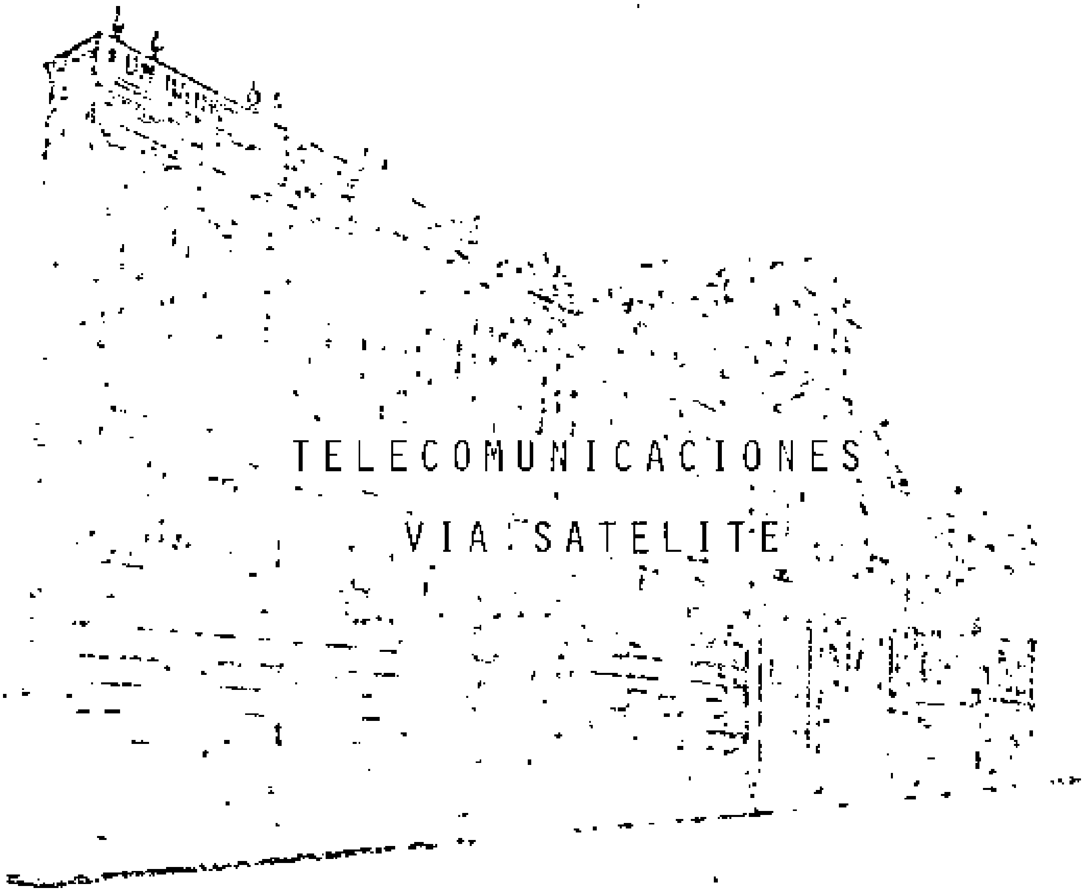
DIVISION DE EDUCACION CONTINUA
FACULTAD DE INGENIERIA U.N.A.M.



TELECOMUNICACIONES VIA SATELITE

SEPTIEMBRE, 1983

HISTORIA

Los orígenes de la idea de satélites para comunicaciones no están muy claros, sin embargo, no hay duda de que la idea de satélite geoestacionario fué propuesta por primera vez por Arthur C. Clarke en un artículo titulado Estaciones Extraterrestres, en la revista "Wireless World". Para esta idea, él tomó comb referencia los cohetes utilizados por los alemanes durante la Guerra y la gran ventaja de la órbita geoestacionaria. Proféticamente, su propuesta fué para usar estos satélites en cadenas de FM y no para servicio telefónico. Y por si esto fuera poco, Clarke también visualizó el uso en el espacio, de potencia eléctrica generada por paneles de celdas solares. La implementación de su idea tenía que esperar todavía hasta la era espacial (Sputnik - 1957) y la tecnología del estado sólido.

Han pasado 31 años desde su profecía y ya existen 22 programas de satélites de comunicaciones ya sea con satélites en órbita ó bajo construcción.

LAS DIFERENTES ETAPAS.

Al final de los 40's y principios de los 50's se demostraron reflexiones en la luna para su aplicación a radar y a sistemas de comunicaciones, Fig. 1. En julio de 1954, los primeros mensajes de voz fueron transmitidos por la Marina de los Estados Unidos sobre la trayectoria de lá tierra a la luna y viceversa. En 1956, se estableció un enlace haciendo uso de la luna, entre Washington D.C. y Hawaii. Este circuito operó hasta 1962, ofreciendo cumunicación segura a larga distancia, teniendo como única limitación la disponibilidad de la luna en los sitios de transmisión y de

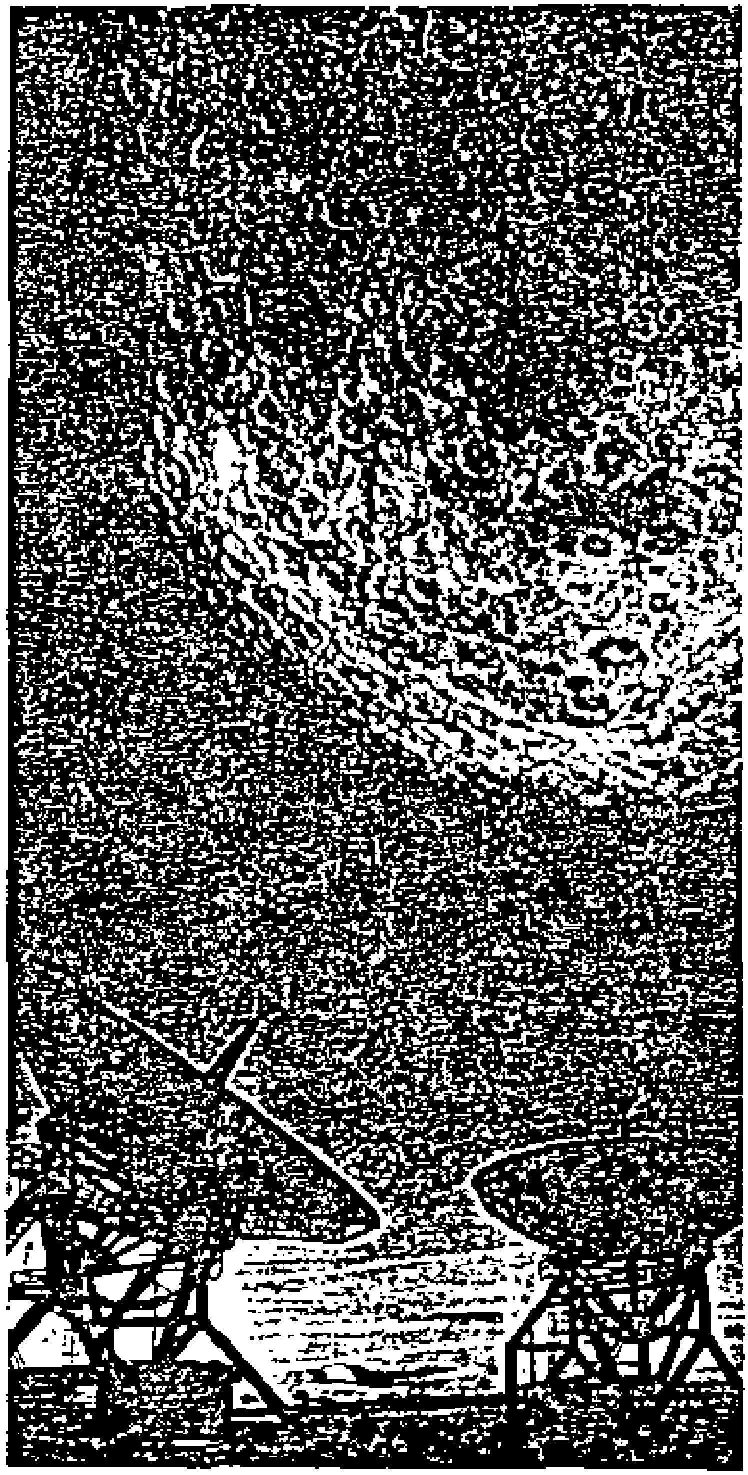


Fig. 1. Reflexiones en la luna.

recepción. La potencia usada fué de 100 kw, con antenas de 26-m de diámetro a una frecuencia de 430 MHz.

En 1958 surge el proyecto SCORE, el cual consistía en saté-lites del tipo grabación y retransmisión con un peso de 150 libras y a una órbita entre 110 y 920 millas, Fig. 2.

Dos años después, conjuntamente con los laboratorios Bell, NASA y JPL, se realizó el experimento ECHO, Fig. 3. Exitosas comunicaciones se establecieron a la largo de los Estados Unidos, primeramente entre Goldstone, CA, y Holmdel, NJ, a frecuencias de 960 MHz y 2290 MHz. El balón ECHO, hecho de plástico y cubierto de aluminio con un diámetro de 100 pies, estaba a una órbita inclinada de 1500 kms. de altitud y era visible al ojo humano, el ECHO II se instaló entre 1000 y 1200 kms. Más tarde y en el mismo mes, ocurrió la primera transmisión transatlántica entre Holmdel, N.J. y una estación receptora en Francia. Este proyecto alertó a todo el mundo sobre la prosperidad del nuevo medio de comunicación, aunque el método específico, nunca fué explotado comercialmente.

Aunque los satélites pasivos tienen capacidad infinita para comunicaciones de acceso múltiple, existe la inconveniencia del ineficiente uso de la potencia de transmisión. En el experimento ECHO, por ejemplo, solamente una parte de la potencia transmitida (10 Kw) era reflejada a la tierra ($\frac{1}{10^8}$).

El primer satélite repetidor activo fué el Courier (1960), Fig. 4. Aceptaba y almacenaba hasta 360,000 palabras de teletipo. Operó por 17 días con 3 watts de potencia de salida y trabajando a una órbita entre 600 y 700 millas.

De los años experimentales, tal vez, el proyecto más conocido es el Telstar, Fig. 5, posiblemente porque fué el primero capaz de retransmitir a través del Atlántico programas de T.V. Este proyecto fué iniciado por la ATT y desarrollado por los laboratorios Bell, quienes habían adquirido considerable experiencia y conocimiento de los trabajos anteriores, como el proyecto ECHO. El primer Telstar fué lanzado de Cabo Cañaveral el 10 de julio de 1962. Era una esfera de aproximadamente 87 cm. de diámetro, con un peso de 80 kg. El vehículo de lanzamiento era un cohete Thor-Delta el cual colocó al satélite en una órbita elíptica con apogeo de 5600 km, y un período de 2 1/2 horas. Telstar II fué hecho con más resistencia a la radiación por la experiencia con el Telstar I, de lo demás fué exactamente igual al anterior. Fué lanzado exitosamente el 7 de mayo de 1963.

La potencia de Telstar I y II de 2,25 w fué suministrada por un TWT con un ancho de banda de radiofrecuencia de 50 MHz en las bandas de 4 y 6 GHz. Ambos satélites fueron estabilizados por giro. La capacidad total fué de 600 canales telefónicos ó un canal de T.V. Estos satélites estaban a una órbita de 682 a 4030 millas.

Por los mismos años (1963), RCA y NASA orbitaron el satélite RELAY (Fig. 6) con frecuencias de operación de 1.7 y 4.2 GHz, con 10 watts de salida y órbitas de 942 y 5303 millas.

En 1963, la fuerza aérea de los Estados Unidos logró poner en órbita un cinturón orbital compuesto de pequeños dipolos a 2300 millas, el cual actuaba como un reflector pasivo, se transmitió voz en forma digital de una forma inteligible. Este proyecto fué el famoso W EST FORD, Fig. 7.

En este mismo año se lanzó el primer satélite de comunicaciones en órbita geoestacionaria. Este satélite fué puesto en órbita por la NASA y se utilizó para múltiples experimentos.

Transmitió señales de TV en los juegos olímpicos de Tokio en 1964, fué el SYNCOM, Fig. 8.

Las comunicaciones comerciales por satélite comenzaron oficialmente en 1965, cuando se lanzó el primer satélite comercial en el mundo, fué el INTELSAT I (Pájaro Madrugador), - Fig. 10.

En el mismo año la Unión Soviética pone en órbita el Molniya, que fué el primero de muchos satélites de comunicaciones, puesto a gran altitud con órbita elíptica, Fig. 9.

En enero de 1966 INTELSAT I fué puesto fuera de servicio cuando la cobertura en el Atlántico y en el Pacífico fué lograda por INTELSAT II e INTELSAT III, Figs.11 y 13.

El LES-6, un pequeño satélite de banda lateral en la banda de UHF y el TACSAT I, un poderoso satélite de UHF y SHF formaron el programa TACSATCOM para operaciones militares en los Estados Unidos a lo largo del mundo, esto fué en 1968 y 1969. Un satélite TACSAT tenía 1000 watts de potencia y transmitía 10,000 canales de voz. Fig. 12.

La fase de madurez total en los satélites de comunicaciones probablemente arribó con la llegada del INTELSAT IV en 1971. Estas naves del espacio pesan aproximadamente 730 Kg. (Fig.14) en órbita y proveen no solamente cobertura de la tierra sino también dos rayos dirigidos a un punto específico de Europa, Norte ó Sudamérica. INTELSAT IV es un satélite de giro, como sus predecesores, pero con todo un ensamble de antenas, consistente de 13 diferentes antenas, se ajusta continuamente hacia un punto de la tierra. Los dos rayos dirigidos se forman por dos antenas parabólicas. Cada satélite provee aproximadamente 6000 circuitos de voz, o más, dependiendo de cómo se divida la potencia en el satélite entre los rayos dirigidos y los de cobertura terrestre. El sistema INTELSAT IV puede conducir 12 canales de color de TV al mismo tiempo.

En 1972, Telstar de Canadá pone en órbita el primer satélite doméstico en el mundo. Este satélite es el famoso ANIK, - - Fig. 15, con capacidad de 5000 canales de voz y 300 watts de potencia.

Estados Unidos lanza su primer satélite doméstico en 1974, el Westar (Fig. 16), el cual inicia una nueva era en las comunicaciones de ese país.

El sistema INTELSAT IV fué puesto en órbita en 1980 en la región del Oceano Atlántico con una capacidad de casi 25000 canales.

1958: SCORE (NASA)

Fig. 2 Un satélite de 150 lbs. que radia un mensaje de navidad grabado del Presidente Eisenhower. Altura de la órbita: de 110 a 920 millas.

1960 ECHO (NASA)

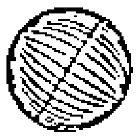


Fig. 3 Un balón de plástico de 100 pies de diámetro con revestimiento de aluminio el cual refleja pasivamente las señales de radio desde una inmensa antena terrestre. Altura de la órbita: 1000 millas; ECHO II: de 600 a 800 millas.

1960 COURIER
(Department of Defense)

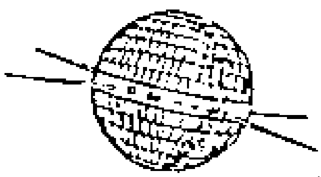


Fig. 4 El primer satélite repetidor activo. Aceptaba y almacenaba hasta 360,000 palabras de teletipo. Operó por 17 días con 3 watts de potencia de salida. Altura de órbita: de 600 a 700 millas.

1962 TELSTAR (AT&T)

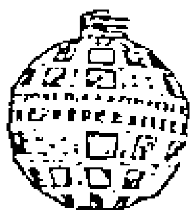


Fig. 5 El primer satélite para recibir y -- transmitir simultáneamente. 4/6 GHz. Utilizado para telefonía, televisión, facsímil y datos. 3 watts de potencia de salida. Altura de la órbita: de 682 a 4030 millas.

1962: RELAY (RCA and NASA)

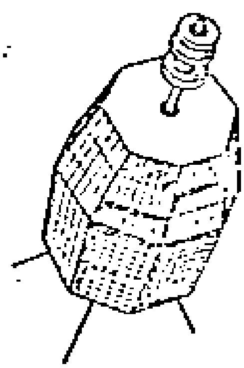


Fig. 6 Satélite de 4.2/1.7GHz, con potencia de salida de 10 watts. Altura de la órbita: de 942 a 5303 millas

(U.S. Air Force)
1963: PROJECT WEST FORD

Fig. 7 Un cinturón orbital de pequeños dipolos fué lanzado a una altura de 2300 millas para actuar como un reflector pasivo. La voz en forma digitalizada se transmitió inteligiblemente.

1963: SYNCOM (NASA)

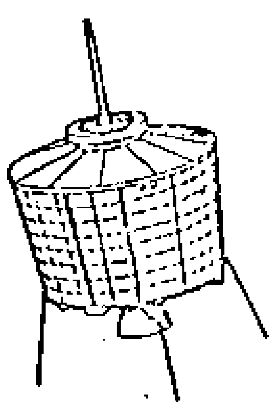


Fig. 8 El primer satélite de comunicaciones en órbita geoestacionaria. Utilizado para muchos experimentos. Transmitió televisión en los Juegos Olímpicos de Tokio en 1964.

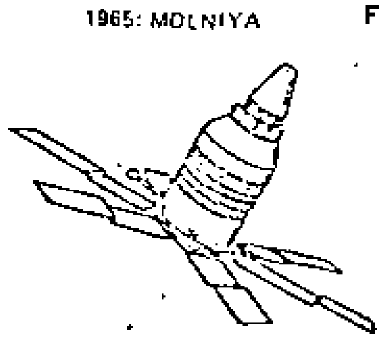


Fig. 9 El primero de muchos satélites de comunicaciones rusos a una órbita elíptica de gran altitud.

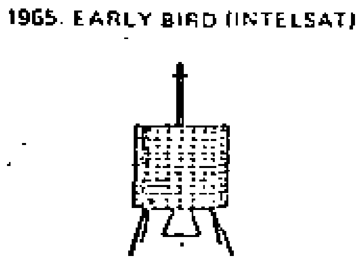


Fig. 10 El primer satélite de comunicaciones comerciales en el mundo, operado por COMSAT. 240 canales de voz. 40 watts de potencia de salida.

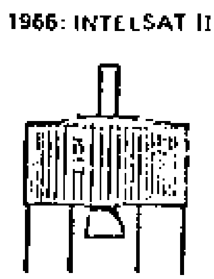


Fig. 11 El segundo satélite de COMSAT. El primer satélite comercial de acceso múltiple con capacidad de multidestino. 240 circuitos de voz. 75 watts de potencia de salida.

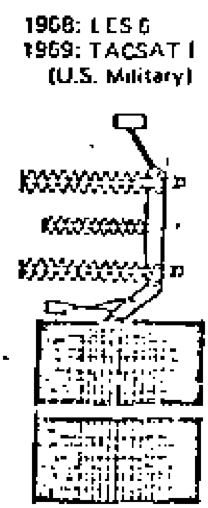


Fig. 12 El LES-6, es un pequeño satélite de UHF de banda única y el TACSAT I es un poderoso satélite en las bandas de UHF y SHF. Ambos formaron el programa TACSATCOM para múltiples operaciones militares en los Estados Unidos. Un satélite TACSAT tenía 1000 watts de potencia y transmitía 10,000 canales de voz.

1968: INTELSAT III

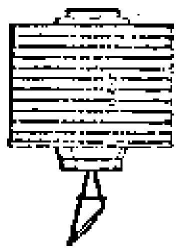


Fig. 13 La tercera generación COMSAT. 1200 circuitos de voz. Una antena direccional de cobertura terrestre. 120 watts de potencia de salida.

1971: INTELSAT IV

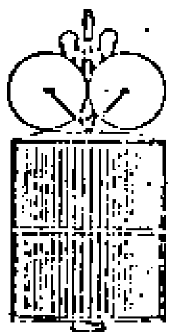


Fig. 14 La cuarta generación COMSAT. 6000 circuitos de voz. Una antena de cobertura terrestre y dos de haces dirigidos. 400 watts de potencia de salida.

1972: ANIK (Telesat Canada)

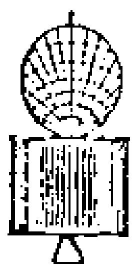


Fig. 15 El primer satélite doméstico en el mundo diseñado por Canadá. 5000 circuitos de voz. 300 watts de potencia.

1974: WESTAR (Western Union)

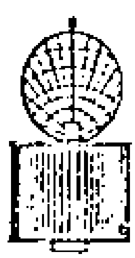


Fig. 16 El primer satélite doméstico en los Estados Unidos. El inicio de una nueva era en las comunicaciones Norte-Americanas.

LAS ERAS DE LOS SATELITES DE COMUNICACIONES

ERA SUBSINCRONA	1958 - 1963
ERA SINCRONA GLOBAL	1964 - 1972
ERA GLOBAL Y DOMESTICA REGIONAL	1973 - 1981
ERA DE NEGOCIOS Y DE ESTACIONES TERMINALES PEQUEÑAS	1981 - 1935
ERA DE SATELITES DE RADIODIFUSION	1935 - 1990
ERA DE PLATAFORMAS ESPACIALES Y SATELITES INTELIGENTES-PROCESADORES EN EL CIELO	1990 ——

ÓRBITAS

Los satélites de comunicaciones modernos tienen órbitas muy diferentes de sus predecesores en el ámbito experimental, tales como Telstar de ATT y Relay de RCA. Estos últimos viajaban rápidamente alrededor de la tierra a una relativamente baja altura. Los satélites Telstar tenían órbitas elípticas muy altas, Telstar I de 600 a 3800 millas y Telstar II de 600 a 6200 millas. El apogeo de la elipse fué puesto en posición tal que el satélite estaba dentro de la línea de vista de ciertas estaciones tanto tiempo como fuera posible. Como con los primeros vuelos espaciales tripulados y muchos otros satélites lanzados en la primera década de los vuelos espaciales, los satélites viajaban alrededor de la tierra en pocas horas: Telstar I, 2 horas y 38 minutos, y Telstar II, 3 horas y 45 minutos. Aquí aparece entonces la desventaja para las telecomunicaciones; estaban dentro de la línea de vista de la estación de rastreo, por sólo un breve período de tiempo, algunas veces menos de media hora.

Los rusos también usaron órbitas elípticas para sus satélites de comunicación Molniya, pero sus órbitas son mayores tal que los satélites están dentro de línea de vista por largos períodos de tiempo. La Fig. 17 grafica el tiempo que toma un satélite para viajar alrededor de la Tierra contra su altura. La órbita a una altura de 22300 millas es especial en la que un satélite a esa altura toma exactamente 24 horas para viajar alrededor de la Tierra (el tiempo de rotación de la Tierra). Si su órbita está sobre el Ecuador y lleva la misma dirección que la superficie de la Tierra, entonces se puede ver como estacionario desde un punto sobre la Tierra. Esta órbita es llamada órbita geoestacionaria.

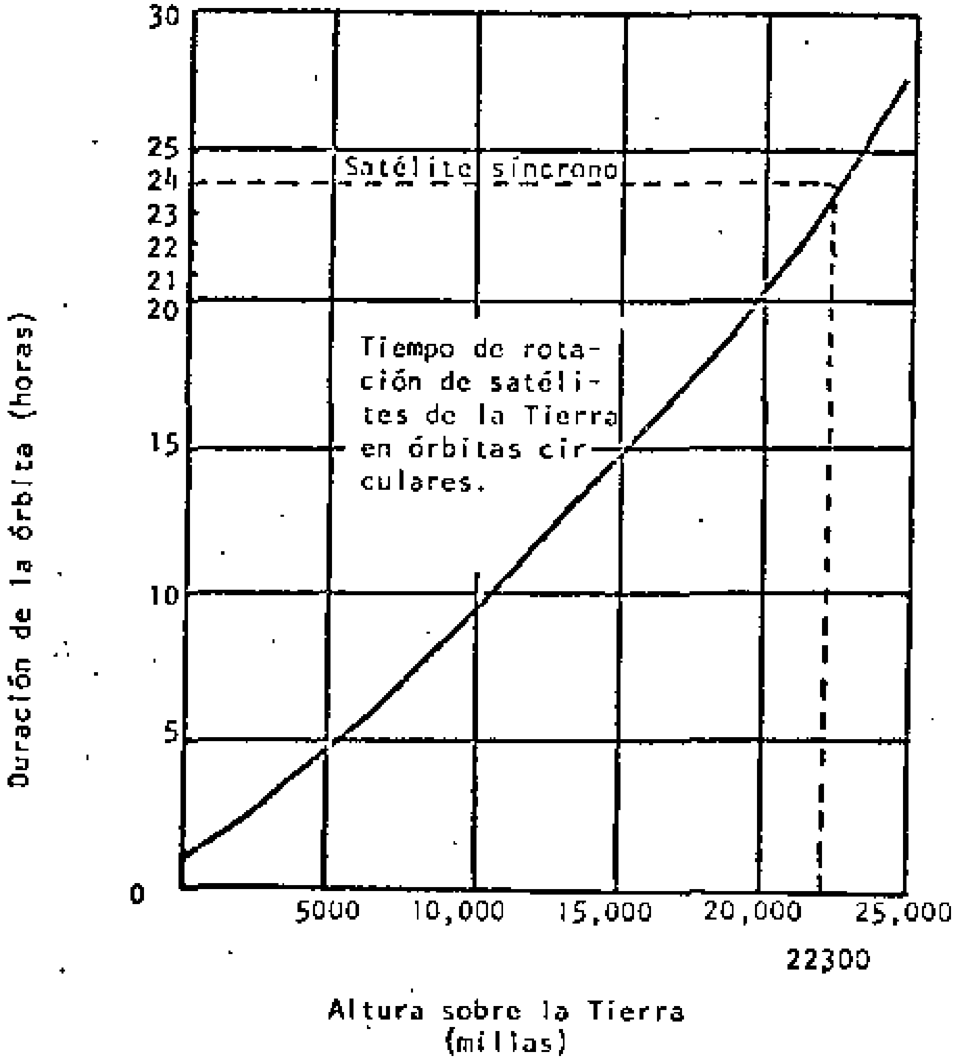

30
25
24
23
22
21
20
15
10
5
0
Satélite síncrono
Tiempo de rotación de satélites de la Tierra en órbitas circulares.
Duración de la órbita (horas)
5000
10,000
15,000
20,000
25,000
22300
Altura sobre la Tierra
(millas)

Fig. 17

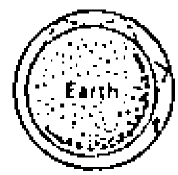
Los satélites INTELSAT son estacionarios sobre el Atlántico y el Pacífico. Los satélites domésticos de E.U. se encuentran también estacionarios en Sudamérica o el Pacífico, sobre el Ecuador. La Fig. 18 muestra las órbitas de los satélites.

El área de vista de un satélite en esa órbita es aproximadamente un tercio del globo.

Para comunicaciones intercontinentales, los satélites se colocan en órbita geoestacionaria sobre cada una de las tres áreas de los oceanos (Fig. 19). Algunas órbitas típicas se detallan también en la Fig. 20.

La Fig. 21 muestra el máximo espaciamiento entre estaciones terrenas para diferentes alturas de satélites, considerarndo que 5° es el ángulo mínimo de elevación de las estaciones terrenas. En la Fig. 22 se muestran las distancias y tiempos de propagación de un satélite geoestacionario.

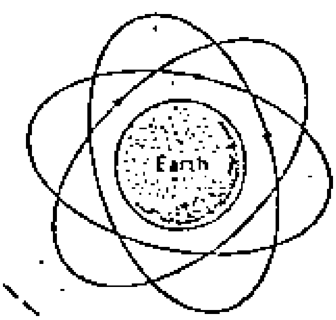
1



Satélite de órbita baja

Altura : 100-300 millas
Período de rotación: 1 1/2 horas aprox.
Tiempo en línea de vista: $\frac{1}{4}$ hr. ó menos

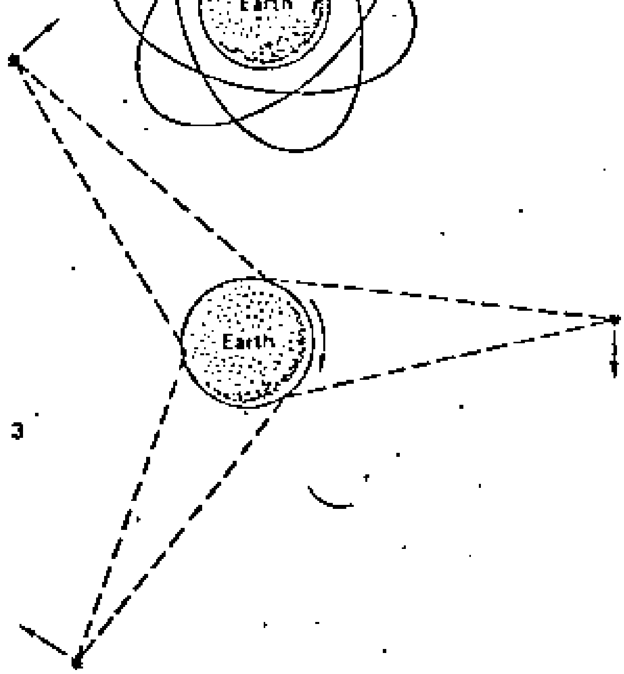
2



Satélite de altitud media

Por ejemplo, el satélite ruso de comunicaciones Molniya y los satélites AT y T's.
Altura típica: 6000-12000 millas
Período de rotación típico: 5-12 hrs.
Tiempo típico en línea de vista 2-4 hrs

3



Satélite Geoestacionario

Por ejemplo, INTELSAT, WESTAR

Altura: 22300 millas
Período de rotación: 24 hrs.
Tiempo en línea de vista:
Toda la vida del satélite la órbita es sobre el Ecuador.

Fig. 18 Orbitas del satélite.

