N_o; this rate leads to an optimum trade-off between the effects of quantization and bit error noise.

Figure 7 plots TT/N vs C/N_o for 40, 32 and 20 kbps Δ-mod/CPSK systems, and for bandwidth optimized FM systems with and without threshold extension demodulation. The Carson's Rule bandwidth corresponding to each TT/N in the FM case can be obtained from Figure 6b.

As a theoretical comparison of FM and Δ-mod/CPSK SCPC transmission efficiencies, our approach is somewhat unfair to the digital scheme because the encoding rate is not optimized at each C/N_o.

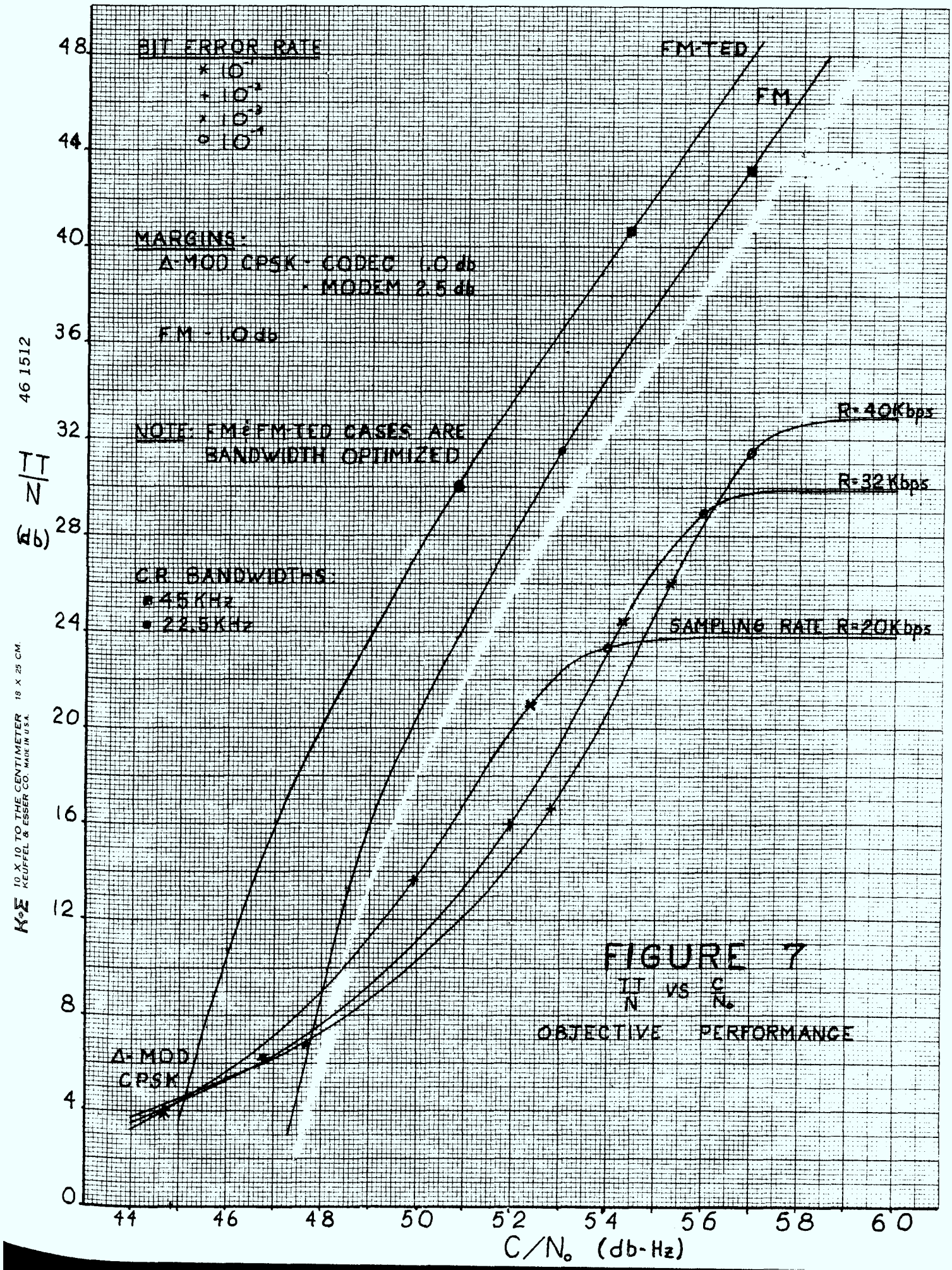


FIGURE 7

$\frac{ST}{N}$ VS $\frac{C}{N_o}$

OBJECTIVE PERFORMANCE

The effect of such an optimization can be observed by comparing TT/N performances for three encoding rates presented in Figure 7. At sufficiently low C/N_o's a lower rate system offers better performance; at sufficiently high C/N_o's no further gain is achievable without increasing the encoding rate. However, using the upper envelope of TT/N performance for three rates considered leads to a reasonable comparison of FM and Δ-mod.

3.1.2 Subjective Noise Performance

The previous section evaluates the test tone-to-noise ratio of the candidate SCPC systems, FM and Δ-mod/PSK. For a non-compandored (i.e. constant noise) system the test tone-to-noise ratio defines the additive noise power relative to an arbitrary signal level, which determines the quality of the circuit. For a compandored system, however, the noise power is signal level dependent, and it is somewhat more difficult to relate test tone-to-noise ratio to subjective performance.

3.1.2.1 Effects of Companding

Both the FM and variable slope Δ-mod systems employ syllabic companding to enhance their performance. While the means of implementation are quite different in the two cases, the effects are similar.

The syllabic compandor is comprised of two devices, a compressor at the transmitter and an expandor at the receiver. The improvement due to syllabic compandors arises from two factors, namely:

1. The speech compression at the transmitter increases the rms value of the signal for the same peak level. This permits an increase in the rms speech level without any accompanying increase in noise, resulting in an improvement in objective signal-to-noise ratio.

2. The expandor at the receiver output suppresses noise during speech pauses which gives a subjective improvement in signal-to-noise ratio to the listener.

The net sum of these effects defines what we call a compandor improvement, which relates the TT/N of a compandored system to the TT/N of a non-compandored system having the same subjective performance. Following the addition of the compandor advantage, we have a "subjectively equivalent test tone-to-noise ratio". The compandor improvement depends on speech characteristics as well as listener response, and is therefore statistical in nature. Compandor improvement also depends on the pre-emphasis network employed and the signal-to-noise ratio into the expandor in the case of FM, and the encoding rate and bit error probability in the case of Δ-mod.

3.1.2.2 Compandor Improvement

3.1.2.2.1 FM Case

The syllabic compressor is a non-linear device with memory which distorts the incoming signal. This distortion will be cancelled by the expandor if:

a. the transmission link produces no distortion

b. the link noise is low enough to cause negligible effects on the expandor control circuit

c. the compressor and expandor have exactly complementary characteristics

The power transfer characteristic of a 2:1 syllabic compressor* is given by:

$$P' = \frac{P}{2} + A \qquad \text{(in dB)}$$

where

P' = long term average power of compressed speech at compressor output

---

* The choice of a companding ratio of 2:1 is a compromise between two conflicting requirements. A high ratio would give rise to greater noise improvements, but would also reduce stability and produce higher distortions due to equipment imperfections.

P = long term average power of uncompressed speech signal at input to compressor

A = compandor long term gain which determines the peak signal at the input to the modulator

Now the highest talker power (99% case) is 2.56σdB above the average talker power, where σ is the standard deviation of talker power.

$$\therefore P_{max} = 2.56\sigma + P_{avg}$$

$$= 2.56\ (5) + -11\ \text{dBmO}$$

$$= 1.8\ \text{dBmO}$$

Then since

$$P'_{max} = \frac{P_{max}}{2} + A \quad ,$$

for A = 0.9 dB

$$P'_{max} = 1.8\ \text{dBmO}$$

i.e. the compressor will not modify the level of the loudest talker.

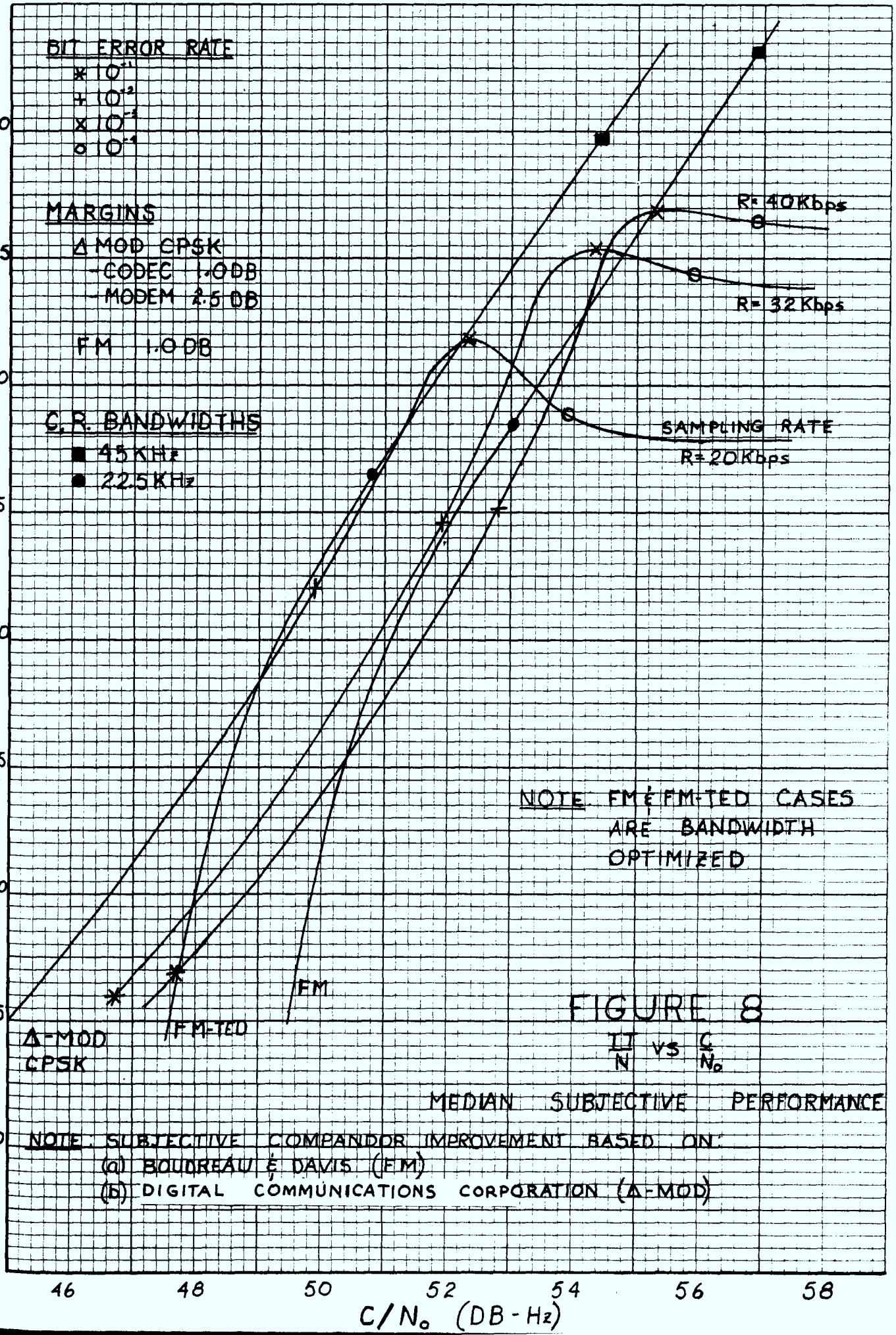
The change in average talker level through the compressor is given by

$$\Delta P_{avg} = P'_{avg} - P_{avg}$$

$$= 6.4 \text{ dB}$$

The second advantage of subjective improvement due to expandor suppression of noise during speech pauses is a function of the input signal-to-noise ratio (SNR) to the expandor.

Boudreau and Davies (8) indicate that this improvement ranges from 5.8 dB to 11.2 dB depending on the input SNR. With systems whose subjective noise objective is 10,000 pWpO, the syllabic compandor will give a total improvement of 15 to 17 dB for average talkers (6). Softer talkers will have more compandor improvement, louder talkers



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less (8, 20). Systems operating above threshold but at lower test tone-to-noise ratios will have less compandor improvement because of the subjective nature of the compandor action. For systems operating at or below threshold, compandor gain may be reduced further if the de-emphasis gain is held fixed at 5.8 dB.

Since bandwidth optimized FM systems (i.e. operation above threshold) only are considered we do not allow any margin for this additional degradation.

Figure 8 plots subjectively equivalent TT/N vs C/N_o for the bandwidth optimized FM system (ref. Figure 9) with and without threshold extension demodulation.

Boudreau and Davies (8) results on compandor improvement vs TT/N are assumed.

3.1.2.2.2 Variable Slope Δ-Mod Case

Compandor improvement in the digital case will depend on the bit error rate into the Δ-decoder as well as characteristics of the speech, the compandor operation, and the listener. Unfortunately, a comprehensive study of this problem has not been undertaken. Subjective testing of a 32 kbps system by thirty listeners at bit error rates from 10^{-1} to 10^{-5} (4) did, however, reveal the following:

1. On the average listeners did not attribute significant noise contribution due to bit errors below an error rate of 10^{-3}; however, as the error rate increased to 10^{-3} listeners had more difficulty in deciding on the equivalent noise value.

2. For error rates below 10^{-3}, where quantization noise dominates, listeners judge 32 kbps as capable

of providing a 10,000 pWpO equivalent voice channel.

Assuming identical compandor vs error rate improvements at 40, 32, and 20 kbps rates, Figure 8 plots the subjectively equivalent TT/N vs C/N_o performance of these systems.

3.2 Transponder Capacity vs G/T

This section evaluates the capacity of an Anik transponder vs earth station G/T for companded FM with and without threshold extension and 20, 32, and 40 kbps variable slope Δ-mod/CPSK. The increase in capacity provided by voice activation and demand assignment are also considered.

3.2.1 Simultaneous Link Equations for a Non-Homogeneous Network

An SCPC satellite communications system, employing either FM or Δ/PSK modulation techniques, must meet specified C/N_o requirements. As indicated in section 3.1 these requirements are related to test tone-to-noise ratio performance of the circuits. Since the G/T's of the participating earth stations, link fade margins, and C/N_o ob-

jectives (e.g. toll vs bush voice circuit performance) may vary, there can, in general, be several transmitted carrier EIRP's. To achieve the most economic system design we wish to choose the carrier levels so that the required C/N_o performance on each link is just met and the total capacity of the satellite transponder is used. This leaves an additional parameter to be determined - either a link capacity (i.e., number of carriers into a specific G/T at a specific C/N_o) or a required G/T given the link capacities and required C/N_o's. The former corresponds to maximizing system capacity assuming fixed (i.e. existing) earth stations, the latter minimizes the G/T required to provide a given number of circuits. Simultaneous link equations for a non-homogeneous network are now presented using matrix notation.

Suppose

$$g = (g_1, \ldots, g_m)$$

and

$$c = (c_1, \ldots, c_n)$$

are vectors whose components denote all distinct G/T's and desired C/N_o's in the SCPC multiple access system.

$$\vec{k} = \begin{pmatrix} k_{11} & \cdots & k_{m1} \\ \cdot & & \\ \cdot & & \\ \cdot & & \\ k_{1n} & \cdots & k_{mn} \end{pmatrix}$$

is a capacity matrix with

k_{ij} = number of carriers into an earth station having a G/T = g_i at a desired $C/N_o = c_j$

The matrix

$$\vec{\Delta} = \begin{pmatrix} \Delta_{11} & \cdots & \Delta_{m1} \\ \cdot & & \\ \cdot & & \\ \cdot & & \\ \Delta_{1n} & \cdots & \Delta_{mn} \end{pmatrix}, \quad \Delta_{11} = 1,$$

represents relative up-link carrier powers for each G/T, C/N_o combination; i.e.

Δ_{ij} = relative power of carrier into an earth station having a G/T = g_i at a desired $C/N_o = c_j$

Relative powers are used here because it will be assumed that the satellite input backoff given by

IBO = satellite I/P backoff

is fixed.

For a sufficiently large number of carriers, and certainly in all realistic thin route examples, the multi-carrier satellite input signal is gaussian, and therefore

OBO = multi-carrier satellite O/P backoff

and

$\frac{C}{I}_{cc}$ = centre channel carrier-to-IM produce interference for equal and equally spaced carriers are unique

deterministic functions of IBO and the amplitude/amplitude and phase/amplitude (AM-to-PM) characteristics of the particular satellite TWT under consideration.

These input/output characteristics of the travelling wave tube (TWT) in the satellite transponder may be predicted from Westcott (21) and McClure (1). The multi-carrier

output and centre channel C/I vs input power transfer curves of an Intelsat 4 TWT are presented in Figure 9. These are quite close to the multi-carrier characteristics of an HS 333 (Anik), and are suitable for use in a comparative evaluation study such as this.