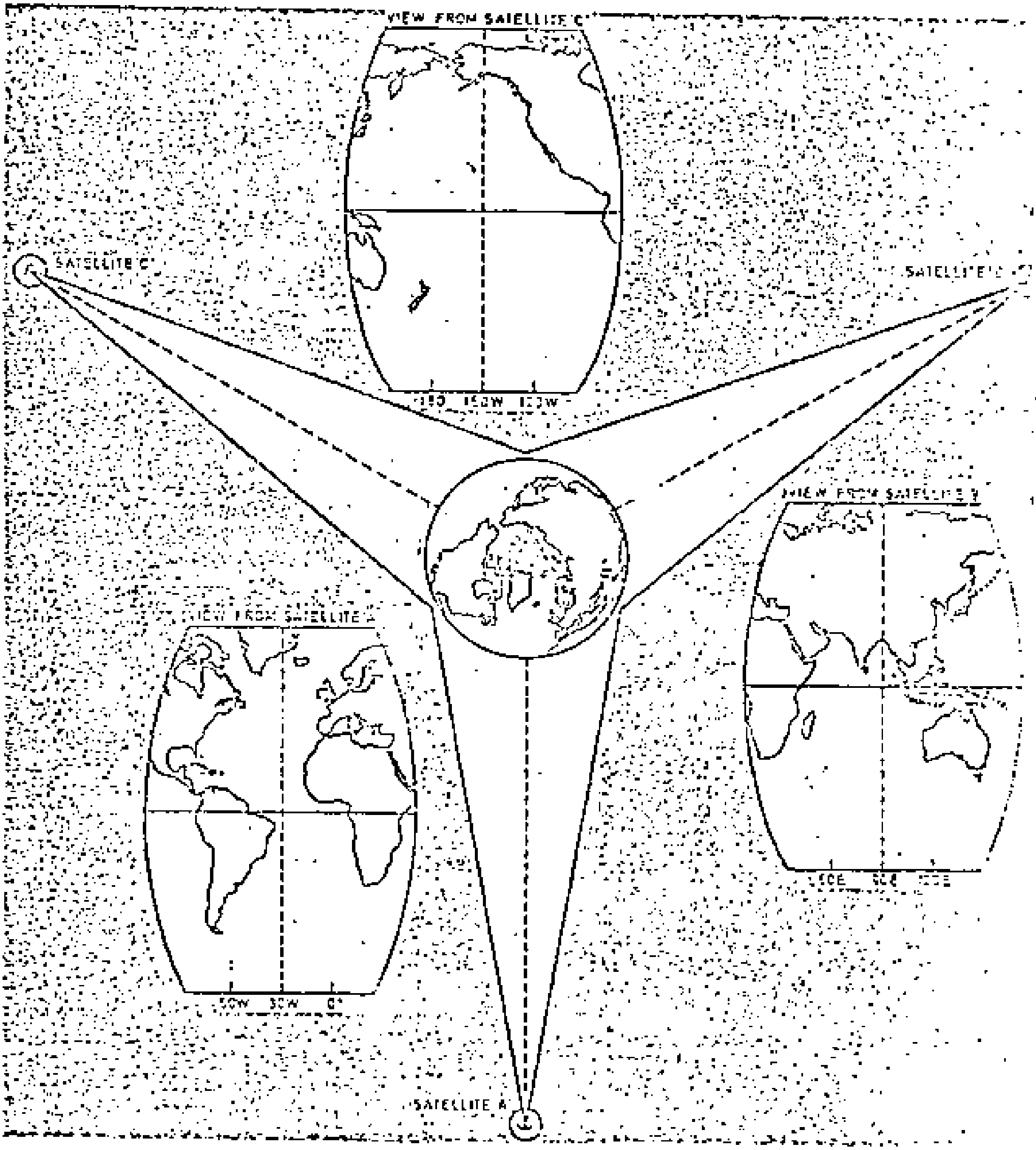


FIGURE 19 WORLD WIDE COVERAGE

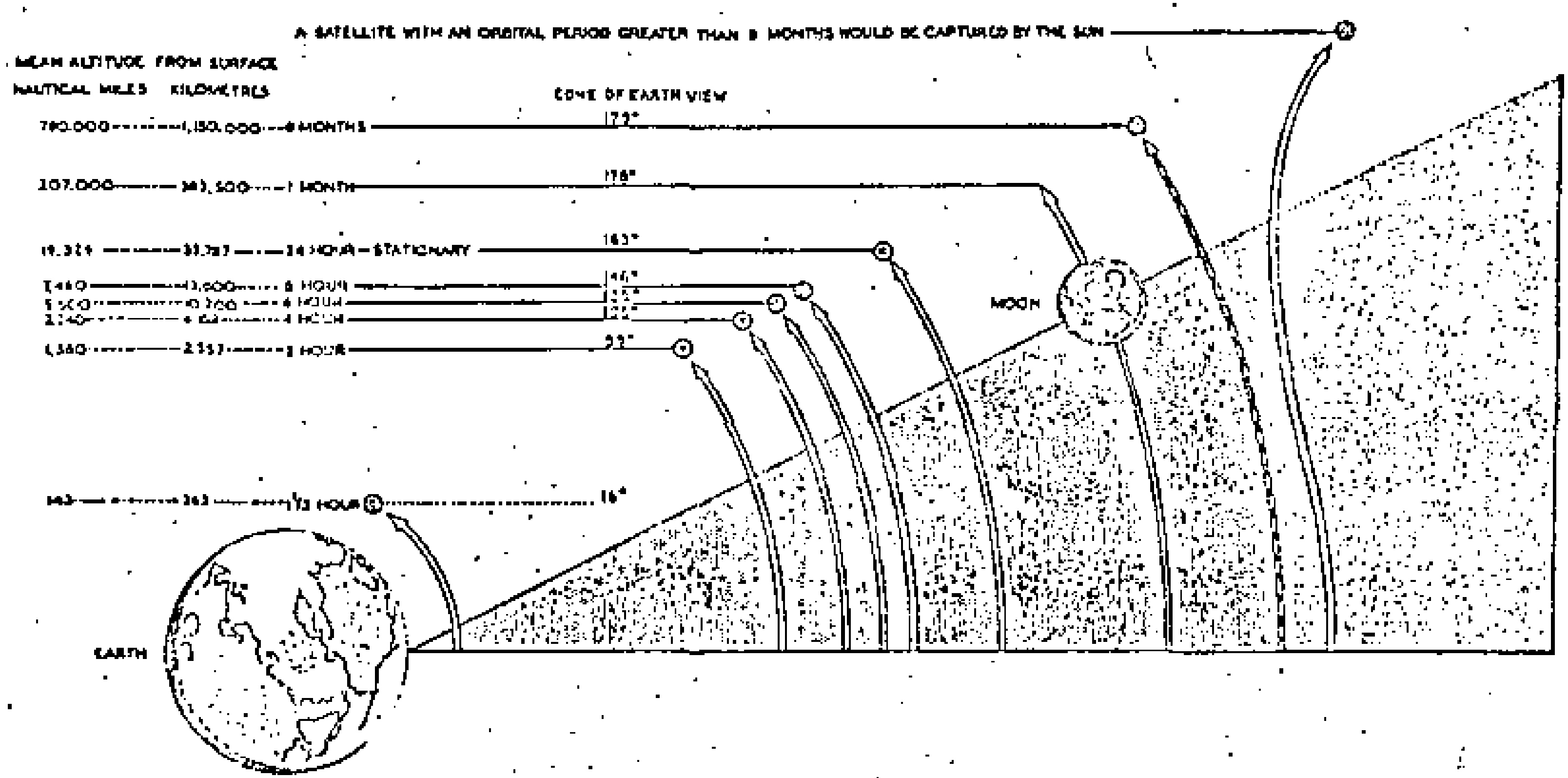


FIGURE 20 ORBIT RELATIONSHIPS

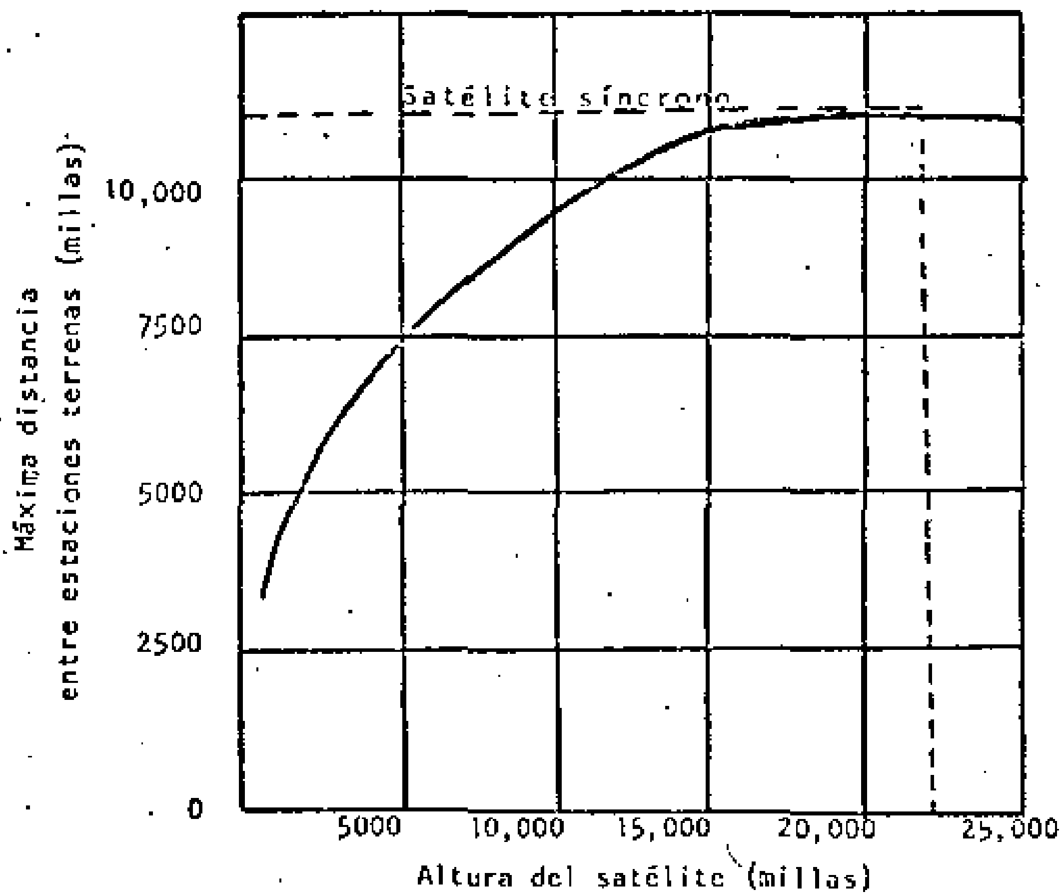


Fig. 21 Máxima separación de las estaciones terrenas

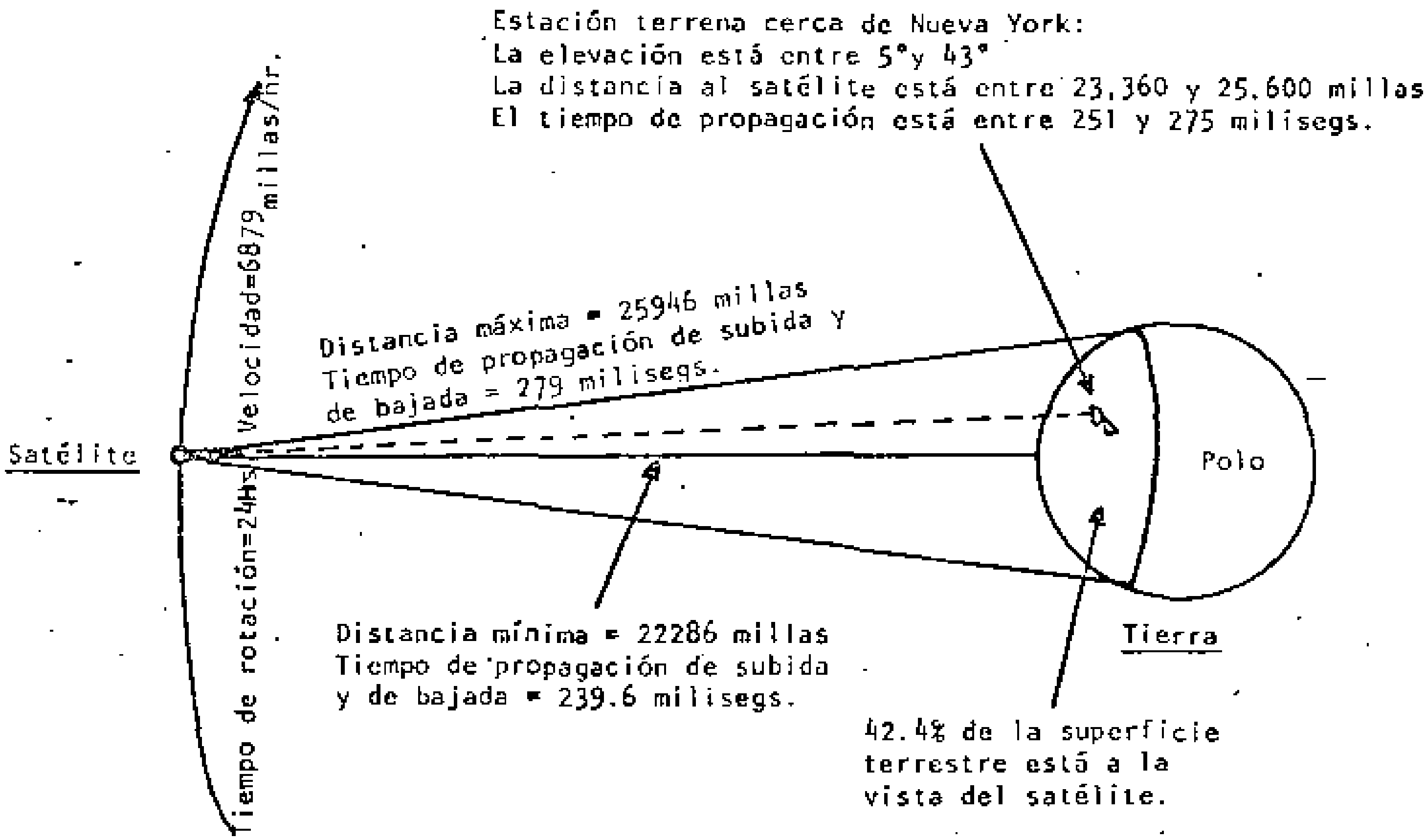


Fig. 22 Distancias y tiempos de propagación con un satélite geoestacionario.

VENTAJAS DE LA ORBITA GEOESTACIONARIA

1. El satélite se mantiene casi estacionario con respecto a las antenas de la Tierra.

2. No hay interrupciones en la transmisión.

3. Debido a la distancia, un satélite geoestacionario está en línea de vista en el 42.4% de la superficie de la Tierra, por lo tanto un gran número de estaciones terrenas pueden intercomunicarse.

4. Tres satélites son suficientes para una cobertura total de la Tierra.

5. Casi no existe el efecto Doppler.

DESVENTAJAS DE LA ORBITA GEOESTACIONARIA

(*)

1. No se cubren las latitudes mayores de 81.25° norte y sur.

2. Existe una atenuación considerable y el tiempo de retraso. es de 270 milisegundos.

List of Geostationary Space Stations by Orbital Positions

The following list includes both satellites already in orbit and those planned for future launching into the geostationary satellite orbit.

The table is based on, and limited to, information supplied to the International Frequency Registration Board (IFRB) by ITU Member administrations under the provisions of the Radio Regulations paragraphs 639AA, 639AJ, 639BA. The designations of the satellites are those officially notified and may not always correspond to the name in general use.

Orbital Position			Space station		Frequency Bands GHz <1	<3	4	6	7	11	12	14	>15
8	E		F/GEO	GEOS-2	1	3							
9	E		F/MET	METEOSAT	1	2							
10	E		F/OTS	OTS	1					11		14	
13	E	•	F	EUTELSAT I-2	1					11		14	
17	E	•	ARS	SABS						11	12	14	
19	E	*	ARS	ARABSAT I		2	4	6					
26	E	*	ARS	ARABSAT II		2	4	6					
26	E	*	IRN	ZOHREH 2						11		14	
29	E		F/GEO	GEOS-2	1	3							
34	E	*	IRN	ZOHREH 1						11	12	14	
35	E		URS	STATSIONAR-2			4	6					
40	E	*	F/MRS	MARECS-B	1	3	4	6					
40	E		F/MTS	MAROTS	1	2							
40	E	*	URS	STATSIONAR-12			4	6					
45	E	*	URS	GALS 2					7				
45	E	*	URS	LOUTCH P2						11		14	
45	E		URS	STATSIONAR-9			4	6					
45	E	*	URS	VOLNA 1	1	2							
47	E	*	IRN	ZOHREH 3						11		14	
53	E		URS	LOUTCH 2						11		14	
53	E		URS/IK	STATSIONAR 5			4	6					
53	E		URS	VOLNA 4		2							
[illegible]	E		USA/IT	INTELSAT5 INDN1			4	6					
57	E	*	USA/IT	INTELSAT4 INDN2			4	6					
57	E	*	USA/IT	INTELSAT4A INDN2			4	6					
57	E	•	USA	INTELSAT5 INDN3			4	6		11		14	
57	E	•	USA	INTELSAT MCS INDN C3			4	6					
60	E		USA/IT	INTELSAT4 INDN2			4	6					
60	E		USA/IT	INTELSAT4A INDN2			4	6					
60	E	*	USA/IT	INTELSAT5 INDN2			4	6		11		14	
60	E	*	USA/IT	INTELSAT MCS INDN B3			4	6					
60	E		USA	USGCSS PHASE2 INDN					7				
60	E	•	USA	USGCSS PHASE3 INDN					7				
63	E		USA/IT	INTELSAT4 INDN1			4	6					
63	E		USA/IT	INTELSAT4A INDN1			4	6					
63	E	*	USA/IT	INTELSAT5 INDN1			4	6		11		14	
63	E	*	USA/IT	INTELSAT MCS INDN A3			4	6					
64.5	E	*	F/MRS	MARECS-C	1	3	4	6					
66	E	*	USA/IT	INTELSAT4 INDN4			4	6					
66	E	*	USA/IT	INTELSAT4A INDN4			4	6					
66	E	•	USA	INTELSAT5 INDN4			4	6		11		14	
66	E	•	USA	INTELSAT MCS INDN D3			4	6					
73	E		USA	MARISAT INDN	1	3*	4*	6*					
74	E		IND	INSAT 1A	1	3	4	6					

Orbital Position			Space station		GHz <1	<3	4	6	7	11	12	14	>15
75	E		USA	FLTSATCOM INDN	1				7				
77	E		INS	PALAPA 2			4	6					
80	E		URS	STATSIONAR-1			4	6					
80	E	*	URS	STATSIONAR-13			4	6					
83	E		INS	PALAPA 1			4	6					
85	E	*	URS	GALS 3					7				
85	E	*	URS	LOUTCH P1						11		14	
85	E		URS	STATSIONAR-3			4	6					
85	E	*	URS	VOLNA 5	1	3							
90	E		URS	LOUTCH 2						11		14	
90	E		URS	STATSIONAR 4			4	6					
90	E	*	URS	VOLNA 4		3							
94	E	*	IND	INSAT 1B	1	3	4	6					
95	E	*	URS	STATSIONAR-14			4	6					
99	E		URS	STATSIONAR T	1			6					
99	E	*	URS	STATSIONAR T2	1			6					
102	E	*	IND	INCOM	1		4	6					
108	E	*	INS	PALAPA B1			4	6					
110	E		J	BSE		3						14	
113	E	*	INS	PALAPA B2			4	6					
118	E	*	INS	PALAPA B3			4	6					
125	E	*	CHN	STW-1			4	6					
130	E		J	ETS 2	1	3				11			34
130	E	*	J	CS-2A		3	4	6					20/30
130	E	*	URS	STATSIONAR 15			4	6					
135	E	*	J	CS-2B		3	4	6					20/30
135	E		J	CS1		3	4	6					18/29
140	E	*	J	GMS	1	3							
140	E		URS	LOUTCH 4						11		14	
140	E		URS	STATSIONAR 7			4	6					
140	E		URS	VOLNA 6		3							
172	E		USA	FLTSATCOM W PAC	1				7				
174	E		USA/IT	INTELSAT4 PAC3			4	6					
174	E	*	USA/IT	INTELSAT4A PAC1			4	6					
174	E	*	USA	INTELSAT5 PAC1			4	6		11		14	
175	E		USA	USGCSS PHASE2 W PAC					7				
175	E	*	USA	USGCSS PHASE3 W PAC					7				
176.5	E		USA	MARISAT PAC	1	3	4	6					
179	E		USA/IT	INTELSAT4 PAC2			4	6					
179	E	*	USA/IT	INTELSAT4A PAC2			4	6					
179	E	*	USA	INTELSAT5 PAC2			4	6		11		14	
177	W	*	EMRS	MARECS B	1	3	4	6					
170	W	*	URS	GALS 4					7				
170	W	*	URS	LOUTCH P4						11		14	
170	W		URS	STATSIONAR 10			4	6					
170	W	*	URS	VOLNA 7	1	3							
149	W		USA	ATS 1	1		4	6					
136	W		USA	US SATCOM 1			4	6					
135	W		USA	GOES WEST	1	3							
135	W	*	USA	USGCSS PHASE3 E PAC					7				
135	W		USA	USGCSS PHASE2 E PAC					7				
132	W	*	USA	US SATCOM 3			4	6					
128	W		USA	COMSTAR D1			4	6					
123.5	W		USA	WESTAR 5			4	6					
122	W		USA	USASAT 6A						11		14	
119	W		USA	US SATCOM 2			4	6					
116	W	*	CAN	ANIK C2						11		14	
114	W		CAN	ANIK A3			4	6					
112.5	W	*	CAN	ANIK C1						11		14	
109	W		CAN	ANIK A2			4	6					
109	W		CAN	ANIK B1			4	6		11		14	
106	W		USA	USASAT 6B						11		14	
105	W		USA	ATS 5	1	3							
104	W	*	CAN	TELESAT D1			4	6					

Orbital Position			Space station		GHz <1	<3	4	6	7	11	12	14
104	W		CAN	ANIK A1			4	6				
103	W	■	MEX	SATMEX-1			4	6				
100	W		USA	FLTSATCOM E PAC	1				7			
99	W		USA	WESTAR-1			4	6				
95	W		USA	COMSTAR D2			4	6				
91	W	*	USA	WESTAR-3			4	6				
87	W		USA	COMSTAR D3			4	6				
86	W		USA	ATS-5	1							
73.4	W		CLM	SATCOL-2			4	6				
75	W		USA	GOES EAST	1	2						
75	W		CLM	SATCOL-1			4	6				
34.5	W	■	USA	INTELSAT MCS ATL E		2	4	6				
34.5	W		USA/IT	INTELSAT4 ATL5			4	6				
34.5	W		USA/IT	INTELSAT4A ATL4			4	6				
34.5	W	*	USA/IT	INTELSAT5 ATL4			4	6		11		14
31	W	*	USA/IT	INTELSAT4A ATL4			4	6				
29.5	W		USA/IT	INTELSAT4 ATL2			4	6				
29.5	W		USA/IT	INTELSAT4A ATL5			4	6				
29.5	W		USA/IT	INTELSAT5 ATL3			4	6		11		14
27.5	W	*	USA/IT	INTELSAT4A ATL3			4	6				
27.5	W	*	USA/IT	INTELSAT5 ATL3			4	6		11		14
27.5	W	* *	USA/IT	INTELSAT MCS ATL B	3		4	6				
25	W	*	URS	GALS-1					7			
25	W	*	URS	LOUTCH P1						11		14
25	W		URS	STATSIONAR-8			4	6				
25	W	*	URS	VOLNA-1	1	2						
24.5	W	■	USA	INTELSAT MCS ATL D		2	4	6				
24.5	W		USA/IT	INTELSAT4A ATL1			4	6				
24.5	W		USA/IT	INTELSAT5 ATL1			4	6		11		14
23	W		USA	FLTSATCOM ATL	1				7			
21.5	W	■	USA	INTELSAT MCS ATL C	1		4	6				
21.5	W	■	USA	INTELSAT5 ATL5			4	6		11		14
21.5	W	*	USA/IT	INTELSAT4 ATL2			4	6				
21.5	W	*	USA/IT	INTELSAT4A ATL1			4	6				
19.5	W		USA/IT	INTELSAT4 ATL3			4	6				
19.5	W		USA/IT	INTELSAT4A ATL2			4	6				
18.5	W	*	USA/IT	INTELSAT4 ATL3			4	6				
18.5	W	*	USA/IT	INTELSAT4A ATL3			4	6				
18.5	W	*	USA/IT	INTELSAT5 ATL2			4	6		11		14
18.5	W	*	USA/IT	INTELSAT MCS ATL A		2	4	6				
18	W	*	BEL	SATCOM III ATL					7			
11	W	*	BEL	SATCOM III					7			
11	W		BEL	SATCOM II					7			
15	W	*	F/MRS	MARECS-A	1	2	4	6				
15	W		I	SIRIO	1					11		
15	W		USA	MARISAT ATL	1	2	4	6				
14	W		URS	LOUTCH 1						11		14
14	W		URS/IK	STATSIONAR-4			4	6				
14	W		URS	VOLNA 3		2						
13	W		USA	USGCSS PHASE2 ATL					7			
12	W	*	USA	USGCSS PHASE2 ATL					7			
12	W	■	USA	USGCSS PHASE2 ATL					7			
11.5	W		F/SYM	SYMPHONIE-2	1		4	6				
11.5	W		F/SYM	SYMPHONIE-1	1		4	6				
10	W	■	F	TELECOM 1A		2	4	6	7		12	14
8.5	W	■	URS	STATSIONAR-11			4	6				
7	W	■	F	TELECOM 1B		2	4	6	7		12	14
4	W		USA/IT	INTELSAT4 ATL1			4	6				
1	W		USA/IT	INTELSAT4 ATL4			4	6				

* Under coordination as per RR639AJ

■ Advanced publication only under RR639AA

(courtesy of International Telecommunication Union)

FRECUENCIAS.

Las bandas de frecuencias que se han determinado para comunicaciones vía satélites se ilustran en la tabla I. La tabla II muestra las frecuencias para sistemas móviles marítimos y aeronáuticos.

Las frecuencias que más se usan actualmente son esas que están abajo de 14.5 GHz. Arriba de 10 GHz, la propagación a través de la atmósfera de la tierra es afectada por la lluvia, la cual produce una atenuación suficientemente grande para afectar el comportamiento del sistema. Las componentes para sistemas de comunicación de banda amplia se obtienen fácilmente en las bandas de 2,4 y 6 GHz, habiéndose desarrollado éstas para los sistemas L OS. Consecuentemente, los sistemas existentes han operado principalmente en las bandas de 4 y 6 GHz, para satélites civiles y en las bandas de 7-8 GHz para sistemas militares. INTELSAT V usa las vandas de 11.7/14,0 GHz y desde luego las de 4/6 GHz. Los sistemas OTS y ECS usarán 11,7/14,0 GHz exclusivamente.

La congestión en las bandas de 4/6GHz ha forzado que en los nuevos sistemas se consideren mayores frecuencias, pero, hasta la fecha, las bandas de 4/6 GHz han sido las más atractivas. La órbita geoestacionaria, vista desde la Tierra, se extiende hasta 120° en longitud; la separación típica de los satélites es de 3°, para prevenir que las estaciones terrenas causen interferencia a satélites adyacentes, así que todo el sector de la órbita geoestacionaria puede acomodar a 40 satélites operando en la misma frecuencia. En la región visible a los Estados Unidos existen cerca de 20 satélites operando en las bandas de 4 y 6 GHz. Los satélites marítimos MARISAT Y MAROTS usan las bandas 1535-1542.5 y 1636-1644 GHz para enlaces barcos-satélites pero 4/6 o 12/14 para enla

Servicio de Satélite Fijo	Frecuencia	Satélites de Radiodifusión	Notas
	606 - 790 MHz	Bajada	
	2500 - 2535	Bajada doméstica	S
	2535 - 2550	Bajada y	S
	2550 - 2655	Bajada comunitaria	S
Subida	2655 - 2690	Subida	S
Bajada	3400 - 3600		S, también usada para radar
Bajada	3600 - 4200		S
Bajada	4500 - 4800		S
Subida	5725 - 5850		P
	5850 - 7075	(áreas limitadas solamente)	S
Bajada	7250 - 7450		S
Bajada	7300 - 7450		S
Bajada	7450 - 7750		S
Subida	7900 - 8025		S
Subida	8025 - 8215		S
Subida	8215 - 8400 MHz		S
Bajada	10.70 - 11.7 GHz	Enlaces de sub., región I unic.	S
	11.70 - 12.5	Bajada	S
Subida	12.75 - 13.25		S
Subida	14.0 - 14.25		P
Subida	14.25 - 14.3		P
Subida	14.3 - 14.8		S
	17.3 - 18.10	Subida	S
Bajada	17.7 - 19.7		S
Bajada	19.7 - 21.2		P
Subida	27.5 - 29.5		S
Subida	29.5 - 31		P
Bajada	37.5 - 40.5		S
	40.5 - 42.5	Bajada	S
Bajada	42.5 - 43.5		S
Subida	47.2 - 51.4		S
Bajada	81 - 84		S
	84 - 86	Bajada	S
Subida	92 - 95		S
Bajada	102 - 105		S
Subida	140 - 142		S
Bajada	149 - 164		S
	202 - 217		S
	231 - 238 GHz		S
	265 - 275		S

N.B : P = Servicio Primario S = Compartida con otros servicios.

Tabla 1 : Distribuciones de frecuencias para servicios fijos de satélites y satélites de radiodifusión.

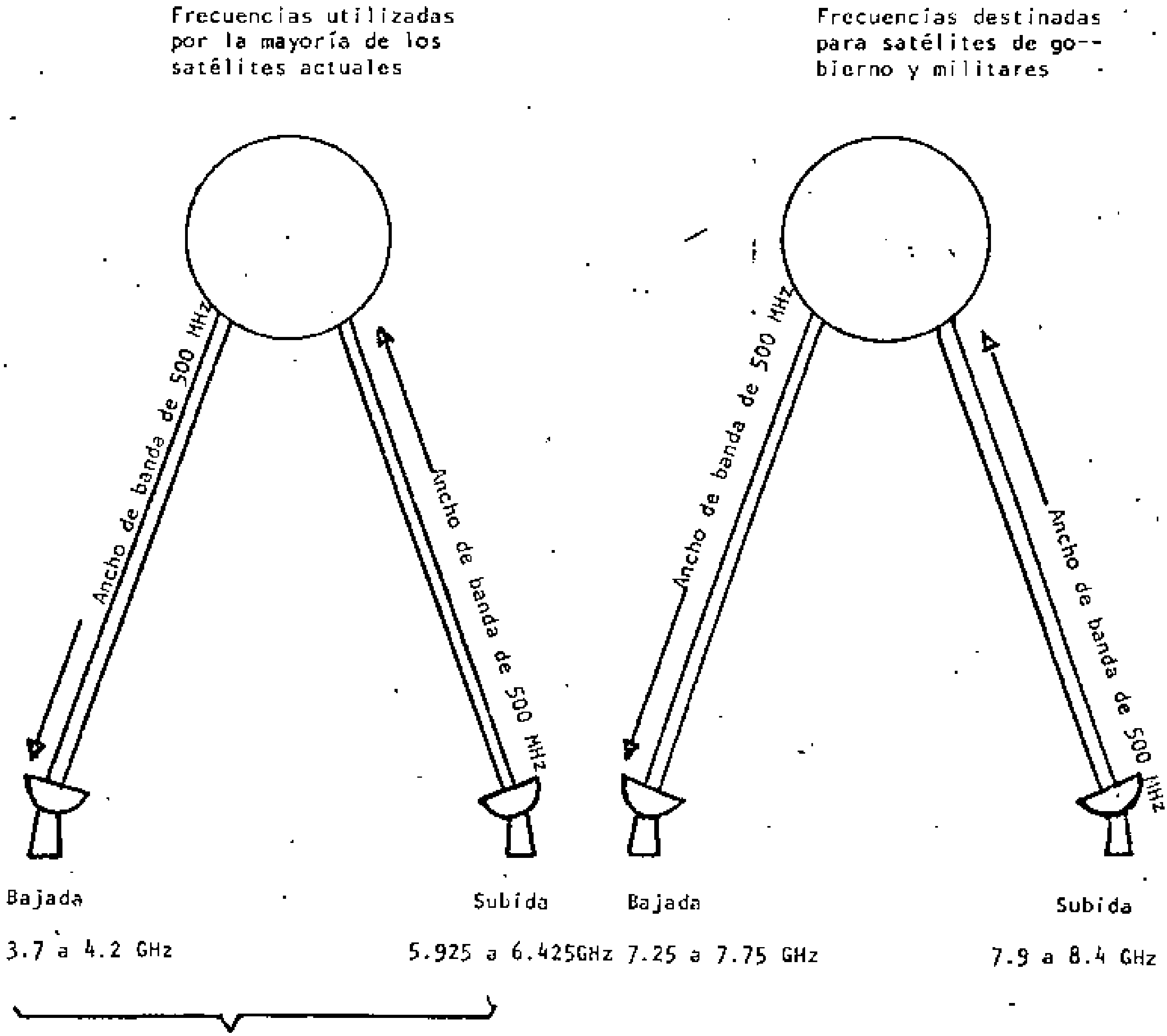
FRECUENCIA	MOVIL AERONAUTICO	MOVIL MARITIMO	MOVIL TERRESTRE
1530 - 1535 MHz		S	
1535 - 1544		E	
1544 - 1545	S	S	
1545 - 1549	E		
1626.5 - 1644.5		E	
1644.5 - 1646.5	S	S	S
1646.5 - 1660	E		
1660 - 1660.5	S		
14.0 - 14.5 GHz			S
43.5 - 47.0			S
66.0 - 71.0			S
95 - 100			S
134 - 142			S
252 - 265			S

Compartida Exclusiva

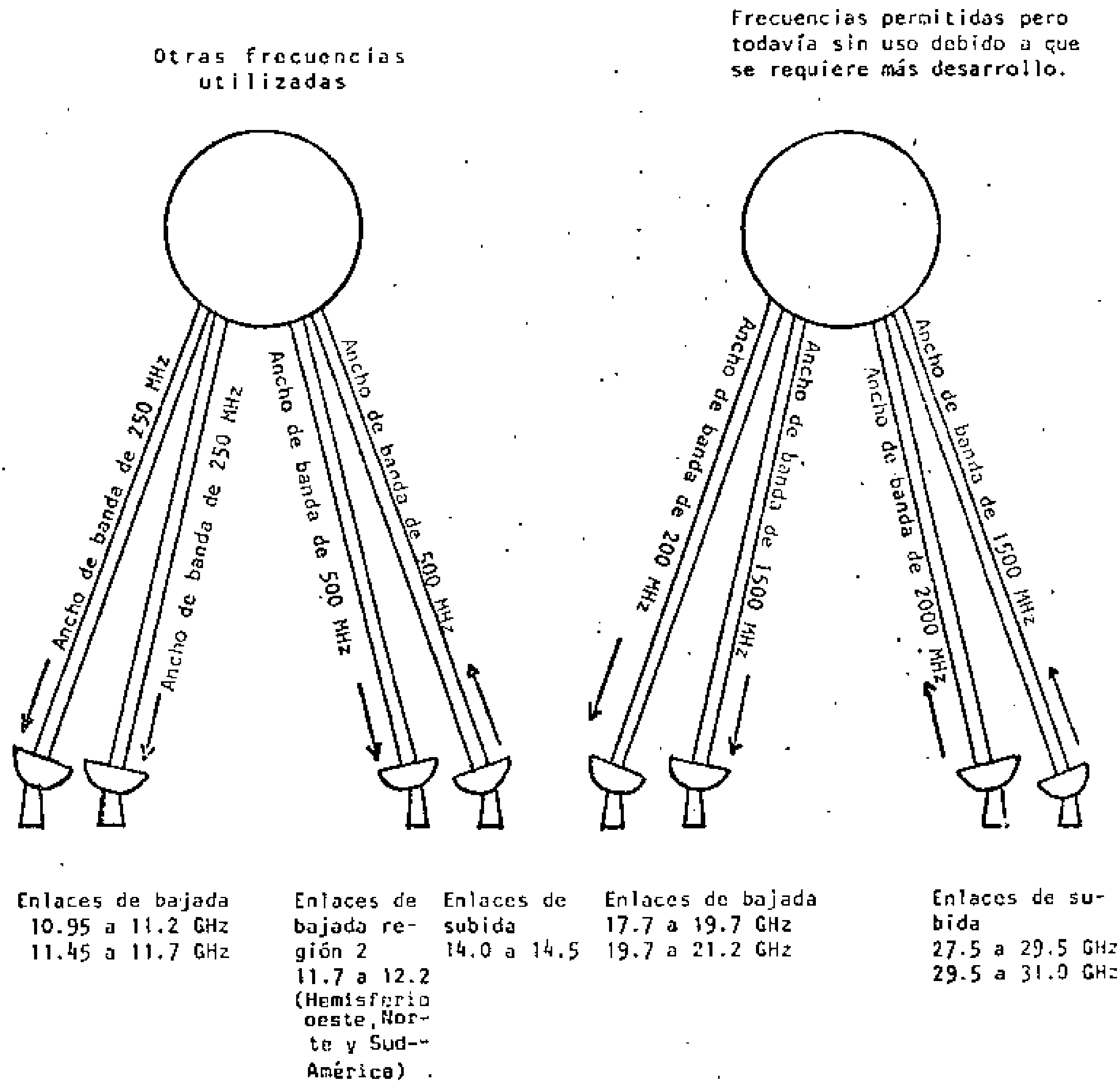
La mayoría de las bandas de servicio de satélites fijos arriba de 3.4 GHz pueden también ser usadas por terminales móviles, para enlaces de subida y de bajada.

Tabla II

Frecuencias para servicios móviles.



Existe una fuerte congestión en estas bandas de frecuencias, debido a que son las mismas utilizadas en microondas terrestres.


Otras frecuencias utilizadas
Frecuencias permitidas pero todavía sin uso debido a que se requiere más desarrollo.
Ancho de banda de 250 MHz
Ancho de banda de 250 MHz
Ancho de banda de 500 MHz
Ancho de banda de 500 MHz
Ancho de banda de 200 MHz
Ancho de banda de 1500 MHz
Ancho de banda de 2000 MHz
Ancho de banda de 1500 MHz
Enlaces de bajada 10.95 a 11.2 GHz 11.45 a 11.7 GHz
Enlaces de bajada región 2 11.7 a 12.2 (Hemisferio oeste, Norte y Sud-América)
Enlaces de subida 14.0 a 14.5
Enlaces de bajada 17.7 a 19.7 GHz 19.7 a 21.2 GHz
Enlaces de subida 27.5 a 29.5 GHz 29.5 a 31.0 GHz

ces satélites-Tierra. Las estaciones terrenas grandes pueden lograr muy buen desempeño a las frecuencias altas, dejándose las bandas de 1500 MHz para los enlaces barcos-satélites, donde es difícil instalar grandes antenas en lugares precisos.

En general, es más fácil y barato usar las frecuencias más bajas. Sin embargo, se limita el ancho de banda y la interferencia es mayor, ya que estas frecuencias se usan para sistemas terrestres. Las frecuencias más altas ofrecen la ventaja de mayores anchos de banda pero como ya se había dicho se presentan dificultades en la propagación. Además, las antenas de un diámetro dado pueden producir haces más angostos a frecuencias mayores, así que los satélites regionales pueden ser hechos más directivos en la parte alta del rango de frecuencias. La frecuencia no aparece en la ecuación de enlace si se fija el tamaño de la antena transmisora en el satélite.

VENTAJAS DE FRECUENCIAS MENORES A 10 GHz

- MENOR ABSORCION ATMOSFERICA.
- MENOR RUIDO.
- SE TIENE UNA TECNOLOGIA BIEN DESARROLLADA.
- MENOR ATENUACION.

DESVENTAJAS DE FRECUENCIAS MENORES A 10 GHz

- LAS BANDAS SON COMPARTIDAS CON SERVICIOS TERRESTRES.
- CONGESTIONAMIENTO DE LA ORBITA.

VENTAJAS DE FRECUENCIAS MAYORES A 10 GHz

- MENOR INTERFERENCIA.
- SE PUEDEN COMPARTIR CON SERVICIOS TERRESTRES.
- FACILIDAD EN LA ORBITA.

DESVENTAJAS DE FRECUENCIAS MAYORES A 10 GHz

- MAYOR ATENUACION.
- MAYORES EFECTOS POR LLUVIA Y GASES ATMOSFERICOS.

El Sistema de Comunicaciones.

Un modelo general que representa un sistema de comunicaciones vía satélite se ilustra en la Fig. 23. La señal se genera por un usuario y entra al sistema terrestre. En algunos sistemas, el sistema terrestre es simplemente un enlace dedicado a la estación terrena, mientras en otros casos es una red telefónica de conmutación. En la estación terrena se procesa la señal de banda base y se transmite a una frecuencia de radio-frecuencia (RF) al satélite donde se procesa y se retransmite a la estación terrena receptora. La estación terrena procesa la señal hasta la señal de banda base la cual se envía al usuario a través de la red terrestre, la Fig. 24 muestra un diagrama simplificado de la estación terrena.

El primer modelo del sistema de comunicación de interés es el modelo FI a FI mostrado en la Fig. 25. Este modelo considera que existe un solo haz desde la antena del satélite, el cual ilumina todas las estaciones terrenas del sistema.

Las características de interés en el modelo que serán discutidas son:

1. La técnica de acceso múltiple.
2. La potencia de la estación terrena y la ganancia de la antena.
3. El ruido y la interferencia.
4. El canal de subida.
5. La ganancia de la antena receptora del satélite.
6. El receptor del satélite.
7. El transpondedor del satélite.
8. La potencia del satélite y la ganancia de la antena.
9. El canal de bajada.
10. La estación terrena y el receptor.

El amplio reflector del satélite ATS-6 se agrega grandemente al EIRP.

El canal de bajada se muestra en la Fig. 28. La potencia de recepción en la estación terrena será:

$$P_b = EIRP_{sat} - L_b - L_e + G_T$$

donde G_T es la ganancia de la estación terrena.

Primero, considérese la trayectoria de transmisión desde una estación terrena a otra vía satélite. El primer paso es obtener las ecuaciones de enlace. El modelo del enlace de subida se muestra en la Fig. 26. La densidad de flujo en el satélite está dada por

$$F_s = EIRP - 10 \log (4\pi d^2) \quad w/m^2$$

donde EIRP es la potencia efectiva radiada isotrópicamente y es la potencia del transmisor tomando en cuenta la ganancia de la antena y las pérdidas en las líneas.

d es la distancia de la trayectoria.

La potencia de la señal recibida en el satélite es:

$$P_s = EIRP - L_s - L_e + G_{ss}$$

donde L_s es la atenuación en el espacio libre, L_e otras atenuaciones y G_{ss} es la ganancia de la antena del satélite. La Fig. 27 muestra la relación entre el tamaño de la antena y su ganancia.