Consider the actual C/N_o performance of an i, j link carrier:

$$(1)\quad \left(\frac{C}{N_o}\right)_{up} = (EIRP)_{up} - (FSL)_{up} + (G/T)_s + 228.6 - IBO + 10 \log_{10}\left(\frac{\Delta_{ij}}{\sum_{i=i}^{m}\sum_{j=1}^{n} k_{ij}\,\Delta_{ij}}\right)$$

where

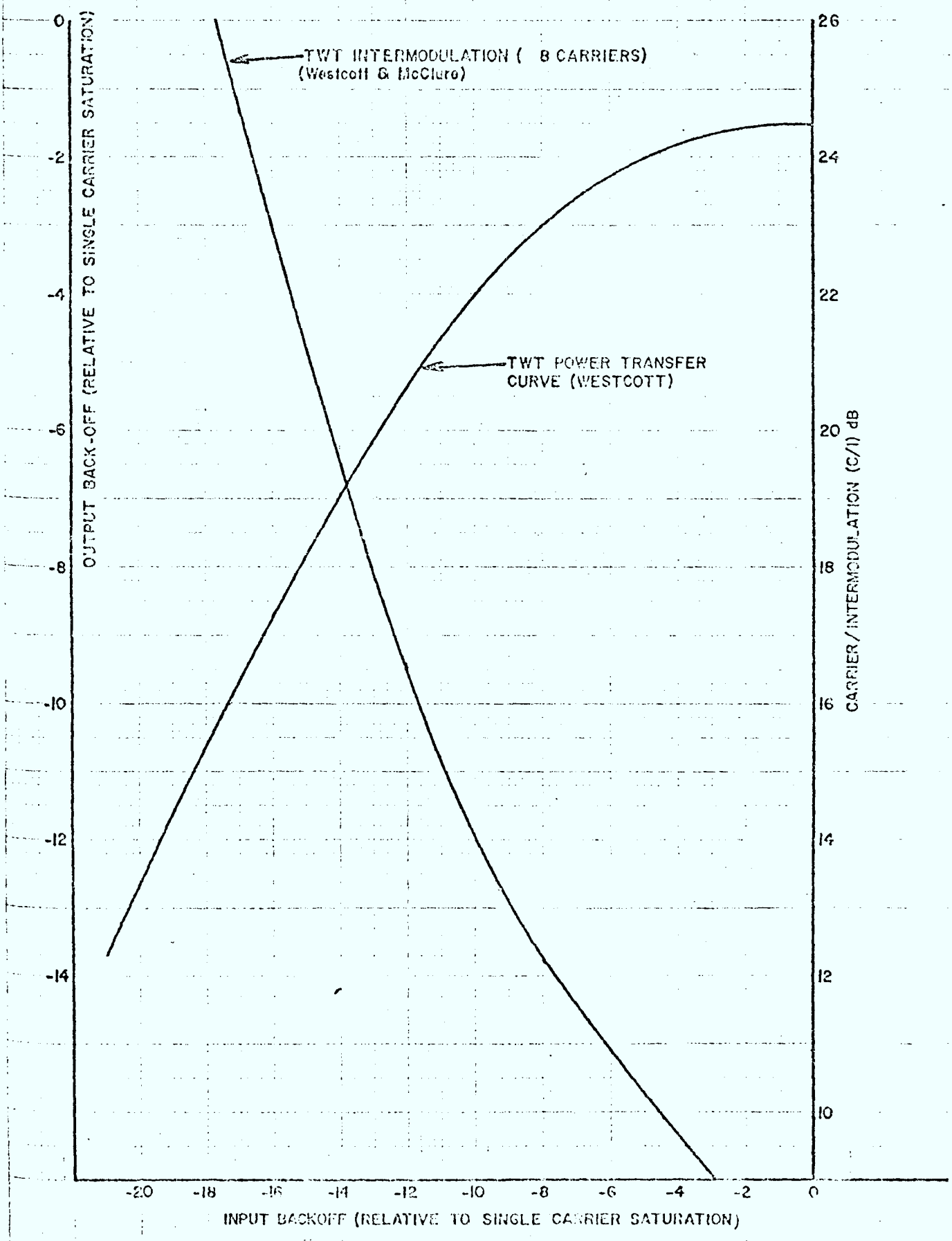
$(EIRP)_{up}$ = saturating EIRP from earth station
= 81.7 dBW

$(FSL)_{up}$ = up-link (6 GHz) free space loss
= 200.7 dB

$(G/T)_s$ = satellite input G/T
= -5.4 dB/^oK

FIG. 9

SATELLITE TRANSPONDER TRANSFER AND INTERMODULATION



$$(2)\quad \frac{C}{I_o} = \left(\frac{C}{I}\right)_{cc} + I_{ij} - 10 \log (W)$$

where

I_{ij} = difference between centre channel C/I for equal/equally spaced carriers and minimum C/I on ij link (dB) (2)

W = carrier noise bandwidth

$$(3)\quad \left(\frac{C}{N_o}\right)_{down} = (EIRP)_{down} - (FSL)_{down} + (G/T)_i + 228.6 - OBO + 10 \log_{10}\left(\frac{\Delta_{ij}}{\sum_{i=1}^{m}\sum_{j=1}^{n} k_{ij}\,\Delta ij}\right)$$

where

$(EIRP)_{down}$ = saturated EIRP from satellite
= 35 dBW from Anik

$(FSL)_{down}$ = down-link (4 GHz) free space loss
= 197.1 dB

$(G/T)_i$ = earth station G/T (dB/^oK)
= g_i

Then the received C/N_o is given by

$$r_{ij} = \left[\left(\frac{C}{N_o}\right)^{-1}_{up} + \left(\frac{C}{I_o}\right)^{-1} + \left(\frac{C}{N_o}\right)^{-1}_{down} \right]^{-1}$$

defining a matrix $\vec{r}$ of received C/N_o's.

3.2.2 Optimization of Static Non-homogeneous Network

Statement of the Problem:

Suppose there is one unknown parameter in the $\vec{g}$ or $\vec{k}$ matrices. Choose the $\vec{\Delta}$ matrix to optimize (maximum capacity or minimum G/T) the unknown parameter while meeting the C/N_o objectives defined in the $\vec{c}$ matrix. This specifies the optimum system design under the constraint of a fixed satellite backoff.

Re-evaluating for successively chosen backoffs leads to an overall system optimization.

Iteration Algorithms:

Initial estimates for the unknown parameter and relative carrier powers ($\vec{\Delta}$ matrix) are made. For a down-link

noise dominated link a reasonable initial estimate of Δ_{ij} is given by:

$$\Delta_{ij} = \frac{c_i}{g_i} \quad \frac{g_1}{c_1}$$

The above equation implies a dB for dB relationship between G/T and received C/N_o, true when down-link noise dominates.

Given the initial estimates the C/N_o's on each of the mn distinct <u>types</u> of links are computed using the equations previously given.

Received C/N_o's are compared to required C/N_o's and the $\vec{\Delta}$ matrix and unknown parameter readjusted according to the following unverse interpolation rules.

1. Adjusting the $\vec{\Delta}$ matrix

$$(\Delta_{ij})_{new} = (\Delta_{ij})_{old} \quad \frac{c_j}{r_{ij}} \quad \frac{r_{11}}{c_1}$$

balances the carrier power to obtain a constant c_j/r_{ij} ratio. (Note that the r_{11}/c_1 adjustment factor here serve as a reference to constrain the apportionment of power to all links).

Adjusting the unknown parameter forces $r_{ij} = c_j$; i.e. the C/N_o's obtained on all links approach the desired C/N_o's.

2. Adjusting the Unknown Parameter

Three important cases of an unknown design parameter are identified here; others are possible. The three cases are:

- an unknown capacity on a given type of link

  i.e., $k_{\ell 1, \ell 2}$, $1 \leq \ell_1 \leq m$, $1 \leq \ell_2 \leq n$, is variable

- an unknown total capacity; i.e. the $\overline{k}$ matrix specifies a fixed set of relative capacities, but

  $$K = \sum_{i=1}^{m} \sum_{j=1}^{n} k_{ij}$$ is variable

- an unknown G/T for a given type of link; i.e.

  g_ℓ, $1 \leq \ell \leq m$, is variable

The following define iteration equations for each of the three cases.

Case 1

$$k_{\ell 1,\ell 2\ \text{new}} = k_{\ell 1,\ell 2\ \text{old}} + \left(1 - \frac{\Sigma\Sigma\ k_{ij}\ \Delta_{ij}\ \frac{c_j}{r_{ij}}}{\Sigma\Sigma\ k_{ij}\ \Delta_{ij}}\right)$$

$$\left(\frac{r_{\ell 1,\ell 2}}{c_{\ell 2}}\right)\left(\frac{\Sigma\Sigma\ k_{ij}\ \Delta_{ij}}{\Delta_{\ell 1,\ell 2}}\right)$$

Case 2

$$(K)_{\text{new}} = (K)_{\text{old}} \left(\frac{\Sigma\Sigma\ k_{ij}\ \Delta_{ij}\ \frac{c_j}{r_{ij}}}{\Sigma\Sigma\ k_{ij}\ \Delta_{ij}}\right)$$

Case 3

$$(g_\ell)_{\text{new}} = (g_\ell)_{\text{old}} + 2\left(\frac{\Sigma\Sigma\ k_{ij}\ \Delta_{ij}}{\Sigma\ k_j\ \Delta_{\ell j}}\right) \log_{10}\left(\frac{\Sigma\Sigma\ k_{ij}\ \Delta_{ij}\ \frac{c_j}{r_{ij}}}{\Sigma\Sigma\ k_{ij}\ \Delta_{ij}}\right)$$

The inverse interpolation equation in case 3 is non-linear and empirical; however, it will give satisfactory convergence.

Hypothetical Examples:

In order to understand the application of the theoretical design procedure outlined here, consider the following hypothetical qualitative examples.

1. Homogeneous System, Fixed Capacity

We wish to construct a homogeneous network (i.e. one G/T size and performance requirement) that provides N one-way circuits with a $C/N_o = Q$, What is the required G/T?

With $m = n = 1$, $k_{11} = N$, and $c_1 = Q$, the minimum G/T = g_1 can be evaluated.

2. Homogeneous System, Fixed G/T

How many simultaneous one-way circuits can an existing network of earth stations all having G/T = X support? The required C/N_o on all links is Q.

With $m = n = 1$, $c_1 = Q$, and $g_1 = X$, the unknown capacity = k_{11} can be evaluated.

3. North-To-South Model, Unknown Total Capacity

In this case 2 G/T's and required C/N_o performances are assumed. We necessarily have the constraint of equal numbers of northbound and southbound carriers. What is the total capacity?

Here $m = n = 2$, c_1, c_2, g_1, (southern terminus) and g_2 (northern station) are given, and (assuming north-to-south traffic only).

$$\vec{k} = K \begin{pmatrix} 0 & \frac{1}{2} \\ \frac{1}{2} & 0 \end{pmatrix}$$

The unknown capacity k can be evaluated.

4. as in 3. with a 50% split between north-to-south and north-to-north traffic.

in this case the traffic matrix is written in the form

$$\overline{k} = k \begin{pmatrix} 0 & \frac{1}{4} \\ \frac{1}{4} & \frac{1}{2} \end{pmatrix}$$

5. The north-to-south model described in 3. has not reached its ultimate growth capacility., and is operating at a capacity $\hat{k}<k$. Transportable earth stations have a G/T $g_3 < g_2$, the standard thin route station. How many circuits can be provided among transportables?

   In this case we have an unknown link capacity, and the traffic matrix becomes:

$$\vec{k} = \begin{pmatrix} 0 & \frac{\hat{k}}{2} & 0 \\ \frac{\hat{k}}{2} & 0 & 0 \\ 0 & 0 & k_{33} \end{pmatrix}$$

Earth Station Cost Minimization:

The theoretical design procedure presented here minimizes the G/T required to achieve a required transponder capacity and link noiwe performance. It is assumed that this will lead to a minimum cost design, i.e. there remains only to select the minimum cost antenna - LNA - HPA combination that gives the required receive G/T and transmit EIRP. This procedure is not strictly optimum, as illustrated in the following paragraph.

The minimum G/T is obtained at a multi-carrier backoff which optimizes the trade-off between up-link and down-link thermal and intermodulation noise. For power limited systems, in which excess bandwidth allows the selection of a frequency plan which minimizes the effects of IM product noise, the "optimum" backoff can be quite low. While this backoff is optimum with regard to the G/T required to receive carriers, it also has a cost impact on the EIRP that stations must transmit. By introducing further backoff, more G/T but less EIRP is required. Therefore, when HPA costs are controlling, the minimum cost earth station may be designed to operate at a backoff somewhat greater than the so-called optimum. For the -81 dBW/m^2 saturating flux density of the present Anik, experience indicates that this can become an important consideration for low capacity SCPC systems (less than about 80 simultaneous one-way circuits). Future Canadian domestic satellites will probably reduce the cost impact of the HPA by having a lower saturating flux density and higher EIRP. In any event, this effect is noted as generally being a second order one which is not accommodated in the model.

3.2.3 Pre-assigned Capacity vs G/T for a Homogeneous Network

Capacity vs G/T calculations for a homogeneous network are presented as the simplest and most appropriate means of quantifying the trade-off between transponder capacity, required C/N_o, and earth station figure of merit. The homogeneous case facilitates both the computation and unified presentation of results, and is the most likely realization of a fully variable demand assigned SCPC system*.

For m equal carriers the previously stated link equations reduce to the following:

$$\left(\frac{C}{N_o}\right)_{up} = 104.2 - IBO - 10 \log (m)$$

$$\left(\frac{C}{N_o}\right)_{down} = 66.5 + G/T - OBO - 10 \log (m)$$

and

$$\frac{C}{I_o} = \frac{C}{I_{cc}} + I + 10 \log (W)$$

---

* If the system uses demand assignment the design must take into account the stations with the lowest G/T's and those in the lowest EIRP area of the satellite or have the earth station transmit EIRP selected by the DAMA control unit.

$$\geq \frac{C}{I}_{cc} + \left\{10 \log\left(\frac{712}{m}\right) - 1.5\right\} + 10 \log (45 \times 10^3)$$

$$= 73.52 + \frac{C}{I}_{cc} - 10 \log (m)$$

assuming a maximum channel spacing of 45 kHz**, random frequency assignment, and a minimum useable bandwidth 32 MHz (21). The resultant C/N_o at the demodulator is then:

$$\left(\frac{C}{N_o}\right)_{tot} = \left[\left(\frac{C}{N_o}\right)_{up}^{-1} + \left(\frac{C}{N_o}\right)_{down}^{-1} + \left(\frac{C}{I_o}\right)^{-1}\right]^{-1}$$

Allowing 3 dB of fade margin, which relates the net fairweather C/N_o computed here to that which is available for about 99.9% of the time, and equating this to the required C/N_o at the demodulator input, we have

$$\left(\frac{C}{N_o}\right)_{required} = -10 \log \left[\left(\frac{IBO}{2.63 \times 10^{10}}\right) + \left(\frac{OBO}{4.46 \times 10^6 \times \frac{G}{T}}\right) + \left(\frac{1}{2.25 \times 10^7 \times \frac{C}{I}_{cc}}\right)\right] - 10 \log m$$

** 22.5/45 kHz channel spacing is emerging as an SCPC standard; it is employed in SPADE and in most conceived domestic systems (except Telesat and G.E. (22) systems which employ 60 kHz). In fact the choice of spacing has obvious importance in determining the bandwidth limited capacity of the transponder, but only a slight effect on its power-limited capacity (via the C/I staggering improvement term).

which is an expression which directly relates at a given backoff the number of allowable simultaneous carriers to the earth station G/T and desired link C/N_o. This also defines the power-limited capacity (no. of simplex circuits) of a pre-assigned SCPC system not employing voice activation.

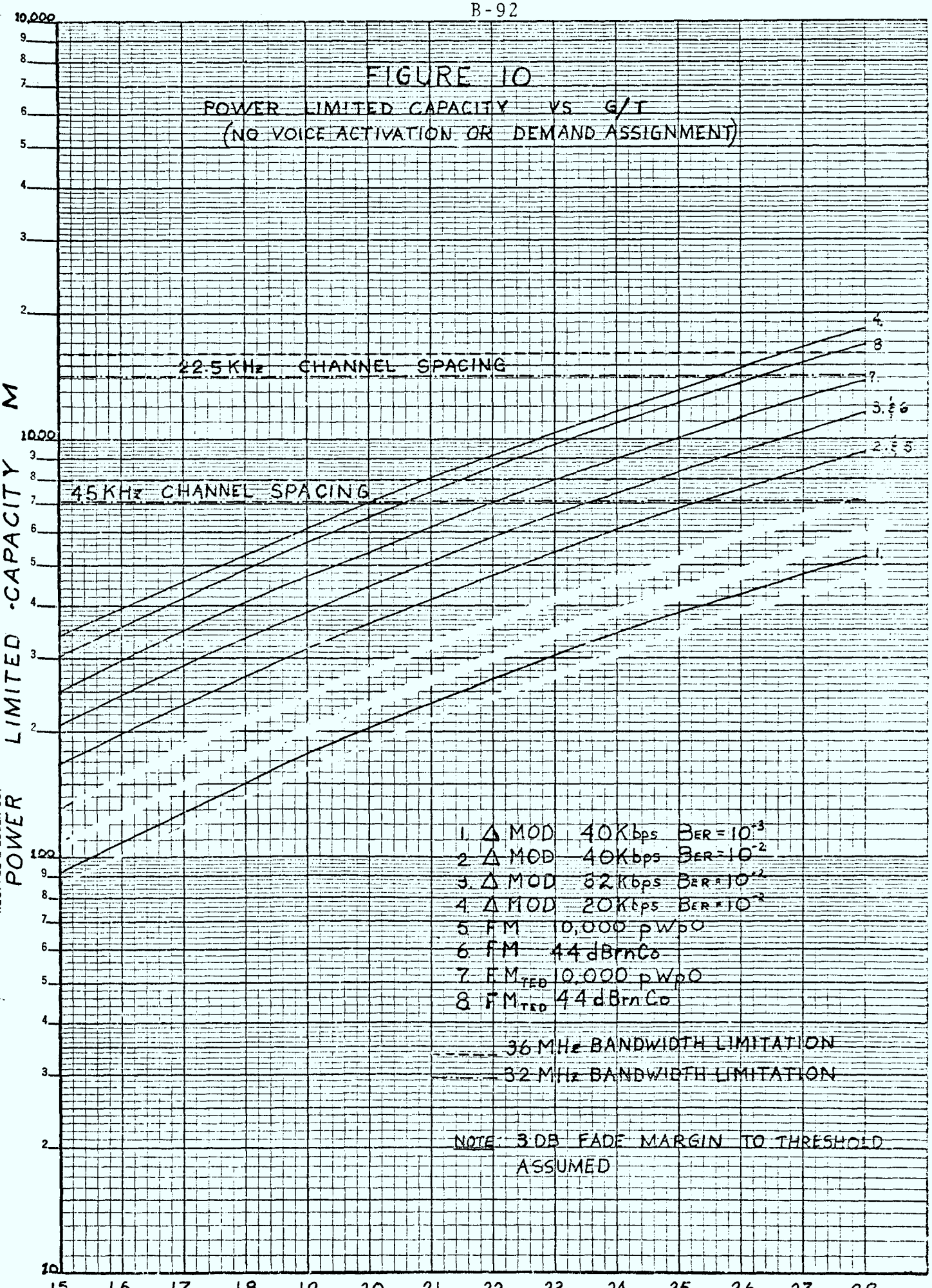
Figure 10 plots capacity vs G/T for the candidate systems presented in the following table.

SYSTEM	THRESHOLD	FADE MARGIN TO THRESHOLD	TEST TONE-TO-NOISE AT THRESHOLD*		REQUIRED UNFADED C/N_o
			OBJECTIVE	SUBJECTIVE EQUIVALENT	
40 kbps Δ	BER = 10^{-3}	3 dB	26.0	1100 pWpO	58.3
	BER = 10^{-2}	3 dB	16.7	11200 pWpO	55.8
32 kbps Δ	BER = 10^{-2}	3 dB	15.8	14100 pWpO	54.9
20 kbps Δ	BER = 10^{-2}	3 dB	13.6	22400 pWpO	52.9
FM	C/N = 10 dB	3 dB	37.2	10000 pWpO	55.8
	C/N = 10 dB	3 dB	34.5	22400 pWpO	54.9
FM with TED	C/N = 7.5 dB	3 dB	37.3	10000 pWpO	54.1
	C/N = 7.5 dB	3 dB	34.4	22400 pWpO	53.2

* A 2.5/2.0 dB (i.e. flat noise) psophometric/C-message weighting advantage has been assumed. Measurement and theory indicate that for SCPC circuits the noise density increases with frequency; therefore somewhat greater weighting advantages may actually be obtained.