El EIRP de los satélites se ha incrementado con el tiempo, ya que tanto la potencia del transmisor como la ganancia de la antena han aumentado conforme los satélites generan más y más potencia y conforme tengan mayor facilidad para desplegar las antenas. La siguiente tabla muestra el incremento

AÑO	SATELITE	EIRP(watts)
1965	INTELSAT I	14
1967	INTELSAT II	36
1968	INTELSAT III	200
1971	INTELSAT IV	6400
1974	ATS-6	140000

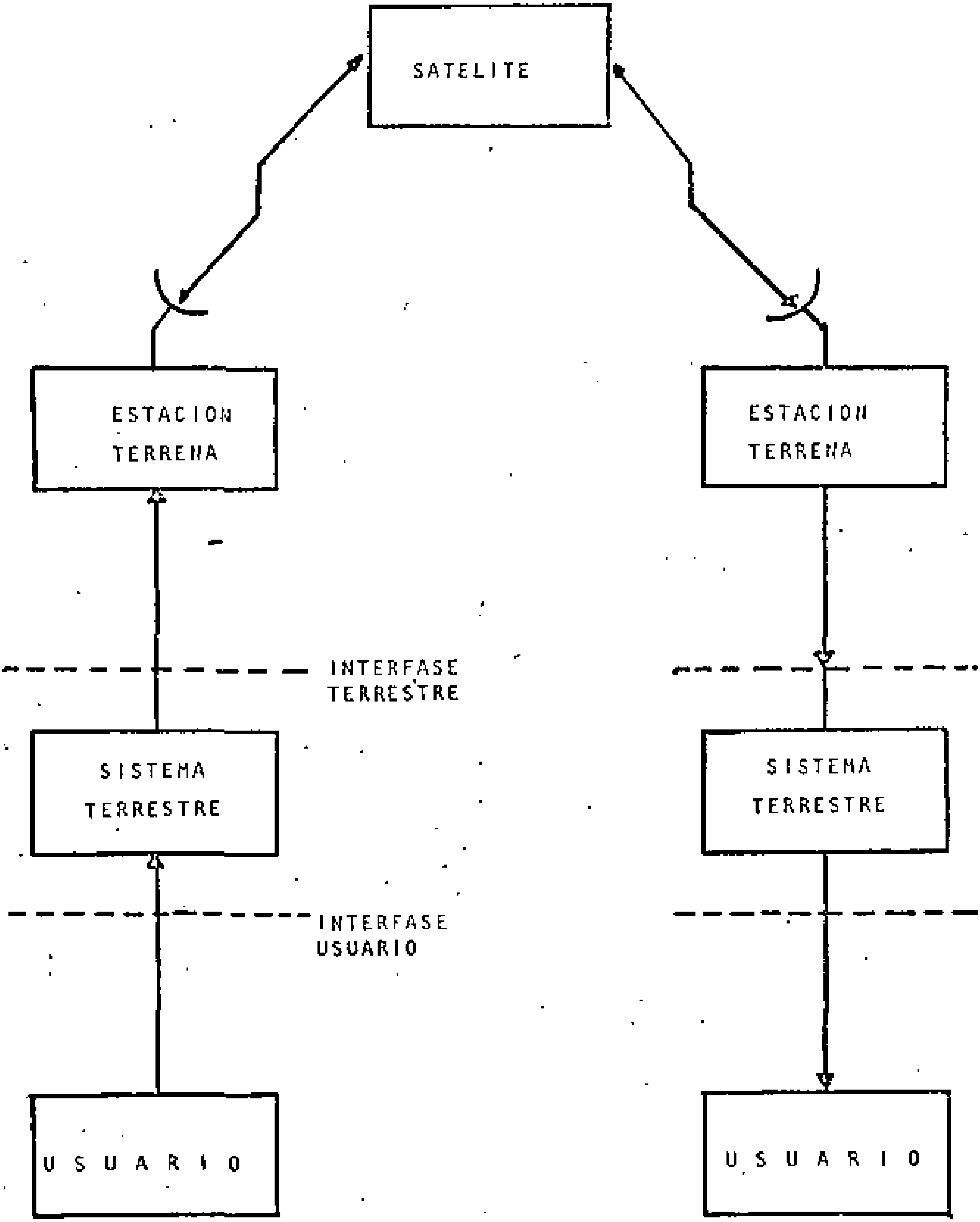


Fig. 23 SISTEMAS DE COMUNICACION POR SATELITE

Estación Terrena

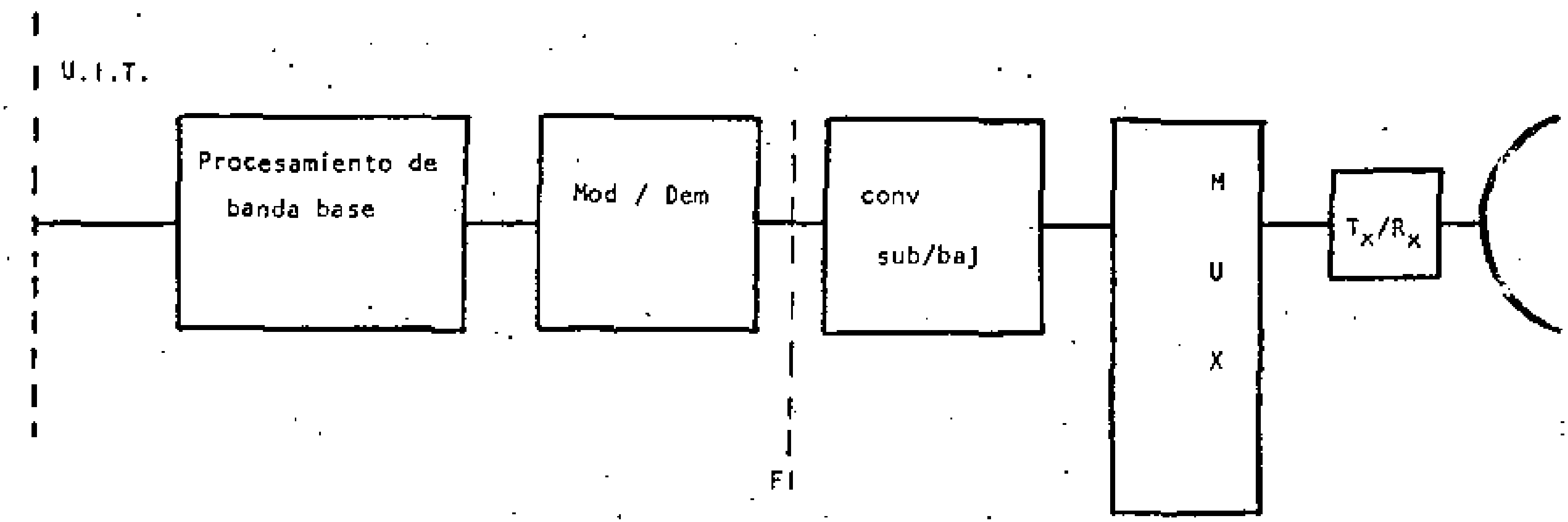


Fig. 29 . Estación Terrena

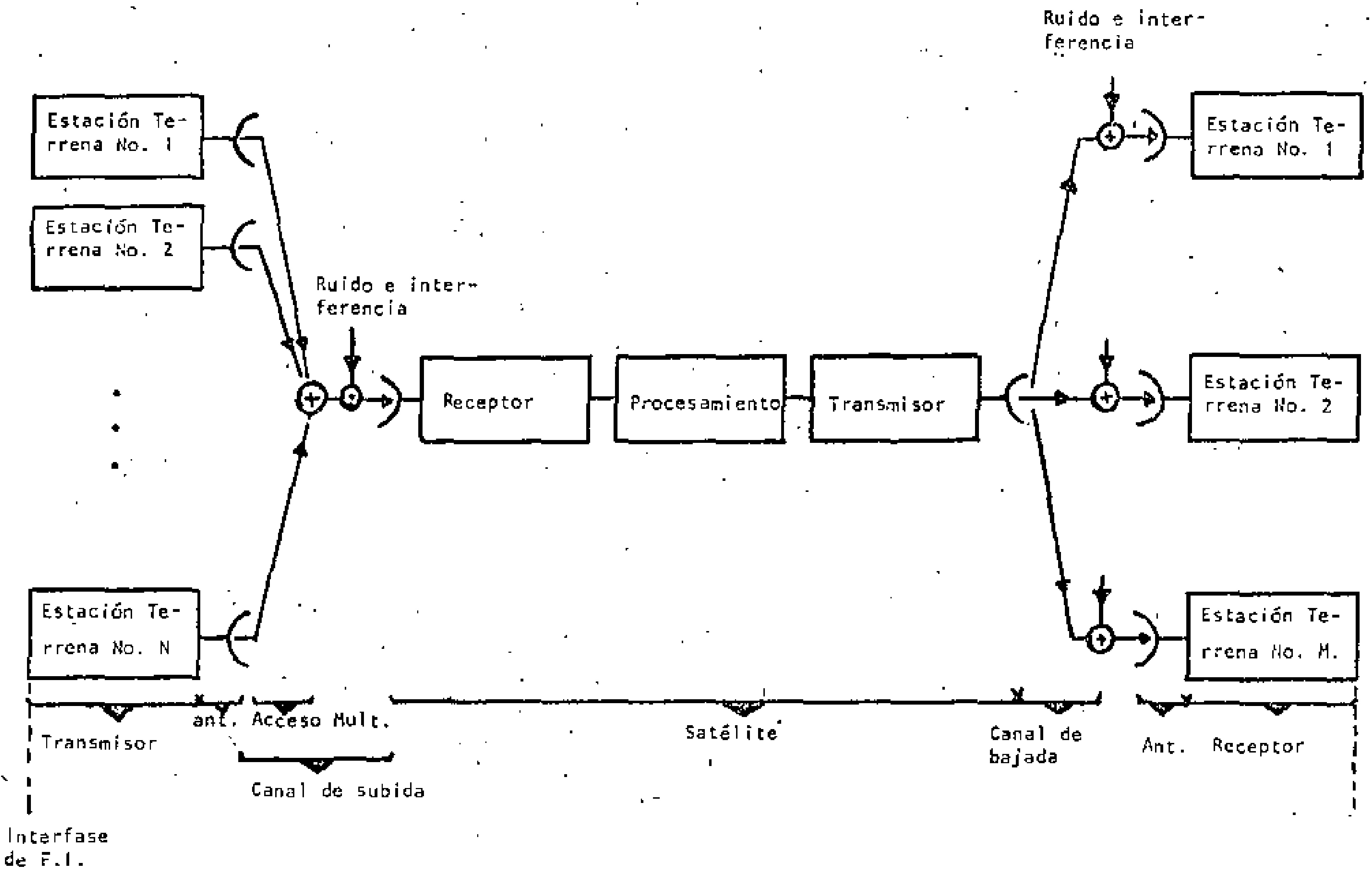


Fig. 25 Modelo de un sistema de comunicación (FI- a FI)

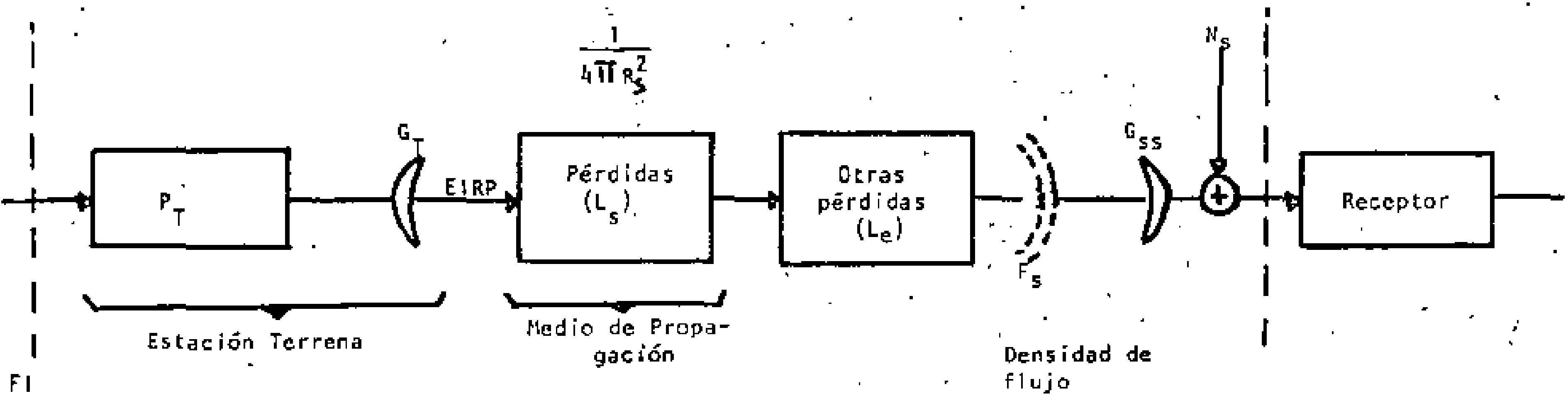


Fig. 26 Modelo de subida.

$$EIRP\ (dBw) = P_T\ (dBw) + G_T$$

$$Pérdidas\ (L_s) = 20 \log f + 20 \log d + 32.46$$

$$\left(\frac{C}{N}\right)_s = EIRP - (L_s + L_e) + G_{SS} - 10 \log (KTB)$$

$$\left(\frac{C}{N}\right)_s = EIRP - (L_s + L_e) + \frac{G_{SS}}{T} - K - 10 \log B$$

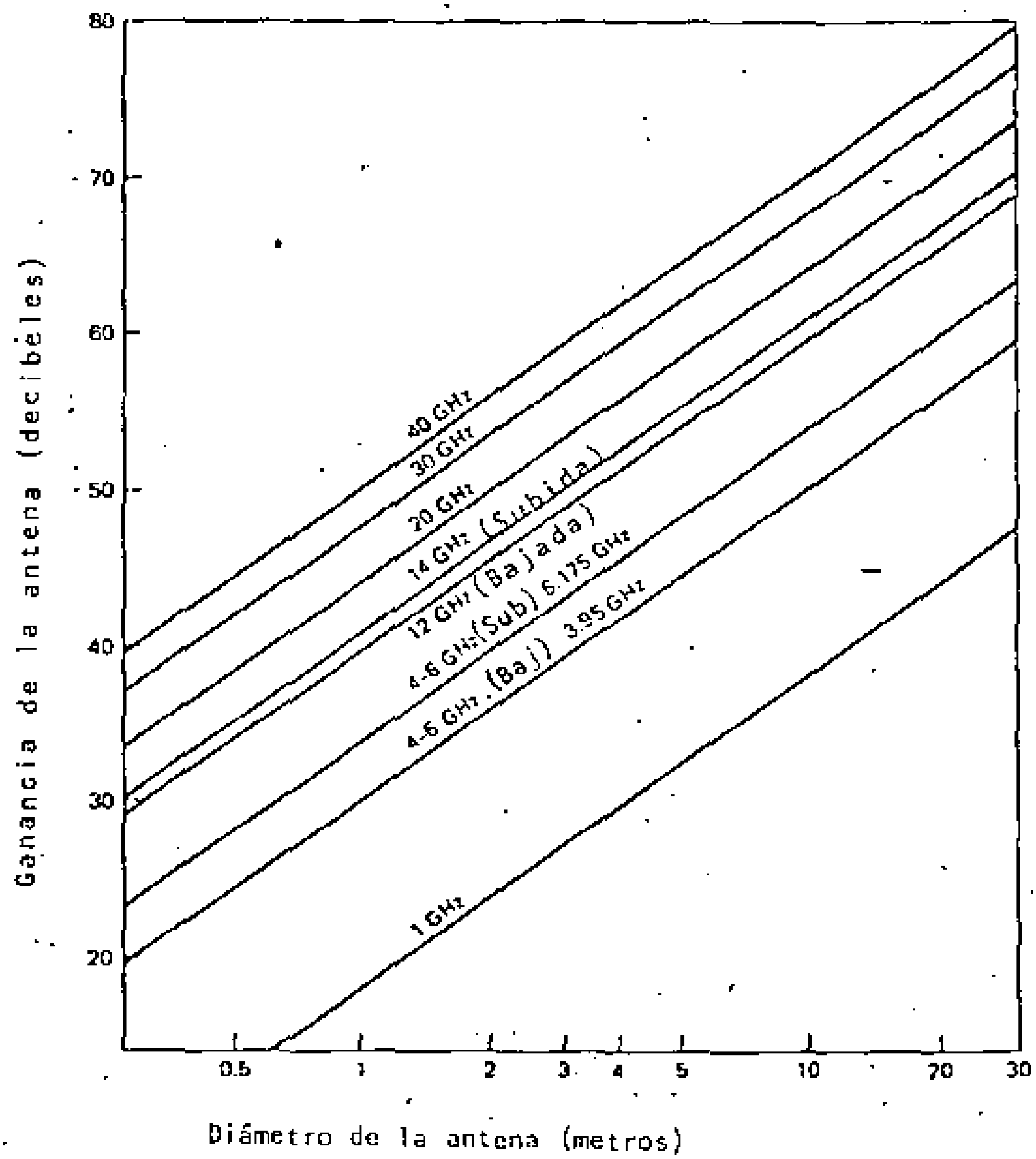

80
70
60
50
40
30
20
Ganancia de la antena (decibeles)
40 GHz
30 GHz
20 GHz
14 GHz (Subida)
12 GHz (Bajada)
4-6 GHz (Sub) 6.175 GHz
4-6 GHz (Baj) 3.95 GHz
1 GHz
0.5
1
2
3
4
5
10
20
30
Diámetro de la antena (metros)

Fig. 27

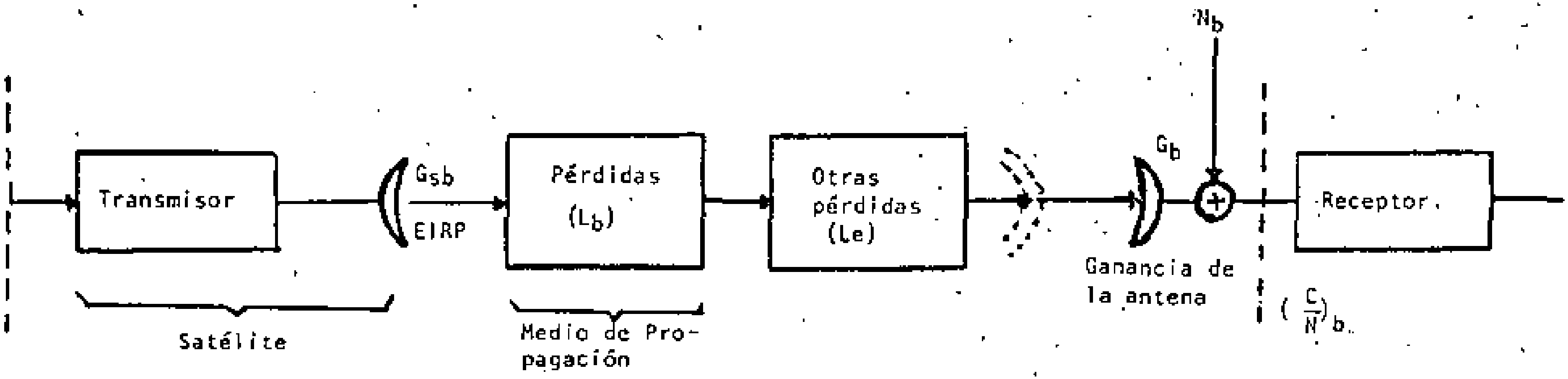


Fig. 28 Modelo de bajada.

$$EIRP\ (dBW) = Ps\ (dBW) + G_{sb}\ (dB)$$

$$\text{Pérdidas}\ (L_b) = 20 \log f + 20 \log d + 32.46$$

$$\left(\frac{C}{N}\right)_b = EIRP - (L_b + L_e) + G_b - 10 \log (KTB)$$

$$\left(\frac{C}{N}\right)_b = EIRP - (L_b - L_e) + \frac{G_b}{T} - K - 10 \log B$$

EFECTOS ATMOSFERICOS

Esta parte trata con varios efectos de propagación que influencían el desempeño de sistemas de comunicación vía satélite. Para bajas frecuencias (100 MHz - 4 GHz), la cintilación ionosférica es un verdadero problema. El fenómeno troposférico, tal como absorción molecular, atenuación por lluvia y dispersión debe también considerarse para frecuencias arriba de 4 GHz. El efecto dominante arriba de 4 GHz es la atenuación debida a lluvia. Las otras degradaciones causadas por lluvia, tal como depolarización, interferencia entre sistemas debido a dispersión, aumento en el ruido de las estaciones terrenas y degradación en el desempeño de la antena de la estación terrena, también son discutidas aquí.

Conforme los sistemas de satélites utilizan mayores frecuencias y usan más sofisticados sistemas de procesamiento de señales, será importante tener un adecuado desarrollo en el campo de la propagación tanto teórico como experimental.

CINTILACION IONOSFERICA.

Las primeras observaciones en esta área revelaron que las señales de radio, frecuentemente presentaban fluctuaciones en intensidad, las cuales eran causadas por irregularidades ionosféricas y subsecuentemente, tales fluctuaciones venidas de fuentes estelares y de satélites fueron estudiadas extensivamente. Estas fluctuaciones de amplitud, fase y ángulo de arrivo, comúnmente llamada cintilación ionosférica, han sido observadas a frecuencias entre 10 MHz y 6GHz.

EFECTOS DE PROPAGACION

- CINTILACION IONOSFERICA

- ABSORCION ATMOSFERICA
  - Oxígeno y vapor de agua
  - Lluvia
  - Niebla y Nubes
  - Nieve y Granizo
  - Electrones libres en la atmósfera.

- DEPOLARIZACION DEBIDO A LLUVIA

- PROTECCION DE ESTACIONES TERRENAS (RADOME).

La cintilación depende de varios factores tales como localización geográfica, frecuencia, trayectoria de propagación, condiciones geofísicas y la medida usada para describir la cintilación. Estas fluctuaciones son realmente producidas por pequeñas irregularidades en la densidad de electrones en la capa F de la ionósfera. Recientes mediciones han motivado esta situación en altas y ecuatoriales altitudes. Existe una región irregular a latitudes altas cuya frontera baja alcanza 57° cerca de la media noche. Durante tormentas magnéticas la frontera desciende a latitudes más bajas y el desvanecimiento es mayor.

Las irregularidades producen cintilaciones profundas en el rango de VHF a ± 15° del Ecuador. Para minimizar el efecto de este fenómeno en transmisiones de satélites, el diseñador del sistema puede utilizar las distribuciones de amplitud, razones de desvanecimiento y profundidad al diseñar la modulación.

Por ejemplo, a 137 MHz la cintilación con desvanecimiento arriba de 6dB ocurren en las trayectorias del Zenith en menos del 20% del tiempo cerca del Ecuador, menos del 2% en las regiones aurales y menos de 0.1% en latitudes medias.

Observaciones experimentales de cintilaciones fueron realizadas en Millstone usando fase coherente a 150 y 400 MHz del sistema de satélites naval de los Estados Unidos (NNSS) y receptores en las facilidades de Millstone Hill. Un ejemplo de fluctuaciones de nivel de la señal recibida y de la fase diferencial (variación de fase a 150 MHz relativa a la referencia de fase de 400 MHz) para un segmento de 1 min. se da en la Fig.29 .

Para este segmento de datos, los valores de σx fueron 1.27 dB en UHF y 5.50 en VHF.

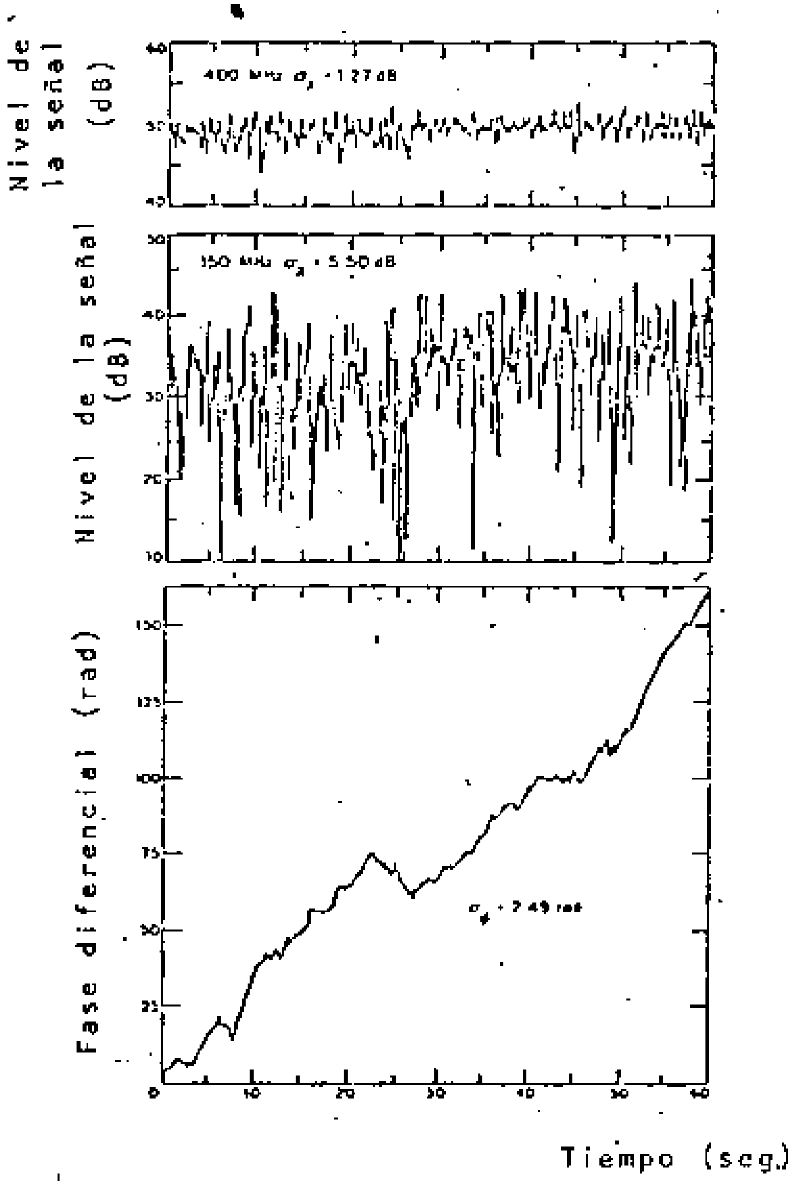


Fig. 29 Nivel de la señal y fluctuaciones de fase para un intervalo de observación de 1 min.

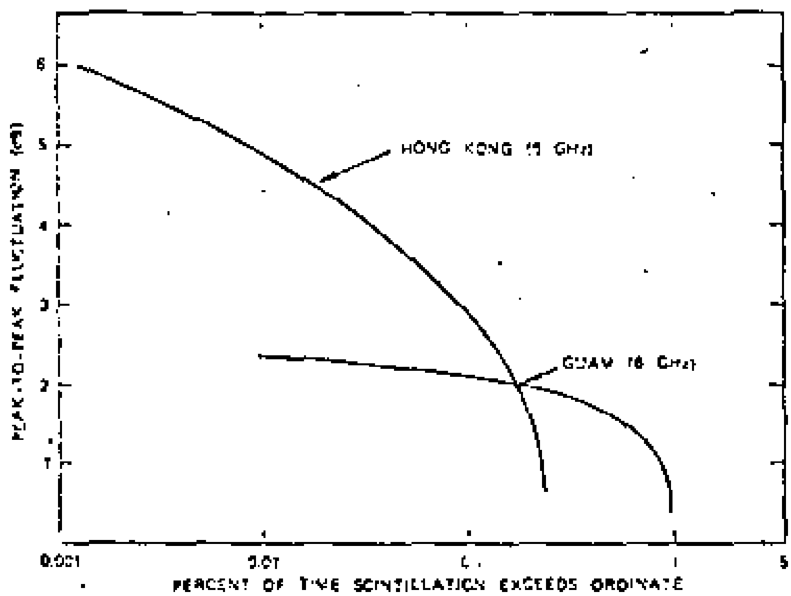


Figure 5. *Cumulative Amplitude Distributions at Guam (6 GHz) and Hong Kong (6 GHz)*

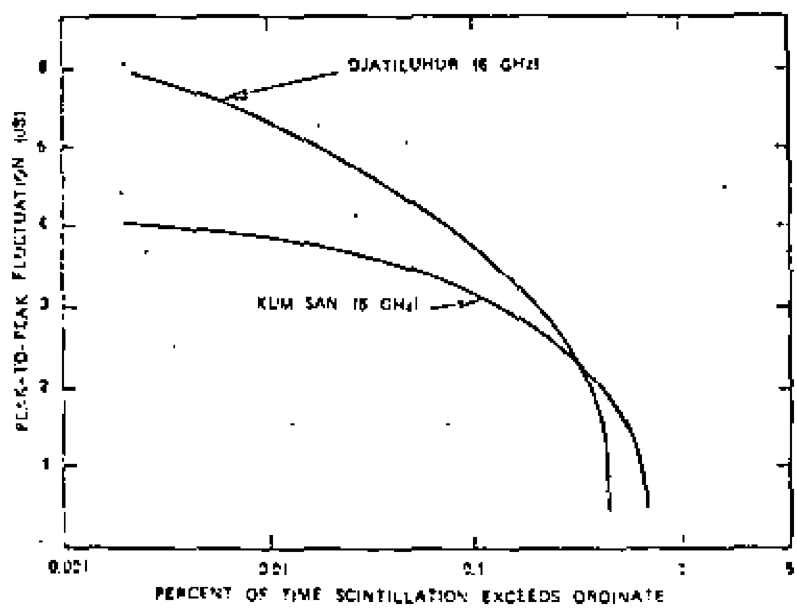


Figure 6. *Cumulative Amplitude Distributions at Djatiluhur (6 GHz) and Kum San (6 GHz)*

ABSORCION ATMOSFERICA

- OXIGENO MOLECULAR
- VAPOR DE AGUA
- LLUVIA
- NIEBLA Y NUBES
- NIEVE Y GRANIZO
- ELECTRONES LIBRES EN LA ATMOSFERA

OXIGENO MOLECULAR, VAPOR DE AGUA Y ELECTRONES.

El oxígeno molecular y el vapor de agua son relativamente constantes. Las Figs. 30 y 31 muestran la absorción debida al oxígeno y al vapor de agua. La absorción es mayor para grandes ángulos de elevación, debido a que el haz recorre mayores trayectorias a través de la atmósfera.

La absorción por oxígeno molecular tiene un pico a 60 GHz, y la aborción por moléculas de agua tiene un pico a 21GHz. La absorción es causada por la onda de radio que cambia los niveles de energía rotacional de las moléculas, y los efectos de resonancia ocurren a esas frecuencias. Cuando hay electrones libres en la atmósfera de la Tierra las ondas de radio chocan con ellos. Esto causa absorción debido a que la energía de radio se transfiere a los electrones. La densidad de electrones de la ionósfera se reduce grandemente durante las horas de la noche. Los principales efectos por absorción de electrones son a frecuencias abajo de 100 MHz y tiene efectos despreciables en las bandas UHF y SHF.

La Fig. 32 muestra un diagrama de la absorción causada por electrones, oxígeno y vapor de agua.

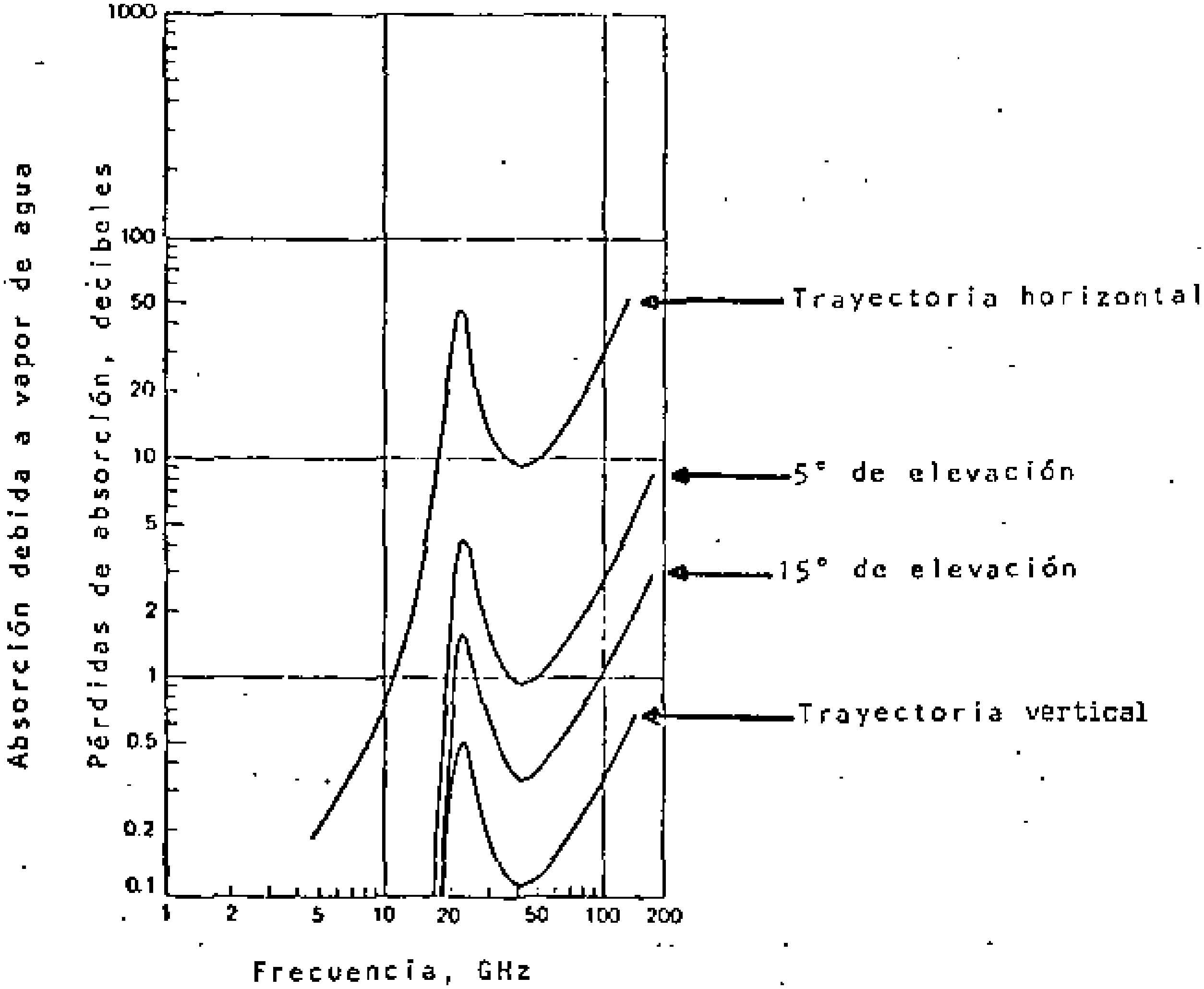


Fig. 30 Absorción en la atmósfera causada por vapor de agua sin condensar.

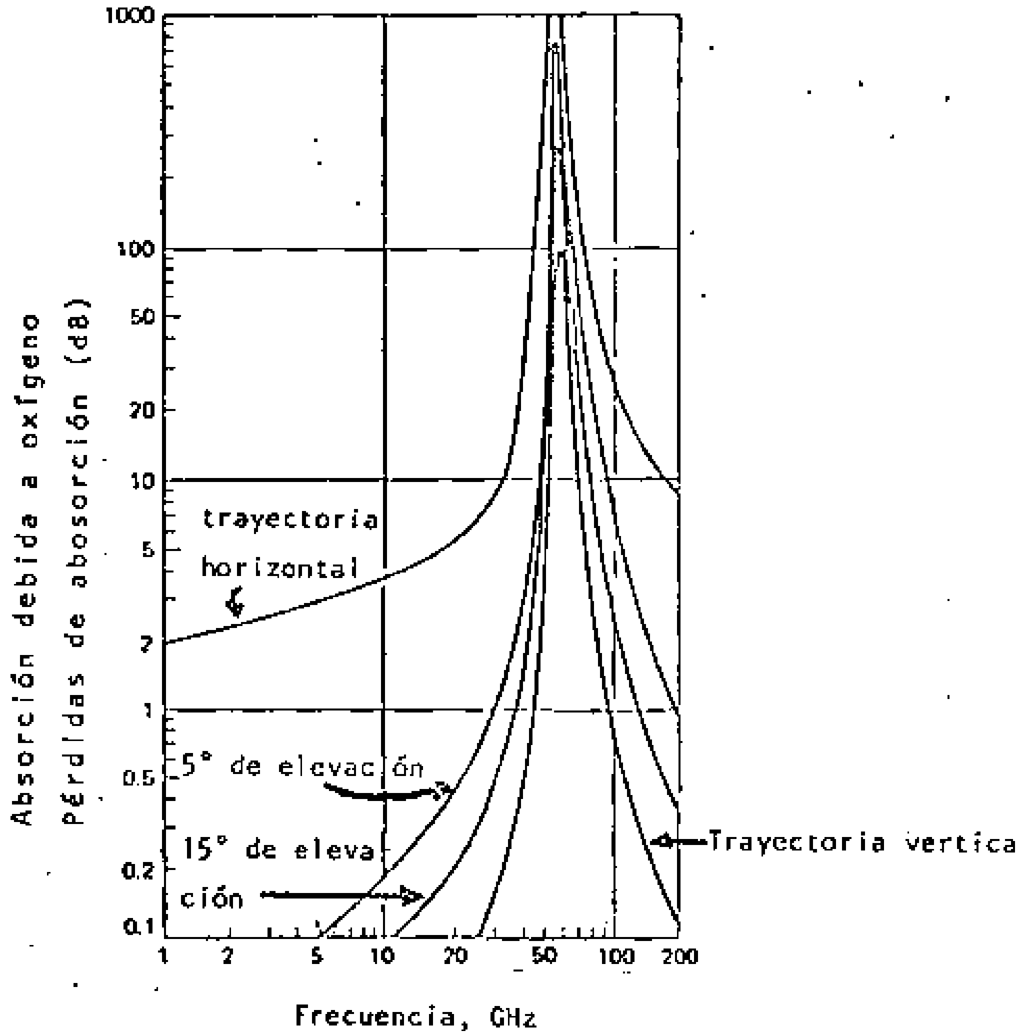


Fig. 31 Abosorción molecular..

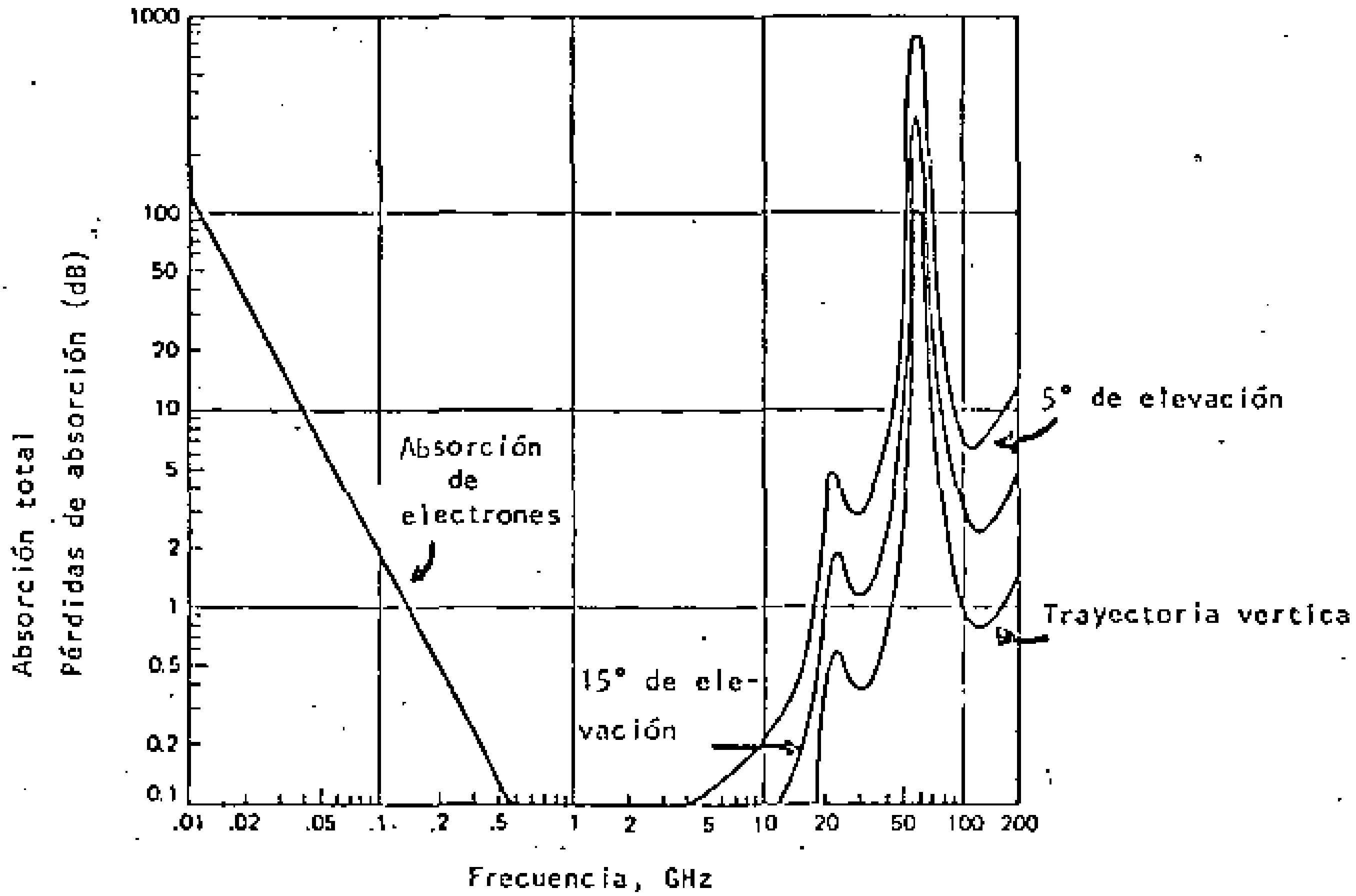
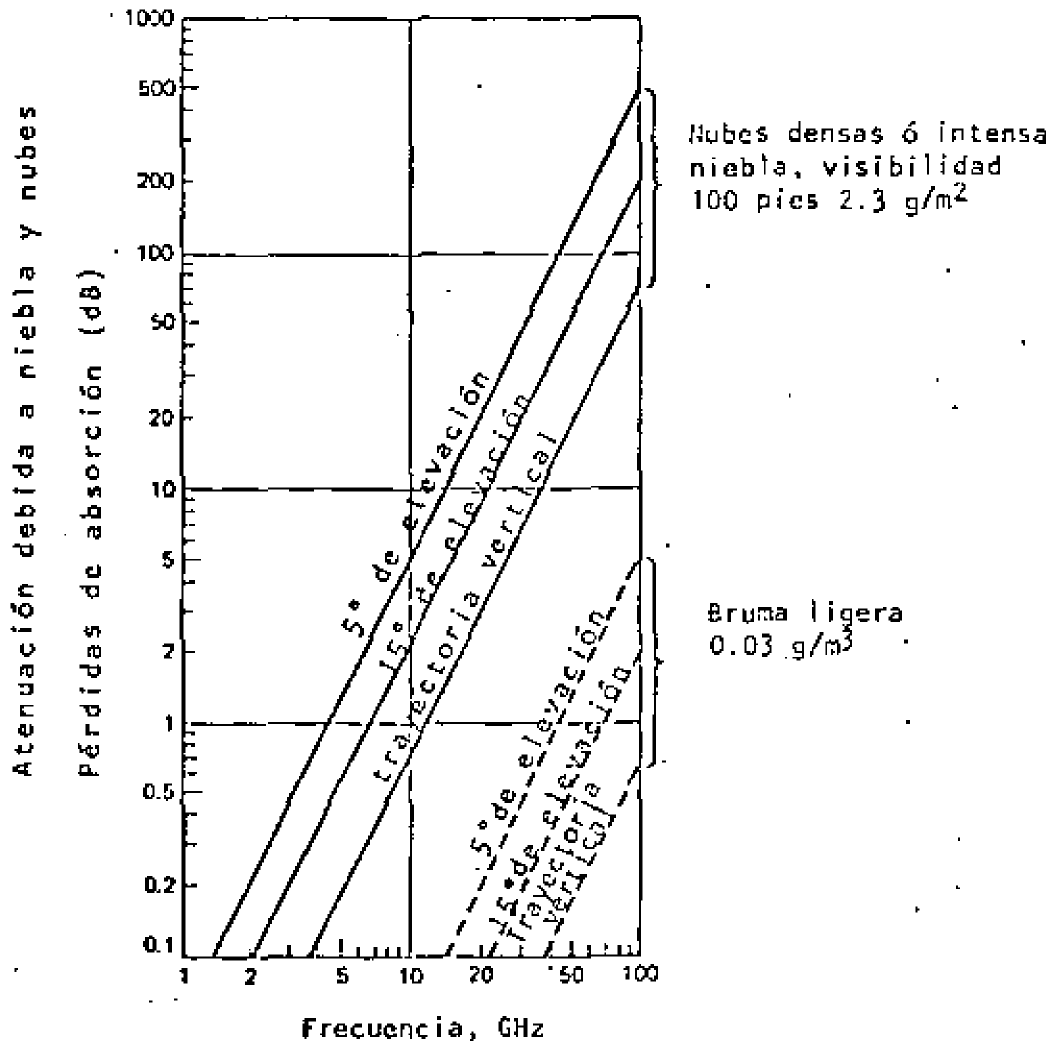


Fig. 32 Absorción en la atmósfera causada por electrones, oxígeno molecular y vapor de agua sin condensar.



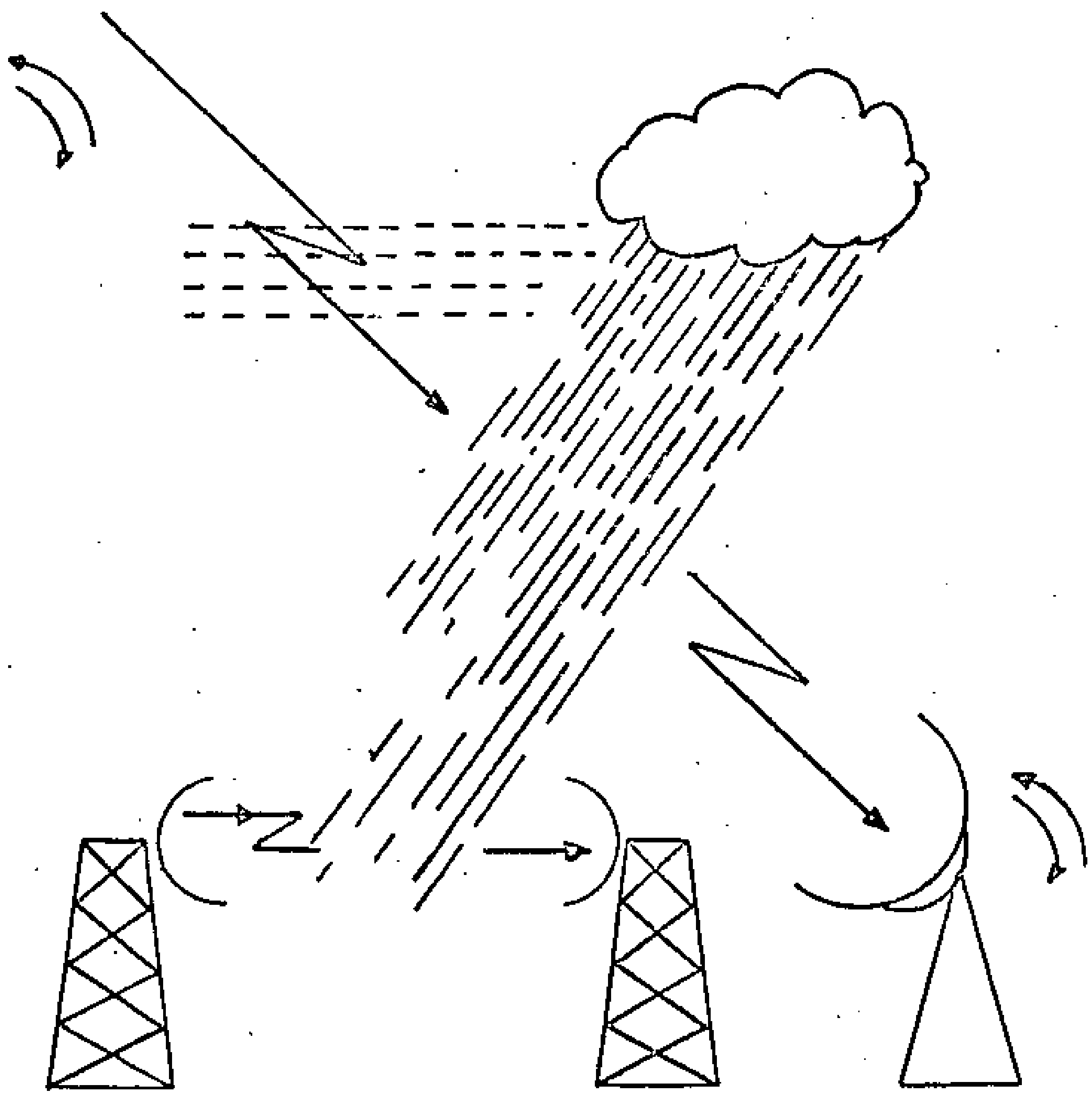
Típica absorción debida a niebla, bruma y nubes.

EFECTOS DE LA LLUVIA.

El mayor obstáculo encontrado en el diseño de sistemas de comunicación vía satélite a frecuencias arriba de 10 GHz es la atenuación por lluvia. La potencia de microondas radiada hacia una estación terrena, limitada por factores, tales como la potencia primaria disponible y el tamaño de la antena en el satélite, es insuficiente con la tecnología presente para contrarrestar la atenuación tan grande producida por intensa lluvia. Otras degradaciones producidas por lluvia, tales como depolarización, interferencia, aumento en el ruido de la estación terrena y deterioro en el desempeño de la estación terrena, se discuten también en esta parte.

Actualmente, las frecuencias de 4 y 6 GHz para satélites de comunicaciones son algo muy común de utilizarse, siendo importante tener cuidado en la calidad de las antenas de las estaciones terrenas, así como en la selección de sitios para evitar interferencia con los sistemas terrestres que comparten en estas mismas bandas de frecuencias. Sin embargo, la saturación a estas frecuencias es ya un hecho. Más aún, el ancho de banda designado a estas frecuencias es de 500 MHz, el cual se puede agotar muy pronto con el incremento de la demanda telefónica, de televisión y de datos. Para contrarrestar esta saturación, se propuso el uso de bandas de mayor frecuencia mayores de 2 GHz, tales como 19 y 29 GHz. Desafortunadamente, la lluvia es un factor mucho más serio a estas frecuencias que a 4 y 6 GHz y recientemente ha habido muchas reuniones internacionales sobre este asunto.

Antes de entrar a los detalles de los diversos efectos que tiene la lluvia en la propagación, deben recordarse



cuatro aspectos también a considerar: la nieve, la niebla, la tropósfera y la ionósfera. Cuando el agua se congela, como en el caso de las partículas en muchas nubes, la resonancia ocurre a longitudes de onda mayores. El resultado neto es que hielo y nieve seca presenta muy bajas pérdidas en la banda de microondas y por lo tanto no se considerará más en esta discusión.

Por otro lado, la niebla está, de seguro compuesta de pequeñas gotas, pero la densidad de agua líquida en intensa niebla es menos que $\frac{1}{20}$ con respecto a intensa lluvia tal que las atenuaciones encontradas son pequeñas y aquí serán despreciadas. Otro efecto de propagación, muy familiar para aquellos concernientes con sistemas terrestres, es el desvanecimiento de la señal causado por capas y otras aberraciones del contorno de refractividad de la tropósfera. Estas atenuaciones de la señal ocurren sobre haces de microondas que son esencialmente horizontales, interactuando de este modo a ángulos de rozamiento cercano con las capas. Pero para trayectorias típicas Tierra-satélite a ángulos de elevación de algunos grados, este tipo de desvanecimiento no es significante y no será discutido en detalle. Las trayectorias Tierra-satélite que operan a las frecuencias bajas en la banda de microondas, es decir a 4 GHz están sujetas a alguna rotación de polarización por la ionósfera vía el efecto de Faraday.

ATENUACION POR LLUVIA.

El decremento en la magnitud S del vector de Poynting al pasar a través de una capa de precipitación de grueso Δl es

$$-\Delta S = S\Delta l \int_0^\infty n(a)\, Q(a,\lambda)\, da$$

donde $Q(a,\lambda)$ es la sección transversal de extinción (centímetros cuadrados) de una gota esférica con radio a(centímetros) y n(a)da es el número de gotas por unidad de volumen (metros cúbicos) en el rango da. Integrando (I) se obtiene

$$S = So \exp (- \int \alpha \, dl)$$

donde

$$\alpha = \int_0^\infty n(a).Q(a,\lambda)\, da$$

la atenuación en decibeles por kilometros es simplemente 0.434α con los parámetros dados en las unidades indicadas anteriormente. Las secciones transversales de extinción Q pueden calcularse usando la solución de dispersión de Mie para esferas y los índices de refracción de agua líquida medidos por Saxton. En el cálculo de atenuación se utilizan las distribuciones de Law y Parson. Durante la segunda guerra mundial Ryde y Ryde llevaron a cabo cálculos de atenuación por lluvia en el rango de microondas, esto ha sido extendido por otros con la ayuda de computadoras modernas. Estos datos son mostrados en la Fig. 33.

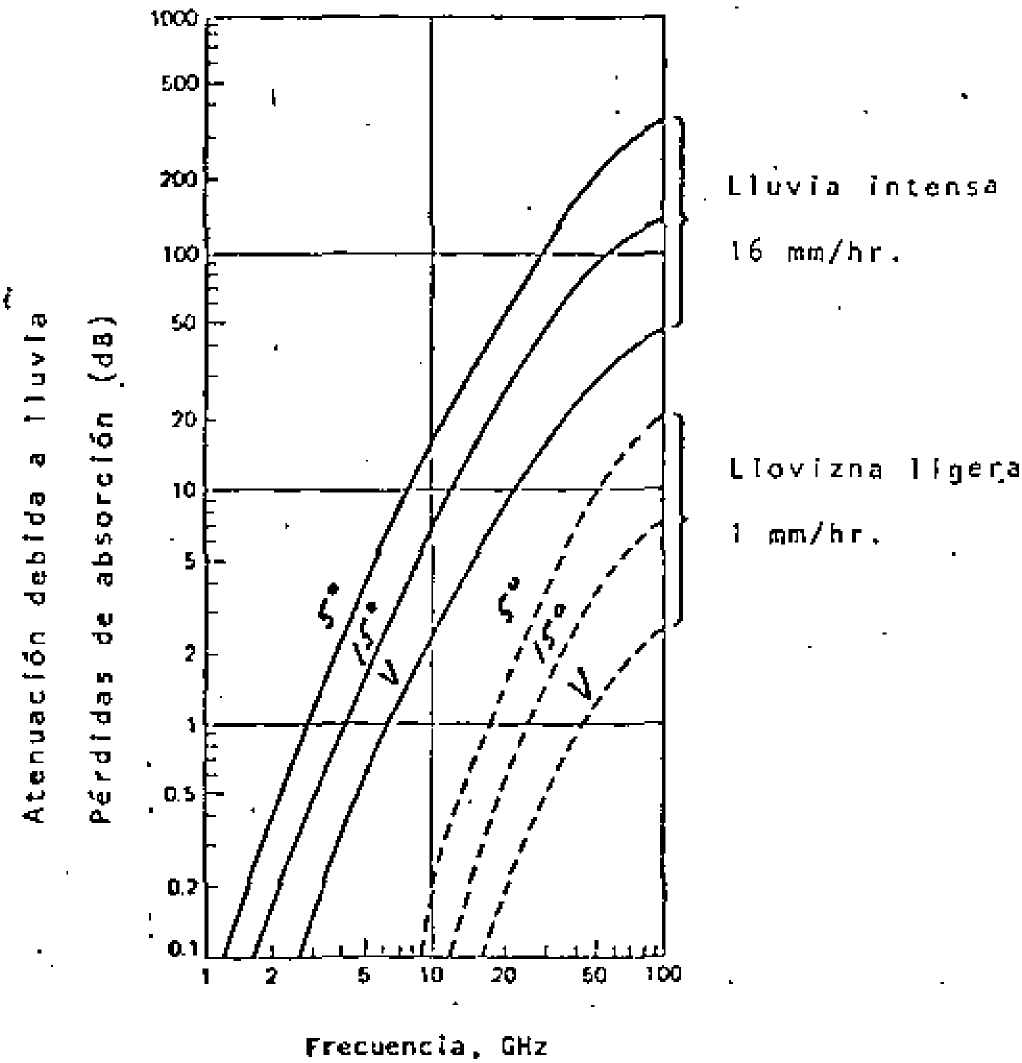
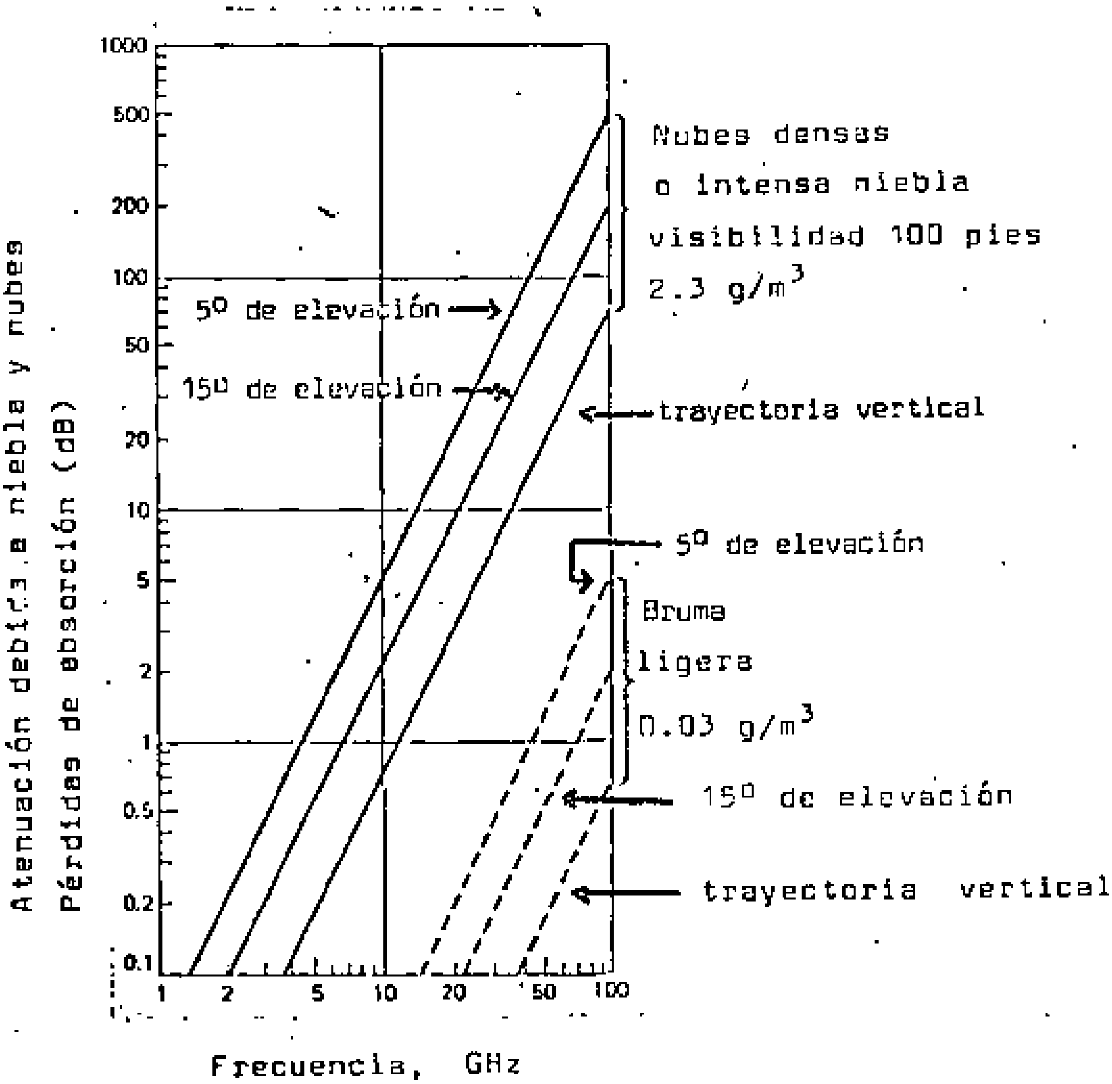


Fig. 33 Absorción típica debida a lluvia.



Típica absorción debida a niebla, bruma y nubes

ATENUACION DEPENDIENTE DE LA POLARIZACION Y LA DEPOLARIZACION.

La consideración de formas de gotas esféricas usada en el cálculo de Mie discutido en la sección anterior es solamente una aproximación de primer orden. Examinación por fotografías revelan que muchas de las grandes gotas son mejor representadas por esferoides achatados por los polos. --- Oguchi fué el primero en investigar el efecto de estas gotas en propagación de microondas usando cálculos de perturbación. La atenuación inducida por lluvia y el defasamiento obtenido son:

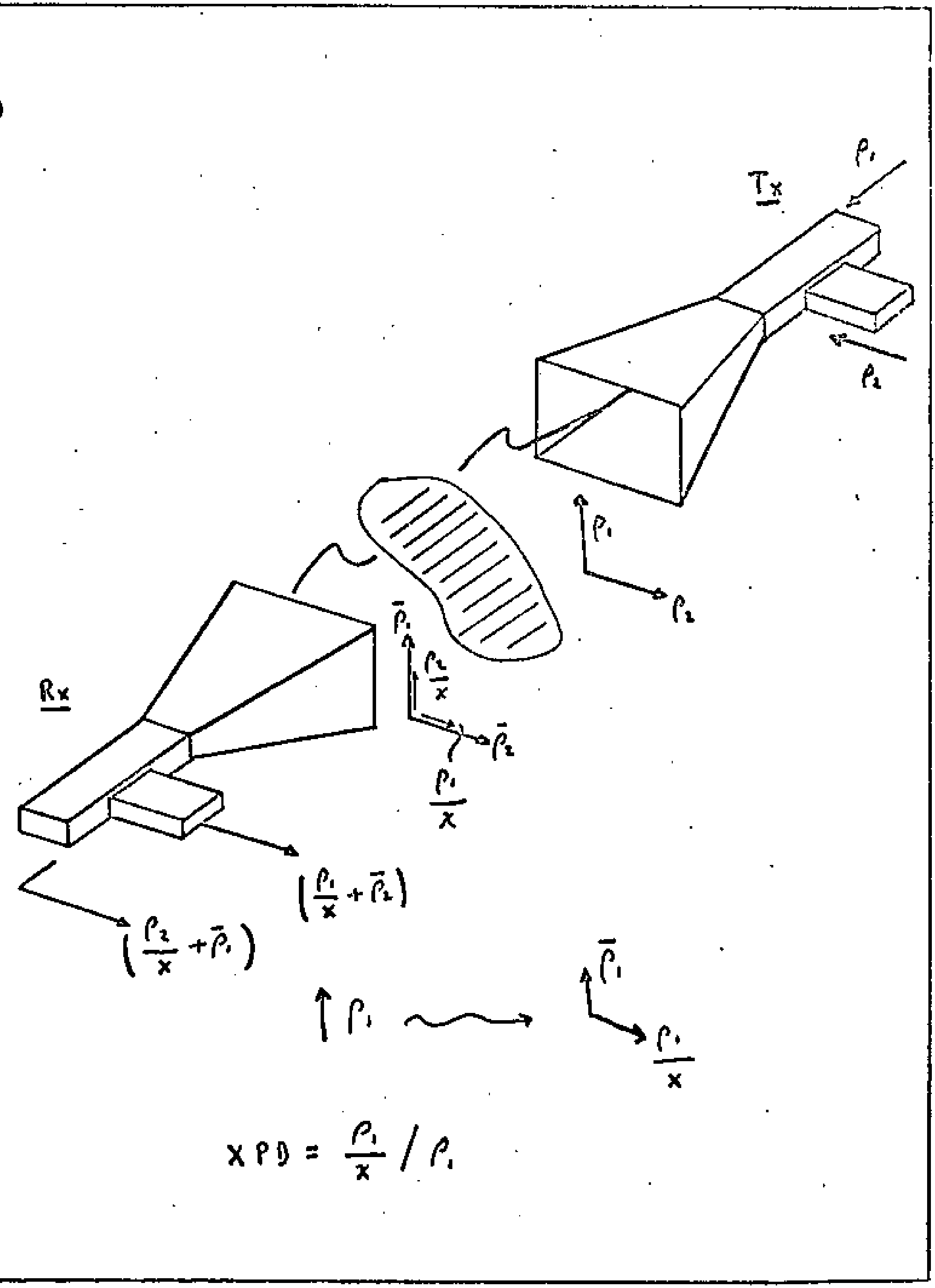
$$A_{I,II} = 0.434 \ \frac{\lambda^2}{\pi} \sum \text{Re } S_{I,II}(o) \ n(\bar{a}) \quad \text{dB/km.}$$

$$\phi_{I,II} = -36 \ \frac{\lambda^2}{\pi} \sum \ \text{Im } S_{I,II} \ (o) \ n \ (\bar{a}) \quad \text{gr/km.}$$

donde $n(\bar{a})$ es el número de gotas con radio esférico de -- igual volumen $\bar{a}$ por metro cúbico, los índices I y II designan los campos eléctricos paralelos y perpendiculares a los planos que contienen los ejes de simetría de las gotas y la dirección de propagación de la onda incidente y la sumatoria es tomada sobre todos los tamaños de las gotas.

La atenuación y fase diferencial entre las polarizaciones II y I para varios valores de precipitación y frecuencias entre 4 y 100 GHz se muestran en la Fig. 34.

La depolarización por gotas inclinadas es un resultado -- tanto de la atenuación diferencial como del defasamiento entre los componentes de las polarizaciones I y II. La


Tx
ρ1
ρ2
ρ1
ρ2
Rx
ρ̄1
ρ2/x
ρ̄2
ρ1/x
(ρ1/x + ρ̄2)
(ρ2/x + ρ̄1)
ρ1
ρ̄1
ρ1/x
XPD = (ρ1/x) / ρ1

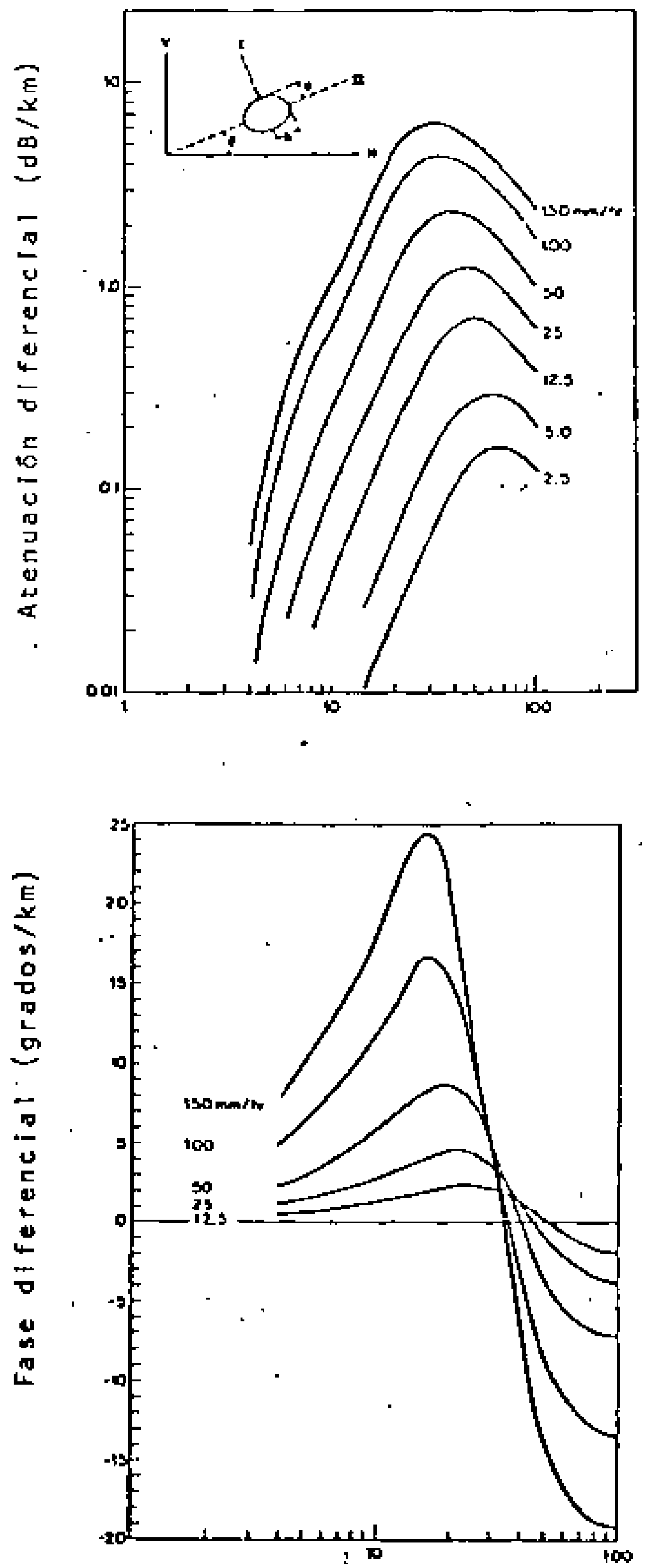


Fig. 34 Atenuación y fase diferencial entre polarizaciones I y II para varias intensidades de precipitación.

descriminación de polarización cruzada, medida para polarización horizontal a 11, 17.71 y 60 GHz se grafica en la Fig. 35.

Debido a que la atenuación por lluvia para polarización vertical es menor que para horizontal, la descriminación de polarización cruzada a una atenuación dada para una polarización vertical es mejor que para polarización horizontal, como ha sido confirmado experimentalmente. La depolarización medida de una onda polarizada linealmente orientada a 45° con respecto a la dirección vertical y de una onda polarizada circularmente, se ha encontrado que es muy severa como se muestra en la Fig. 36.

Predicciones teóricas de polarización cruzada inducida por lluvia se ven impedidas por incertidumbre en la distribución del ángulo de inclinación de las gotas. Se encontró por medio de fotografías de gotas de lluvia que el ángulo de inclinación no esta lejos de una distribución igual a la dirección vertical (gravedad). En el caso de una onda polarizada horizontal o verticalmente, la polarización cruzada producida por ángulos de inclinación positivos ó negativos tiende a cancelarse. Pero se ha encontrado que predicciones sistemáticas puedenobtenerse al comparar atenuaciones diferenciales medidas y polarizaciones cruzadas a una frecuencia, con valores calculados, para determinar dos parámetros empíricos: Un promedio efectivo del valor absoluto del ángulo de inclinación y el desequilibrio en el número de gotas con ángulos de inclinación positivos y negativos. Tales medios empíricos se muestran en la Fig. 37 para la polarización cruzada de ondas polarizadas horizontalmente a varias frecuencias. Comparando las Figs. 35 y 37 se muestra una clara concordancia para frecuencias abajo de 30 GHz.

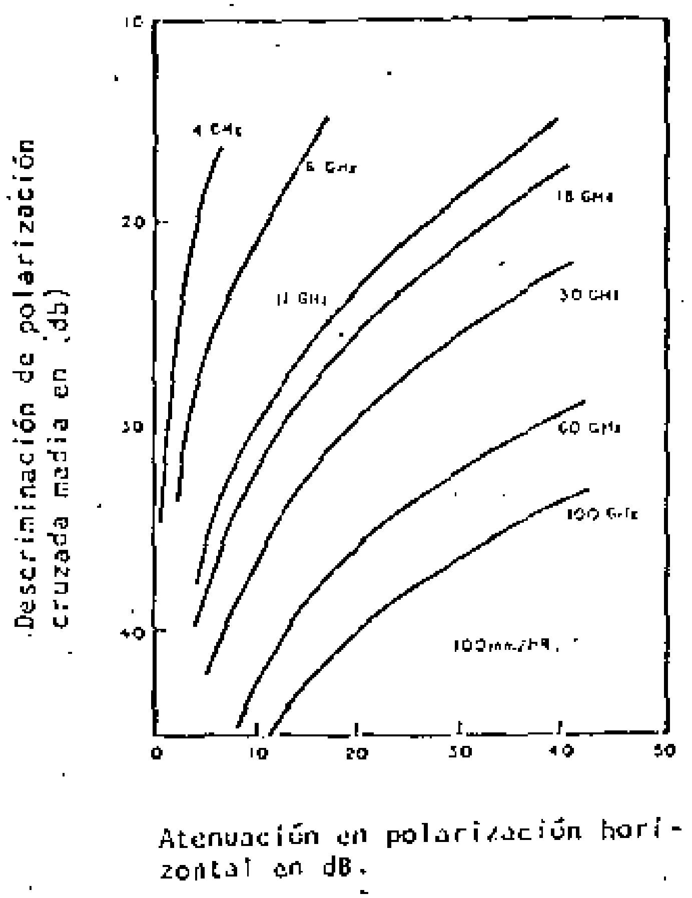

Descriminación de polarización cruzada media en (db)
4 GHz
6 GHz
11 GHz
18 GHz
30 GHz
60 GHz
100 GHz
100mm/HR
10
20
30
40
0
10
20
30
40
50
Atenuación en polarización horizontal en dB.

Fig. 37

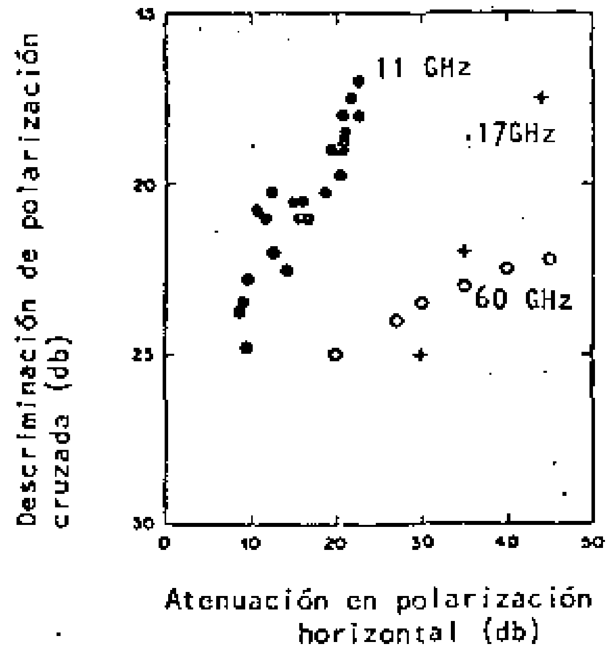


Fig. 35 Polarización cruzada inducido por lluvia para polarización horizontal.

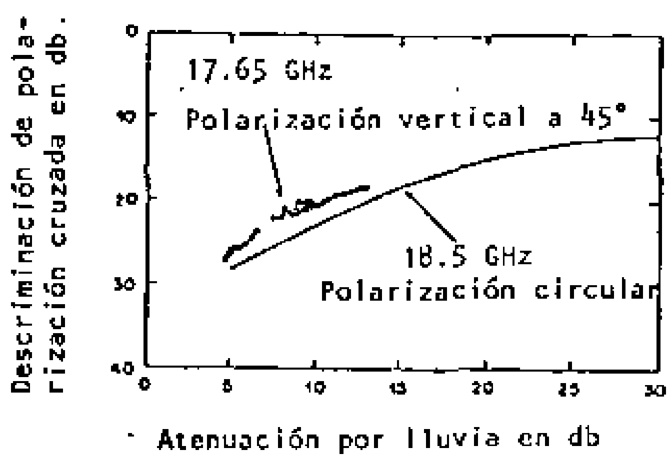


Fig. 36 Depolarización medida contra atenuación para polarización circular y polarización lineal a 45°.

DEGRADACION DEL RENDIMIENTO DE LA ANTENA POR LLUVIA.

Un análisis de la caída de lluvia sobre una cubierta (Radome) hemisférica, mostró que se forma una capa de lluvia de espesor constante sobre la cubierta. Una onda electromagnética que incide sobre una capa de agua, experimenta tanto pérdidas de absorción como de reflexión, la degradación total resultante en la transmisión se muestra en la Fig. 38, como una función del espesor para muchas frecuencias de interés. Por ejemplo a 18.5 GHz, se introduce una atenuación de 10 dB para una capa de agua de un cuarto de milímetro de espesor. Por lo tanto si capas de ese espesor se forman por la caída de lluvia en una cubierta, el diseñador del sistema se encuentra con otra atenuación del mismo orden que la producida por lluvia en la trayectoria de propagación. Desafortunadamente, se sabe poco respecto al espesor de las capas formadas para un dado índice de lluvia, ya que depende de la geometría y las propiedades de fricción y humedad de la superficie. Si no se coloca la cubierta en la antena, se introduce pequeña atenuación por las capas de lluvia en la superficie reflectora en cambio lluvia y nieve producen polarización cruzada.

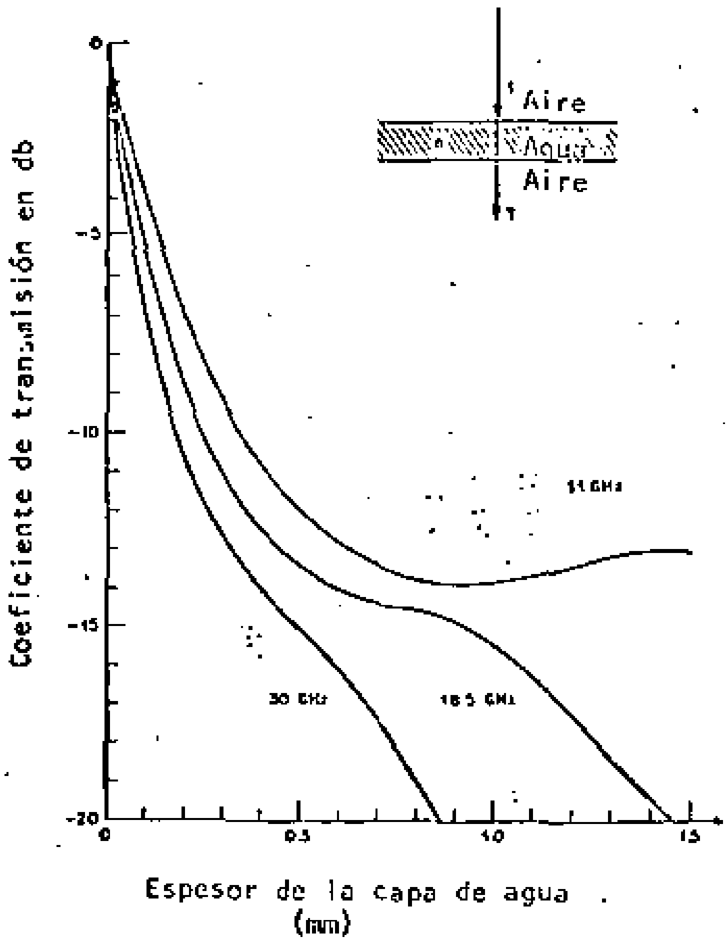

Aire
Agua
Aire
0
-5
-10
-15
-20
0
0.5
1.0
1.5
11 GHz
30 GHz
18.5 GHz
Coeficiente de transmisión en db
Espesor de la capa de agua
(mm)

Fig. 38

RUIDO.

Temperatura de Ruido. La potencia de ruido es usualmente cuantificada en términos de su temperatura de ruido. Si el equipo electrónico estuviera perfectamente aislado de interferencias externas, de todos modos habría ruido en dicho equipo, debido al movimiento aleatorio de los electrones. Este ruido es llamado ruido térmico.