FIGURE 10

POWER LIMITED CAPACITY VS G/T
(NO VOICE ACTIVATION OR DEMAND ASSIGNMENT)



K&E SEMI-LOGARITHMIC 46 5493
3 CYCLES X 70 DIVISIONS MADE IN U.S.A.
KEUFFEL & ESSER CO.

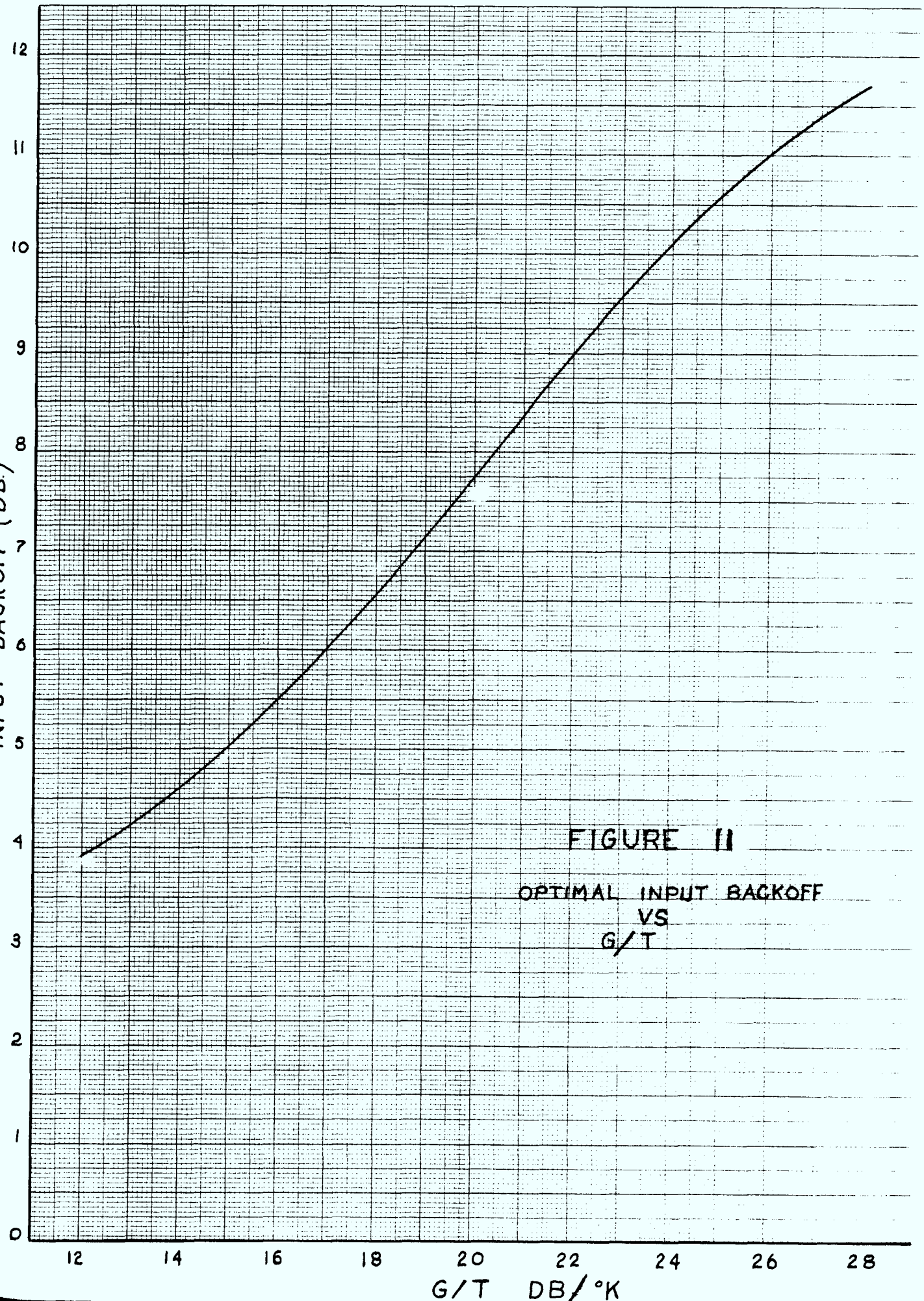


FIGURE 11

OPTIMAL INPUT BACKOFF
VS
G/T

46 1512
K&E 10 X 10 TO THE CENTIMETER 18 X 25 CM
KEUFFEL & ESSER CO. MADE IN U.S.A.

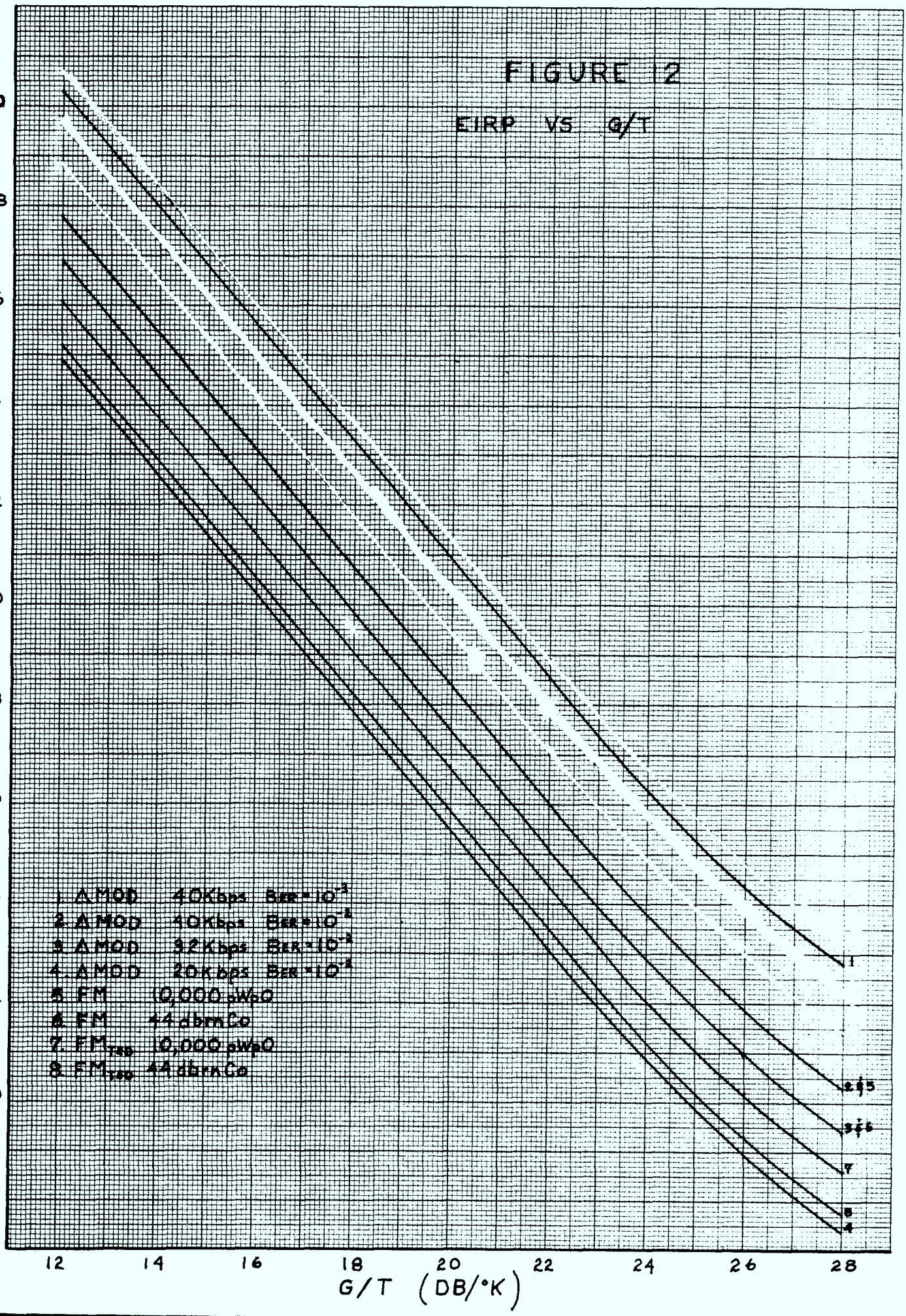


FIGURE 12
EIRP VS G/T
EIRP (DBW)
G/T (DB/°K)
60
58
56
54
52
50
48
46
44
42
40
38
12
14
16
18
20
22
24
26
28
1. ΔMOD 40Kbps BER = 10⁻³
2. ΔMOD 10Kbps BER = 10⁻⁴
3. ΔMOD 32Kbps BER = 10⁻³
4. ΔMOD 20Kbps BER = 10⁻⁴
5. FM 10,000 pWpO
6. FM 44 dbrnCo
7. FM TSB 10,000 pWpO
8. FM TSB 44 dbrnCo
1
2 & 5
3 & 6
7
8
4
46 1512
K+E 10 X 10 TO THE CENTIMETER 18 X 25 CM.
KEUFFEL & ESSER CO. MADE IN U.S.A.

The corresponding optimum input backoff and transmit EIRP per carrier vs G/T for each of these systems is plotted in Figures 11 & 12. Note that the lower capacity systems operate optimally nearer to saturation than do the higher capacity systems. This is because under the assumption that carrier frequencies are assigned randomly across the band and the resulting IM spectrum is spread more or less uniformly, the excess bandwidth has been used proportionately to reduce the contribution of IM noise at a given backoff. Extensive simulation testing (21) has demonstrated that the staggering improvement assumed here is valid and even conservative over a wide range of capacities and frequency assignments. However, specific frequency plans do exist which will not provide the improvement assumed; the worst of these is equal spacing, in which case there is no staggering improvement at all.

A voice activated SCPC system is one in which the transmitted carrier is up only when speech activity is actually present. Using the statistics of speech activity in a full duplex channel and considering large numbers of voice channels, an improvement in power sharing of the satellite transponder between voice channels can be accomplished. The amount of power sharing is dependent on the voice activity factor

assumed for speech and the burst transmission overhead (pre-amble plus hangover). A factor of .4 is commonly used and is the value assumed in this study.

Voice activation increases the randomness of IM products and is sometimes assumed to provide a C/I improvement (1).

In our case, however, a staggering improvement has already been assumed through the random selection of carrier frequencies. When voice activation is added the number of active channels is increased but the average number of carriers present in the transponder and their levels remain virtually unchanged; therefore the link equations previously developed continue to apply. If anything voice activation makes it more difficult to realize the staggering improvements assumed. However, since all IM products and carriers will be modulated we do achieve the following real additional advantages:

1. Some IM energy over the 45 kHz allocated channel bandwidth will be attenuated by the SCPC receive filter (i.e. the system noise bandwidth is less than the channel spacing). The demultiplexing advantage obtained here is typically .5 to 1.0 dB.

2. The elimination of unmodulated carriers and interfering RF tones ensures that the FM advantage assumed in link calculations applies to the IM noise as well as the white thermal noise;i.e. any error which could be incurred by adding the IM noise as flat noise in the non-activated system does not exist for the voice activated systems.

In summary, it seems reasonable to conclude that the power limited capacity of an SCPC transponder employing voice-activation techniques is simply increased by a factor 2.5. The voice-activated capacity of the candidate systems previously defined are presented in Figure 13.

A system factor that affects the improvement that can be gained using voice activation and should be carefully considered when designing an operation system is the amount of data or facsimile traffic the system is expected to carry. As there will be no voice activation advantage while these types of signals are present, a system designed to carry only voice traffic could become overloaded if data or facsimile traffic is not accounted for in the system loading.

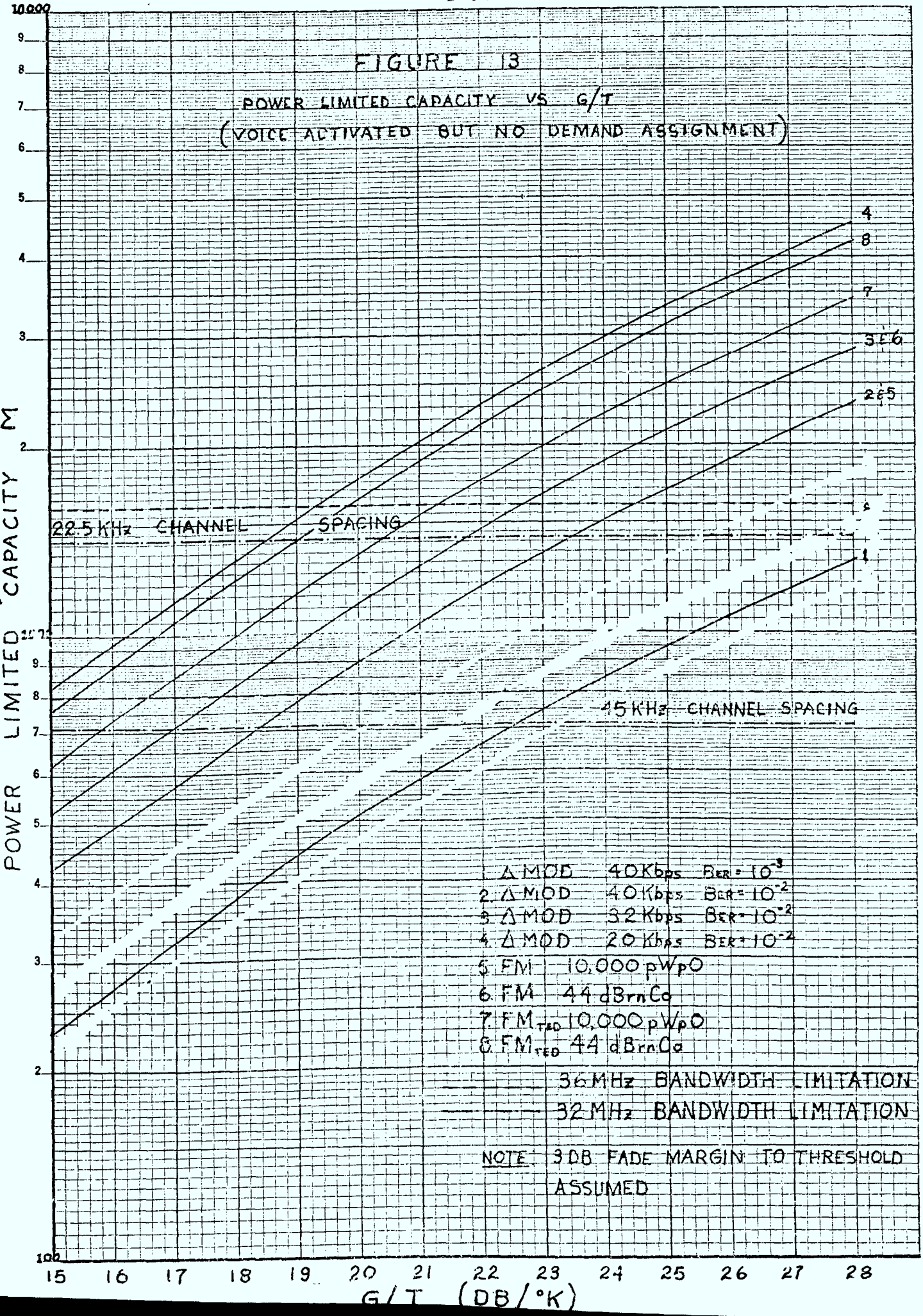


FIGURE 13
POWER LIMITED CAPACITY VS G/T
(VOICE ACTIVATED BUT NO DEMAND ASSIGNMENT)
POWER LIMITED CAPACITY M
G/T (DB/°K)
22.5 KHz CHANNEL SPACING
45 KHz CHANNEL SPACING
4
8
7
3 & 6
2 & 5
1
1 ΔMOD 40Kbps BER = 10⁻³
2 ΔMOD 40Kbps BER = 10⁻²
3 ΔMOD 32Kbps BER = 10⁻²
4 ΔMOD 20Kbps BER = 10⁻²
5 FM 10,000 pWpO
6 FM 44 dBrnCo
7 FM TED 10,000 pWpO
8 FM TED 44 dBrnCo
36 MHz BANDWIDTH LIMITATION
32 MHz BANDWIDTH LIMITATION
NOTE: 3DB FADE MARGIN TO THRESHOLD ASSUMED

3.2.4 Demand Assigned Capacity vs G/T for a Homogeneous Network

In a demand assignment system, the pool of available satellite voice channels can be accessed by any community (i.e. earth station) that requires a voice channel on a first-come-first-served basis.

It is not the intent of this Appendix to delve into all the operational/design aspects of a demand assignment system (these are examined in detail in Appendix D) but to provide estimates of the traffic that a demand assignment system can support.

A suitable demand assignment configuration for serving remote areas is the centralized control system developed by Campbell and Lambert (23). In this system three one way signalling channels (fixed in frequency) must be assigned between the central and remote stations.

These three channels are:

Request Channel - from remote station to central station requesting service and frequency (channel) assignments.

Assignment Channel - from central station to remote terminals providing frequency

assignment information, calling party and called party information for billing, etc.

Hang-up Channel - information from remote terminals to central terminal on call termination for centralized billing and system status purposes.

To arrive at the total number of access channels a demand assignment system can support, traffic engineering is required and a few assumptions regarding access trunk loading and blocking probabilities must be made. Figure 14a portrays a demand assignment layout showing earth station access trunks and satellite access trunks. The satellite transponder has N_S available simultaneous trunk (i.e. two way circuit) accesses. Since any one of these trunks is available to any earth station on demand, the transponder can be regarded as one large trunk group accessed by many more users (i.e. earth terminals). The traffic the satellite transponder can support (i.e. N_M access trunks) is dependent solely on the number of available channels, the traffic statistics assumed and the blocking probabilities (i.e., grade of service) that can be tolerated in the system design.

The assumptions made in this report on traffic statistics are:

- Erlang B traffic statistics which assumes infinite source (i.e. Poisson arrival distribution) and that lost calls are cleared

- A blocking probability of 3% (P(03)) which allows for 3 calls in a 100 calls to be lost.

To determine the total number of earth station access trunks and/or earth stations the system can support, an assumption must be made regarding the efficiency (or loading) of the earth station access trunks. If we are considering remote area applications, the traffic requirements are generally small and would be expected to require from 1 - 6 access trunks per community. For such small access trunk groups and to maintain a P(03) blocking probability, the efficiency ranges from 0.2 to 0.4 Erlangs per trunk.

As an example assume that N_s (the total available number of simultaneous transponder accesses) is 350 (i.e. 700 simultaneous one-way channels) and assume that the access trunk efficiency is 0.3 Erlang per trunk. Thus using

Erlang B tables, the total traffic the transponder can support for P(03) blocking is 342.95 Erlangs. Using a 0.3 Erlang/trunk access results in approximately 1142 trunk accesses. Thus the system could support 1142 communities (earth stations) with one access per station or approximately 190 communities with 6 access trunks each or a mixture of earth stations/access trunks between the two limits.

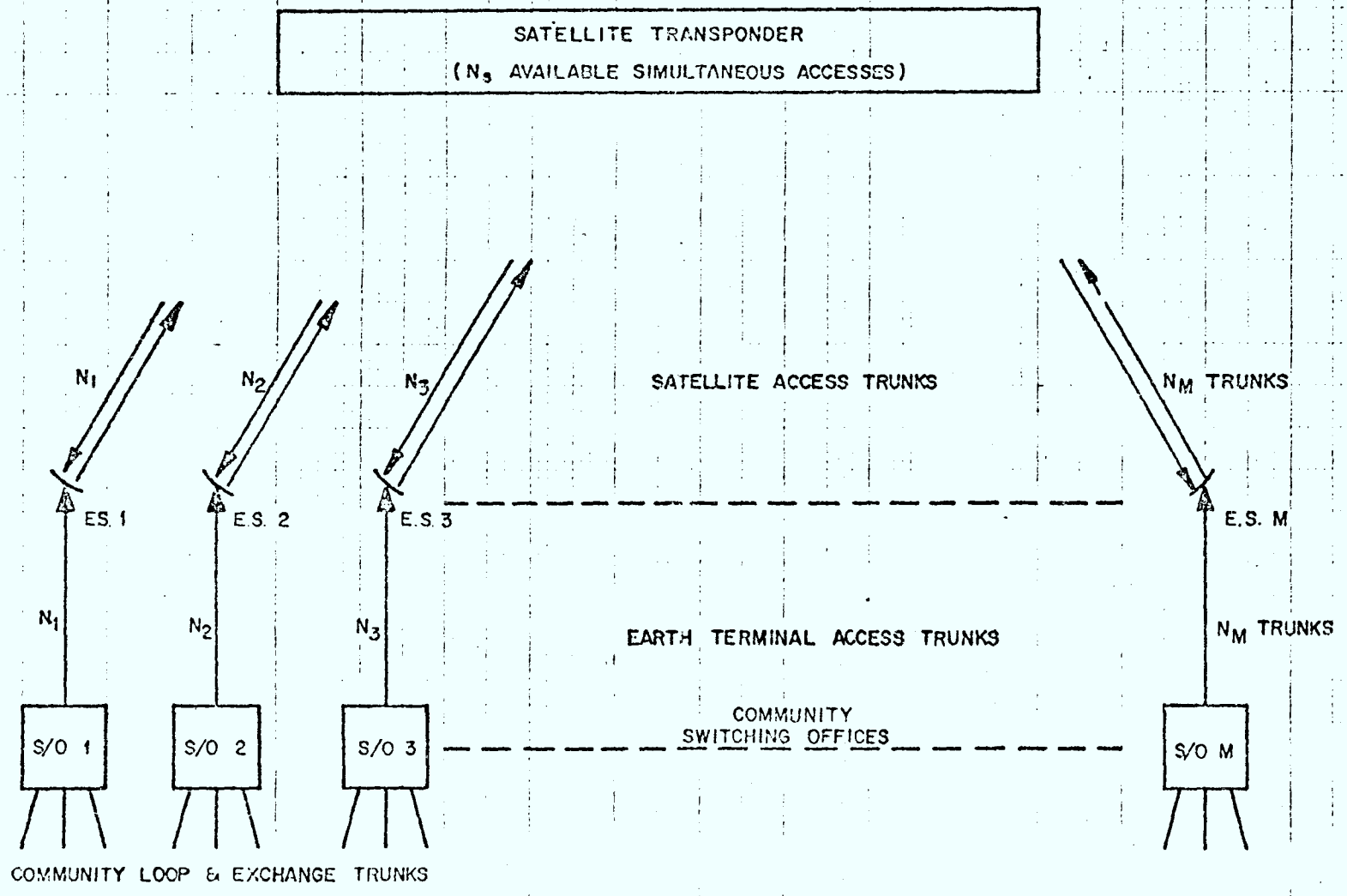
Since the traffic loading on the local access trunks is not known, the capacity of the demand assignment system will be expressed in terms of Erlangs of traffic using a call blocking probability of P(03) grade of service.

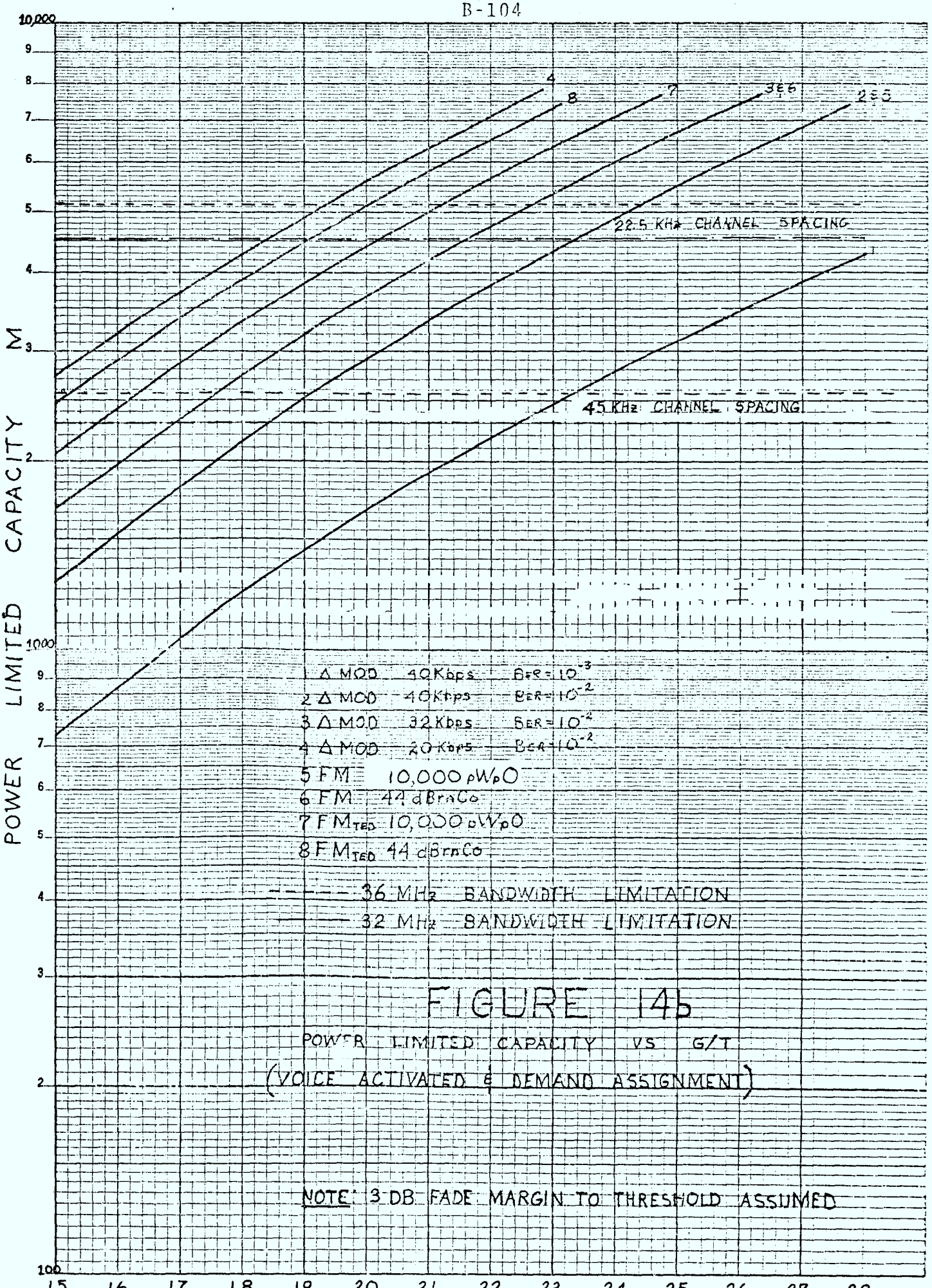
Figure 14b plots the capacity of the previously described candidate systems with voice activation and demand assignment.*

* Considering the higher costs of implementation of demand assignment it is highly unlikely that voice activation would not be used to obtain the maximum efficiency in the system.

FIGURE- 14a

DEMAND-ASSIGNMENT TRUNK ACCESS ARRANGEMENT





K+E SEMI-LOGARITHMIC 46 4973
2 CYCLES X 70 DIVISIONS MADE IN U.S.A.
KEUFFEL & ESSER CO.

FIGURE 14b

POWER LIMITED CAPACITY VS G/T

(VOICE ACTIVATED & DEMAND ASSIGNMENT)

NOTE: 3 DB FADE MARGIN TO THRESHOLD ASSUMED

3.3 Optimizing a Dynamic Growth Model

3.3.1 Dynamic Model

If future traffic requirements are completely known, successive earth station G/T's in a developing network can be chosen such that:

1. the network ultimately utilizes the full capability of the allocated transponder(s) and at no time tries to exceed it.

2. the total cost of the network expressed in starting year dollars is minimized.

If space segment costs are utilization independent, it is clear that such a minimum-cost network will initially employ cheaper lower G/T earth stations and evolve through addition and up-grading to include larger G/T's as the space segment becomes filled. If the traffic growth pattern is known a priori, a cost-minimization algorithm such as that presented here can be applied to determine the optimum G/T's.

The following growth-cost model and corresponding cost

minimization algorithm are presented to offer insight into the problem and indicate the kind of analysis that can be applied. Numerical results are not included here since no realistic estimate of traffic growth is available.

3.3.2 Cost-Minimization Algorithm

A network will ultimately interconnect M earth stations over an N year growth period. In the i'th year the j'th station must terminate n_{ij} circuits. (If $n_{ij} = 0$, the earth station need not exist up to the i'th year). The optimum earth station G/T to provide ℓ circuits in the i'th year is $(G/T)_{i\ell}$.

For a given C/N_o requirement the cost of an earth station is assumed to be given as a time invariant function of G/T and number of circuits terminated; i.e.,

Earth station Annual Charge = f(G/T,n)

Note that f(G/T, o) = 0. This equation assumes as before that HPA costs do not dominate in the selection of antenna diameter, and further ignores the changing cost of hardware over the N-year growth period of the

network. Suppose:

S_i = space segment (i.e. transponder) annual charge in the i'th year

and

= value of money expressed as a percentage over 1 year (typically 10%),

then the total cost of meeting the network requirements up to the N'th year expressed in equivalent starting year dollars is:

$$C = \sum_{i=1}^{N} (1 - \varepsilon)^{i-1} \left(\sum_{j=i}^{M} f(G/T_{i,n_{ij}}, n_{ij}) \right) +$$

$$+ \sum_{i=i}^{N} (1 - \varepsilon)^{i-1} S_i$$

The objective, then, stated mathematically, is to minimize the above cost function with respect to the variables

$G/T_{i,n_{ij}}$ = G/T of the j'th station in the i'th year

under the constraint that $\left(\frac{C}{N_o}\right)_{kj} \geq \left(\frac{C}{N_o}\right)$ required

Considering down-link thermal noise as being the prime contributor at the desired operating point, this constraint can be expressed approximately (see previously derived link equations):

$$\sum_{j=i}^{m} \frac{n_{ij}}{(G/T)_{ij}} \leq \frac{(EIRP\text{-}OBO)\quad (k)}{(FSL)\left(\frac{C}{N_o}\right)_{required}} \quad , \text{ for } i=i, \ldots, N$$

Minimization of the cost function under this constraint ensures minimum expenditure over and following the N-year growth period. The cost equation, however, neglects the non-capital costs of up-grading an earth station. A reasonable additional constraint might be to not alter earth station G/T in a year no additional circuits are provided; this constraint on the variables is written:

$$(G/T)_{1,n} = (G/T)_{2,n} = \ldots. = (G/T)_{N,n}$$

The further constraint that no earth station once installed with have its G/T altered even if circuits are added can be expressed as:

$$(G/T)_{i,n_{ij}} = (G/T)_{i_o,\, n_{i_o j}} \quad \text{for } i \geq i_o$$

where i_o is the year of installation.

If space segment is charged on an actual utilization basis, then S_i in the cost function equation depends on G/T and n_{ij}, which in turn means that space segment costs directly influence the selection of earth station G/T's. If, on the other hand, a fixed component of space segment is initially allocated at fixed cost to the SCPC network, the optimum choice of G/T's becomes independent of space segment loading up to the available capacity. The latter approach is clearly more applicable to the Canadian scene, in which space segment must presently be purchased on a per transponder basis. There is nothing to be gained in this case by underutilizing the assigned satellite resources.

The problem considered here, then is simply to minimize the cost of an earth segment that meets a required traffic demand using a given space segment*.

---

* The demand itself may depend on the total annual charge per access, which includes a space segment charge. In addition, the required G/T depends in part on satellite EIRP, which is related to the initial cost of the satellite (i.e. if the annual space segment charge per transponder were made a function of the EIRP provided we could simultaneously optimize both ground and space segments to meet a given requirement!) This indirect relationship between earth segment and space segment costs are noted, but again are not applicable to the Canadian situation in which in-orbit satellites already exist (i.e. have been paid for) and excess capacity is available.

To solve this problem, then, there remains only to derive a reasonable earth station cost function f (G/T, n). The cost information required to develop this equation for various modulation techniques and circuit qualities is provided in this study.

3.4 Effects of Adjacent Satellite Interference

This section develops equations required to compute the C/I and corresponding baseband distortion due to the presence of adjacent satellite interference. The effects of various types of interference, namely single and multiple access FDM/FM message, FM television, PSK/TDMA, and SCPC/FDMA on each of the candidate SCPC systems are evaluated vs antenna diameter assuming 5^{o} satellite spacing. The overall issue of orbit-spectrum utilization and its influence in the design of SCPC systems is introduced, and several methods of reducing interference to and from SCPC systems discussed briefly. Since the variables are numerous and involve both national policy and international (CCIR) regulations as well as future co-ordination agreements between administrations, a detailed examination of all aspects of the problem is not undertaken and definitive results are not produced. The objective here is rather to demonstrate how the effects of adjacent satellite interference can be added to the analysis, to compare the sensitivity of the candidate

systems to such interference, and to rationalize a recommended lower limit on antenna diameter.

3.4.1 "Allocated" vs Computed Interference

For FM telephony systems, the CCIR (Rec. 466, Rev. 72) stipulates that 1000 pWpO is the maximum total allowable contribution of adjacent satellite interference and 400 pWpO is the maximum interference from any one adjacent satellite system. Using these criteria and assuming complete homogeneity of satellite systems and antenna side-lobe characteristics, the minimum satellite spacing depends on antenna diameter and FM modulation index (24).

For example, for a 33' station having a receive G/T of 28 dB/^{o}K and employing a modulation index of 1.5, the satellite spacing must exceed 4^{o} (24).

The stated CCIR interference criteria were developed for and are strictly applicable only to multichannel FM systems. While companded SCPC FM systems are not excluded, it is unlikely that they were anticipated at the time recommendation 466 was prepared. Interference criteria for digital links, in particular variable-slope Δ-mod/CPSK systems, have not been finalized and are presently under review in Study Group 4. It appears

likely that the following criteria will be adopted for both analog and digital message systems:

1. For antenna diameters greater than 100λ (λ = receive wavelength) the 1000 pOpO and 400 pOpO limits will apply. For diameters less than 100λ (25' at 4 GHz) interference allocation will increase as 20 log 100λ/D) dB.

2. Off-axis EIRP transmission at an angle θ will be limited to 38.5 -25 log θ dBW/4kHz (Canadian position) or less.

Telesat employs antennas of smaller diameter (as low as 12') than those proposed for use in the Intelsat and U.S. domestic networks. To ensure that its existing communications systems are not subject to unacceptable levels of interference, Telesat assumes existing CCIR interference allocations in its link budgets and has resisted the limiting of antenna diameter or the reduction of orbital spacing to less than 5^{o}. However, as the demand for orbit space grows the pressure to use it efficiently increases. This pressure conflicts with the desirability of low cost 4/6 GHz satellite earth terminals to provide communications into remote parts of Canada.

It appears unlikely that a fixed interference allocation will remain a satisfactory means of effectively designing co-channel satellite systems and at the same time making efficient use of the geostationary orbit. It appears already that the allowable circuit noise due to interference will be modified in the case of small antennas.

Although it does provide a measure of security and protection to administrations, fixed interference limits and corresponding allocations are somewhat artificial, especially in view of the rapid development of domestic and regional satellite systems having quite different objectives.

An alternative to the interference allocation approach is to compute the interference to and from specific anticipated systems, and to design and co-ordinate these systems appropriately. This permits much greater flexibility in the design of satellite systems and the use of geo-stationary orbit, and will ultimately be endorsed at the CCIR (25).

Our approach, then, is not to allocate a fixed noise contribution due to adjacent satellite interference to all candidate systems, which would provide no insight

whatsoever into their interference susceptibility. Rather, an effective comparison is made by analyzing the reduction in SCPC capacity resulting from specific adjacent satellite interferers.

3.4.2 Computing the C/I Ratio

It is desired to compute the C/I ratio at the demodulator input due to the presence of a co-channel adjacent satellite system. This interference can occur both on the up-link (EIRP transmitted off-axis to wanted satellite) and down-link (EIRP received off-axis from adjacent satellite). Both depend on earth station antenna discrimination. In the case of interference into an SCPC system, down-link interference dominates since the 4 GHz SCPC receive antenna gain is expected to be much less than the 6 GHz gain of an antenna pointing at the adjacent satellite.