La potencia del ruido térmico que afecta un rango dado de frecuencias es proporcional a la temperatura absoluta y al ancho de banda de frecuencias en cuestión, es decir:

$$P_r = KTB$$

donde P_r = Potencia del ruido en watts

k = Constante de Boltzman 1.3×10^{-23} watts seg/°k

T = Temperatura en °k

B = Ancho de banda en Hertz.

La temperatura de ruido de una fuente de ruido es la temperatura que produce la misma potencia de ruido sobre el mismo rango de frecuencias.

Así, si una fuente de ruido crea ruido de potencia Pr, su temperatura de ruido, algunas veces llamada temperatura de ruido equivalente, ENT, es

$$T = \frac{Pr}{kB}$$

Densidad de Ruido.

El término densidad de ruido se refiere al ruido por Hertz de ancho de banda:

$$\text{densidad de ruido} = \frac{Pr}{B} = KT$$

Relación Portadora a Ruido.

Una relación frecuentemente usada para establecer la calidad de un satélite es:

$$\frac{\text{Potencia de la portadora recibida}}{\text{densidad de ruido}} = \frac{P_R}{KT}$$

La potencia de la portadora se simboliza frecuentemente con C. La anterior relación $\frac{C}{N}$ es llamada la relación señal a ruido. En la Fig. 39 se grafica la relación $\frac{C}{N}$ contra el EIRP para un enlace de subida de un típico satélite doméstico Norte-Americano. Se puede ver que dicha relación no puede ser mejorada hasta algún cierto nivel debido a que se alcanza una saturación en el canal.

Fuentes externas de ruido.

Las siguientes son fuentes externas de ruido: El sol, la luna, la tierra, ruido galáctico, ruido cósmico, ruido del cielo, ruido atmosférico y ruido hecho por el hombre. Estas fuentes difieren en su intensidad, frecuencias y localización en el espacio.

Si la antena de un satélite apunta hacia el sol, la señal será prácticamente contaminada debido a la temperatura de ruido del sol que es de 100,000°k ó más.

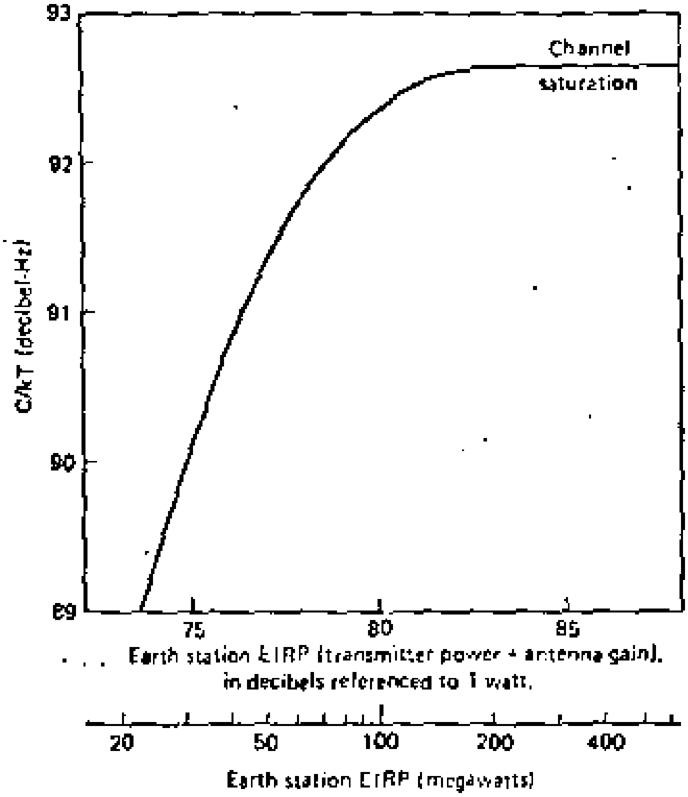


Figure 7.1 C/kT for a typical American domestic satellite.

Fig. 39

El ruido del sol varía con la actividad solar.

La temperatura del ruido del cielo es de aproximadamente 30°K. La directividad de una antena no sólo es para enfocar el haz sino también para proteger la señal recibida, de otras fuentes de ruido.

La temperatura de ruido de la tierra, vista desde el espacio, es en promedio de 254°K. Una antena de satélite con un ancho de haz igual al ancho proyectado de la tierra recibiría esta cantidad de ruido, como fondo a las señales que vienen desde la tierra. Debido a las variaciones del terreno, haces dirigidos a alguna porción de la tierra reciben una temperatura de ruido ligeramente mayor a 254°K.

El ruido galáctico se refiere al ruido de las estrellas en la galaxia. Este ruido decrece rápidamente a altas frecuencias y tiene efectos despreciables arriba de 1 GHz.

El ruido cósmico se refiere a otro ruido del espacio exterior y también es despreciable a frecuencias arriba de 1 GHz.

Los destellos de luz y las descargas electrostáticas en la atmósfera son una fuente mayor de ruido abajo de 30 MHz. Afortunadamente son despreciables a las frecuencias utilizadas en los satélites.

El ruido atmosférico se origina principalmente de las moléculas de oxígeno y vapor de agua, las cuales absorven la radiación. Consecuentemente las frecuencias en las cuales la absorción atmosférica es alta son las mismas en las que el ruido atmosférico es alto. La Fig. 40 muestra las temperaturas de ruido de vapor de agua y oxígeno atmosférico. El ruido hecho por el hombre, el cual es una plaga a frecuencias bajas, tiene un efecto pequeño arriba de 1 GHz. Surge principalmente de la maquinaria eléctrica y es mucho mayor en áreas industriales. Si estuviera pre-

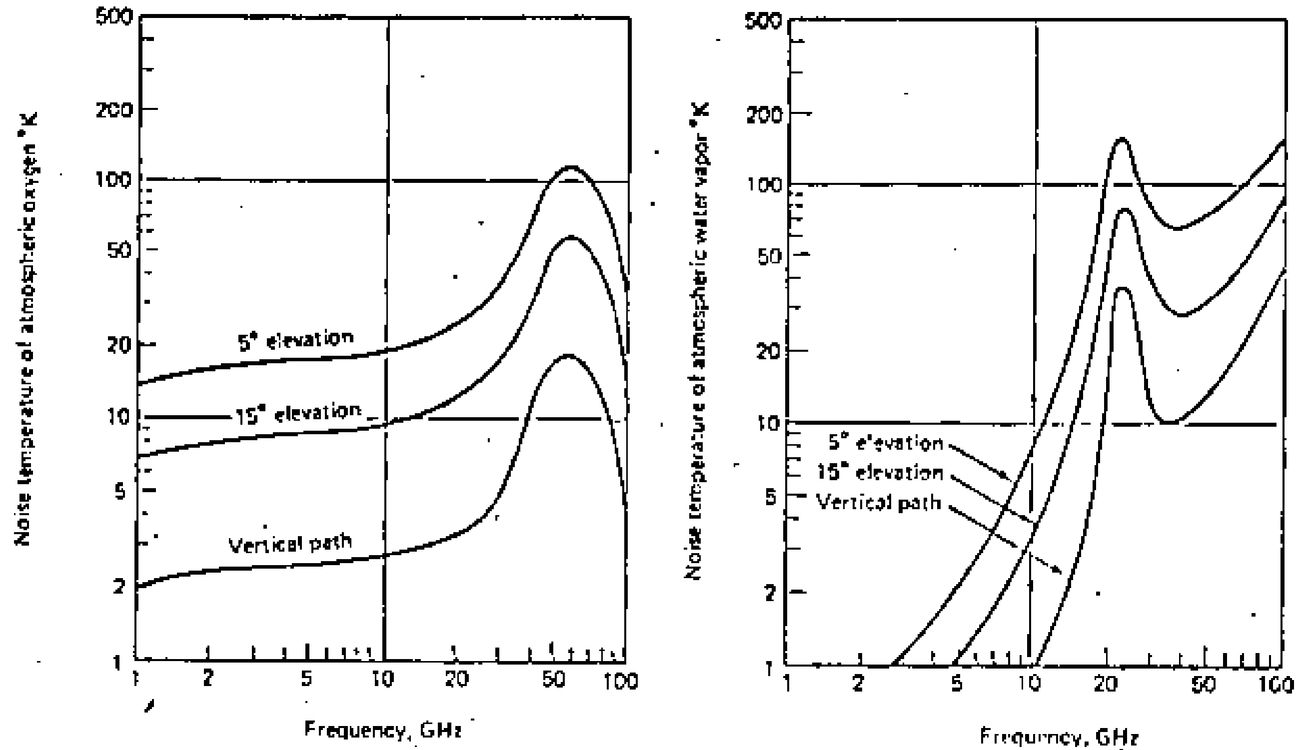

Noise temperature of atmospheric oxygen °K
5° elevation
15° elevation
Vertical path
Frequency, GHz
Noise temperature of atmospheric water vapor °K
5° elevation
15° elevation
Vertical path
Frequency, GHz

Fig. 40

sente podría reducirse al cubrirse la antena. Está virtualmente ausente en el espacio.

La Fig. 41 muestra cómo afecta a la señal la combinación de estos diferentes tipos de ruido. El ruido recibido en el satélite, que domina, es la temperatura de ruido de la Tierra. En la estación terrena hay una ventana de ruido entre el efecto del ruido cósmico y el efecto del vapor de agua.

Mal tiempo.

La lluvia muy intensa causa más ruido en la estación terrena que todas las otras fuentes de ruido combinadas. Como en el caso de la absorción, este efecto es peor a frecuencias mayores. La Fig. 42 muestra el efecto de la lluvia, nubes y niebla intensas. Es recomendable evitar tanto como sea posible ángulos de elevación bajos a estas frecuencias.

Figura de Mérito.

Debido a que la señal recibida es muy débil, tanto en el satélite como en la estación terrena, es importante que la antena receptora y la parte electrónica introduzcan tan poco ruido como sea posible. Para evitar pérdidas y ruido en las líneas que conectan la antena receptora a la electrónica, la antena tiene usualmente el preamplificador construido internamente como se muestra en la Fig. 43 . La eficiencia de tal combinación usualmente se cuantifica como la relación de la ganancia a la temperatura de ruido y se llama la figura de mérito

$$\text{Figura de Mérito} = \frac{G}{T}$$

donde G = antena y ganancia del preamplificador.

T = temperatura de ruido en el sistema receptor.

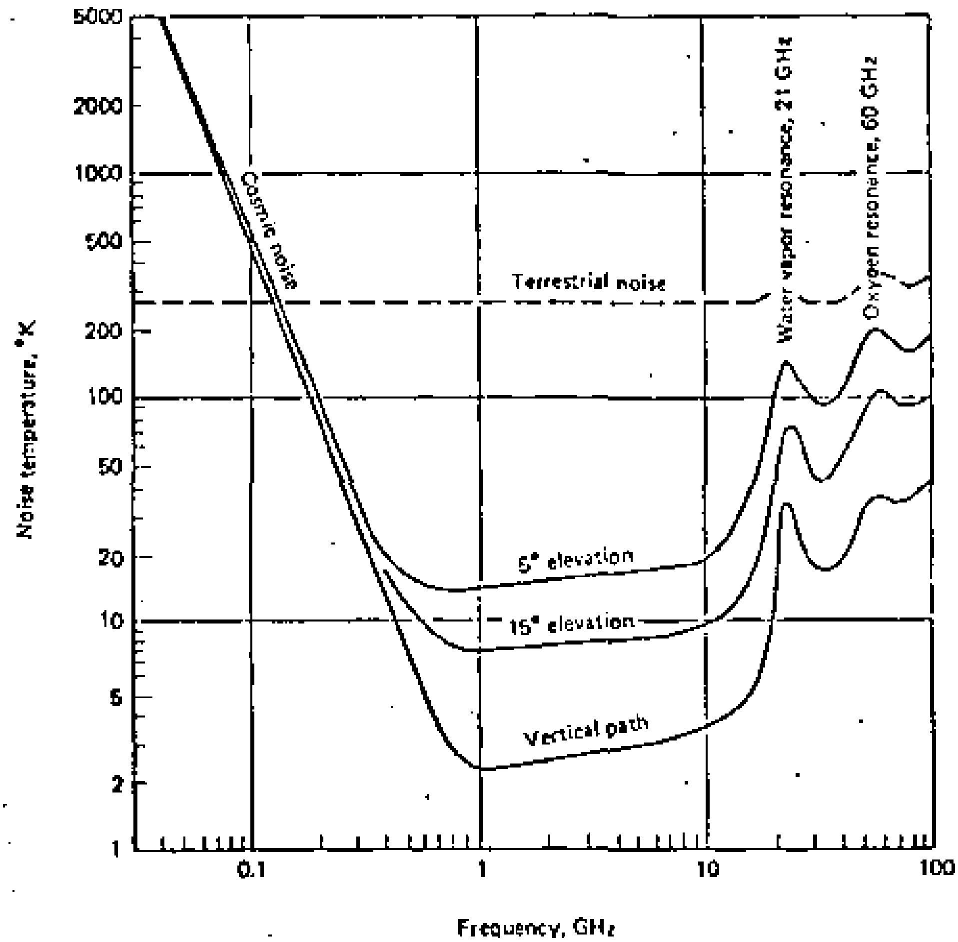

5000
2000
1000
500
200
100
50
20
10
5
2
1
0.1
1
10
100
Frequency, GHz
Noise temperature, °K
Cosmic noise
Terrestrial noise
Water vapor resonance, 21 GHz
Oxygen resonance, 60 GHz
5° elevation
15° elevation
Vertical path

Fig. 41

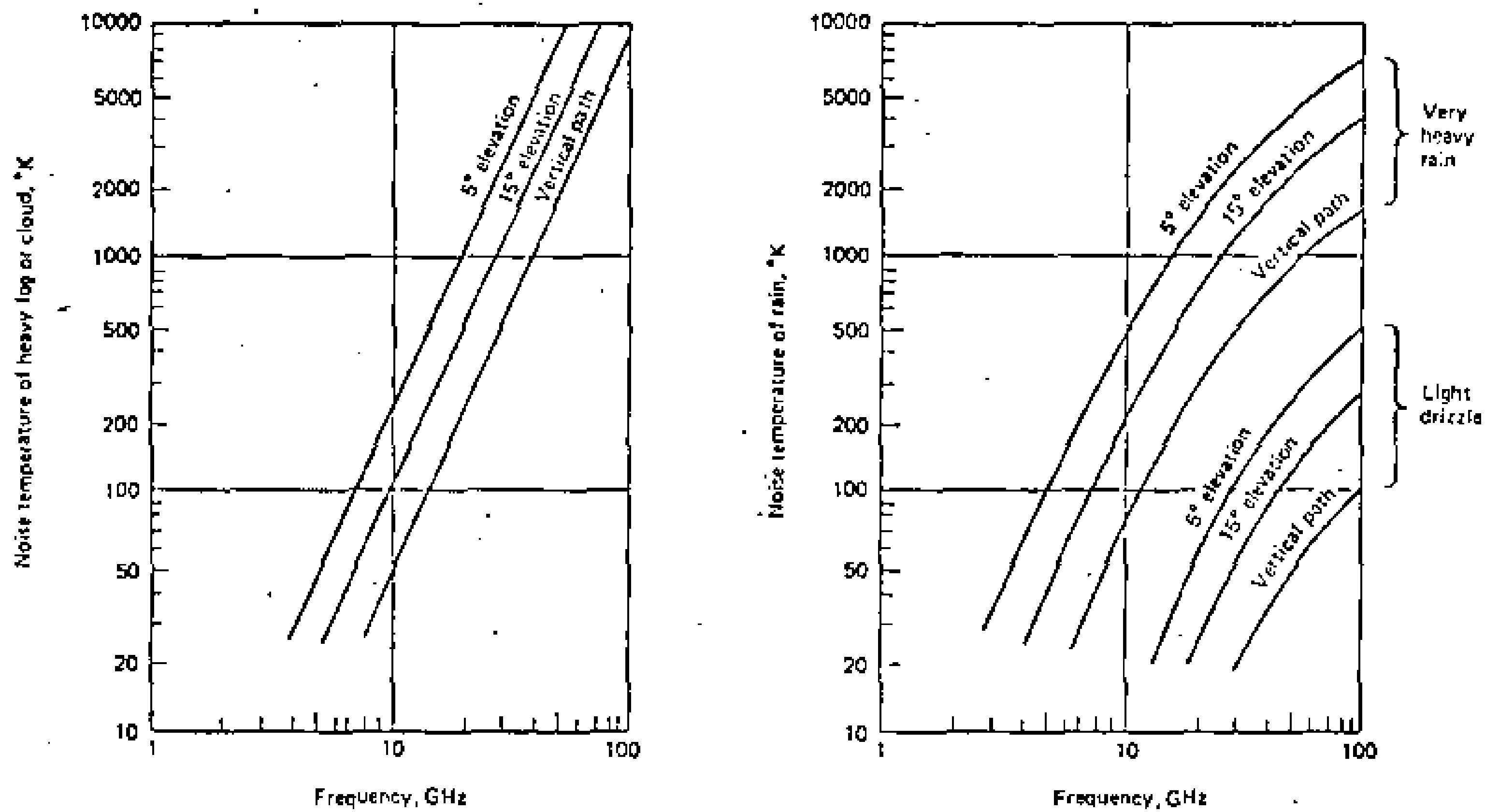

Noise temperature of heavy fog or cloud, °K
10000
5000
2000
1000
500
200
100
50
20
10
1
10
100
Frequency, GHz
5° elevation
15° elevation
Vertical path
Noise temperature of rain, °K
Very heavy rain
Light drizzle

Fig. 42

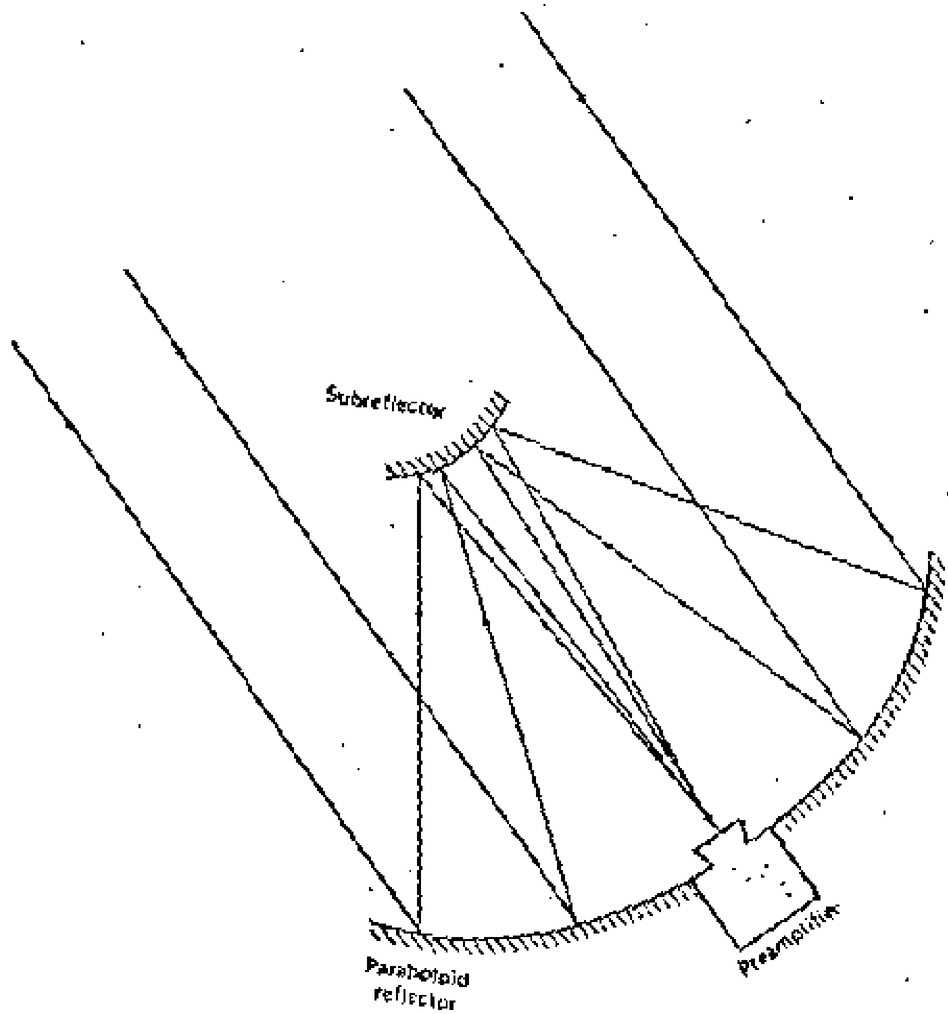


Fig. 43

Esta figura de mérito se relaciona a la relación señal a ruido resultante y por consiguiente indica la capacidad relativa del subsistema receptor para recibir una señal. La Fig.44 grafica algunos valores típicos para receptores con electrónica sin enfriamiento.

Ruido del equipo.

La temperatura de ruido T del equipo receptor es originado tanto por la estructura de la antena como por la electrónica asociada. Las primeras estaciones terrenas usaron preamplificadores enfriados criogénicamente para reducir la temperatura de ruido. Actualmente, con satélites más potentes, se puede usar equipo receptor más barato con una figura de merito menor, teniendo tanto una antena más pequeña como una temperatura de ruido mayor. Así que con una potencia de satélite mayor se puede tener equipo receptor más barato. La Tabla I muestra algunas figuras de temperatura de ruido típicas para diferentes tipos de electrónica. La relación señal a ruido resultante se calcula asumiendo que la temperatura de ruido de antena es 60°K y entra al amplificador una señal de 10 picowatts.

Una relación señal a ruido de 10 dB es típica en enlaces de satélites, mientras que una relación señal a ruido de 30 dB es adecuada para enlaces terrestres. Se pueden insertar estos valores en la ecuación de Shannon y comparar los valores teóricos entre un - satélite típico y un enlace terrestre del mismo ancho de banda.

Las componentes de ruido incluídas en T pueden dividirse en 4 categorías:

- RUIDO DE ANTENA
- RUIDO DE COMPONENTE PASIVA
- RUIDO DE ESCAPE (HPA)
- ETAPAS DE AMPLIFICACION

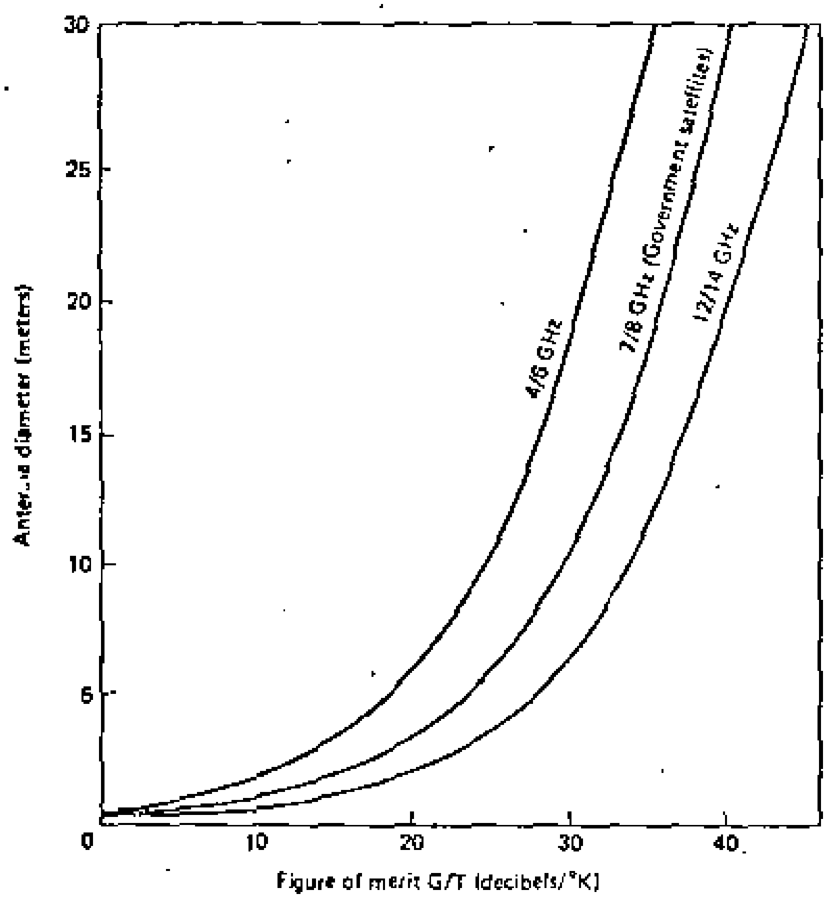

30
25
20
15
10
5
0
10
20
30
40
Anter..a diameter (meters)
Figure of merit G/T (decibels/°K)
4/6 GHz
7/8 GHz (Government satellites)
12/14 GHz

Fig. 44

Tabla 7.1 Noise figures for typical types of receiver equipment and typical resulting signal-to-noise ratios.

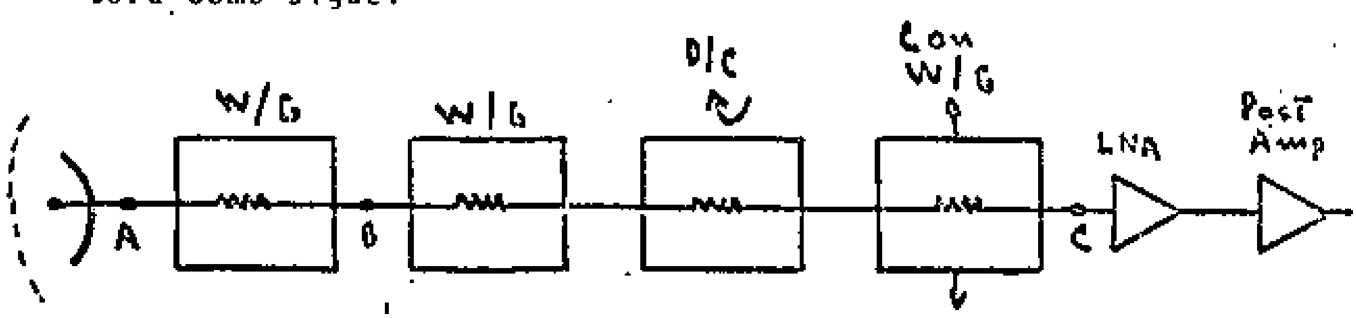
Type of Electronics	Noise Temperature of Electronics	Typical Noise Temperature of Antenna	Combined Noise Temperature of Antenna & Electronics	S/N for a Received Signal of 10 picowatts (10^{-11} watts)
Maser (cooled to 4.2°K)	10	60	70	14.6 decibels
Parametric amplifier (cooled to 25°K)	35	60	95	13.3 decibels
Uncooled parametric amplifier	120	60	180	10.5 decibels
Inexpensive parametric amplifier	300	60	360	7.5 decibels
Tunnel diode amplifier	530	60	590	5.3 decibels
Schottky mixer	1000	60	1060	2.7 decibels

	A 4/6-GHz Link. Satellite Antenna: Earth Coverage. Earth Antenna: 12 meters. Moderately Low-Cost Electronics in Earth Station.	A 12/14-GHz Link. Satellite Antenna: 1.8 Meters. Earth Antenna: 1.8 Meters. Low-Cost Earth-Station Receiver.	A 20/30-GHz Link for Common Carrier Use. Satellite Antenna: 2 Meters. Earth Antenna: 27.5 Meters. Cryogenically Cooled Receiver.	A 12/14-GHz Link on a Broadcast Satellite. Satellite Antenna: 9 Meters. Earth Antenna: 1.8 Meter Receiver-only; 12-Meter Transmit. Low-Cost Earth Station.
Up-Link				
Transmitter power, dBw*	35	25	20	20
Transmitter system loss, decibels	−1	−1	−1	−1
Transmitting antenna gain, decibels	55	46	76	62
Atmospheric loss, decibels	0	−0.5	−2	−0.5
Free space loss, decibels	−200	−208	−214	−208
Receiving antenna gain, decibels	20	46	53	60
Receiver system loss, decibels	−1	−1	−1	−1
Received power, dBw*	−92	−93.5	−69	−68.5
Noise temperature, °K	1000	1000	1000	1000
Received bandwidth, MHz	36	36	350	36
Noise, dBw*	−128	−128	−118	−128
Received SNR, decibels	36	34.5	49	59.5
Loss in bad storm, decibels	2	10	25	10
Received SNR in bad storm, decibels	34	24.5	24	49.5
Down-link				
Transmitter power, dBw*	18	20	8	10
Transmitter system loss, decibels	−1	−1	−1	−1
Transmitting antenna gain, decibels	16	44	49	58
Free space loss, decibels	−197	−206	−210	−206
Atmospheric loss, decibels	0	−0.6	−2	−0.6
Receiver antenna gain, decibels	51	44	72	44
Receiver system loss, decibels	−1	−1	−1	−1
Received power, dBw*	−114	−100.6	−85	−96.6
Noise temperature, °K	250	1000	250	1000
Received bandwidth, MHz	36	36	350	36
Noise, dBw*	−131	−128	−121	−128
Received SNR, decibels	17	27.4	36	31.4
Loss in bad storm, decibels	2	10	25	10
Received SNR in bad storm, decibels	15	17.4	11	21.4

*dBw means decibels referenced to one watt. I. E. 1 watt = 0 dBw; 100 watts = 2 dBw, etc.

La Fig. 45 representa las contribuciones de ruido gráficamente. La Fig. 46 muestra aproximadamente la variación de ruido del cielo con el ángulo de elevación. A un ángulo de 5°vemos que la temperatura de ruido del cielo alcanza el orden de 25°K. Será visto también que el ruido de antena mínimo ocurre cuando la antena está en el Zenith (es decir, un ángulo de elevación de 90°). Los ángulos de elevación son con respecto al horizonte; así, el ángulo de elevación sería de 0°cuando la antena apunta directamente al horizonte. El derramamiento de antena se refiere a la energía radiada de la antena al suelo y dispersada por los elementos metálicos que sostienen los dispositivos de alimentación. La suma total del ruido de antena puede alcanzar 39 ó 40 K, 25 de los cuales es ruido del cielo.

Para calcular toda la temperatura del ruido del sistema Tsys de los diversos elementos en Tandem se hace uso de la cadena receptora como sigue:



donde A, B y C son puntos de referencia o planos de referencia. A es la base del punto radiador, B es la base del pedestal de la antena y C es el punto de entrada al amplificador de bajo ruido (LNA).

Para calcular la temperatura de ruido del sistema Tsis, se puede decir que:

$$Tsis = Tant + Tr$$

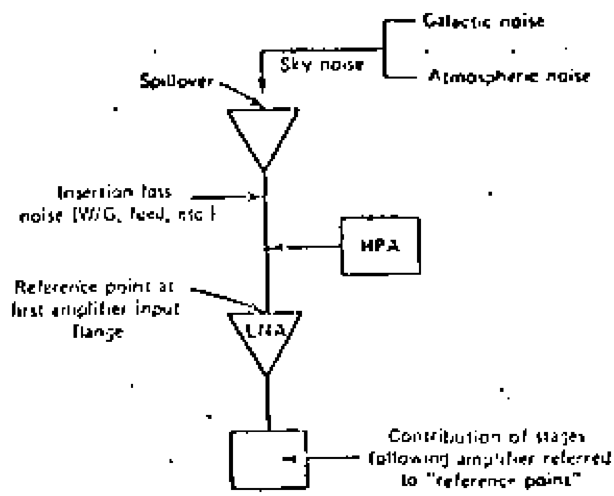
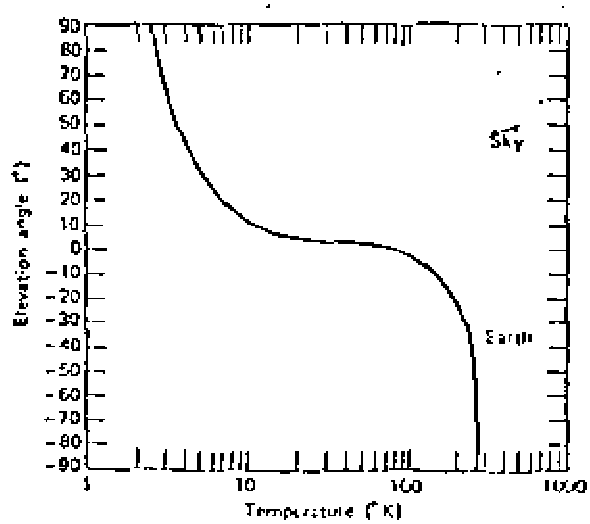


Figure 7.5 Graphical representation of noise contributors. HPA = high-power amplifier; LNA = low-noise amplifier.

Fig. 45



Approximate sky noise variation with antenna elevation angle (4 GHz).

Fig. 46

donde: Tant = Temperatura de ruido de la antena

Tr = Temperatura de ruido del sistema receptor.

El primer paso es establecer un punto de referencia. Este es un punto arbitrario desde donde se calcula la ganancia de la antena también como su temperatura de ruido Tant. La Tsis variará conforme G varíe, dependiendo del punto de referencia. Encontraremos que conforme el punto de referencia se mueva del alimentador de la antena, la ganancia disminuirá y así también la temperatura de ruido. Sin embargo, la G/T para un sistema dado se mantendrá constante, sin importar la referencia. En la anterior figura, cada componente de pérdidas óhmicas es un generador de ruido, como lo es cada componente activa como el LNA y el post-amplificador, el mixer, los amplificadores de FI y así sucesivamente. Las contribuciones de ruido a la izquierda del plano de referencia están incluidas en la temperatura de antena (Tant) en la ecuación anterior y siempre incluye el ruido del cielo. A la derecha del plano de referencia, esto es, hacia el sistema, todas las contribuciones de ruido se incluyen en Tr.

Para diferenciar entre pérdidas óhmicas y no óhmicas, considere que todos los dispositivos con una pérdida de inserción están en la categoría óhmica y todas las pérdidas no asociadas con una pérdida de inserción son no óhmicas. Un ejemplo de pérdidas no óhmicas es el espacio libre.

El análisis para determinar Tsis es una operación de dos pasos, es decir Tant y Tr se calculan separadamente y entonces se realiza la suma.

Cuando se calcula la contribución de ruido de una pérdida -- ohmica, la cual está dada en las unidades tradicionales de medición, el decibel, debemos convertir el valor del decibel a su relación numérica equivalente:

$$\text{Pérdidas (dB)} = 10 \log_{10} \left(\frac{P_1}{P_2} \right)$$

sea $\frac{P_1}{P_2} = L$. Entonces:

$$\text{Pérdidas(Relación)} = \log_{10}^{-1} \left(\frac{L}{10} \right)$$

Supongamos que el punto de referencia fuera el punto B, hubiera una pérdida de cubierta de 1 dB y las pérdidas de la guía de onda a la base del pedestal fueran 1.3 dB. Calcular la relación de pérdidas,

$$\text{Pérdidas(relación)} = \log^{-1} \left(\frac{2.3}{10}\right)$$

$$= 1\ 698$$

Asumiendo que L_T sean las pérdidas totales de la red de la antena, incluyendo la cubierta (radome), expresada como una relación de pérdidas. Entonces:

$$Tant = \frac{(L_T - 1)\ (Tamb + Ts}{L_T}$$

donde T_s = ruido del cielo y Tamb = temperatura ambiente, tradicionalmente dada como 290 K (17°C).

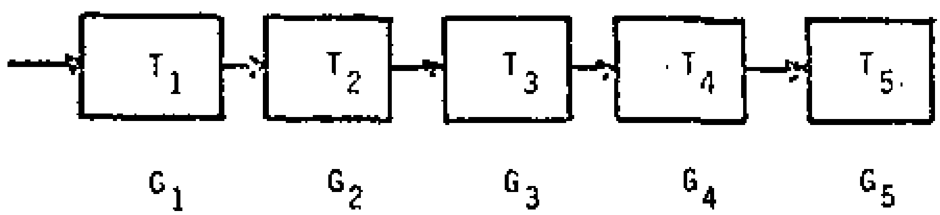
La temperatura de ruido del receptor Tr es el ruido total recibido obtenido al referirnos a los efectos de la contribución del LNA(y subsecuentes amplificadores o mezcladores) y las pérdidas del circuito de entrada al mismo plano de referencia como en el caso de la temperatura de ruido de antena.

En otras palabras, estamos tratando con todas las contribuciones de ruido a la derecha del plano arbitrario de referencia.

Cuando se calcula Tr se debe utilizar la fórmula de cascada tradicional para temperatura de ruido:

$$T_r = T_1 + \frac{T_2}{G_1} + \frac{T_3}{G_1 G_2} + \frac{T_4}{G_1 G_2 G_3} + \frac{T_5}{G_1 G_2 G_3 G_4} + \cdots$$

donde T_1 = Temperatura de ruido del contribuidor de ruido n y G_n = ganancia del contribuidor n, n = 1, 2, ...



Recuerde que las pérdidas de un dispositivo que atenúa una señal pueden ser expresadas como una ganancia equivalente, la cual es menor que 1.

La temperatura de ruido en el receptor Tr se expresa como:

$$T_r = (Li - 1)\ Tamb + T_{LNA}\ Li + \frac{T_{pa}\ L_i}{G_{LNA}} + \ldots$$

donde L_i = suma de las pérdidas desde el plano de referencia a la entrada del LNA, donde estas pérdidas se expresan como una relación, T_{LNA} = Temperatura de ruido en grados Kelvin del LNA, G_{LNA} = ganancia del LNA, y T_{pa} = temperatura de ruido en grados Kelvin del postamplificador, donde sea necesario, o bien del mezclador.

Ejemplo.

Dado un ruido del cielo de 50°K, pérdidas en guía de onda de 0.2dB a la base del pedestal de antena, y otras pérdidas de inserción de conmutación de guía de onda de 0.07 dB, T_{LNA} = 105k, G_{LNA} = 30dB y T_{pa} = 600 k.

1. ¿Cuál es el valor de Tsis cuando el plano de referencia es la base del pedestal de la antena?.

2. Si la ganancia de la antena es de 40dB, ¿cuál es la relación G/T usando el mismo plano de referencia como en el punto 1?.