The following parameters for a homogeneous SCPC system and an interfering adjacent satellite system are defined:

Homogeneous SCPC System:

m = number of simultaneous carriers

OBO = satellite output backoff (dB)

$G(\theta)$ = receive gain at an off-axis angle θ degrees (dB)

T = receive noise temperature (°K)

B = SCPC noise bandwidth (Hz)

k = Boltzmann's constant

Interfering System:

P(f) = equivalent isotropic radiated flux density (watts/Hz) from interfering satellite

For a satellite spacing of θ degrees*, under clear weather conditions, the down-link carrier-to-noise and carrier-to-interference ratio are given by

$$\left(\frac{C}{N}\right)_{down} = EIRP - OBO - 10 \log (m) - 10 \log (B) + G (0) - 10 \log (T) + k - FSL$$

---

* In principle θ is the tropocentric angle, but since it approximates the geocentric angle, these two angles are used interchangeably.

and

$$\left(\frac{C}{I}\right)_{down} = EIRP - OBO - 10 \log (m) + G(0) - G(\theta) - EIRP_i$$

where:

$$EIRP_i = 10 \log \int_{[f_c - 1/2\ B]}^{[f_c + 1/2\ B]} P(f)\, df$$

is the interfering EIRP from the adjacent satellite.

The difference between the interference and thermal noise at the demodulator input is then given by:

$$D = \frac{C}{I} - \frac{C}{N}$$

$$= 10 \log (B) + 10 \log (T) + FSL - k - G(\theta) - EIRP_i$$

$$= 10 \log (T) + FSL - k - I_o - G(\theta) \qquad (1)$$

where I_o is the interfering EIRP per Hz.

On an RF power addition basis, the increase in noise due to interference is simply $1 + 10^{-.1D}$ which depends only on:

1. The noise temperature of the receiving station

2. The EIRP per Hz from the interfering satellite

3. The receiving antenna gain towards the interfering satellite

Assuming in the absence of more precisely known data the reference radiation pattern given in recommendation 465 (rev. 72).

$G(\theta) = 32 - 25 \log \theta$,

Figure 15 plots the increase in noise vs EIRP/kHz of the adjacent satellite interference for an off-axis angle of 5° and noise temperatures ranging from 60°K to 600°K.

3.4.3 Computing the Baseband Distortion

3.4.3.1 Stationary Wide-band Interference

When the interfering carrier has a bandwidth much greater than the SCPC carrier bandwidth, its effect is equivalent to that of thermal noise*.

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* A sufficiently narrow bandwidth of a stationary random process is asymptotically gaussian.

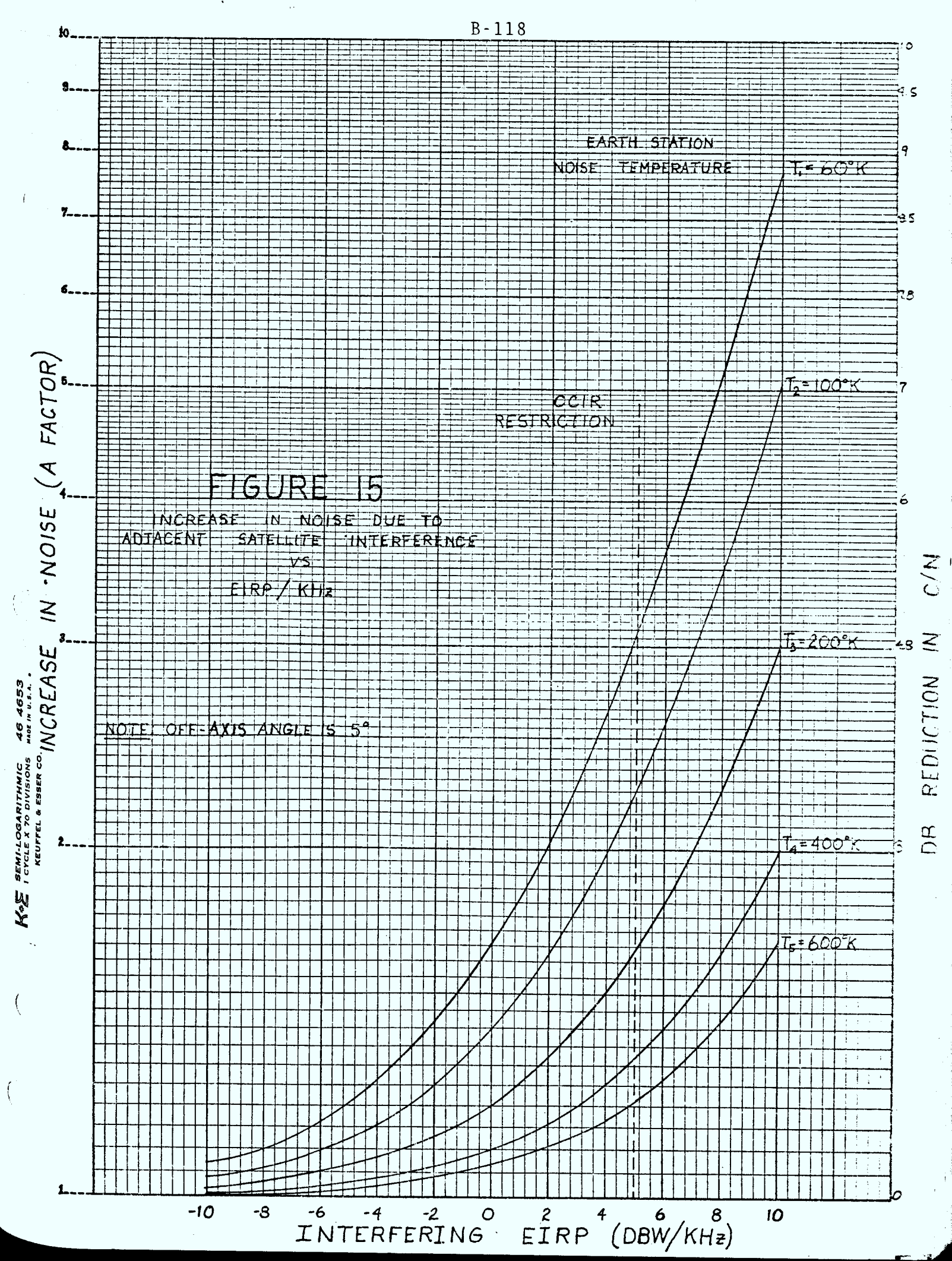


FIGURE 15

INCREASE IN NOISE DUE TO ADJACENT SATELLITE INTERFERENCE VS EIRP / KHz

The reduction in effective down-link carrier-to-noise ratio due to the interference, or the corresponding increase in G/T required to compensate its effect, is therefore given by the results of Figure 15.

The following types of interfering traffic fall into the wide-band category:

- FDM/FM/FDMA (with energy dispersal present during light loading)

- single carrier FDM/FM (e.g. 960 channel message)

- PSK/TDMA (scramblers are normally used to ensure random bit patterns)

FM/TV interference is not immediately included in this category because it possesses some non-stationary characteristics as well as harmonic components.

3.4.3.2 Stationary Narrow-band Interference

For interfering carrier bandwidths approaching the SCPC carrier bandwidth, the interference at the SCPC demodulator input is neither white nor gaussian. Adding it on a power basis to the thermal noise does not necessarily lead to an accurate prediction of its effects on voice circuit quality. The following treatments are required for random phase modulated sinusoidal interference into FM and Δ-mod/CPSK SCPC links.

3.4.3.2.1 Interference into FM/SCPC

Above threshold RF interference into an FM system produces a baseband interference spectrum which depends mostly on (26):

- the spectral distributions of the wanted and interfering carriers

- the separation between the carrier frequencies

A general treatment of FM baseband distortion due to thermal noise and multiple interference is given by Wong (26). In most practical cases of interest, the shape of baseband interference spectrum following FM demodulation and ideal (6 dB/octave) de-emphasis can be determined by convolving the wanted and interfering spectra (27). Under worst case conditions all of the interference power falls within the 300 to 3400 Hz voice channel bandwidth; i.e., there is no bandwidth expansion FM improvement . On the other hand, for sufficient separation between the wanted and interfering carriers, the FM improvement exceeds that for thermal noise. For a multi-channel telephony system subject to an unmodulated or lightly modulated interference*, the test tone-to-interference ratio is generally expressed in the form (27):

---

* Interference from digital or large modulation index FM carriers into analog FM systems is computed by the convolution integral referred to in (27). However, the spectral densities of these signals cannot be easily generalized, and a comprehensive set of B factor computations not easily generated.

$$\frac{TT}{I} = \frac{C}{I} + B$$

where:

B = "interference reduction factor"

$$= 10 \log \left\{ \frac{1}{bf^2} \cdot \frac{2\sqrt{2}\ \Delta tt_{RMS}^2\ \Delta f_{RMS}\ P}{\exp\left(\frac{-(f_o - f)^2}{2\Delta f_{RMS}^2}\right) + \exp\left(\frac{-(f_o + f)^2}{2\Delta f_{RMS}^2}\right)} \right\}$$

f = test tone frequency (in SCPC case)

= 1000 Hz

b = bandwidth of a telephone channel

= 3100 Hz

Δtt_{RMS} = rms test tone deviation of wanted SCPC signal

Δf_{RMS} = rms signal deviation of the wanted signal

P = pre-emphasis advantage (5.8 dB)

f_o = frequency separation of wanted and unwanted channels.

Referring to equation (1) in section 3.1.1.1, the ratio of the FM improvement factors applied to interference and noise can be written:

$$R = 10 \log \left\{ \frac{2\sqrt{2}\ \Delta f_{RMS}\ f_m^{\,3}}{3b\ f^2 B_{CR} \left[\exp\left(\frac{-(f_o - f_m)^2}{2\Delta f_{RMS}^2} \right) + \exp\left(\frac{-(f_o + f_m)^2}{2\Delta f_{RMS}^2} \right) \right]} \right\}$$

Figure 16 plots R vs f_o for a companded FM SCPC system having a peak-to-rms ratio (see section 3.1)

g = 7.6 dB

and a Carson's Rule Bandwidth

$$B_{CR} = 2\ (g\Delta f_{RMS} + f_m)$$

of 20, 40, and 60 kHz

Given the R factor presented in Figure 16, the effects of interference appearing in the voice channel baseband above threshold can be approximated simply by:

1. adding the R factor to the C/I ratio

2. adding the RF interference to the thermal noise on an RF power basis (see equation (1)).

This simple approach, however, possesses the following disadvantages:

1. It assumes stationary gaussian wanted and interfering spectra. While this assumption is valid for FDM/FM carriers, it is clearly not the case for SCPC/FM systems. This disadvantage can be circumvented by considering a reasonable range of wanted deviations Δf_{RMS}.

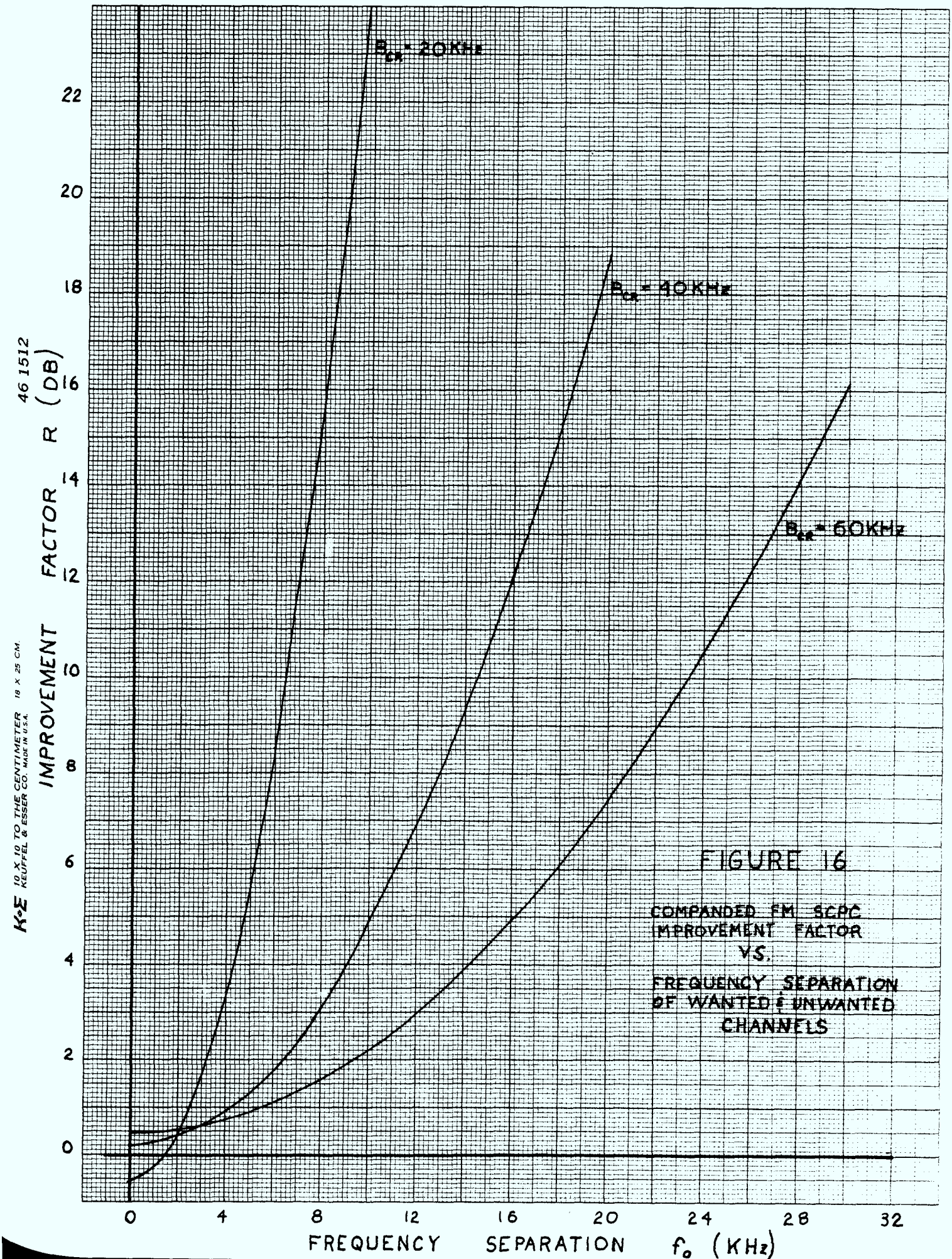


FIGURE 16

COMPANDED FM SCPC
IMPROVEMENT FACTOR
VS.
FREQUENCY SEPARATION
OF WANTED & UNWANTED
CHANNELS

2. It does not predict the performance in the vicinity of threshold or the effects of interference on the C/N threshold characteristic.

3. It assumes a companding advantage identical to that for thermal noise.

4. It neglects possible subjective effects such as intelligible crosstalk, undesirable baseband tones, etc.

A literature search revealed no general treatment of interference into FM/SCPC systems. Further study of the subject is considered beyond the scope of this report. It is conjectured that the stated approach is accurate providing the carrier-to-interference ratio is well above the carrier-to-noise ratio.

This assumption seems reasonable in view of the general CCIR co-ordination policy which typically allocates only a small fraction of a total noise budget to interference.

Narrow-band interferers into SCPC systems include:

- low capacity FDM/FM/FDMA carriers possessing no energy dispersal
- SCPC and radio program carriers
- RF harmonic components of a TV/FM carrier possessing no energy dispersal
- other narrow band traffic.

3.4.3.2.2 Interference into Δ-mod/CPSK/SCPC

RF interference in a digital transmission system is not demodulated directly into baseband interference, but rather increases the bit error rate and the corresponding level of bit error noise. A digital SCPC system is in general less sensitive than an FM/SCPC system to narrow band interference for the following reasons:

1. The effects of interference can be completely compensated by an increase in the carrier-to-thermal noise ratio. Due to its peak-limited nature, the degradation due to sine wave interference is normally less than that of an equivalent power of thermal noise.

2. Unlike FM, no subjective degradation due to baseband tones or intelligible crosstalk will occur.

The bit error rate of a CPSK system in the presence of gaussian noise and interference depends on the carrier-to-noise and carrier-to-interference ratios, as well as the envelope probability density function of the interference. Jeruchim and Lilley (28) have analyzed the case of multiple sine wave (i.e. constant amplitude) interference, and computed the degradation due to adjacent satellite interference for a homogeneous

set of equally spaced satellites employing multi-phase CPSK. Rosenbaum (29) has derived expressions for the error probability of multi-phase CPSK and DPSK systems corrupted by noise and a single sine wave interferer.

Figures 17 and 18 reproduce Jeruchim and Lilley's results of error probability vs C/N at several satellite separation angles for 2 and 4 phase CPSK systems respectively. Note that the curves are parametric with respect to D/λ vs satellite spacing, and therefore specify the C/N degradation vs separation angle for any antenna diameter. Figures 19 and 20 plot required separation angles vs antenna diameter for C/N degradations of .5 dB, 1.0 dB, and 2.0 dB at a 10^{-3} error rate for 2 phase and 4 phase CPSK systems. Note that these results only apply to a homogeneous set of equally spaced satellites with identical SCPC transmission parameters*.

---

* As pointed out in section 2.5 of reference (22), the homogeneous configuration makes the most efficient use of the geostationary orbit.

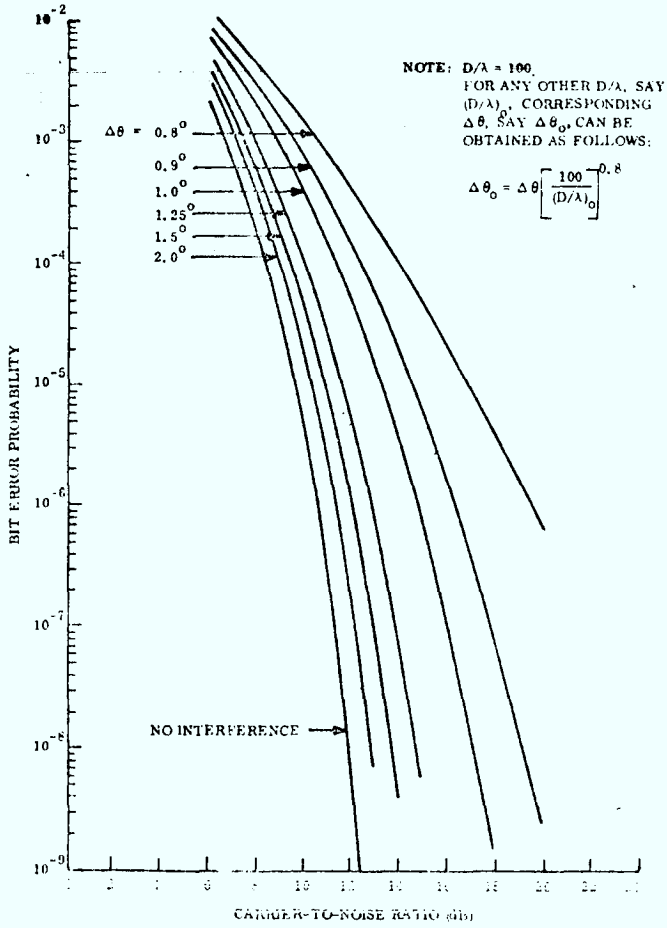


FIG. 17 Error rate for binary CPSK.

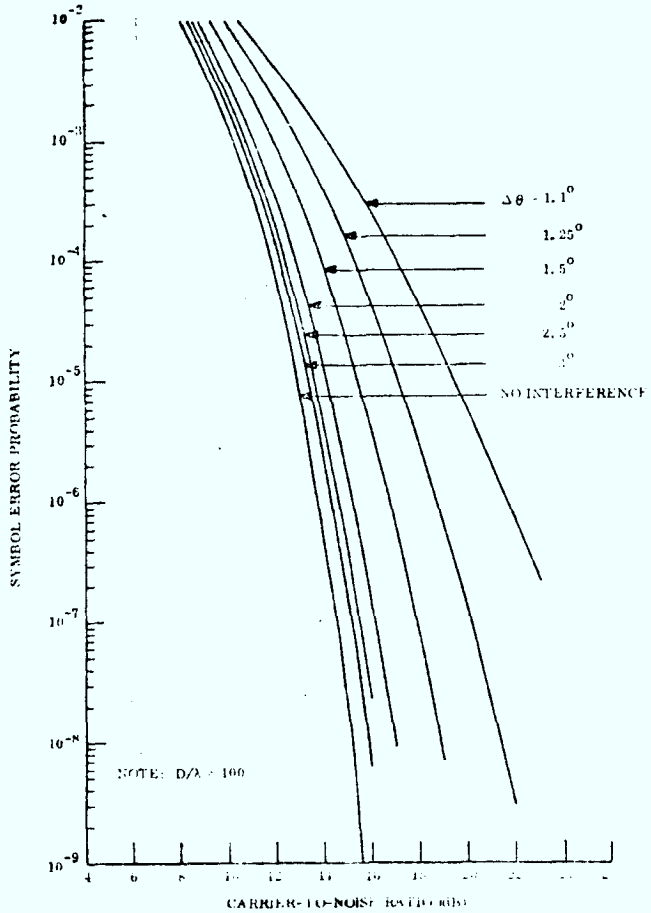


FIG 18 Error rate for 4-phase CPSK.

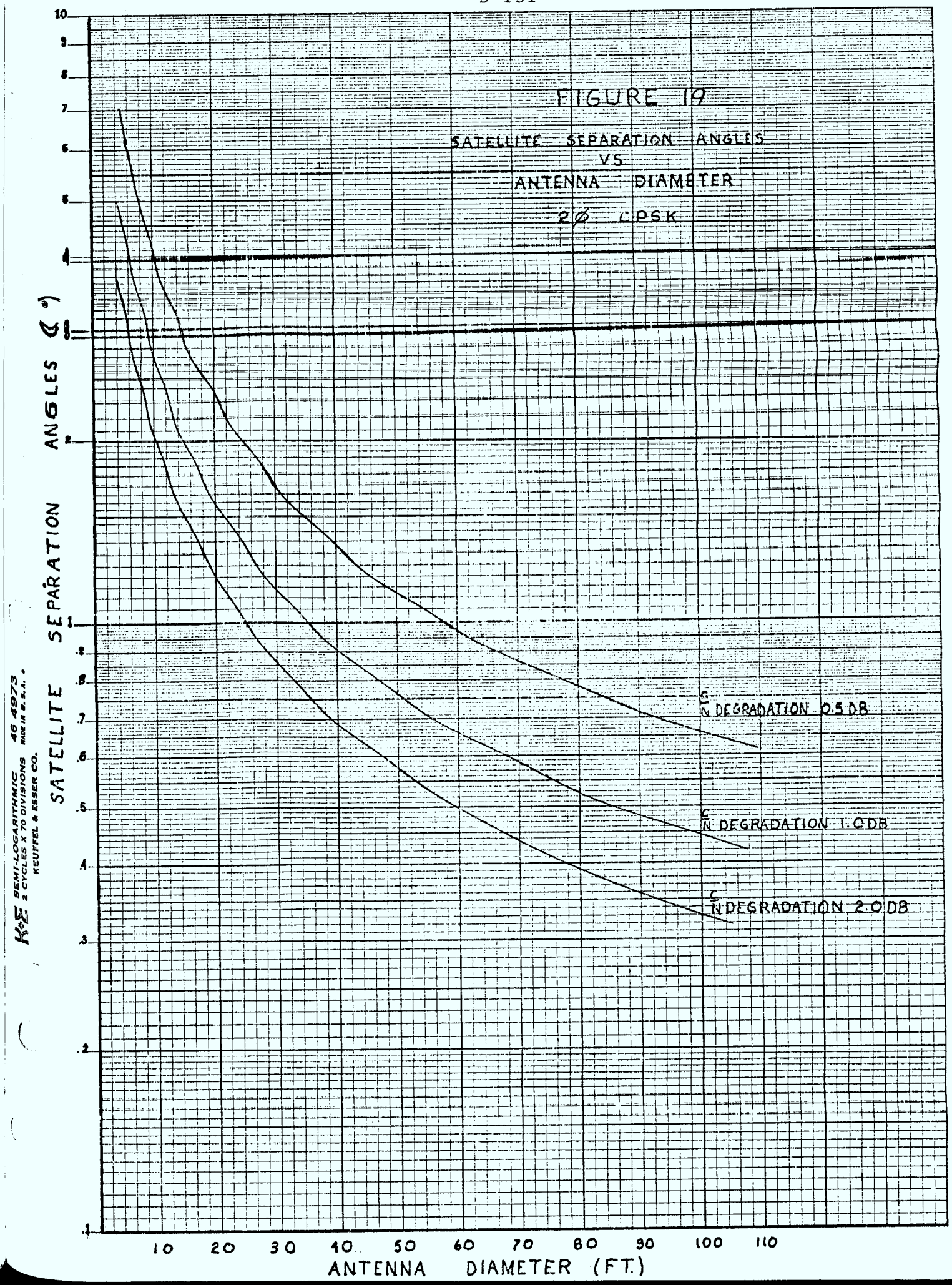


FIGURE 19
SATELLITE SEPARATION ANGLES
VS.
ANTENNA DIAMETER
2Ø BPSK
SATELLITE SEPARATION ANGLES (°)
10
9
8
7
6
5
4
3
2
1
.8
.7
.6
.5
.4
.3
.2
.1
Eb/N DEGRADATION 0.5 DB
Eb/N DEGRADATION 1.0 DB
Eb/N DEGRADATION 2.0 DB
10
20
30
40
50
60
70
80
90
100
110
ANTENNA DIAMETER (FT.)
K&E SEMI-LOGARITHMIC 46 4973
3 CYCLES X 70 DIVISIONS MADE IN U.S.A.
KEUFFEL & ESSER CO.

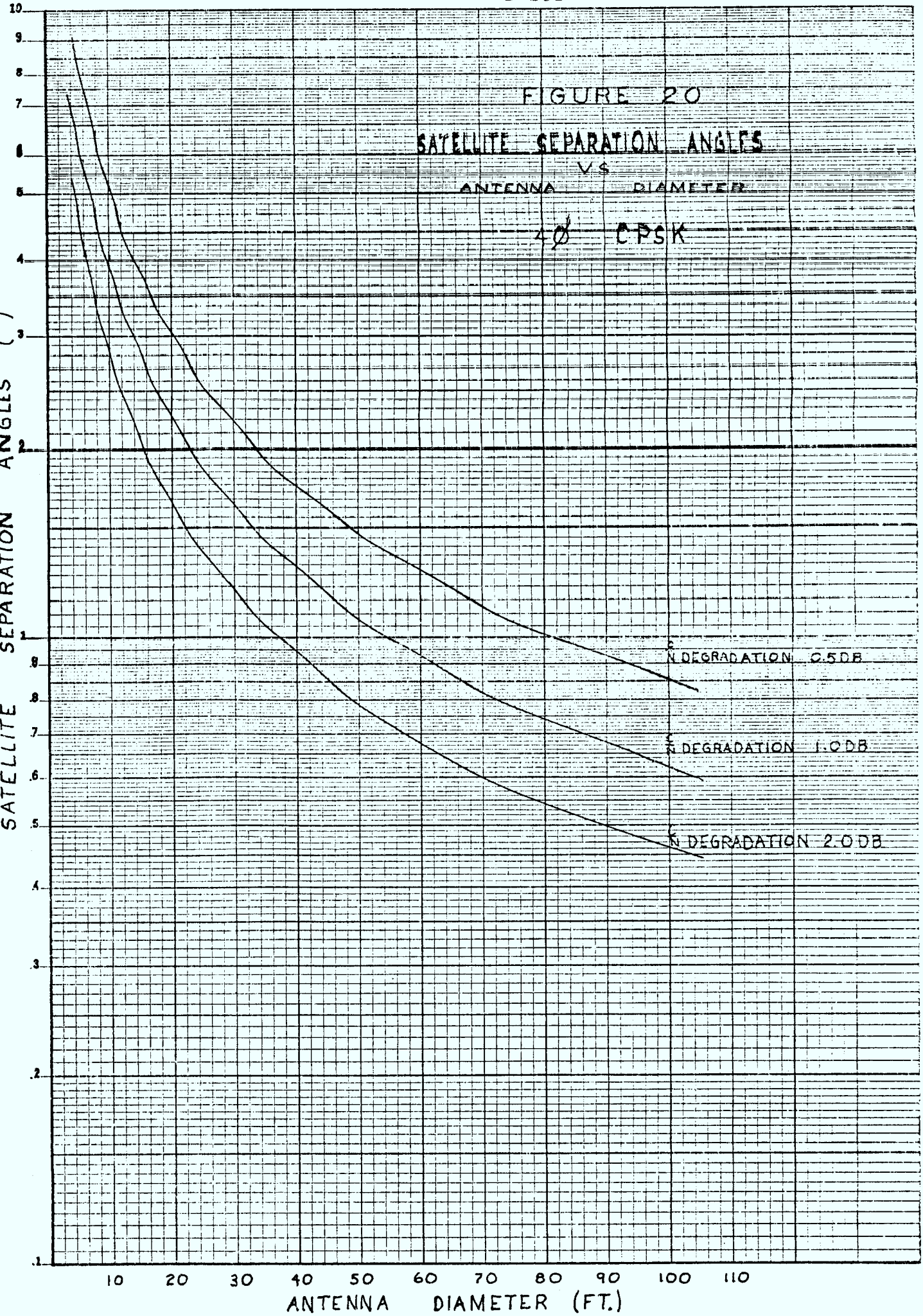


FIGURE 20
SATELLITE SEPARATION ANGLES
VS
ANTENNA DIAMETER
4Ø CPSK
SATELLITE SEPARATION ANGLES (°)
ANTENNA DIAMETER (FT.)
Eb/N DEGRADATION 0.5DB
Eb/N DEGRADATION 1.0DB
Eb/N DEGRADATION 2.0DB
10
9
8
7
6
5
4
3
2
1
.9
.8
.7
.6
.5
.4
.3
.2
.1
10
20
30
40
50
60
70
80
90
100
110
SEMI-LOGARITHMIC
2 CYCLES X 70 DIVISIONS
KEUFFEL & ESSER CO.
MADE IN U.S.A.
46 4973

3.4.3.3 Non-Stationary Interference

For our purposes non-stationary interference is defined as interference which produces variations in performance at the output of the SCPC demodulator. If the variation is sufficiently rapid as to be imperceptible to the ear, the average level of baseband interference can be considered as an objective for voice transmission. For voice-band data transmission the ensemble of interference statistics and the combined effect on bit error rate should be estimated. This analysis is normally circumvented by designing to the maximum level of interference. When the interference varies at a perceptible rate, it is not obvious what the subjective effects will be or what design procedure should be adopted. Designing the system to meet overall noise objectives under worst case interference conditions normally ensures satisfactory performance.

Because they are relatively narrow band and low power, both FM and digital SCPC carriers are volnerable to various types of non-stationary interference that could be generated in an ad-

jacent satellite system. Among these are:

1. Slow periodically time-varying RF spectral components of a TV carrier.

2. A TV carrier modulated only with slow sweep (30 Hz) energy dispersal

3. On/off initial acquisition ranging tone(s) in a TDMA system

4. Components of TDMA carrier at burst repetition frequency

5. Swept frequency carrier generated by link analyzer.

All of these processes vary at a rate less than the bandwidth of the SCPC carrier, and therefore produce transient or time-varying effects not predicted by quasi-stationary analysis. The average carrier-to-interference ratio can be high but the degradation severe!

One case of this type of interference which has caused some concern is a saturating TV carrier possessing energy dispersal modulation only. The objective of the energy dispersal is to uniformly spread the energy of the signal across 1 MHz (or more) of bandwidth in order to meet the CCIR radiated EIRP/4 kHz restriction. This restriction is designed to protect low modulation index multi-channel terrestrial FM systems, not adjacent satellite systems.

While energy dispersal uniformly spreads RF power, the signal spectrum is more accurately described as a single line periodically sweeping a 1 MHz bandwidth at a 30 Hz rate. While this rate is sufficiently high to ensure that individual 4 kHz circuits in a co-channel FDM/FM system will only be degraded by the average interference power, it does not imply that such is the case for a co-channel SCPC system.

Consider an SCPC system having a 45 kHz channel bandwidth. For 30 Hz triangular energy dispersal over 1 MHz peak-to-peak, the interfering carrier will "hit" each SCPC carrier in this 1 MHz band

twice every 1/30 second for a duration of

$$T_i = \frac{45}{60}$$

$$= .75 \text{ msec}$$

Noting that

T_i > time constant of SCPC receive filter (symbol length $\simeq$ .05 msec for digital SCPC),
we conclude that the full interfering carrier power will appear periodically at the SCPC demodulator input. For an FM system, the momentary operation below threshold will produce clicks into the baseband, disrupting both data and voice transmission. For a Δ-mod system a momentary increase in error probability will disrupt data transmission although it <u>may</u> be acceptable for voice.*

Other possible degrading effects of non-stationary interference are:

---

* The Δ-mod system has a more graceful threshold characteristic and furthermore is normally dominated by quantization noise rather than bit error noise. Assuming the interference does not disrupt carrier recovery and bit timing circuits, a digital SCPC system is probably less sensitive to this type of interference.

- false triggering of circuit activation on the receive side of a voice-activated SCPC link

- disruption of critical control information in a demand assignment system.

Until experiments presently underway prove otherwise, it is recommended that no energy dispersal advantage be assumed for TV/FM interference into an SCPC system. For small diameter antennas, this will almost certainly mean that the centre 1-2 MHz of an RF channel cannot be used if an adjacent satellite is carrying co-channel TV.

3.4.4 Methods of Reducing Adjacent Satellite Interference to and from SCPC Systems.

Due to low carrier bandwidth and power and the desirability for low cost (i.e. small diameter) earth terminals to serve remote areas, SCPC systems are especially sensitive to adjacent satellite interference. Alternatively, SCPC systems tend to make rather inefficient use of the geo-stationary orbit by limiting the minimum separation between satellites. Techniques to reduce their sensi-

tivity to interference are therefore desirable (30).

3.4.4.1 Control of Carrier-to-Interference Ratios

3.4.4.1.1 Increase in Antenna Diameter

The most obvious way to reduce interference into and from adjacent satellites is to increase the diameter and hence the gain of the antenna. However, this is often undesirable for low capacity terminals because of the considerable increase in cost that would result.

3.4.4.1.2 Antenna Sidelobe Control

The amount of interference to and from adjacent satellites can be reduced by careful control of the sidelobe level at the angular spacing between satellites as viewed from the earth terminal. Moreover, this control does not have to be maintained everywhere, but only in and near the transverse plane which includes the adjacent satellites.

These sidelobes are best controlled by proper feed design, illuminated taper and careful attention to blockage by feed components. For antennas with $D/\lambda \geq 40$ it appears that one can achieve production sidelobes that follow a law given by

$$G = 34 - 31 \log_{10}\theta, \quad 2° \leq \theta \leq 50°$$

instead of the more conventional $32 - 25 \log_{10}\theta$ set by the CCIR and FCC. Simple sidelobe suppression techniques could allow an even more stringent control of sidelobe levels (31).

3.4.4.1.3 Increase of Satellite Transponder Gain

Future satellites specifically designed for service with small terminals will have gains on the order of 10 dB or more greater than present. The required up-link EIRP will decrease by roughly the amount that the satellite gain is increased, reducing the interference

into the adjacent satellite by the same amount.

3.4.4.1.4 Polarization Isolation.

Whenever feasible, the use of a polarization orthogonal to that of adjacent satellites can provide substantial isolation; however, this technique requires that the off-axis polarization properties of the antenna be maintained.

3.4.4.1.5 Carrier Spreading

Spreading the energy on FDM/FM carriers will result in a significant reduction in the interference into an SCPC channel.

3.4.4.1.6 Reduction in Down-Link Power Requirements

For a particular satellite transponder gain, any reduction in down-link power requirements will be reflected in an up-link EIRP reduction, and lower interference into the adjacent satellite.

Thus more efficient modulation techniques will produce lower interference levels that less efficient techniques. Reduction in down-link EIRP may also be obtained by an increase in G/T of the receiving station, at some increase in system cost. One case where this technizue is particularly useful is in highly asymetrical networks where essentially all of the traffic originates at one large terminal. In this case the G/T of this terminal may be increased significantly without appreciable economic impact on total system cost.

3.4.4.1.7 Satellite Station Keeping and Earth Station Pointing

Worst-case interference levels are obviously a function of the satellite station keeping and earth station pointing tolerances.

3.4.4.1.8 Interference Cancellation

Interference at a particular earth station is significant at only a single direction of arrival and usually at discrete frequencies. Microwave equipment to provide cancellation of an interfering signal from an adjacent satellite is theoretically possible, and an economical implementation of a cancelling device would be highly desirable. This technique has in fact been implemented at Goonhilly, U.K. to reduce the effects of a bothersome radio relay interferer (32). Unfortunately, the changing amplitude and phase of arriving interference make cancellation a non-trivial problem.

3.4.4.2 Coordination of Channel Assignments

Coordination of frequency plans between adjacent satellites can be one of the most satisfactory means of preventing interference. Most of the benefit arises because an interfering carrier spaced several baseband bandwidths away from the

wanted carrier has much less effect than a co-channel interferer for FM systems.

3.4.4.2.1 SCPC into SCPC

Typically, SCPC systems use carrier spacings of 20 to 45 kHz. For FM SCPC systems interleaving of the carriers in adjacent satellites is very beneficial, giving an additional 15 to 30 dB of improvement over the co-channel interference case.