Tant

Suma de las pérdidas óhmicas al plano de referencia

$$L_T = \log_{10}^{-1} \left(\frac{0.2}{10} \right)$$

$$= 1.047$$

considérese Tamb = 290 K, entonces

$$Tamb = \frac{(L_T - 1)\ Tamb + T_S}{L_T}$$

$$= \frac{(1.047-1)290+50}{1.047} = \frac{13.63 + 50}{1.047} = 60.77K$$

Tr

Suma de las pérdidas desde el plano de referencia a la entrada del LNA:

Pérdidas en guía de onda	0.015dB
Pérdidas en acoplador direccional	0.09
Conmutación de guía de onda	0.07
Total	0.175dB

entonces

$$L_i = \log_{10}^{-1}\left(\frac{0.175}{10}\right)$$

$$= 1.041$$

$$T_r = (L_i - 1)\,(T_{amb} + T_{LNA}\,L_i + \frac{T_{pa}\;L_i}{G_{LNA}}$$

$$= (0.041)\,290 + 105\,(1.041) + \frac{600(1.041)}{1000}$$

$$= 121.8\ K$$

Así que:

$$T_{sis} = T_{ant} + T_r = 60.77 + 121.8$$

$$T_{sis} = 182.57$$

por otro lado

$$G = 40\ dB - 0.2\ dB = 39.8\ dB$$

Entonces

$$\frac{G}{T} = G - 10\log_{10} Tsis$$

$$= 17.18\ dB$$

Obsérvese que G_{LNA} debe convertirse de su valor en decibeles a su valor numérico equivalente; con este caso 30 dB es equivalente a 1000. También, para calcular la relación $\frac{G}{T}$, a la ganancia de la antena deben reducirse las pérdidas del elemento radiador de la antena al plano de referencia seleccionado.

MODULACION EN SATELITES.

La técnica de modulación que domina en los sistemas de comunicaciones vía satélite es la modulación en frecuencia (FM). El sistema FM es ampliamente usado en radiodifusión y en microondas terrestres, de tal forma que la teoría era muy bien entendida cuando los satélites de comunicaciones tenían el gran auge y desarrollo, y también se disponía de la tecnología cuando comenzaron a operar. Además, esta técnica provee la suficiente relación señal a ruido, por ejemplo 30 dB en el sistema INTELSAT IV. La alternativa de utilizar modulación digital, PSK por ejemplo, se ha considerado seriamente, sin embargo se sigue manteniendo el uso de FM como en el caso de microondas terrestres. La utilización de modulación digital en satélites se hará con la misma rapidez que se use modulación digital en microondas terrestres, ya que finalmente, éstas son las que alimentan a las estaciones terrenas. Sin embargo, si la información está en forma digital, se puede utilizar modulación digital directamente.

MODULACION ANALOGICA.

En un sistema troncal tal como INTELSAT o COMSTAR las señales telefónicas están en grupos o supergrupos que consisten de 12 ó 60 canales los cuales han sido multicanalizados por división de frecuencia (FDM) Fig. 47 .

Como se puede observar el sistema FM es similar al usado en microondas terrestres, excepto que se requiere ampliar más la desviación de frecuencia para mejorar la relación señal a ruido ($\frac{S}{N}$). Una sola portadora puede ser modulada por hasta 900 canales de banda base y ocupar un ancho de banda de 36 MHz Fig. 48 (no todos los transpondedores tienen ancho de banda de 36 MHz.)

M O D U L A C I O N

ANALOGICA

SCPC

SSB | FDM | FM | ACCESO

DIGITAL

SCPC

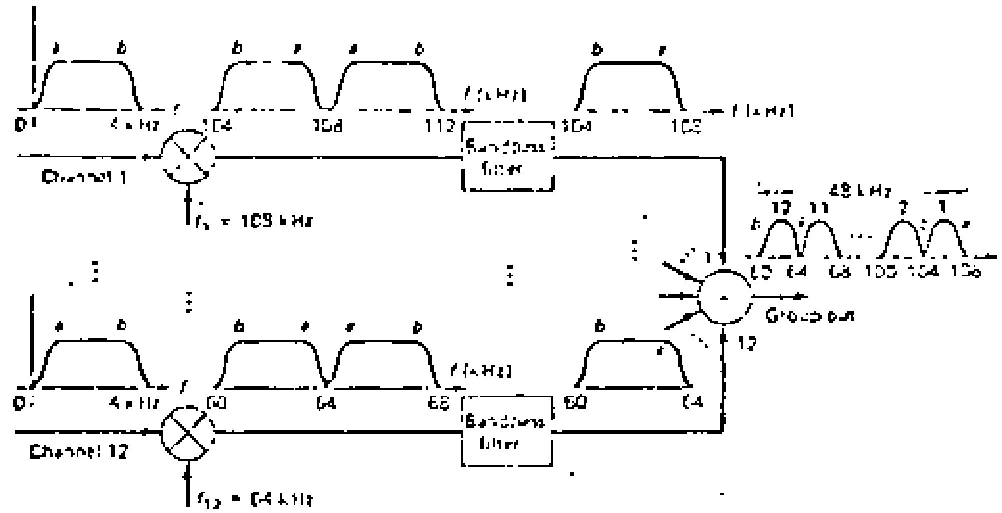
PCM | TDM | PSK | ACCESO

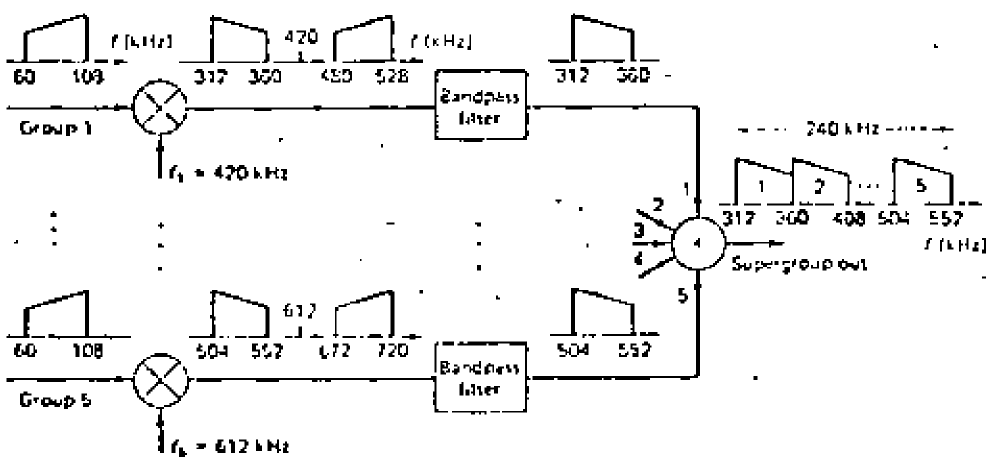
DPCM FSK

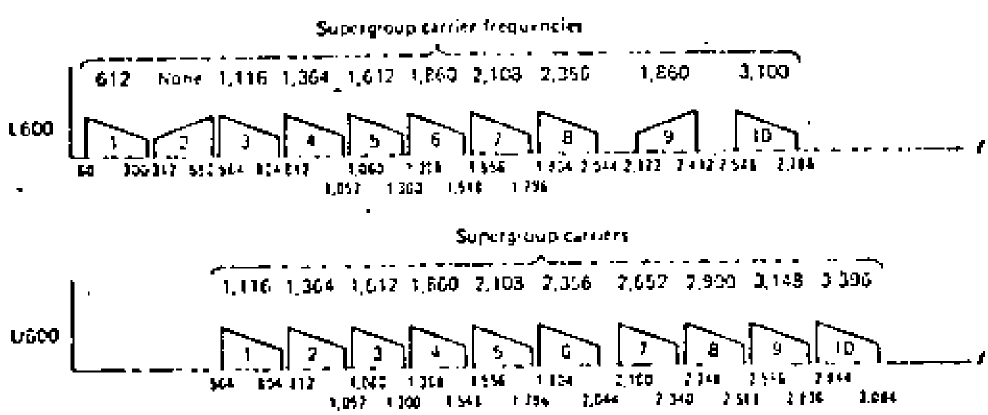
DM ASK

SISTEMA	ANCHO DE BANDA	RESPUESTA A CC.	$\frac{1}{\alpha}\left(\frac{S}{N}\right)_d$	EFICIENCIA	COMPLEJIDAD	APLICACION TIPICA
AM	$B_T = 2fx$	NO	1/3	<50%	MINIMA	RADIODIFUSION COMERCIAL
DBL	$B_T = 2fx$	SI	1	100%	MEDIA	SISTEMAS BAJA FRECUENCIA
BLU	$B_T = fx$	NO	1	100%	MAXIMA	TRANSMISION DE VOZ
BLR	$fx < B_T < 2fx$	SI	1	100%	MAXIMA	SISTEMAS DE GRAN ANCHO DE BANDA
BLR+P	igual a VSB	NO	1/3	<50%	MEDIA	VIDEO DE TV
FM	$B_T \approx 2f\Delta + 2fx$	SI	$\frac{3}{2}\left(\frac{f\Delta}{fx}\right)^2$	--	MEDIA	RADIODIFUSION COMERCIAL
PM	$B_T = 2f\Delta + 2fx$	SI (con ajuste)	$K_p^2/2$	--	MEDIA	TRANSMISION DE DATOS Y GENERACION DE FM

COMPARACION DE LOS SISTEMAS DE MODULACION ANALOGICOS







Note All frequencies in kHz.

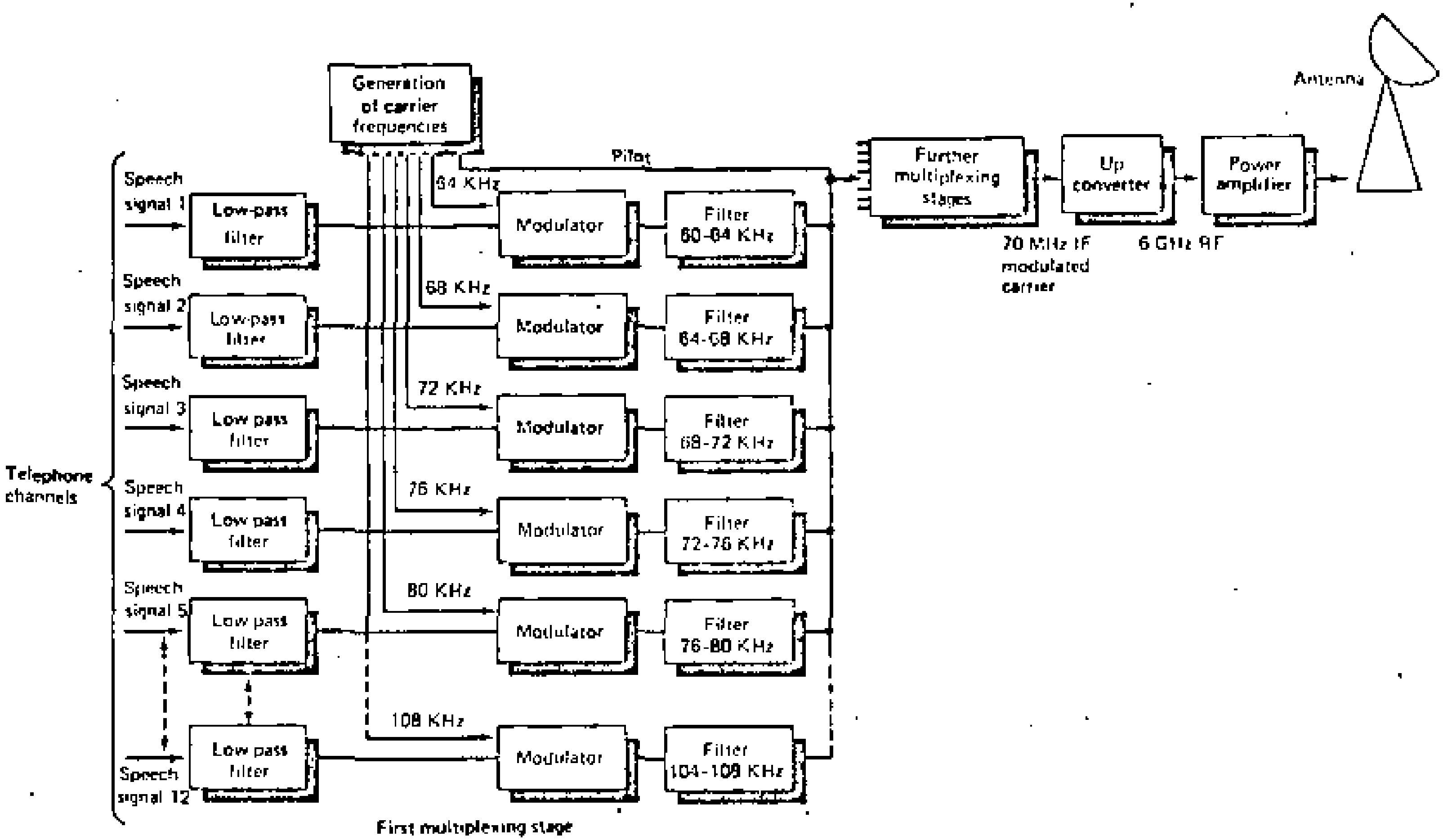
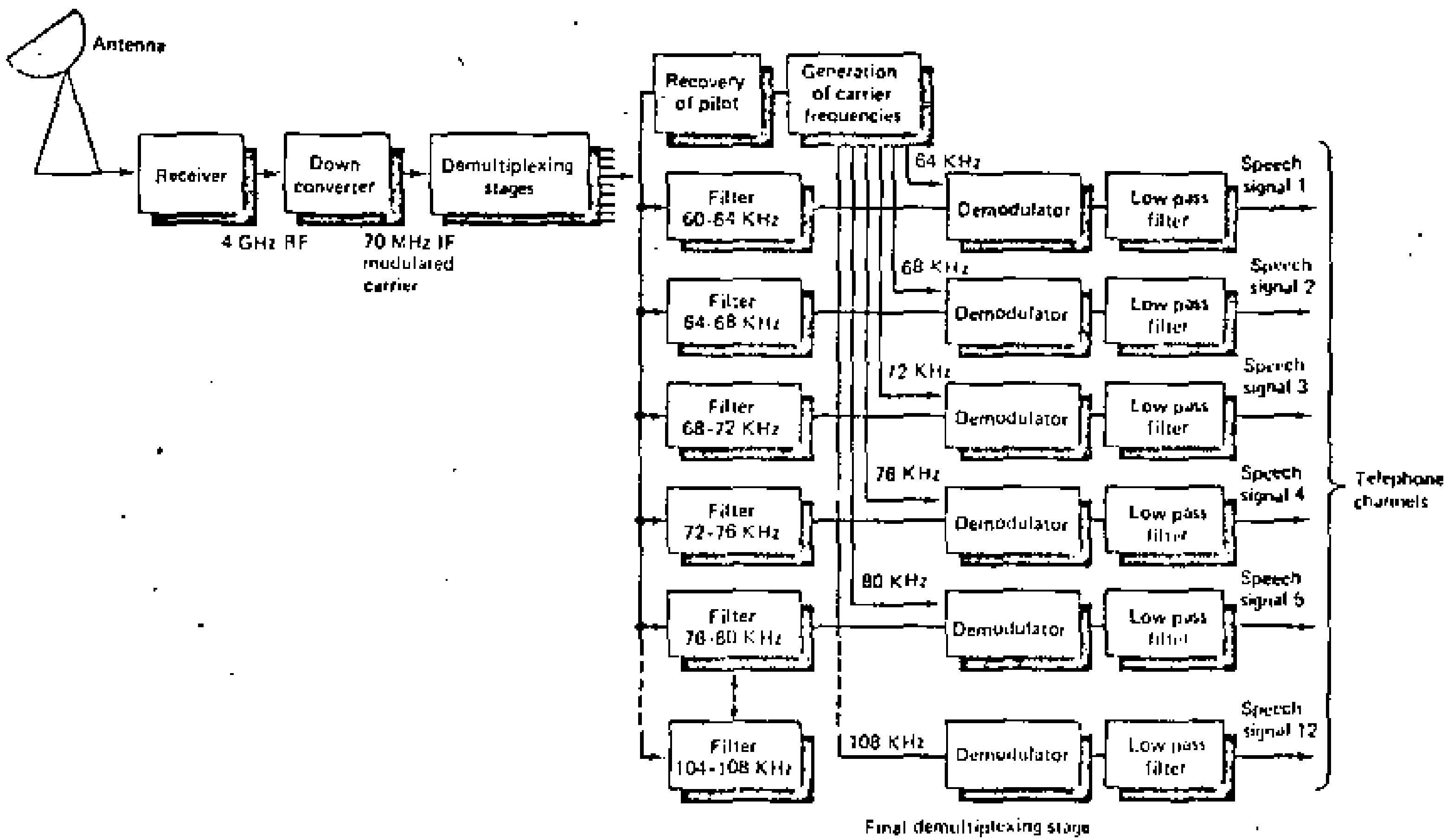

Generation of carrier frequencies
Pilot
Telephone channels
Speech signal 1
Speech signal 2
Speech signal 3
Speech signal 4
Speech signal 5
Speech signal 12
Low-pass filter
Low-pass filter
Low pass filter
Low pass filter
Low pass filter
Low pass filter
64 KHz
68 KHz
72 KHz
76 KHz
80 KHz
108 KHz
Modulator
Modulator
Modulator
Modulator
Modulator
Modulator
Filter 60-64 KHz
Filter 64-68 KHz
Filter 68-72 KHz
Filter 72-76 KHz
Filter 76-80 KHz
Filter 104-108 KHz
First multiplexing stage
Further multiplexing stages
70 MHz IF modulated carrier
Up converter
6 GHz RF
Power amplifier
Antenna

Fig. 47


Antenna
Receiver
4 GHz RF
Down converter
70 MHz IF modulated carrier
Demultiplexing stages
Recovery of pilot
Generation of carrier frequencies
Filter 60-64 KHz
Filter 64-68 KHz
Filter 68-72 KHz
Filter 72-76 KHz
Filter 76-80 KHz
Filter 104-108 KHz
64 KHz
68 KHz
72 KHz
76 KHz
80 KHz
108 KHz
Demodulator
Low pass filter
Speech signal 1
Speech signal 2
Speech signal 3
Speech signal 4
Speech signal 5
Speech signal 12
Telephone channels
Final demultiplexing stage

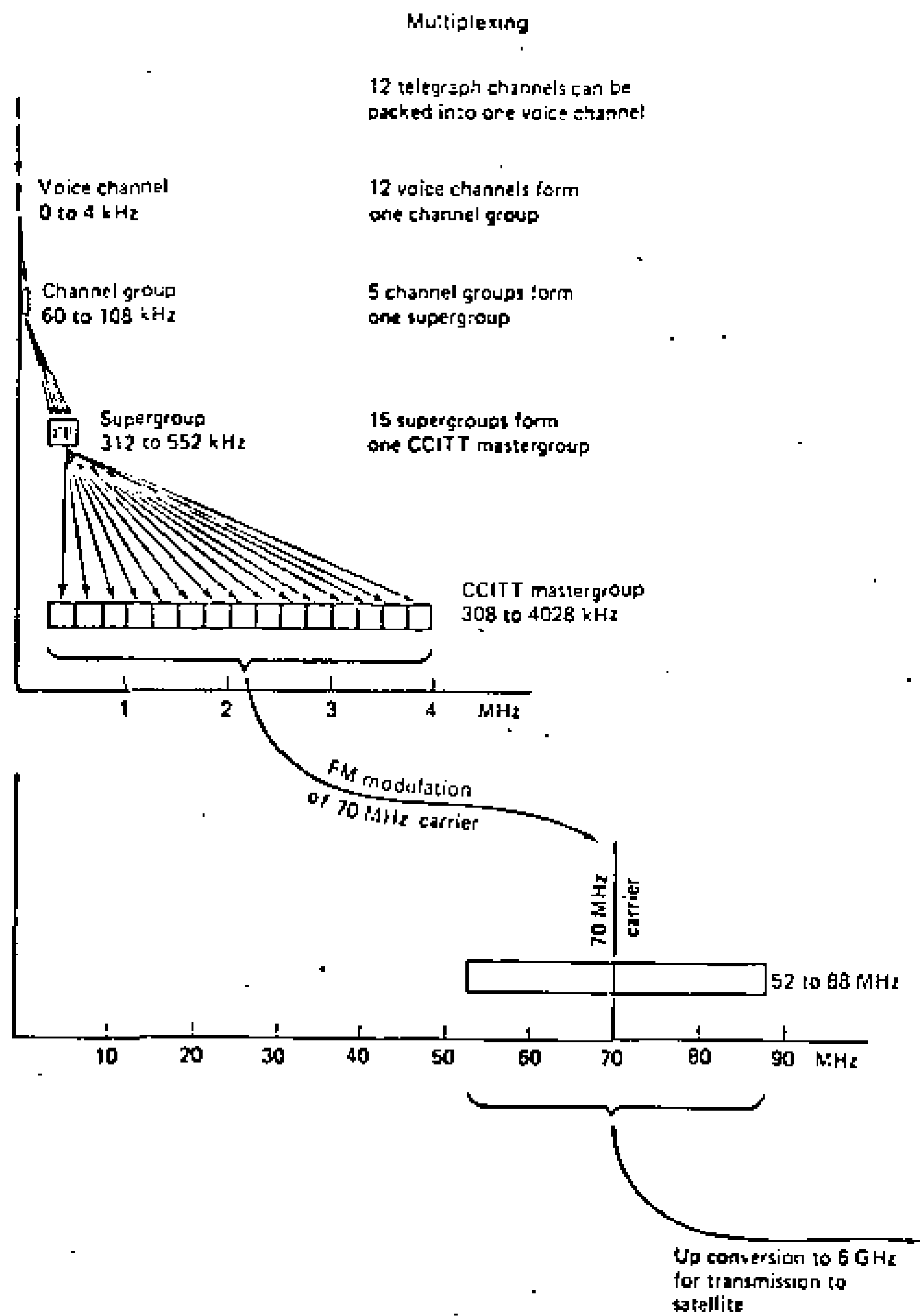

Multiplexing
12 telegraph channels can be packed into one voice channel
Voice channel 0 to 4 kHz
12 voice channels form one channel group
Channel group 60 to 108 kHz
5 channel groups form one supergroup
Supergroup 312 to 552 kHz
15 supergroups form one CCITT mastergroup
CCITT mastergroup 308 to 4028 kHz
1
2
3
4
MHz
FM modulation of 70 MHz carrier
70 MHz
carrier
52 to 88 MHz
10
20
30
40
50
60
70
80
90
MHz
Up conversion to 6 GHz for transmission to satellite

Fig. 48

El uso de modulación en frecuencia en un sistema fm-fdma es más eficiente en términos de ancho de banda y se mantiene comparable al esquema TDM-PCM-CPSK. La mayor desventaja de un sistema FM es la acumulación de ruido a lo largo de los diferentes saltos del enlace, lo que resulta que se mantengan las contribuciones de ruido a un valor mínimo en cada etapa. En las bandas de 4 y 6 GHz se debe agregar ruido por interferencia causado por enlaces terrestres y por satélites que operan con satélites adyacentes.

MODULACION DIGITAL.

Las técnicas de modulación digital más recomendadas son: FDM-PCM-PSK y TDM-PCM-PSK, la primera se basa en esquemas de SCPC y la segunda, en una sola portadora de banda ancha.

Actualmente, el sistema PCM-PSK se usa en el sistema SPADE.

La Fig.49 ilustra el modelo de un sistema de comunicación digital vía satélite, tal como en el caso de modulación analógica, el TWTA en el transpondedor del satélite es un elemento clave en el diseño del sistema digital. La probabilidad de error para varios esquemas PSK y FSK se muestra en la Fig. 50

En la tabla I mostrada, se ve claramente que 4-PSK con detección coherente es la mejor, ya que el ancho de banda requerido es la mitad que para PSK. El ancho de banda ocupado por una señal PCM-PSK depende de la conformación y del número de fases.

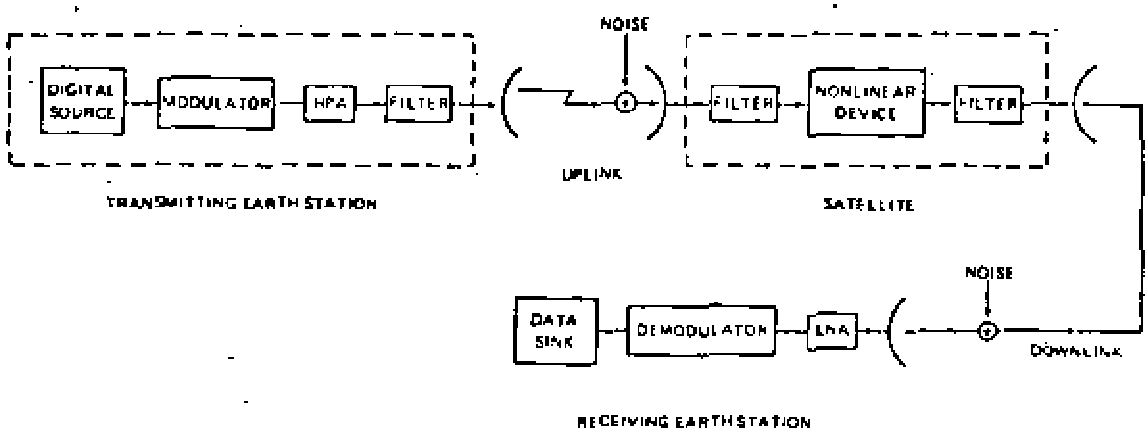

NOISE
DIGITAL SOURCE
MODULATOR
HPA
FILTER
UPLINK
TRANSMITTING EARTH STATION
FILTER
NONLINEAR DEVICE
FILTER
SATELLITE
NOISE
DATA SINK
DEMODULATOR
LNA
DOWNLINK
RECEIVING EARTH STATION

Fig. 49

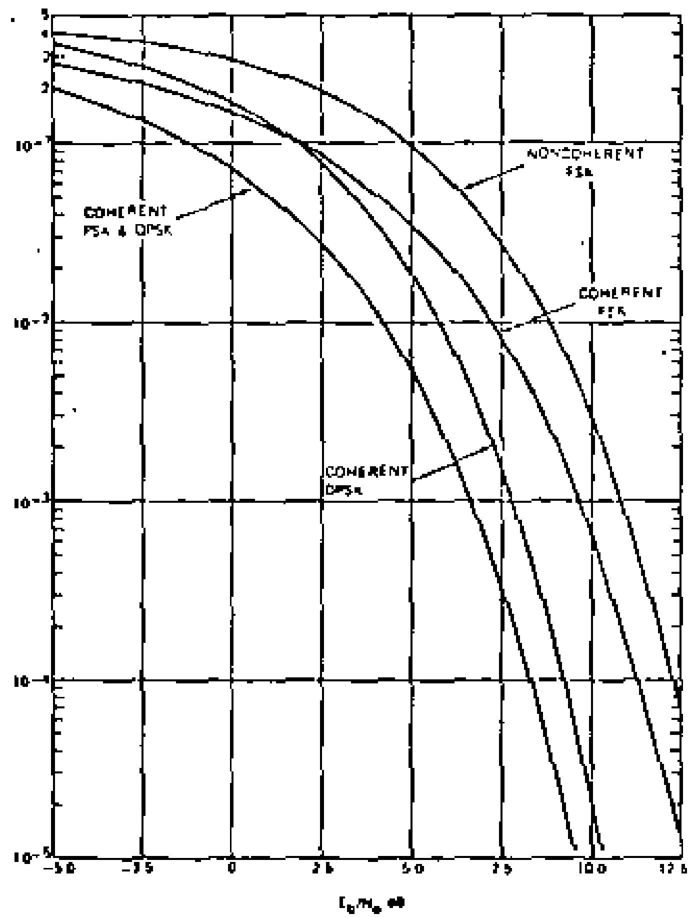
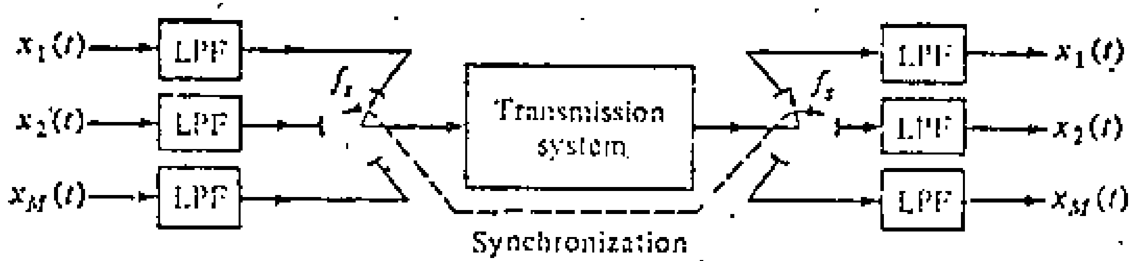


Fig. 2. P_e versus E_b/N_0.

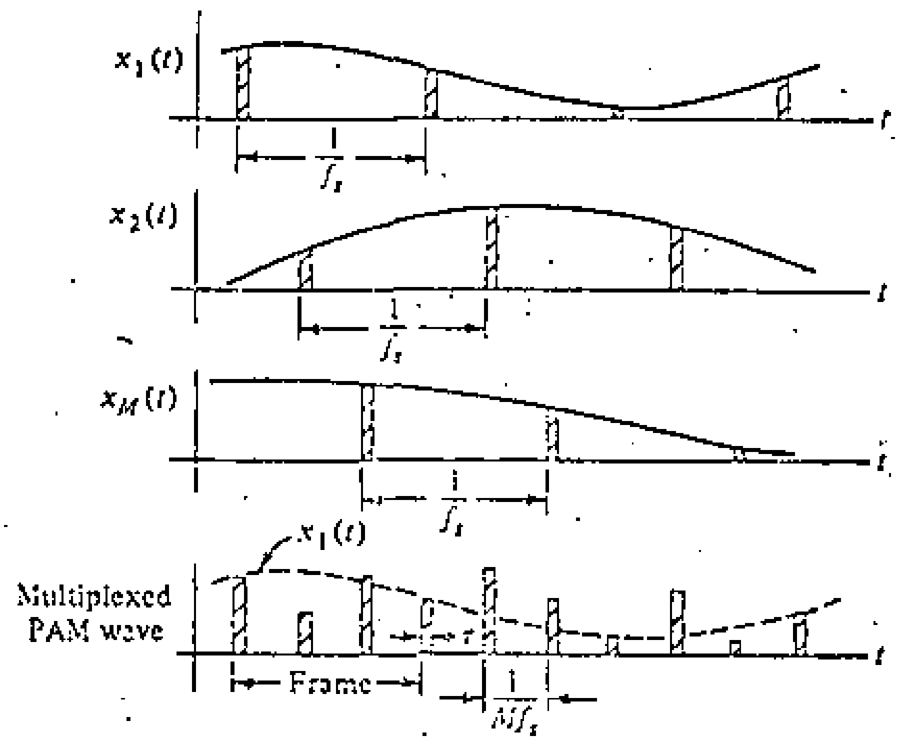
Fig. 50

Por ejemplo conformación de coseno elevado y medio coseno elevado dan el óptimo ancho de banda.

Una comparación del ancho de banda relativo usado por los sistemas FM y PCM-PSK se ilustra en la Tabla II.


$x_1(t)$
$x_2(t)$
$x_M(t)$
LPF
LPF
LPF
f_s
Transmission system
Synchronization
f_s
LPF
LPF
LPF
$x_1(t)$
$x_2(t)$
$x_M(t)$

(a)


$x_1(t)$
$\frac{1}{f_s}$
$x_2(t)$
$\frac{1}{f_s}$
$x_M(t)$
$\frac{1}{f_s}$
$x_1(t)$
Multiplexed PAM wave
τ
Frame
$\frac{1}{Mf_s}$
t

(b)

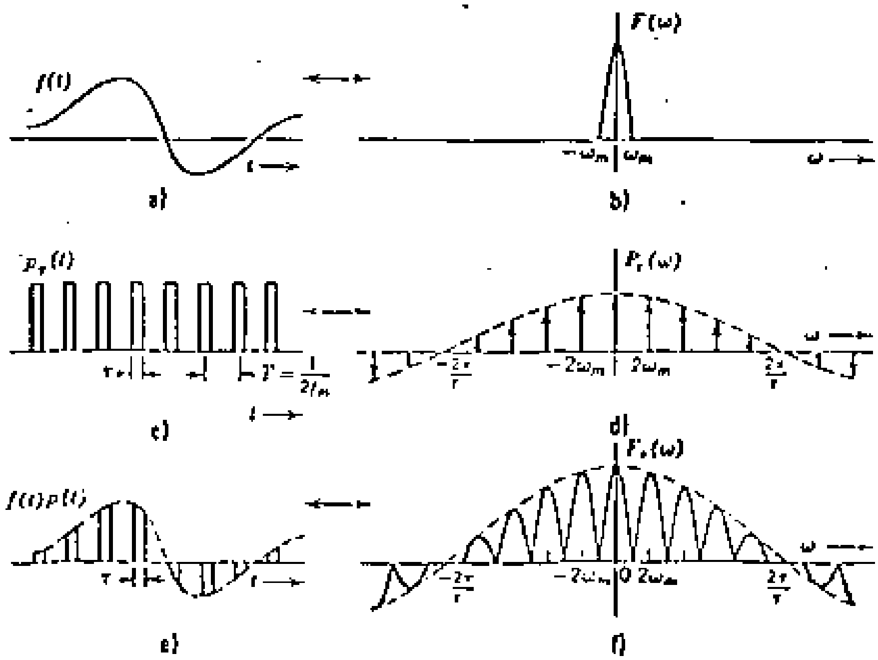


FIGURA N° 3
MODULACION POR PULSOS: Muestreo No Ideal.

$$F_s(w) = \frac{1}{2\pi} F(w) * P_\tau(w)$$

$$T = \frac{1}{2\ fm} = \frac{\pi}{Wm}$$

$$w_0 = \frac{2\pi}{T} = 2\ w_m$$

$$P_\tau(w) = 2\ A_\tau W_m \sum_{n=-\infty}^{\infty} Sinc(n\tau w_m)\ \delta(w - 2\ nw_m)$$

$$F_s(w) = \frac{A_\tau\ w_m}{\pi} F(w) * \sum_{n=-\infty}^{\infty} Sinc(n\tau w_m)\ \delta(w-2nw_m)$$

$$= \frac{A\tau}{T} \sum_{n=-\infty}^{\infty} Sinc(n\tau w_m)\ F(w) * \delta(w - 2\ nw_m)$$

$$= \frac{A\tau}{T} \sum_{n=-\infty}^{\infty} Sinc(n\tau w_m)\ F\ (w-2nw_m)$$

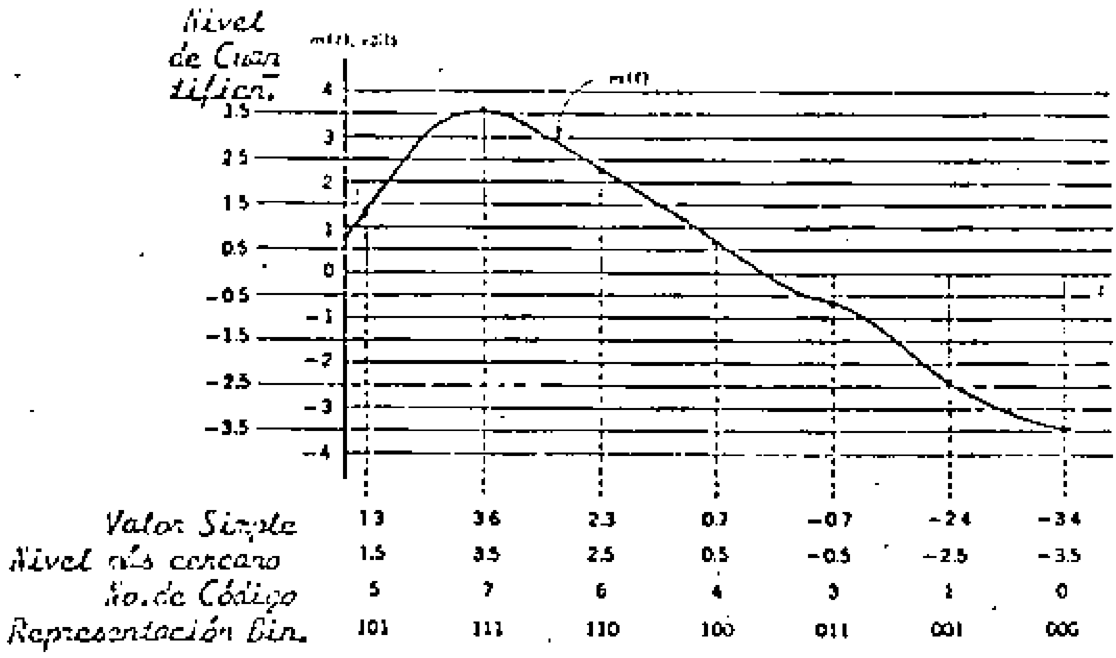
Obsérvese que F(w) se repetirá sin traslaparse siempre que

$$W_o \geq 2 W_m$$

$$\frac{2\pi}{T} \geq 2 (2 \pi f_m)$$

$$T \leq \frac{1}{2 f_m}$$

$$\underline{\underline{fs \geq 2 fm}} \quad \text{Muestras/s}$$



Se muestrea regularmente una señal. En la figura se han indicado los niveles de cuantificación así como su representación binaria. Para cada muestra se da el valor de cuantificación.

FIGURA N° 1-A

Binario				Decimal
b_4	b_3	b_2	b_1	
0	0	0	0	0
0	0	0	1	1
0	0	1	0	2
0	0	1	1	3
0	1	0	0	4
0	1	0	1	5
0	1	1	0	6
0	1	1	1	7
1	0	0	0	8
1	0	0	1	9
1	0	1	0	10
1	0	1	1	11
1	1	0	0	12
1	1	0	1	13
1	1	1	0	14
1	1	1	1	15

Números Equivalentes en Representación Binaria y Decimal

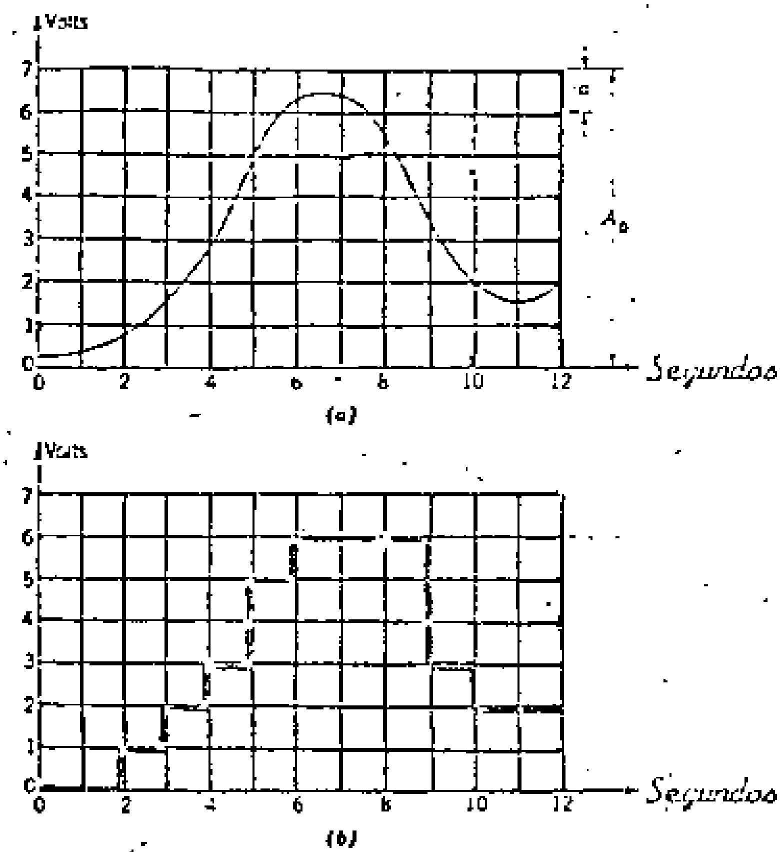


FIGURA N° 1
MUESTREO Y CUANTIFICACION: a) Señal b) versión muestreada y cuantificada.

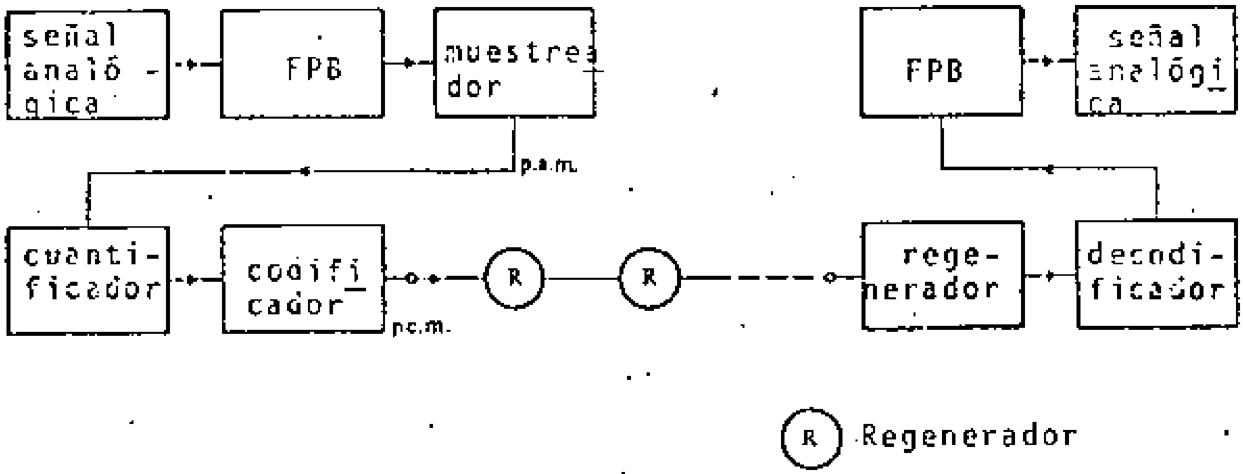


FIGURA N° 2
DIAGRAMA EN BLOQUES DEL SISTEMA PCM

TABLA III
PARAMETROS DE PCM

Tipo de señal	Ancho de Banda	Tasa de muestreo	No.de intervalosde cuantific.	Long. de la palabra en el código
voz	de 300 Hz a 3400 Hz	8 KHz	128 ó 256	7 ó 8
programa de música	15 KHz	32 KHz	2048	11
TV a color	5.5 MHz	13 MHz	512	9

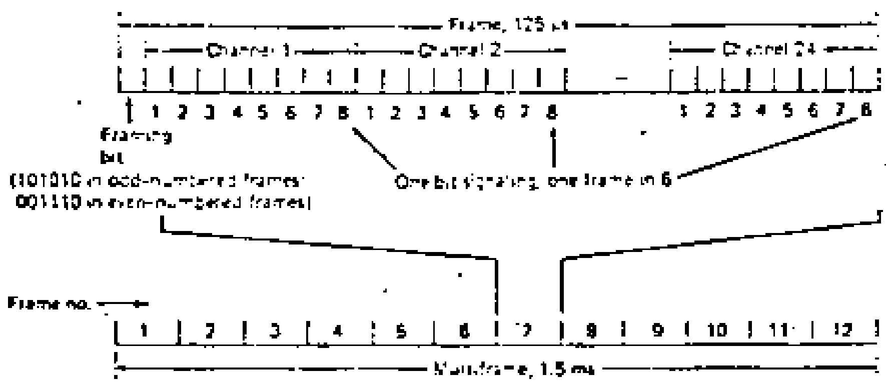


FIGURA N° 12.

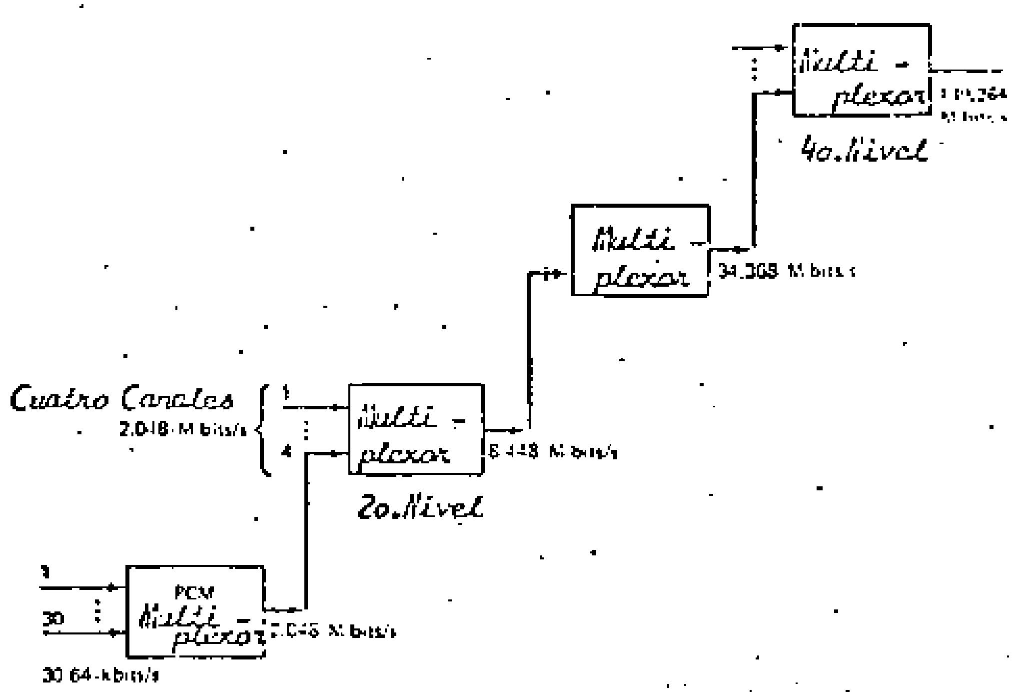


FIGURA N° 13
JERARQUIA DIGITAL: Recomendación CCITT.

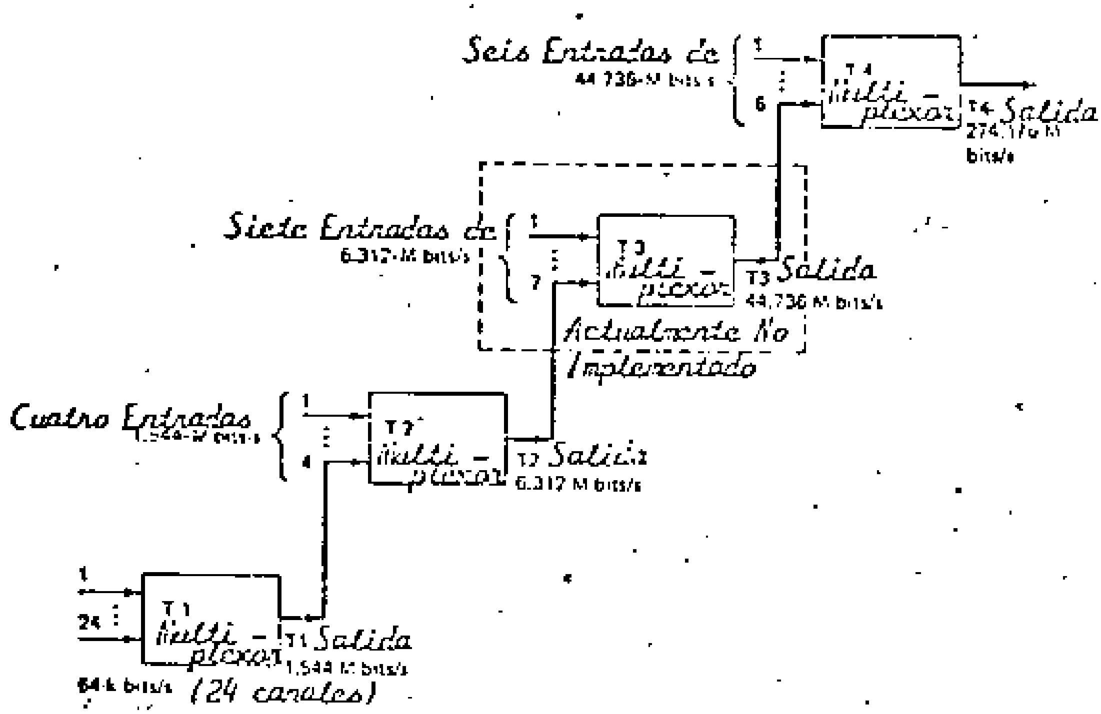


FIGURA N° 14
SISTEMA ATT : Jerarquía Digital.

TABLA IV

Velocidades Estándar de Transmisión en Estados Unidos, Canadá, Japón Europa.

Jerarquía. Nivel No.	EU/Canadá (Mb/s)	Japón (Mb/s)	Europa (Mb/s)
1	1.544	1.544	2.048
2	6.312	6.312	8.448
3	44.736	32.064	34.368
4	274.176	97.728	139.264
5	-	396.200	560-840

TABLA V

Capacidad Estándar de Canales de Voz en Sistemas PCM. Estados Unidos, Canadá, Japón y Europa.

Jerarquía. Nivel No.	capacidad		
	EU/Canadá	Japón	Europa
1	24	24	30
2	96	96	120
3	672	480	480
4	4032	1440	1920
5	-	5760	7680-11520

Ancho de banda (MHz)

No. de Canales	FDM	PCM coseno elevado	PCM 1/2 coseno elevado
24	0.92	3.08	2.28
60	4.23	7.68	5.75
120	5.88	15.36	11.5
600	14.45	76.8	57.5

TABLA II

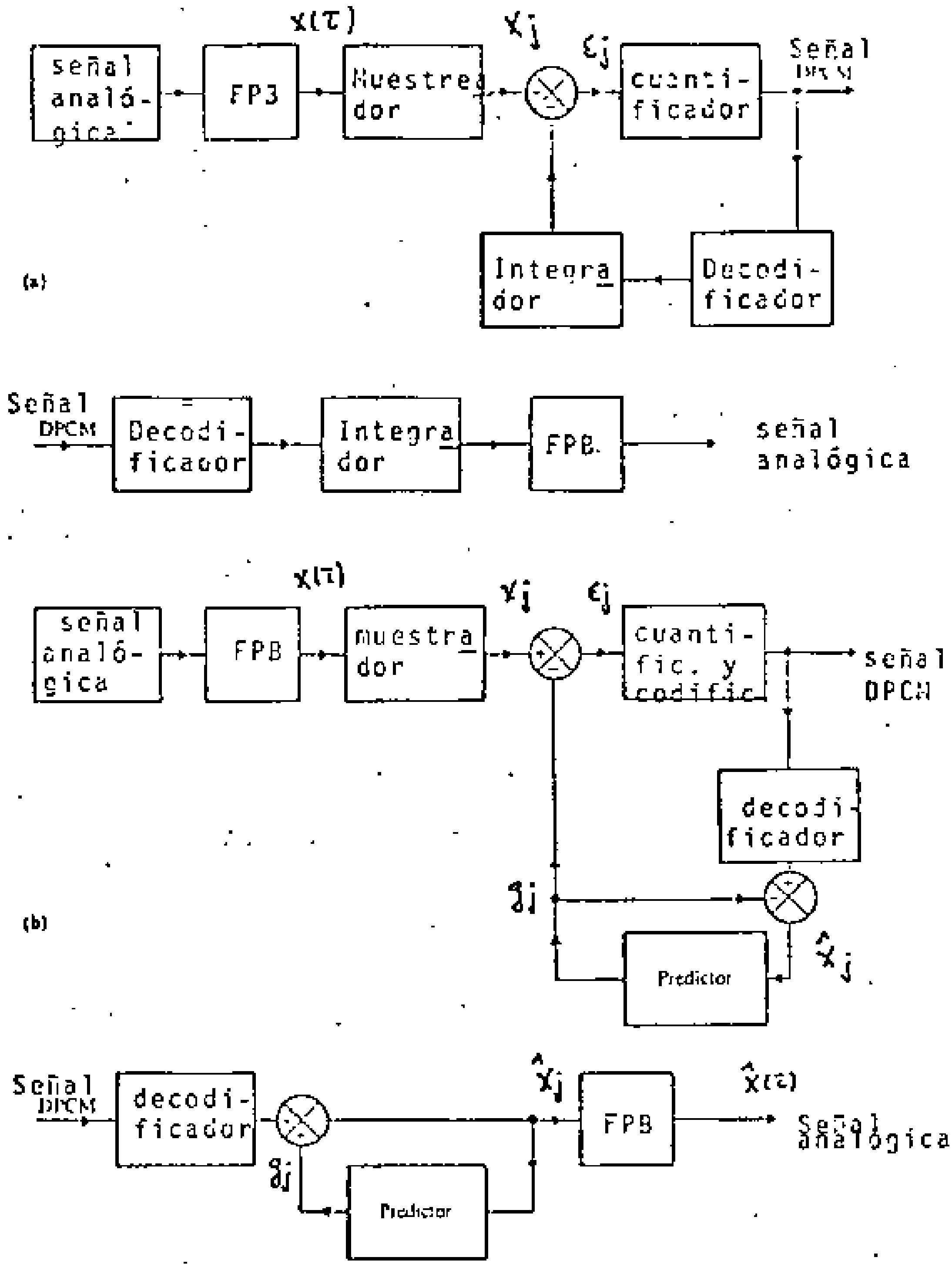


FIGURA N° 15

DIAGRAMA EN BLOQUES DEL SISTEMA PCM DIFERENCIAL: a) Transmisor y receptor de un sistema PCM simple; b) Transmisor y receptor de - de un sistema PCM con predictor.

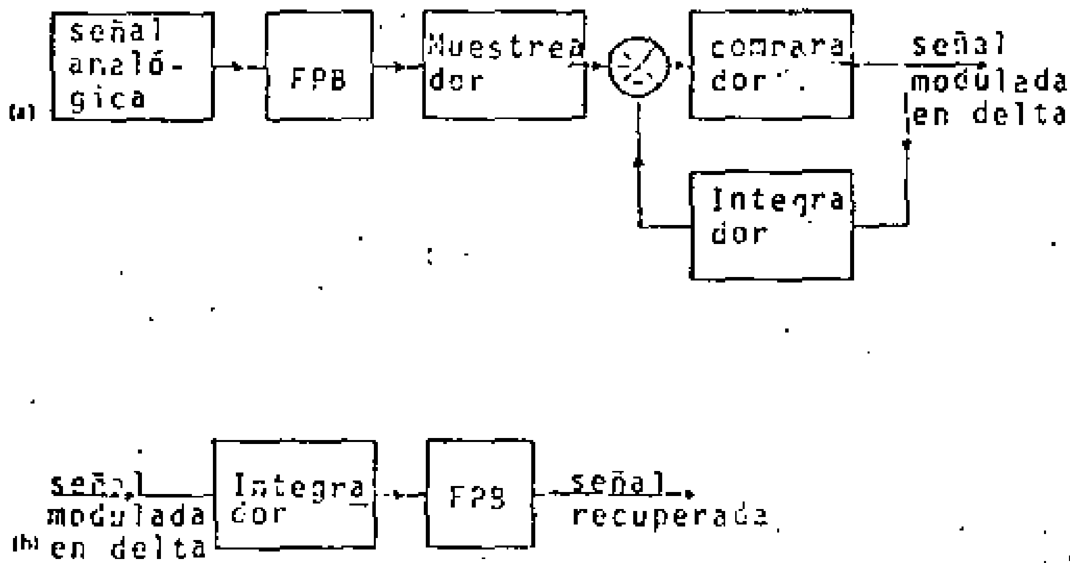


FIGURA N° 16

DIAGRAMA EN BLOQUES DE UN SISTEMA DE MODULACION DELTA SIMPLE a) Transmisor; b) Receptor.

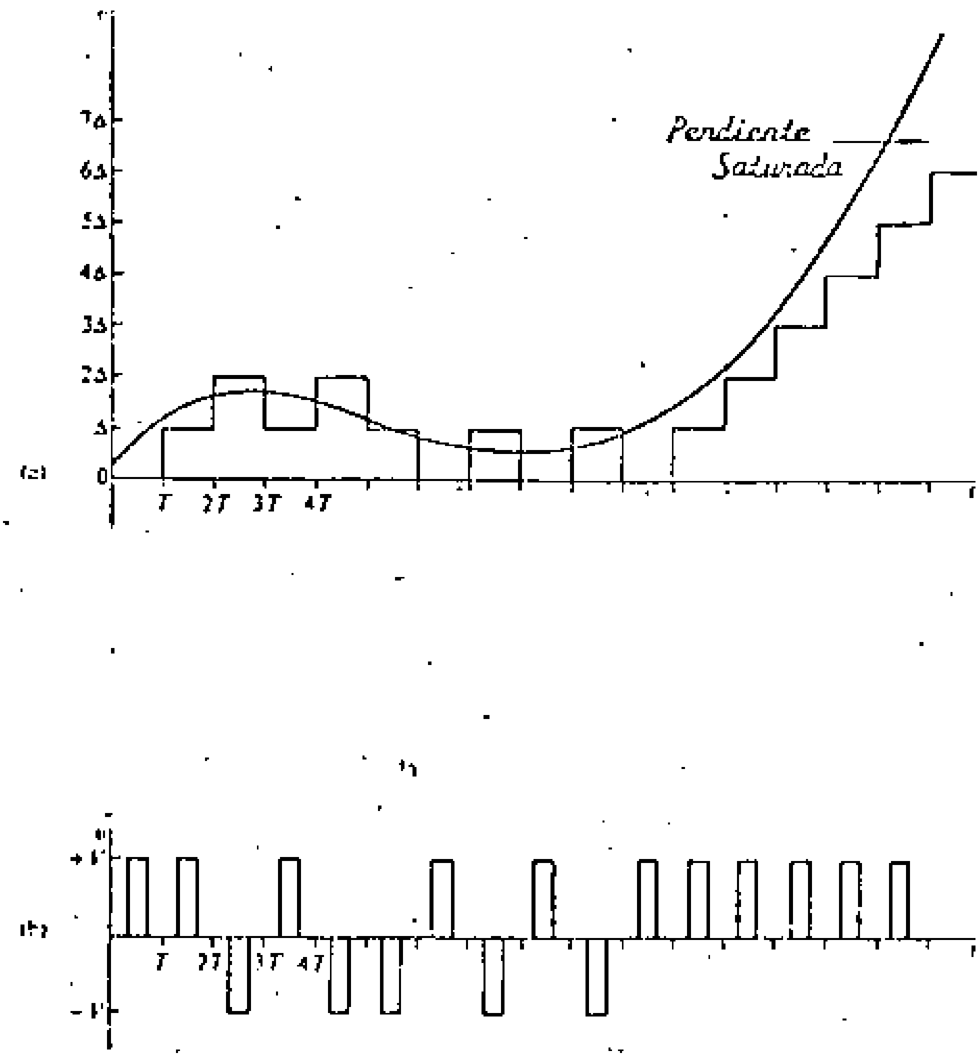


FIGURA N° 17
SEÑALES EN UN SISTEMA DE MODULACION DELTA :
a) Señal analógica de entrada y su reconstrucción.
b) Señal modulada en delta.

MODULACION Y DEMODULACION

Para poder transmitir los trenes de pulsos a través de enlaces por altas frecuencias, una portadora contínua puede modularse en amplitud, fase o frecuencia en el sistema transmisor, ya que las características de transmisión a altas frecuencias son del tipo de banda base. La señal transmitida es primero demodulada en pulsos en la banda de frecuencia de la portadora en el sistema receptor para dar los pulsos PCM en la banda base. Entonces los pulsos digitales binarios, sin distorsión de transmisión en sus formas de ondas, son regenerados por los pulsos demodulados a través del decodificador.

La modulación y demodulación de la portadora de microondas son esenciales en el sistema de radioenlace PCM. Los pulsos binarios antes de la modulación y después de la demodulación son llamados pulsos banda base.

LLAVEO POR CORRIMIENTO DE AMPLITUD (ASK)

Considere una secuencia de pulsos binarios, como se muestra en la Fig. 18. Los 1's hacen que la portadora esté presente y los 0's la hacen ausente.

Es evidente que el espectro de la señal ASK dependerá de la secuencia binaria particular a ser transmitida. La señal ASK es simplemente:

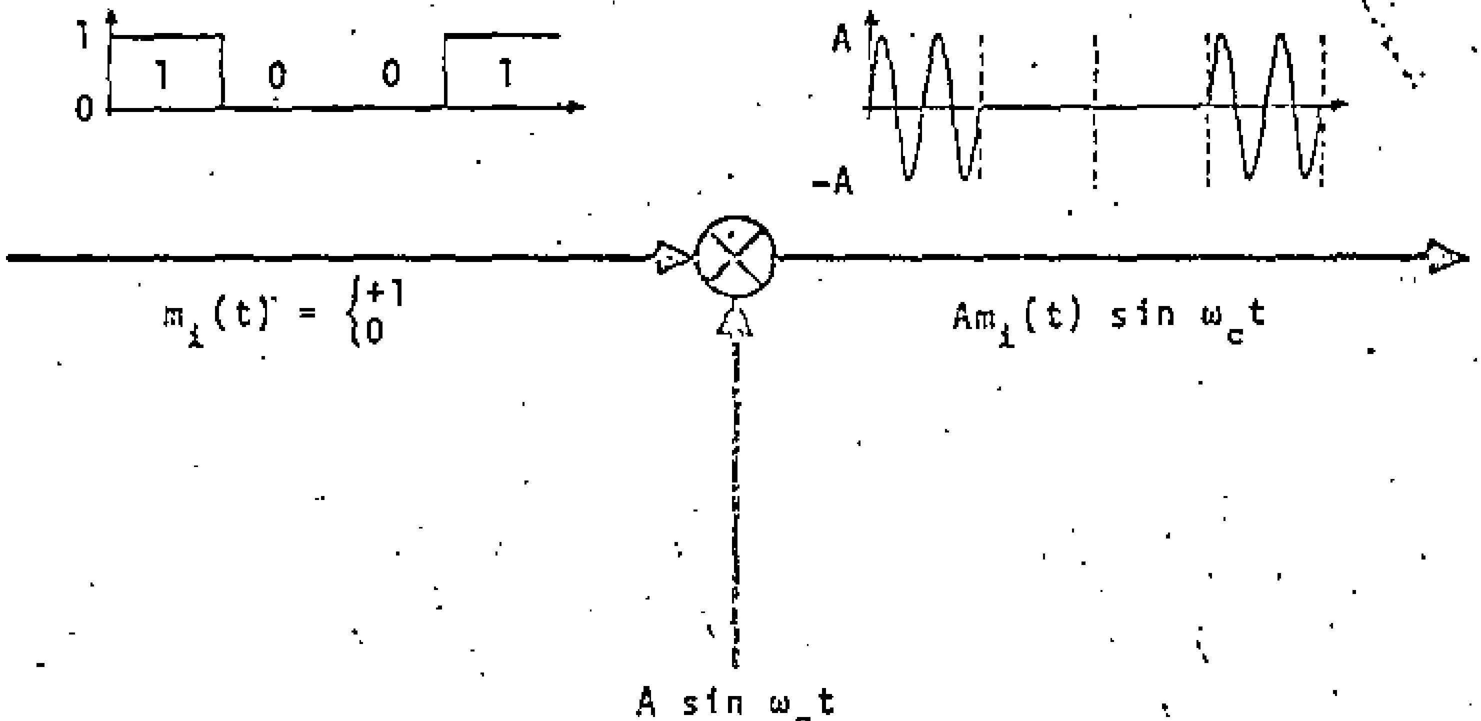


FIGURA N° 18
MODULACION ASK

$$x_c(t) = x(t) \cos w_c t \qquad (1)$$

donde $x_c(t) = 1$ ó o, sobre un largo intervalo T segundos. Note que esto es exactamente la forma de la señal modulada discutida en capitulos anteriores. Como se mostró, al tomar la transformada de Fourier de la señal modulada en amplitud (ASK) y usando el teorema de desplazamiento de frecuencia, tenemos

$$X_c(\omega) = \frac{A}{2}\left[X(\omega-\omega_c) + X(\omega+\omega_c)\right] \qquad (2)$$

El efecto de multiplicar por $\cos w_c t$ es simplemente defasar el espectro original de la señal binaria (señal de banda base) a la frecuencia ω_c (fig. 19). En realidad esto es la forma general de una señal de AM.

El especto de la señal modulada (ASK) se muestra en la fig. 20, ya que como se vió anteriormente, es simplemente el espectro de un tren de pulsos esto es $\frac{\text{Sen } X}{X}$.

LLAVEO POR CORRIMIENTO DE FRECUENCIA

Aquí, si consideramos una forma rectangular por simplicidad,

$$\left.\begin{array}{l} x_c(t) = A \cos \omega_1 t \\ \text{ó} \\ x_c(t) = A \cos \omega_2 t \end{array}\right\} \quad -\frac{T}{2} \leq t \leq \frac{T}{2} \qquad (3)$$

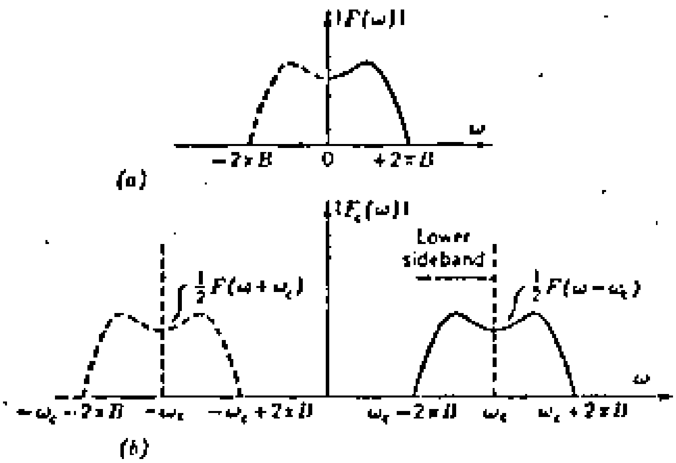


FIGURA N° 19
ESPECTRO DE AMPLITUD:
a) Espectro de la señal moduladora:
b) Espectro de la señal modulada.

Un uno corresponde a la frecuencia x_1, un cero a la frecuencia x_2 (Fig.21). En algunos sistemas, particularmente sobre líneas telefónicas x_1 y $x_0 = \frac{1}{T}$, pero en general x_1 y $x_2 >> \frac{1}{T}$. Una representación alternativa de la onda de FSK consiste de hacer $x_1 = x_c - \Delta x$, $x_2 = x_c + \Delta x$. Las dos frecuencias difieren entonces por $2\Delta x$ hertz. Entonces

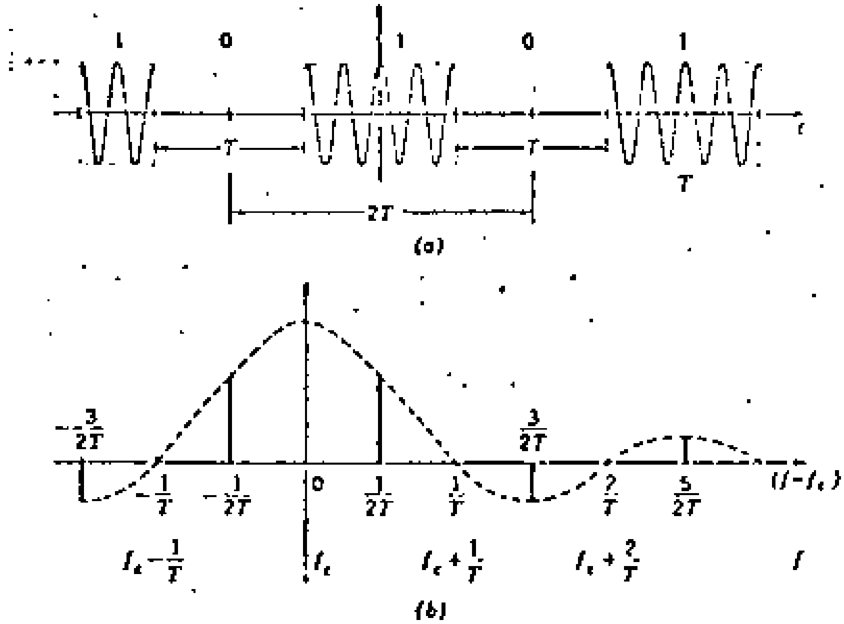
$$x_c(t) = A \cos (\omega_c \pm \Delta\omega)\ t \qquad -\frac{T}{2} \leq t \leq \frac{T}{2} \qquad (4)$$

entonces la frecuencia se desvía $\pm\Delta x$ respecto a x_c. Δx es comunmente la desviación de frecuencia. El espectro de frecuencia para FSK es, en general, difícil de obtener. Debemos de observar que esto es una característica general de señales de FM.

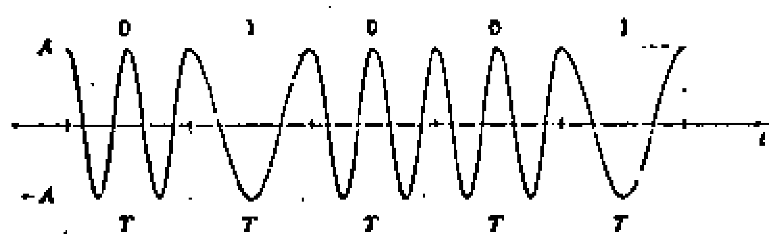
Consideremos que el mensaje binario consiste de una secuencia alternativa de 1's y 0's. Si las dos frecuencias son múltiples por el recíproco del período binario T ($x_1 = m/T$, $x_2 = n/T$, m y n integrados), y son sincronizadas en fase, como se considera en la ecuación (3), la onda FSK es la función periódica de la fig. 22. Note, sin embargo, que esto puede también ser visualizado como la superposición lineal de dos señales periódicas ASK tales como la de la fig. 22, una retrazada T segundos con repecto a la otra.

LLAVEO POR CORRIMIENTO DE FASE

En este caso, tenemos que la señal de llaveo por corrimiento



Espectro de una Señal Periódica OOK: a, Señal Periódica; b) Espectro (frecuencias positivas únicamente).



Señal FSK

FIGURA N° 20

MODULATION - FSK

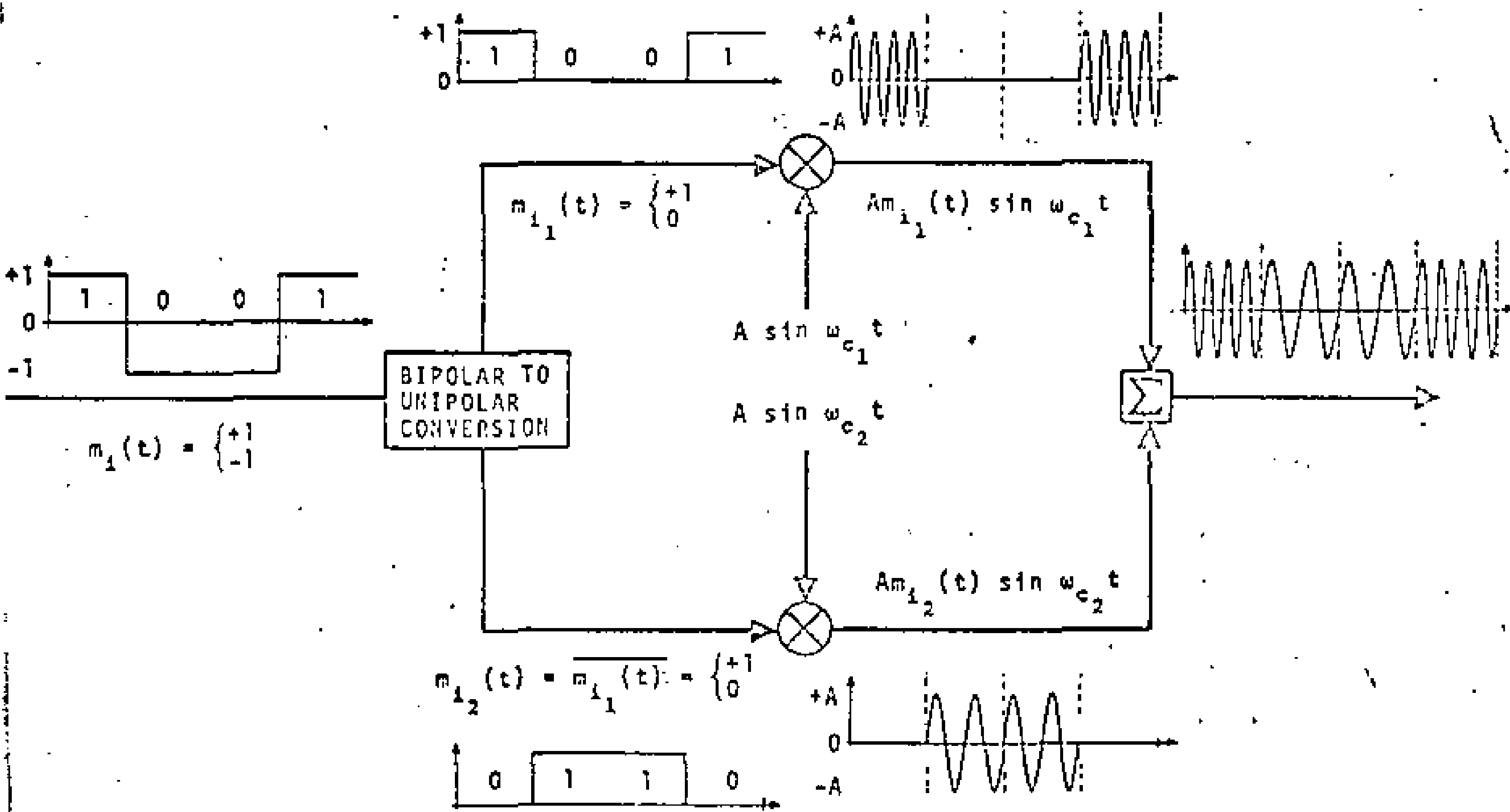


FIGURA Nº 21

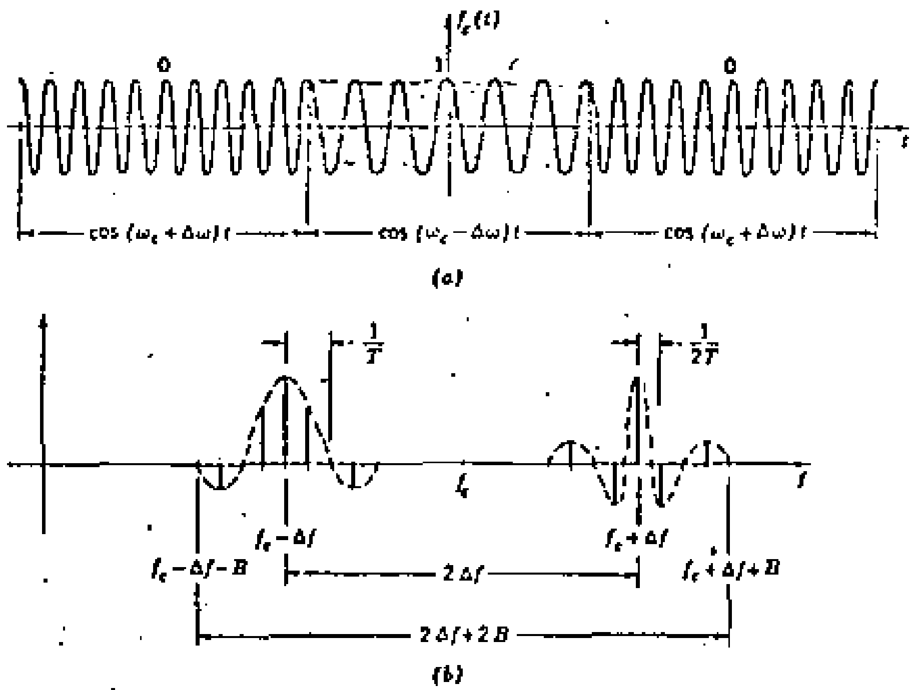

$f_c(t)$
0
1
0
cos $(\omega_c + \Delta\omega)t$
cos $(\omega_c - \Delta\omega)t$
cos $(\omega_c + \Delta\omega)t$
(a)
$\frac{1}{T}$
$\frac{1}{2T}$
f_c
f
$f_c - \Delta f$
$f_c + \Delta f$
$f_c - \Delta f - B$
$2\Delta f$
$f_c + \Delta f + B$
$2\Delta f + 2B$
(b)

FIGURA Nº 22

de fase esta dada por

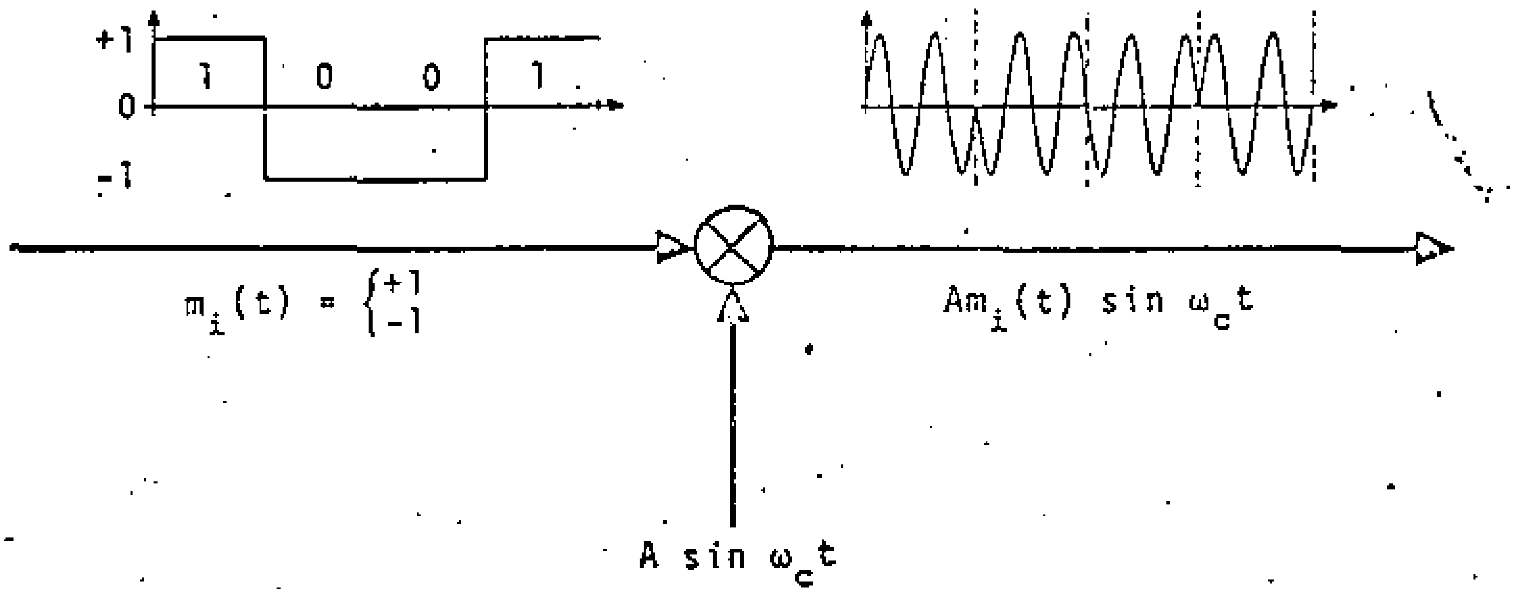
$$x_c(t) = \pm \cos \omega_c t \qquad -\frac{T}{2} \leq t \leq \frac{T}{2} \qquad (5)$$

Si una forma rectangular es asumida. Aquî un 1 en el flujo binario de banda base corresponde a polaridad positiva, y un 0 a polaridad negativa. La señal PSK corresponde esencialmente a un flujo binario sin retorno a cero, como se muestra en la Fig. 23.