3.4.4.2.2 SCPC into FDM/FM

An interfering SCPC carrier has the greatest effect on the FDM telephony channel with baseband frequency equal to the carrier spacing. For carrier spacing greater than one baseband width, SCPC carriers have little effect on FDM/FM performance. Therefore, the carriers in the SCPC system can be located outside of one baseband bandwidth away. Although this arrangement will result in a loss in the number of

assignable frequencies, it is often the case in a small terminal systems that channel capacity is limited by power, rather than bandwidth considerations.

4. OTHER SCPC PERFORMANCE MEASURES AND APPLICATIONS

This section considers in a non-analytical way additional SCPC performance measures and possible applications.

4.1 Circuit Characteristics

4.1.1 Frequency Response

Figure 21 plots the amplitude response of FM_{TED} and 40 kbps Δ-mod SCPC circuits over satellite links. The digital system exhibits a noticeable high frequency roll-off, which limits the use of the circuit for high-speed voice-band data, data above voice, music, and other services requiring high frequency fidelity. Also included for comparison is the back-to-back response of a 32 kbps Δ-codec.† CCITT upper and lower amplitude response limits are also indicated.

Figure 22 plots the group delay response of an FM_{TED} system (over a satellite link) and a 32 kbps Δ-codec. For FM this is a function of several transmission

---

† For a digital SCPC system, unlike FM, the baseband frequency response does not depend on the characteristics of the RF channel. The differences observed in Figure 21 are due to codec differences.

K+E SEMI-LOGARITHMIC 46 4973
2 CYCLES X 70 DIVISIONS MADE IN U.S.A.
KEUFFEL & ESSER CO.

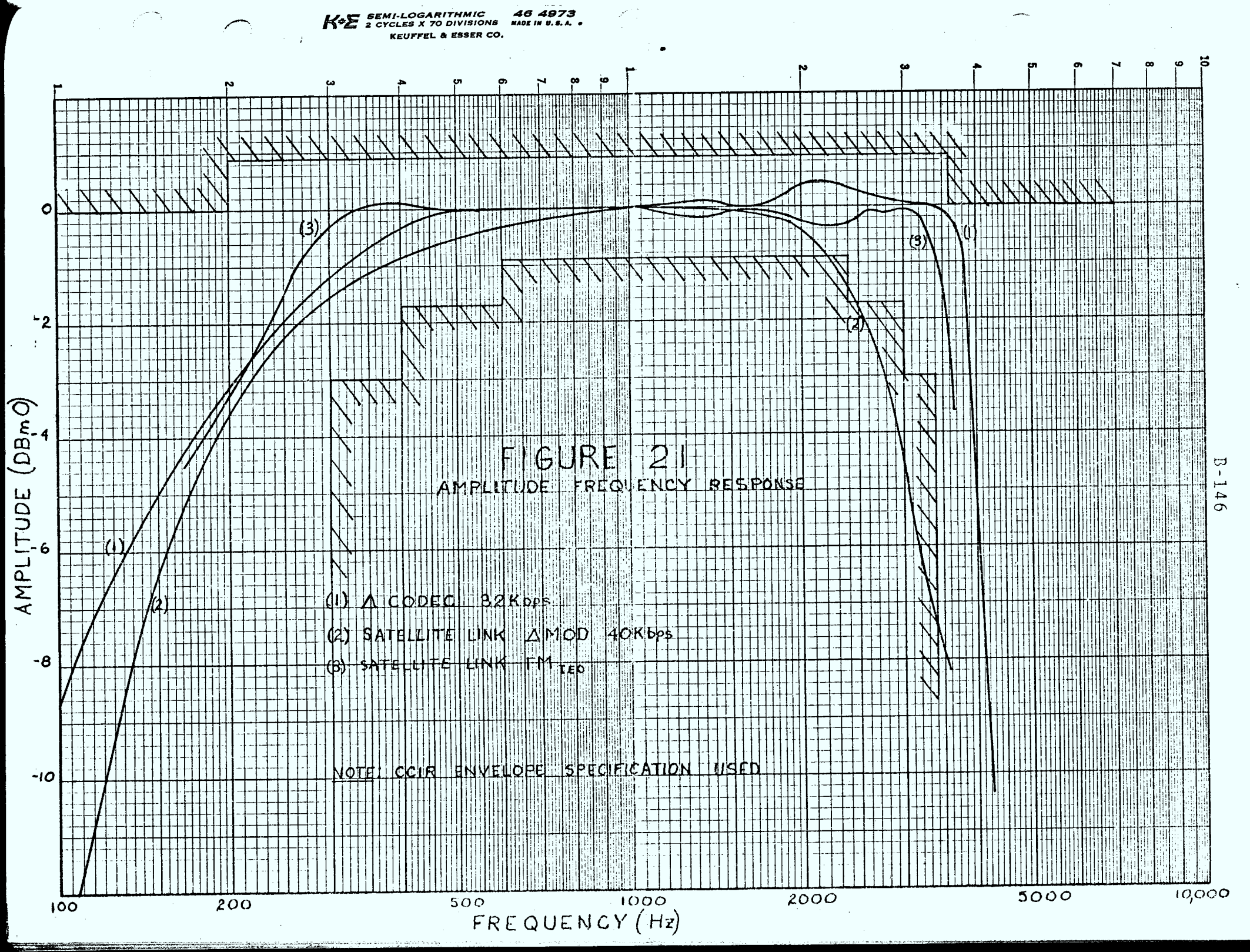
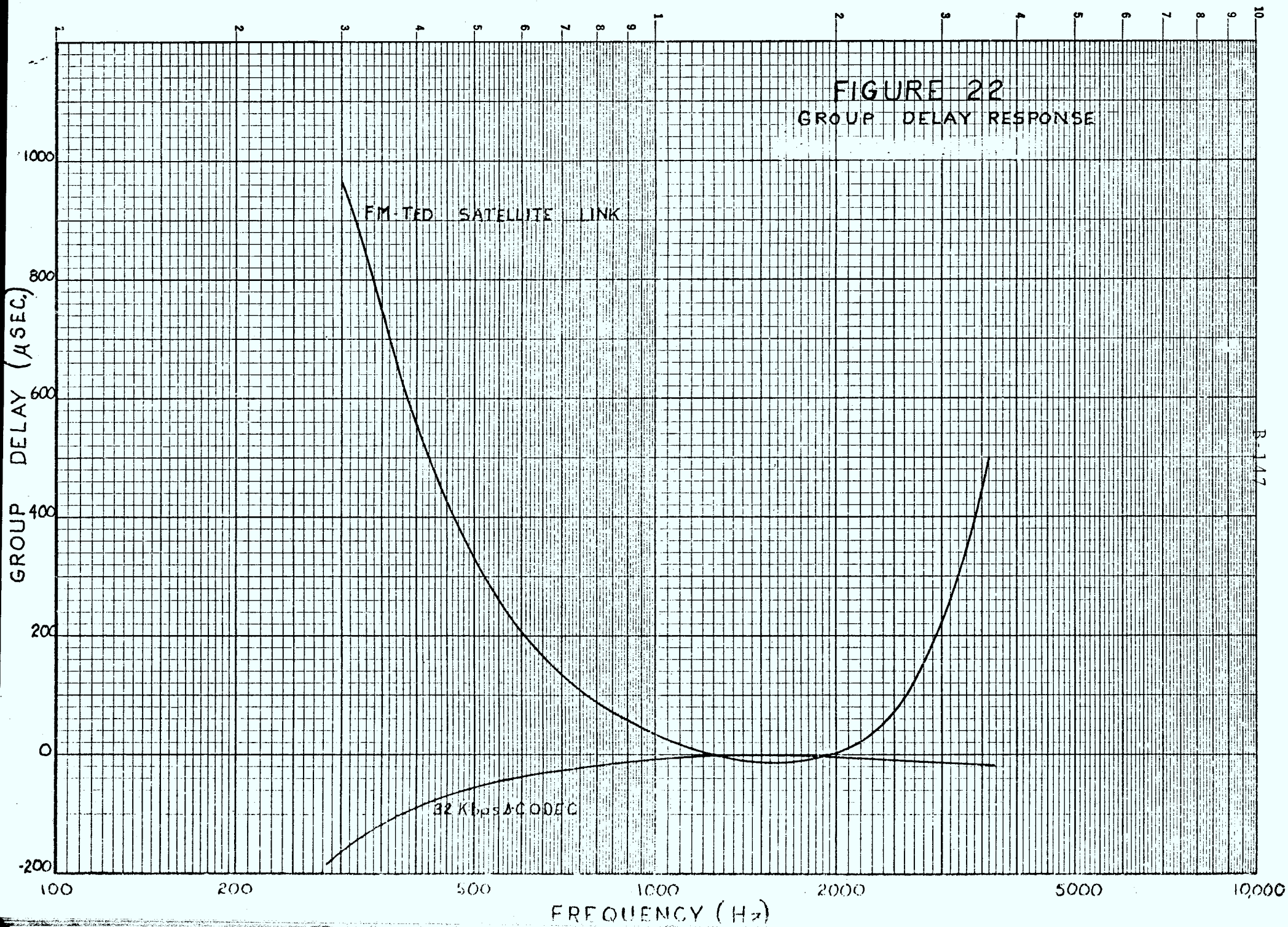


FIGURE 21
AMPLITUDE FREQUENCY RESPONSE

K+E SEMI-LOGARITHMIC 46 4973
2 CYCLES X 70 DIVISIONS MADE IN U.S.A.
KEUFFEL & ESSER CO.



parameters including FM deviation, compandor and emphasis characteristics, and the frequency response of the RF channel. The linear and parabolic components of delay are dominant. The response of both systems plotted easily meet CCITT group delay specifications which permit differences between the minimum group delay (over the frequency band) and the group delay at the lower and upper frequency limits of 15 msec and 7.5 msec respectively.

4.1.2 Signal-to-Quantizing Noise for Δ-mod

In an FM system, the noise which is introduced above 300 Hz is in general thermal (and impulsive when operating above threshold). However, digital systems operating above threshold* are degraded by encoding rather than transmission noise. The signal-to-quantizing noise ratio (S/N_q) for a sinusoid varies with applied input level and frequency. From equation (1) of section 3.1.1.2 it can be seen that the S/N_q degrades with increasing test tone frequency. In general this quantizing noise has two components. Since the

---

* Δ-mod has a threshold advantage over PCM. Unlike Δ-mod, bit errors in a PCM system tend to produce disturbing noise impulses.

output samples may only be discrete values, a granular noise effect occurs. A second type of distortion arises when the applied signal changes level at a rate greater than the maximum average slope of the encoder signal. This is known as slope overload.

For a fixed sampling rate, the encoder step size determines the balance between granular and overload noise. For very large step sizes almost all of the noise will be granular, while decreasing the step size causes slope overload to be the dominant distortion. By varying the step size, i.e, companding, the maximum S/N_q is achieved over a larger dynamic range. This dynamic range is limited at low levels by the minimum coarseness of the quantizer and at high levels by the maximum step size of the quantizer. The typical effect of companding on S/N_q vs input level for a 1 kHz sunusoid is shown in Figure 23. Figure 24 plots S/N_q vs input level for several codecs and sampling rates; the better performance of the 32 kbps codec vs the older 40 kbps codec reflects improved design and fabrication (i.e. operation closer to theoretical.)

K+E 10 X 10 TO ½ INCH 7 X 10 INCHES
KEUFFEL & ESSER CO. MADE IN U.S.A.
46 1327

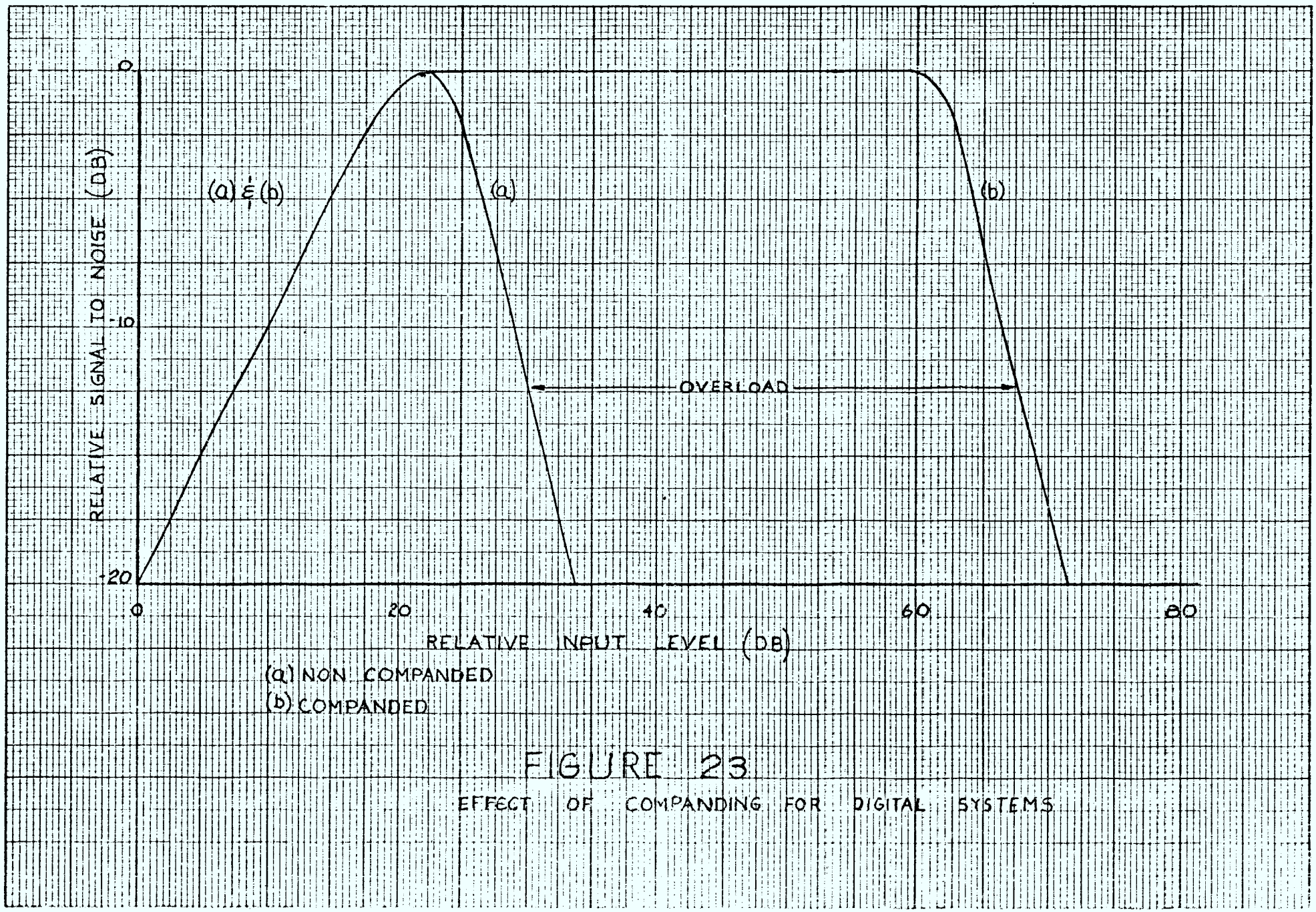


FIGURE 23

EFFECT OF COMPANDING FOR DIGITAL SYSTEMS

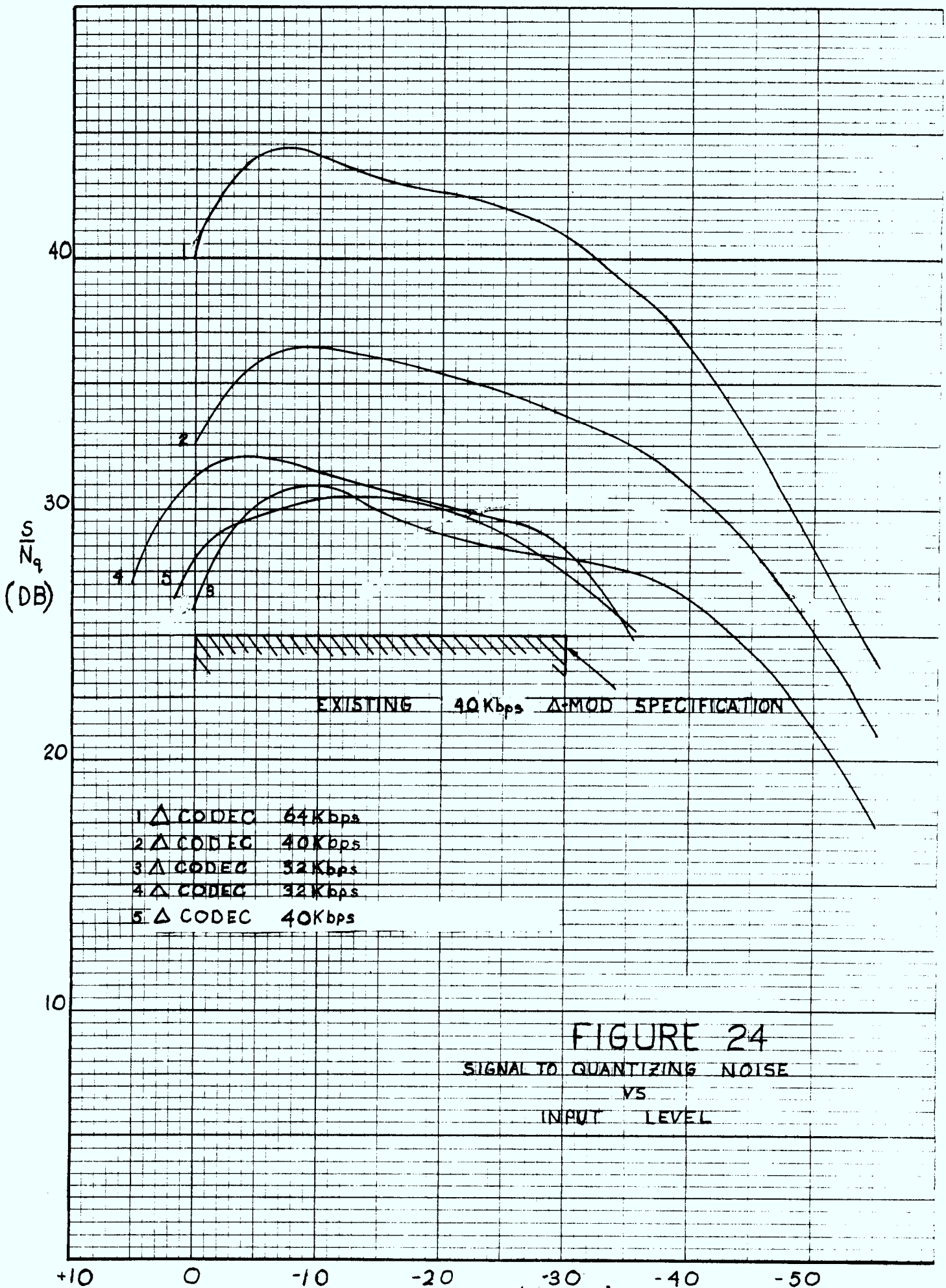


FIGURE 24

SIGNAL TO QUANTIZING NOISE VS INPUT LEVEL

46 0782
10 X 10 TO THE INCH • 7 X 10 INCHES
KEUFFEL & ESSER CO. MADE IN U.S.A.

4.2 Voice Band Data

4.2.1 Voice Plus Data

For more efficient utulization of the telephone channel, teletype carrier(s) may be included above the expected speech spectrum. With appropriate modification of deviation and C/N_o if necessary*, an FM SCPC link can accommodate this arrangement. In general an FM SCPC system can be designed to give the degree of channel transparancy required to transmit any combination of input signals over (and above) the voice band.

Such is not the case with a Δ-mod system, in which performance is limited by the encoding process rather than transmission. Even at a zero bit error rate, the use of a 40 kbps Δ-mod circuit for voice plus data has not been proven. With normal voice and a 3.094 kHz-110 baud input levels into a companded 40 kbps codec optimized for voice, speech quality is adequate but cross-talk during speech peaks may cause bit errors in the teletype channel.

---

* Lowering the deviation improves the linearity and frequency response of the voice channel at the expense of reduced FM noise advantage. This can be compensated for by an increase in carrier power.

4.2.2 Data Only

Voice channels can be used to transmit data at standard line rates of up to 9600 bps. As was the case for voice plus data, FM offers somewhat greater flexibility for such applications. Voice band data generally falls into two broad categories - multi-carrier slow speed (e.g., 24 channel telegraphy) or single carrier high speed (2400 - 9600 bps).

4.2.2.1 Multicarrier Telegraphy

In the multicarrier case, some disabling must be done to the FM system. Since the compressor would bring the instantaneous peaks of such a multichannel telegraph system into clipping level, the compander must be disabled (generally a switch is provided). Further, the pre-emphasis network would tend to cause overload for a full 24 channel input (with each carrier at -20 dBmO) so the individual VFCT signals must be reduced to approximately -26 dBmO. Operation under these conditions has sufficient margin to ensure acceptable telegraph error rate performance.

Again, a fully loaded multi-carrier data signal would probably undergo unacceptable intermodualtion in a Δ-mod circuit. The typical maximum capacity of a 32 kbps Δ-mod system is approximately 10 carriers.

4.2.2.2 High Speed Voice Band Data

Higher speed voice band data applications can be accommodated with the FM SCPC channel. For applications requiring dedicated private line access, the RF level may be adjusted to obtain a required bit error rate. For alternate voice/data operation a data transmission detector must be available to disable the echo suppressor function during the period of data transmission. However, as noted earlier in section 2.3.1.4, the constant nature of such data streams will "lock" the compandor gain. Thus, since the data is normally sent at a low level, the compandor action is both beneficial and necessary to the performance of the data signal.

Testing done with 4800 and 9600 bps data indicates acceptable performance with compandored

emphasized FM. The small difference in BER vs C/N_o performance at the two speeds indicates that even above normal FM threshold the dominant source of errors is click noise. The evaluations considered the effect of pn sequence word length - 63 bits (Bell/ATT standard), 511 bits (CCITT standard) and 2047 bits. In general, as the word length was reduced the bit error rate decreased.* The results are presented in the following Table and graphically in Figure 25. Also included for comparison in Figure 25 is a plot of 2400 bps BER vs C/N_o performance of a 40 kbps Δ-mod/2 phase CPSK satellite link.

An evaluation of Δ-mod as regards to data handling capability has been made in a laboratory (33). Data error rates under error free transmission conditions (i.e. through Δ-codec only) were examined as a function of data rate, Δ-encoding rate, and input level. Some results are plotted in Figure 26.

---

* The explanation for this phenomenon was not given. It may lie in the fact that longer sequences can produce greater "instantaneous" bandwidths. The effects of filter distortion and FM truncation noise appearing at the output of a threshold extension demodulator during specific "bad" sub-sequences within a long pn sequence can increase the susceptibility to bit errors and result in a higher average BER.

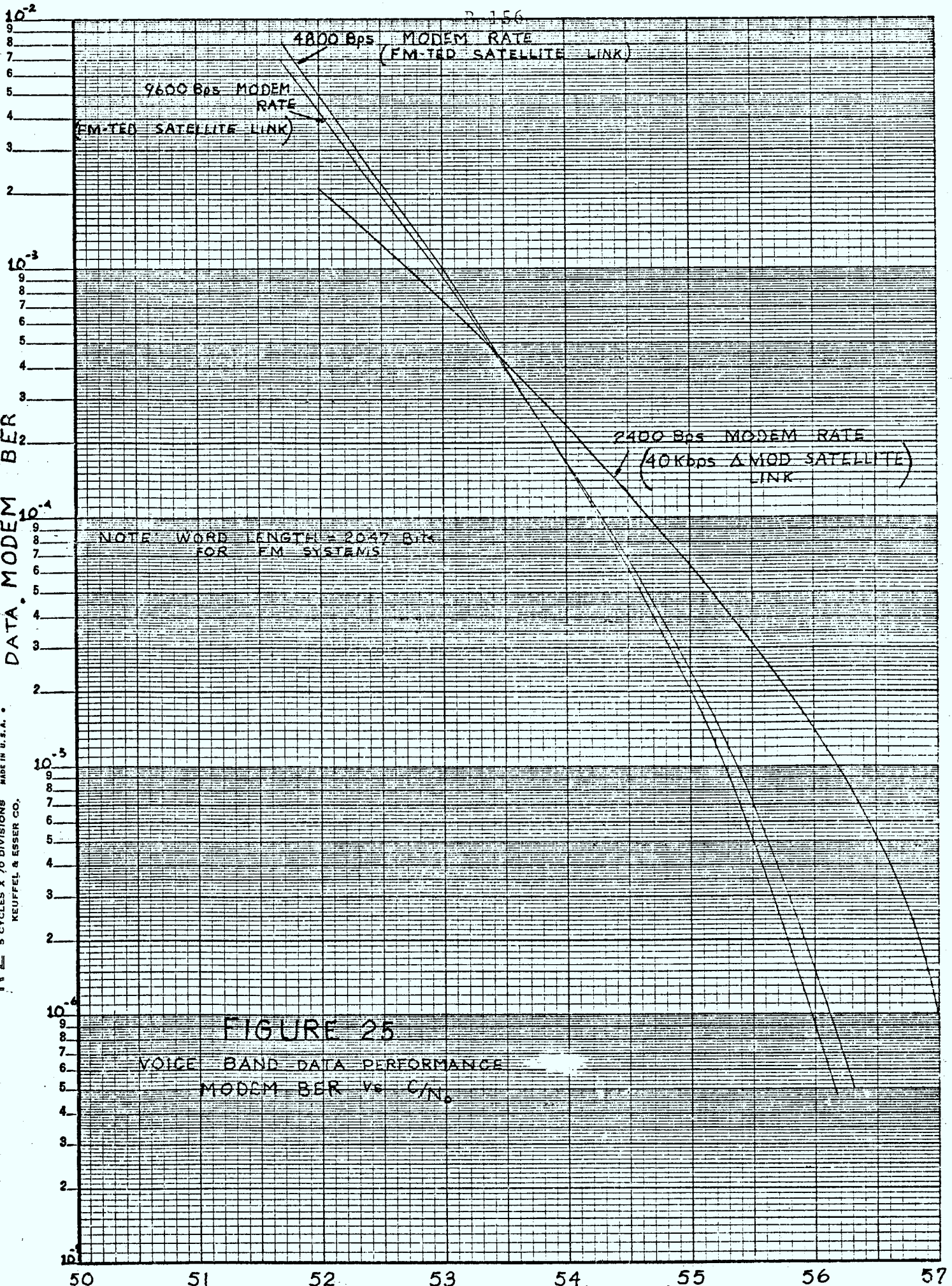


FIGURE 25

VOICE BAND DATA PERFORMANCE

MODEM BER Vs C/N_o

K&E SEMI-LOGARITHMIC 46 6213
5 CYCLES X 70 DIVISIONS MADE IN U.S.A.
KEUFFEL & ESSER CO.

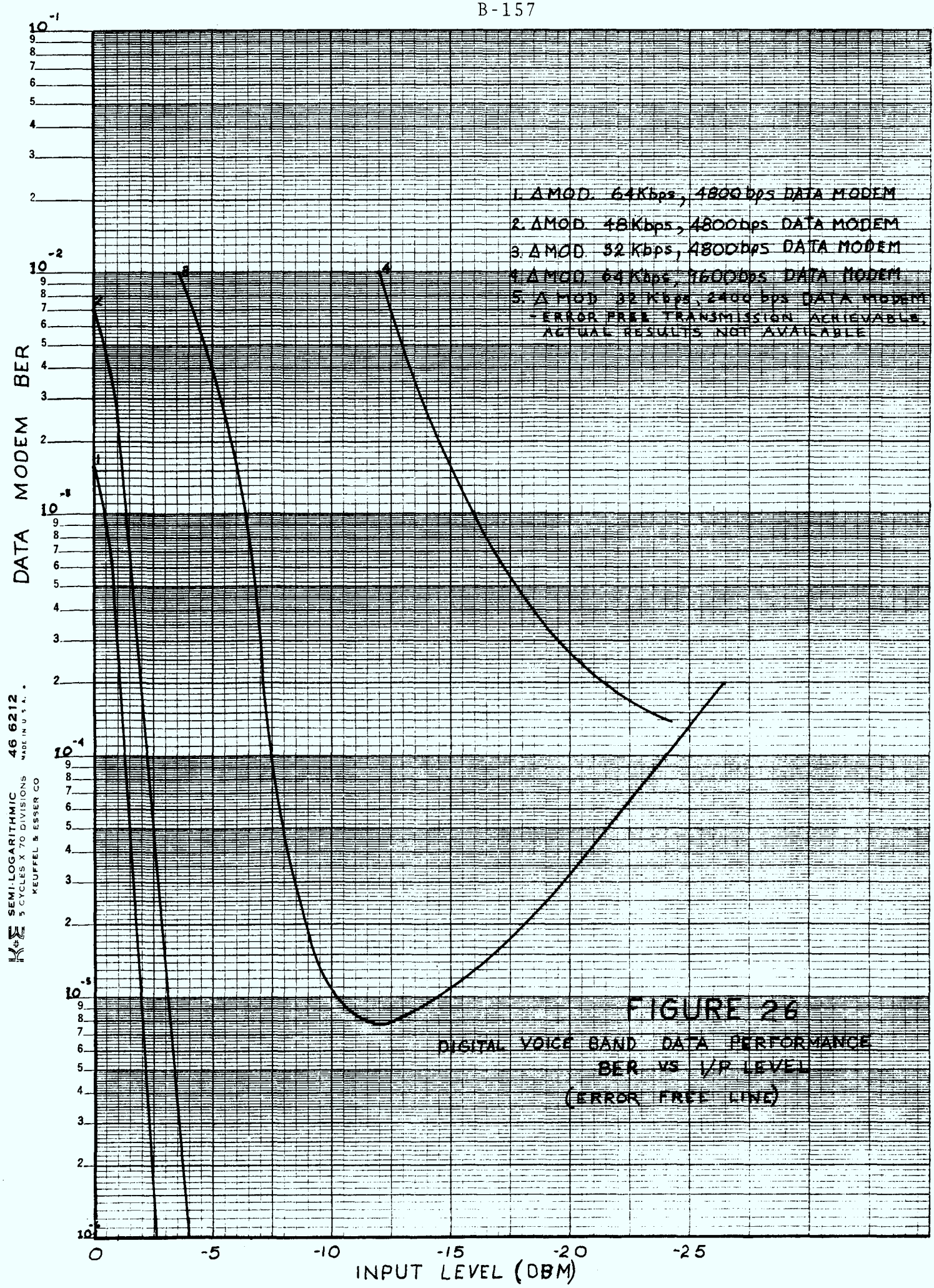


FIGURE 26

DIGITAL VOICE BAND DATA PERFORMANCE
BER VS I/P LEVEL
(ERROR FREE LINE)

Then choosing an input level which yielded a BER ≃ 0 for the 64 kbps system, as established in the error free transmission line tests, random errors were inserted digitally between the Δ-encoder and decoder and data error rate measured as a function of transmission error rate. For comparison purposes, a PCM channel of the same sampling rate and employing the same data modem, was evaluated. The results are shown in Figure 27.

The examination indicates that for an error free condition the Δ-mod channel can transmit successfully 2400 bps at a sampling frequency of 32 kbps; 4800 bps at a sampling frequency of 48 kbps, and that 9600 bps requires a sampling rate in excess of 64 kbps. More generally the data modem BER performance is governed by the sampling rate and the input signal level. Attempting to transmit voice frequency data at too low an encoding rate results in waveform distortion sufficient to cause the occurrence of bit errors. Transmitting at too low or too high an input level produces the same efect.

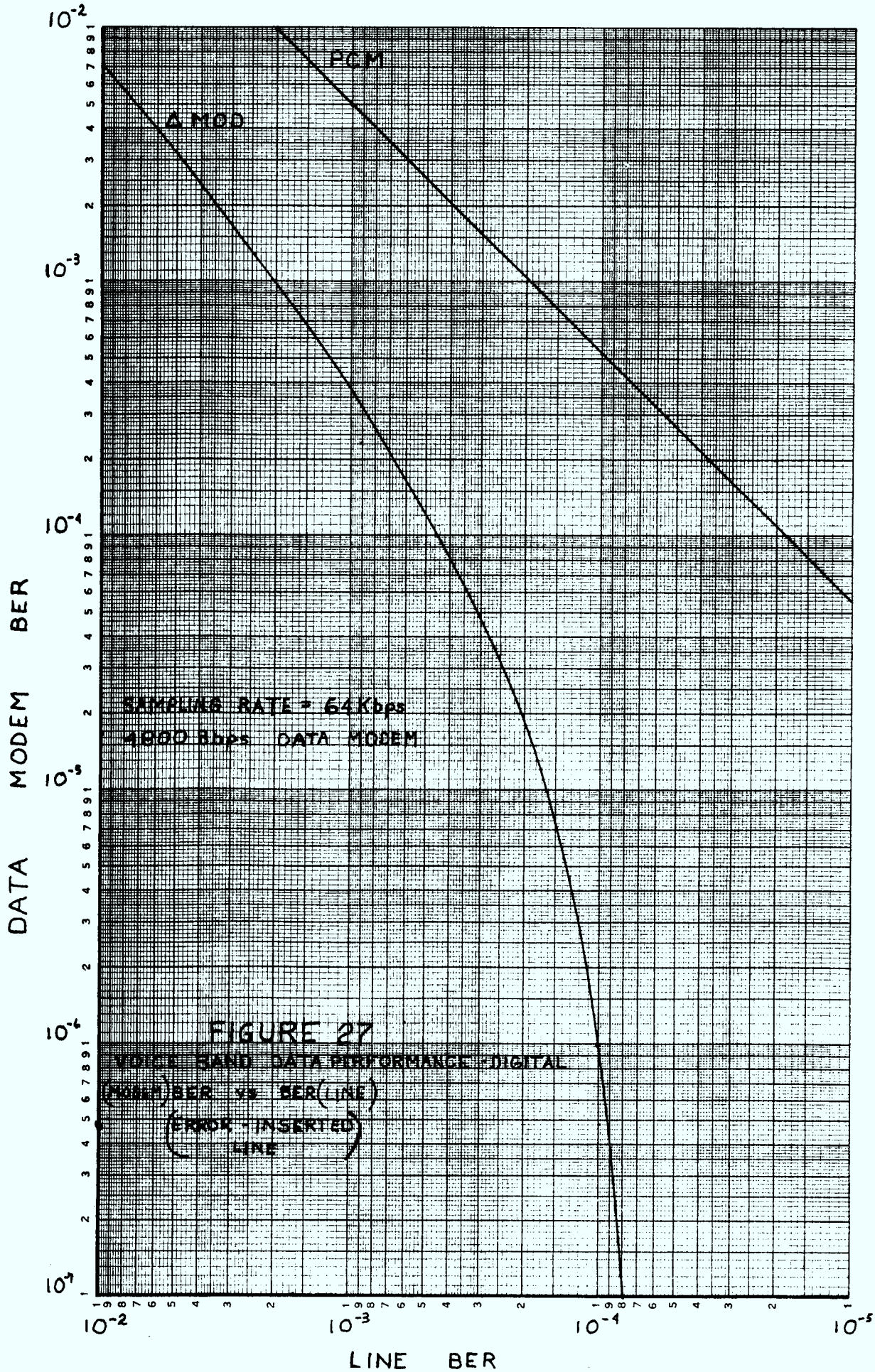


FIGURE 27
VOICE BAND DATA PERFORMANCE - DIGITAL
(MODEM) BER vs BER (LINE)
(ERROR - INSERTED LINE)

K+E LOGARITHMIC 46 7522
3 X 5 CYCLES MADE IN U.S.A.
KEUFFEL & ESSER CO.

In the case where transmission line errors are simulated, the results indicate that for a Δ-mod channel, the modem BER tends exponentially to zero as the line BER decreases. In fact, for line BER's < 10^{-5}, the modem BER is negligibly small. PCM, on the other hand produced modem BER's which were consistently worse than Δ-mod and linearly related to the line BER.

4.3 Radio Program Applications

A further potential use of the SCPC system is the broadcasting of radio program signals.

The FM approach is readily adaptable to the higher quality requirements of the standard 15 kHz voice/music transmission channel. By appropriately modifying the baseband processing hardware and adjusting the deviation and C/N_o, radio program standards are achieved in a straight forward manner. A 13 dB improvement is yielded by using a standard 75 μsec broadcast emphasis network and a further 10 dB from wideband program compandors. Thus, this single channel carrier is similar to that used to carry a telephone voice circuit, possible exceptions being the absence of voice activation, echo suppression and squelch circuitry.

An example of a wide-band (200 kHz) SCPC radio program system is the Telesat Northern Radio Network. In this case the radio program carrier is placed in the same 36 MHz RF channel as the voice signal. The carrier is displaced 16 MHz below the centre frequency of the transponder and approximately 20 dB below the video carrier at the satellite output. The audio S/N achieved with companding is about 53 dB.

The frequency response achievable with current 32 kbps and 40 kbps Δ-mod limits the achievable radio program signal quality. Nonetheless limited tests have indicated that at an error rate of 10^{-4} subjectively acceptable AM radio quality is achievable (for voice and music).

REFERENCES TO APPENDIX B

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