Las señales ASK, FSK y PSK pueden producirse por medio de moduladores digitales. Sin embargo, dichos moduladores pueden ser implementados más simplemente alimentando la entrada de datos directamente a un conmutador el cual puede seleccionar la forma de onda de la señal apropiada de una de las dos fuentes de la señal, para asi, construir la señal modulada. Moduladores de este tipo son mostrados esquemáticamente en la fig. 24. El modulador ASK representada en la fig. 24a simplemente conmuta una portadora en encendido o apagado. El modulador FSK en la fig. 24b conmuta entre dos señales de diferentes frecuencias. El conmutador de PSK, como se muestra en la fig. 24c, introduce un retraso de duración de medio longitud de onda a la señal del oscilador para que así se produzca un cambio de fase de π en la señal modulada.

DEMODULACION

MODULATION - PSK



$$\underbrace{A \cos\left(\omega_c t - m_i(t)\,\frac{\pi}{2}\right)}_{\text{PSK SIGNAL}} = A \cos \omega_c t \,\cos\left[m_i(t)\,\frac{\pi}{2}\right]^{\to 0} + A \sin \omega_c t \,\sin\left[m_i(t)\,\frac{\pi}{2}\right]$$

$$= \underbrace{Am_i(t) \sin \omega_c t}_{\text{DSB SIGNAL}}$$

FIGURA Nº 23.

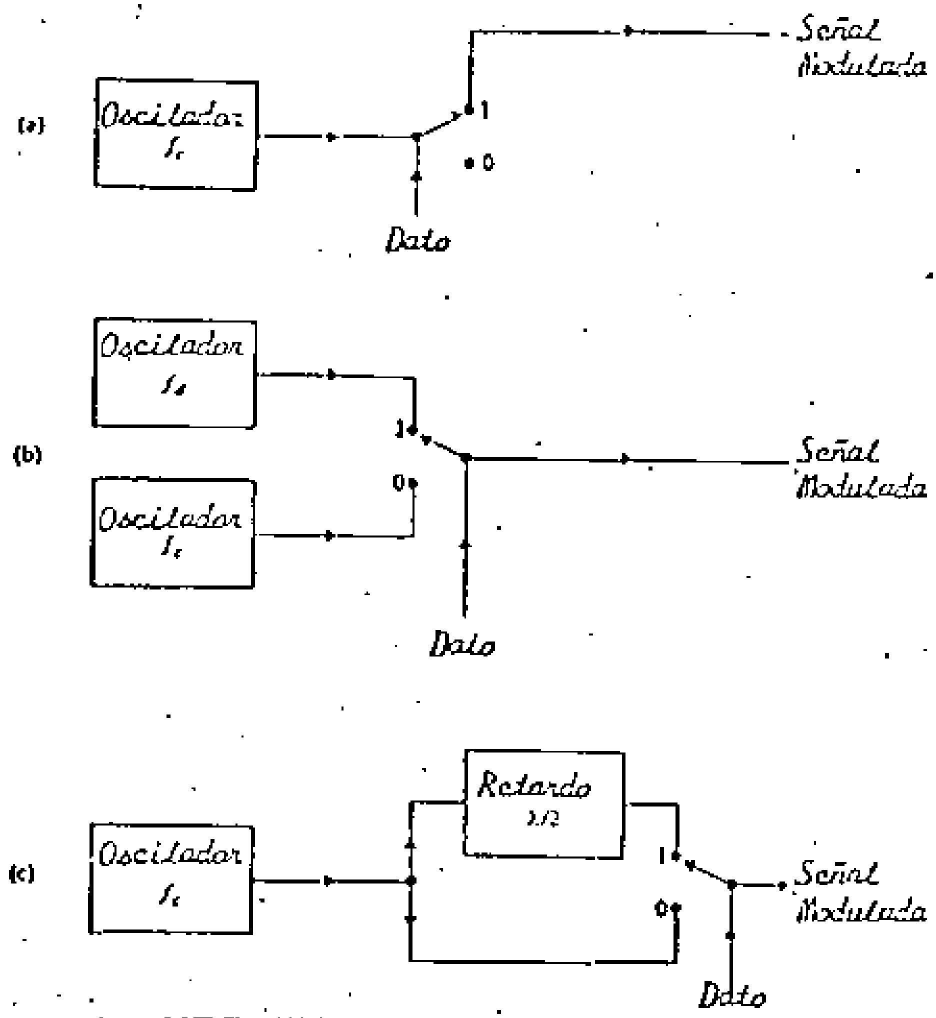
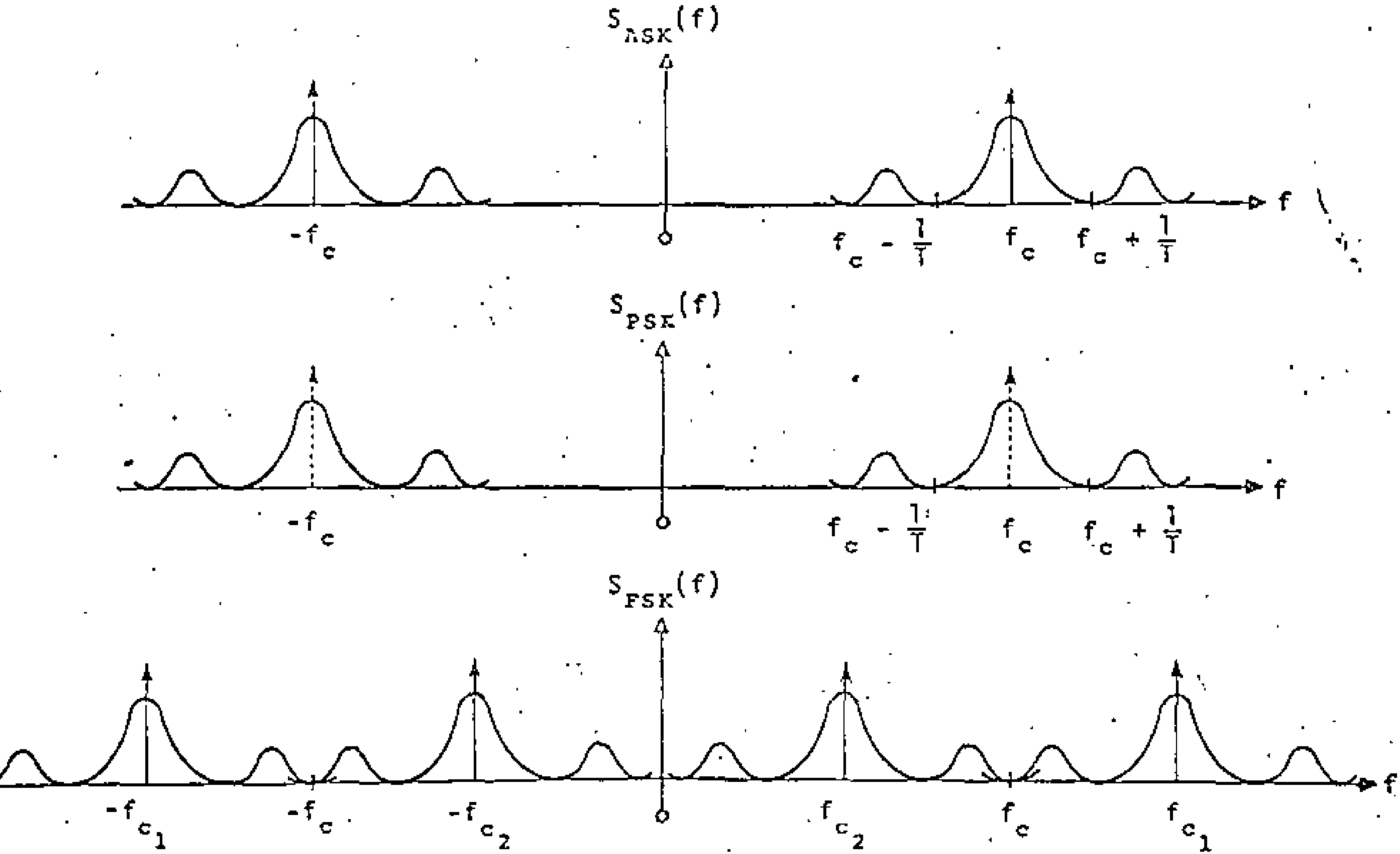


FIGURA N° 24
DIAGRAMA EN BLOQUES DE MODULADORES: a) ASK ; b) FSK ; c) PSK .

ASK, PSK, AND FSK POWER SPECTRA



Cuando la señal modulada es recibida, debe ser demodulada para así recobrar la señal original de dos niveles. Ya que una señal de PSK es tanto $+\cos \omega_c t$ como $-\cos \omega_c t$ en cualquier intervalo, su demodulación puede lograrse al detectar el signo en cada intervalo del tiempo. Esto es enteramente equivalente a detectar su fase. Un demodulador de señales PSK es mostrado esquemáticamente en la figura 24a. Opera al multiplicar la señal de entrada por la señal $\cos \omega_c t$. La señal de referencia debe estar en fase con la portadora sin modular como serie recibida si se transmitiera al receptor. La salida del multiplicador es

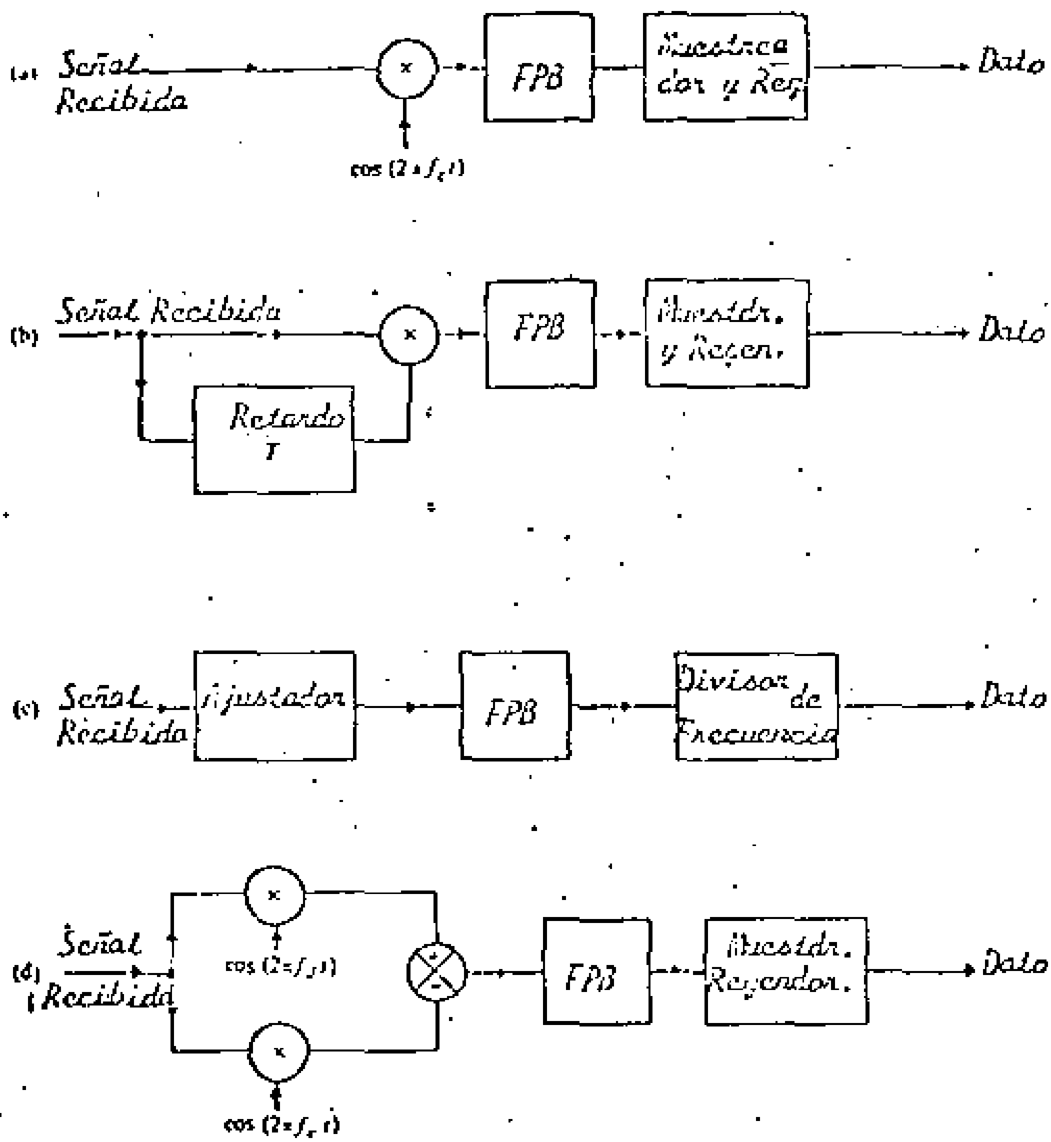
$$\pm x(t) \cos^2 \omega_c t = \pm \frac{x(t)}{2} \left\{ 1 + \cos 2 \omega_c \zeta \right\} \qquad (6)$$

donde el signo depende del signo de la señal modulada. Cuando esta señal de salida es filtrada por un filtro para bajas obtendremos $\pm x(t)$.

Observese que para ASK, $x(t)$ es 1 ó 0 y para PSK es ± 1 por lo que para ASK utilizamos el mismo diagrama. Este tipo de demodulación es llamada detección síncrona o coherente, debido a que la frecuencia local debe ser igual a la frecuencia de la señal recibida.

Un tipo alternativo de demodulador para señales PSK es el demodulador coherente diferencial (fig. 25b). Este tipo de demodulador evita el uso de señal de referencia al comparar la señal en cada intervalo de tiempo con esa del intervalo

FIGURA Nº 25 DEMODULADORES FSK y PSK



de tiempo con esa del intervalo anterior. El diagrama de bloques del demodulador para señales FSK es mostrada en la fig. 25c. Este demodulador requiere dos señales de referencia como se muestra.

En cualquier intervalo de tiempo la señal de FSK es tanto $\cos \omega_d t$ como $\cos \omega_c t$, y un análisis similar al que se hizo para PSK muestra que la entrada al filtro paso-bajas es tanto

$$x(t) \cos^2 \omega_d t = x(t) \cos \omega_d t \cos \omega_c t$$

ó (7)

$$x(t) \cos \omega_d t \cos \omega_c t - x(t) \cos^2 \omega_c t$$

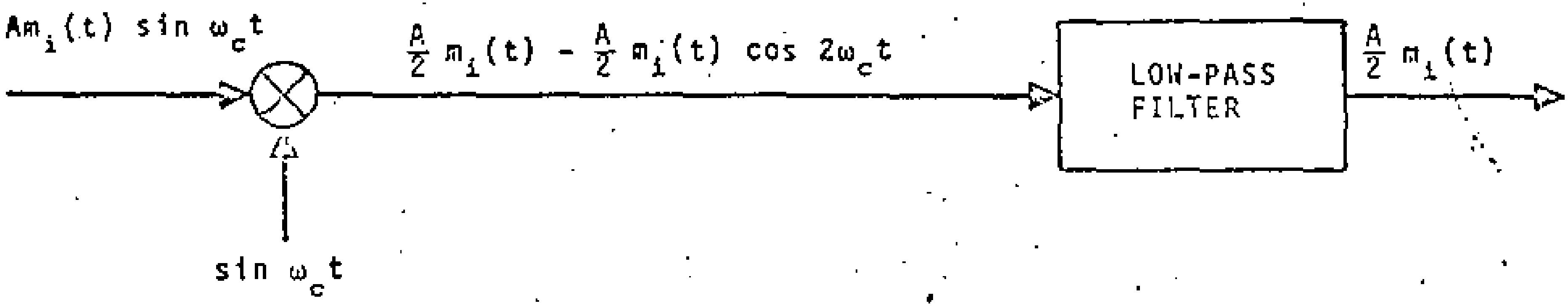
de tal forma que a la salida del filtro tendremos solamente f(t).

La otra forma común de detección, detección de envolvente, evita problemas de tiempo y de fase de la detección síncrona, Aquí la señal de entrada de alta frecuencia pasa a través de un dispositivo no lineal y un filtro para bajas (fig. 26). Sin embargo existe una desventaja. La señal PSK tiene una envolvente constante (fig. 23), tal que no puede usarse un detector de envolvente. Así que el sistema PSK requiere detección síncrona.

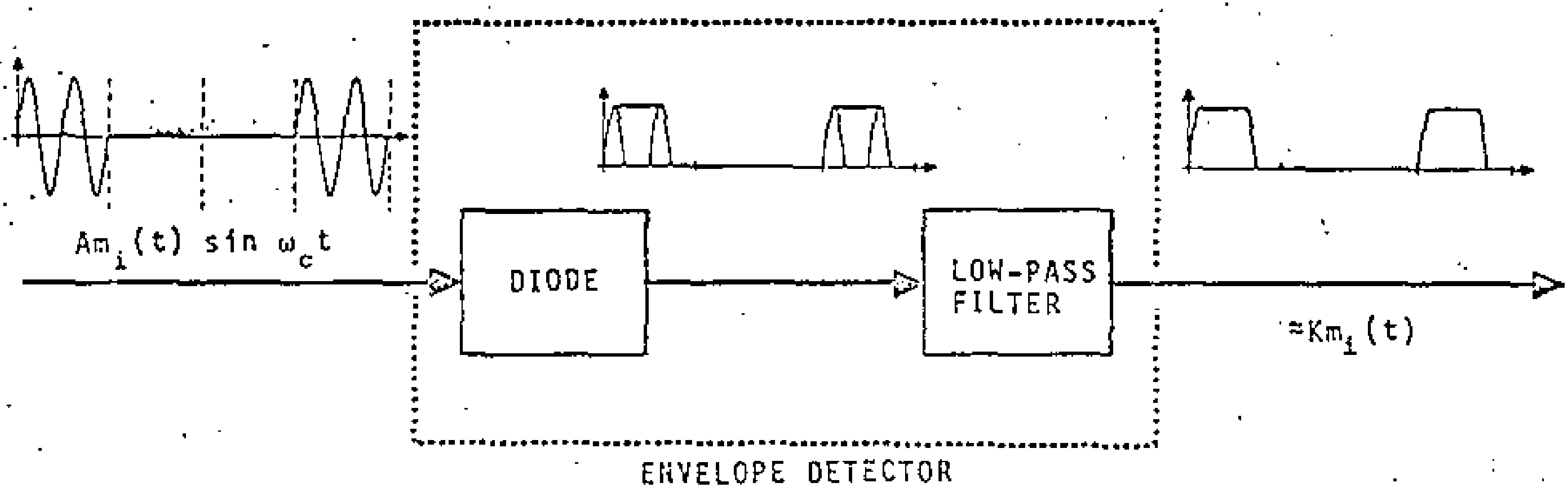
Para concluir la discusión de señalización binaria, mostramos en la fig. 26a un diagrama completo de un sistema PCM.

DEMODULATION - ASK

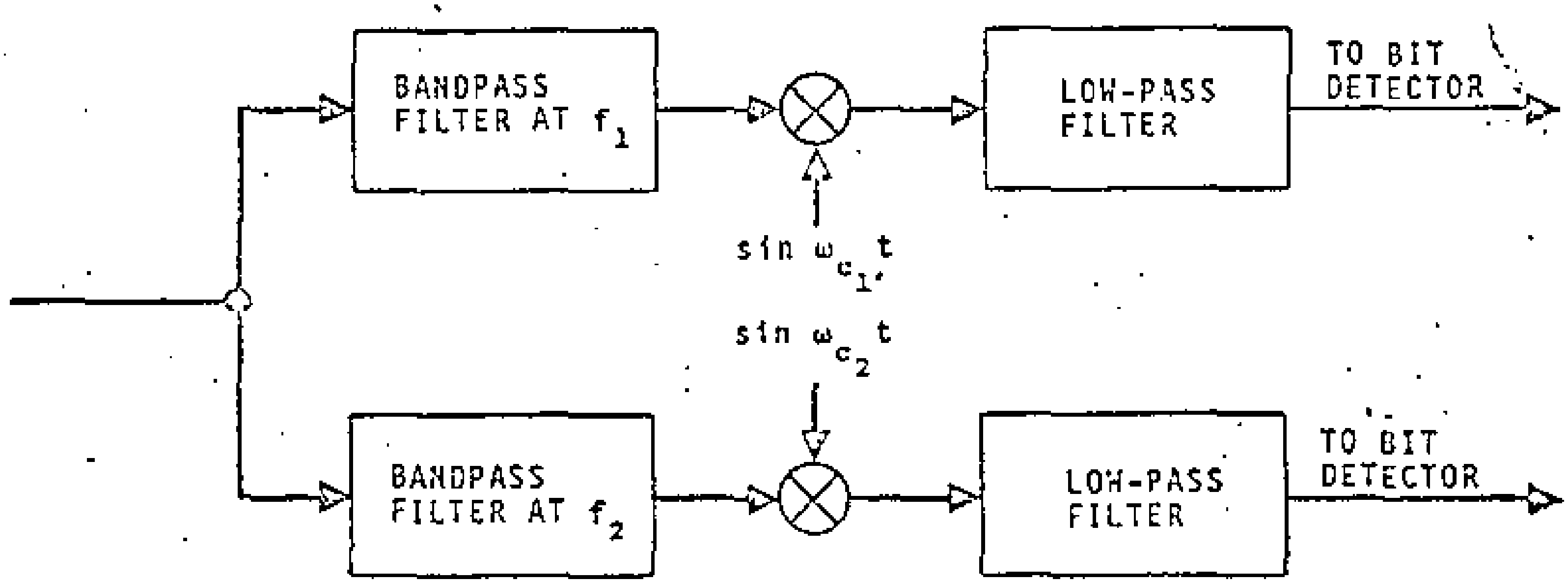
COHERENT



INCOHERENT

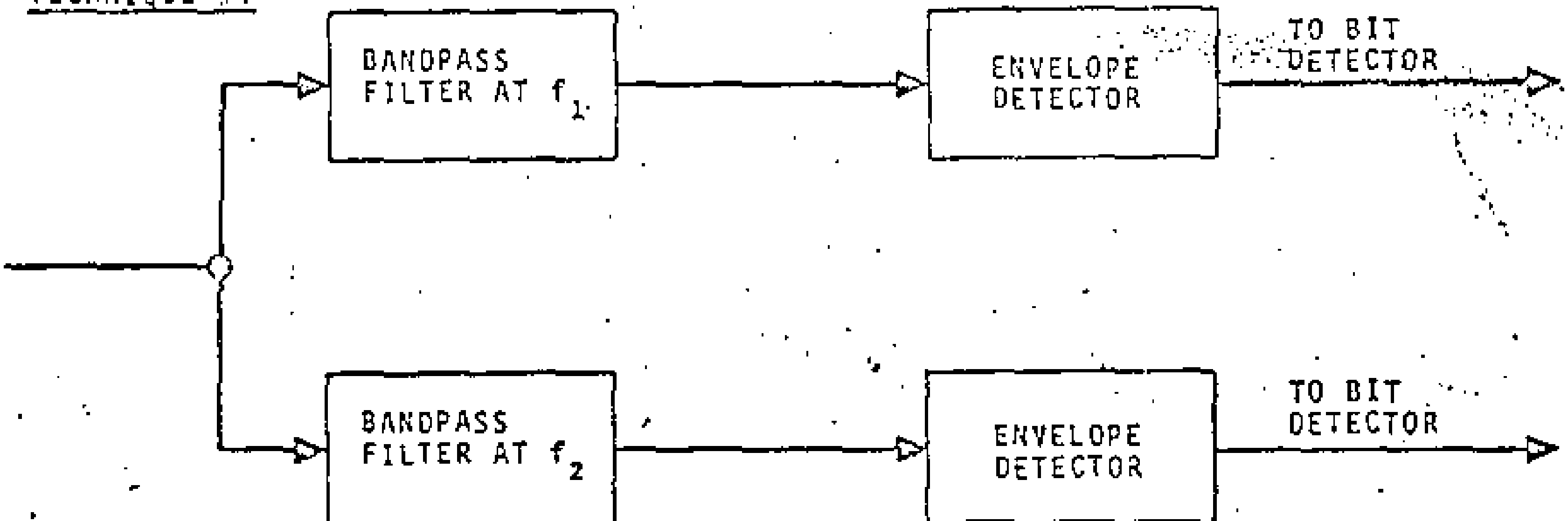


COHERENT DEMODULATION - FSK

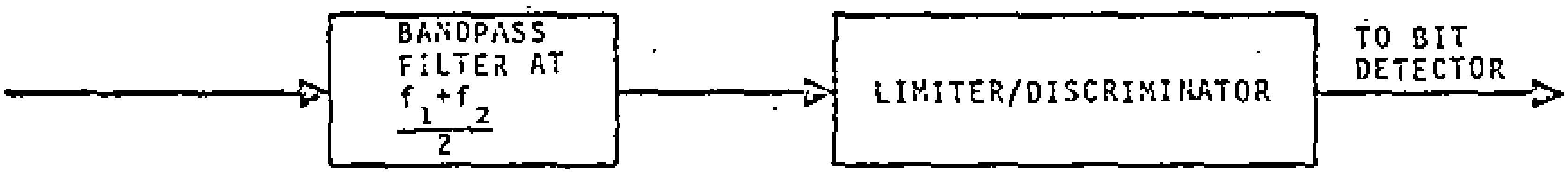


INCOHERENT DEMODULATION -FSK

TECHNIQUE #1

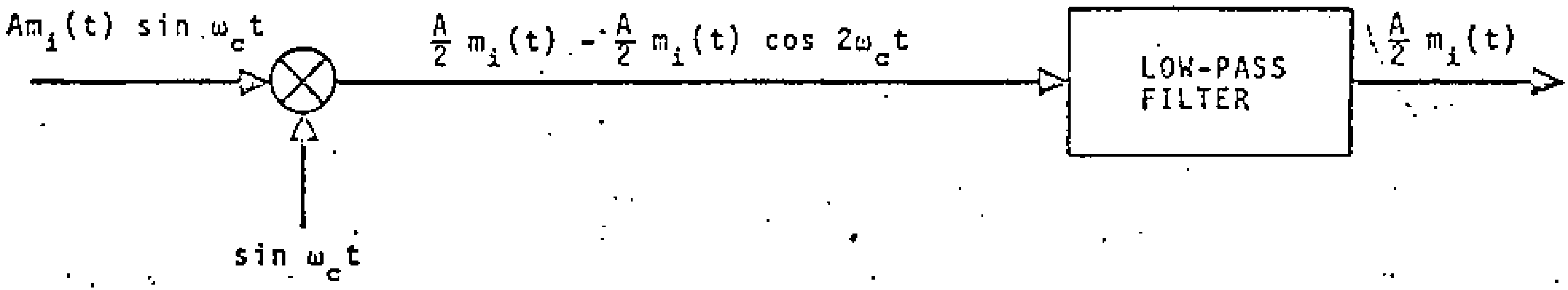


TECHNIQUE #2

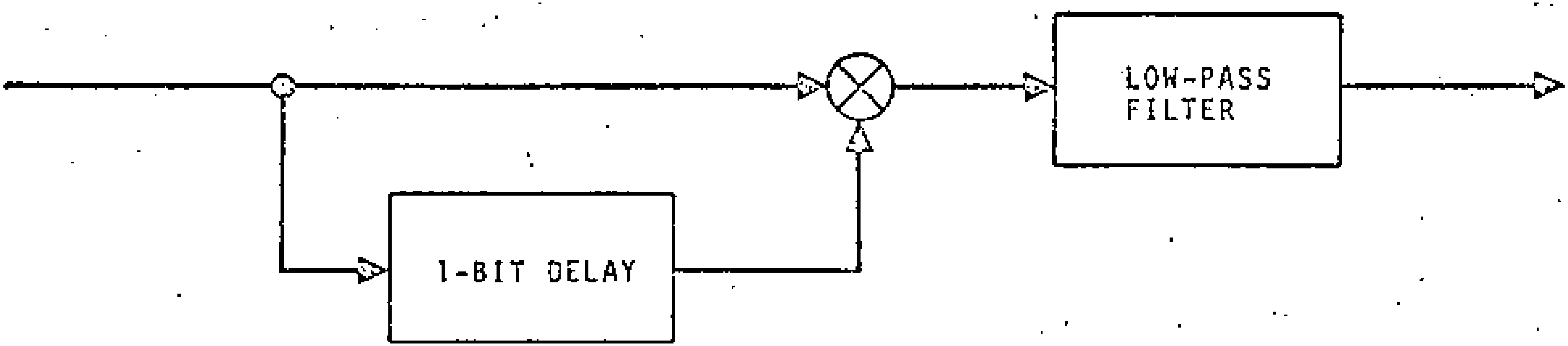


DEMODULATION - PSK

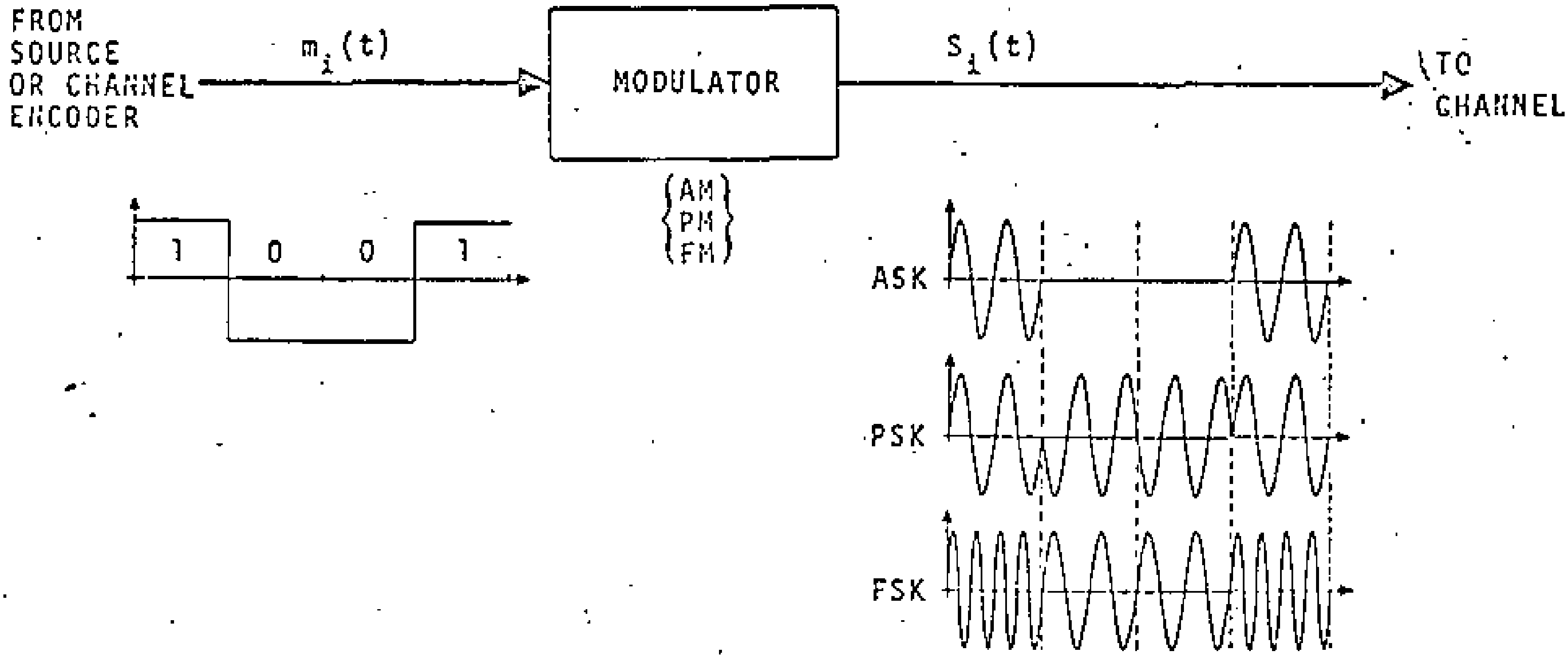
COHERENT



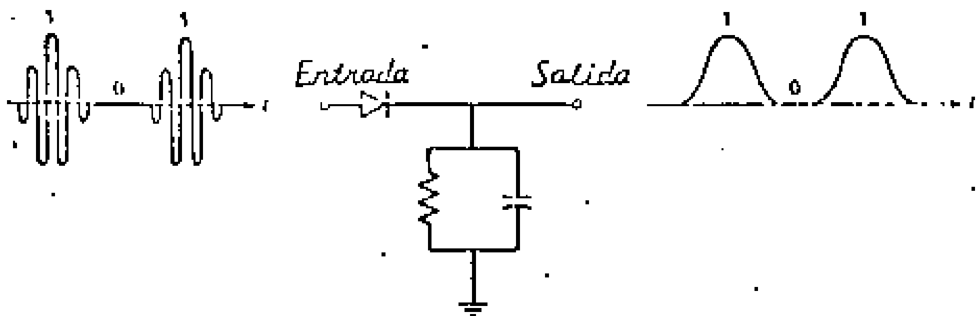
DIFFERENTIALLY COHERENT



BINARY SIGNALING



Detector Síncrono



Detector de Envolvente

FIGURA N° 26

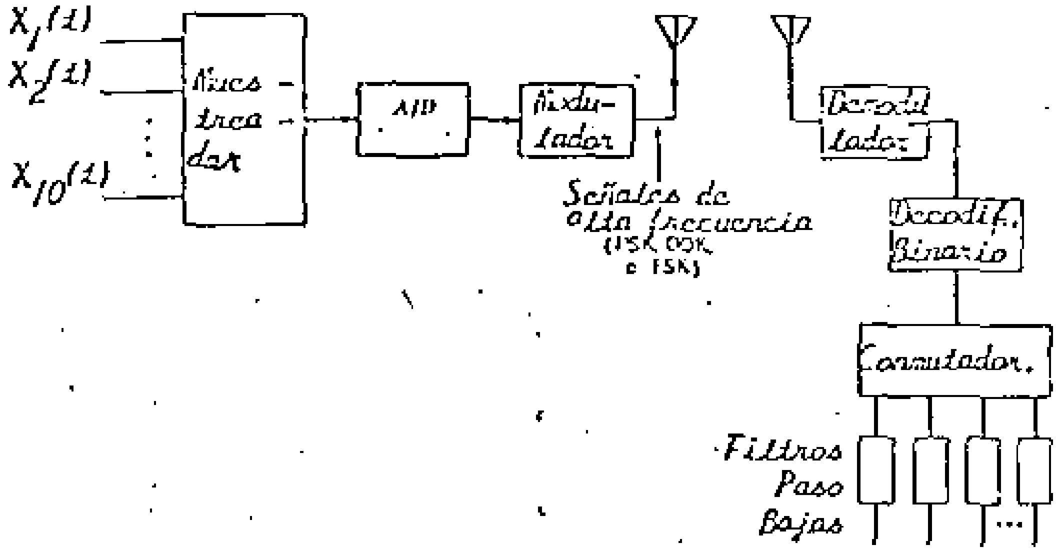


FIGURA N° 26 a
Sistema PCM Completado

Este incluye la circuitería A/D, el modulador, el cual produce las señales binarias de alta frecuencia, en el receptor, el demodulador, el cual incluye un detector síncrono ó de envolvente, un decodificador binario, un conmutador, o circuito de conmutación para ordenar las señales multiplicadas en el tiempo, y finalmente un filtro paso bajas, a la salida de cada canal, para proveer las señales de salida finales.

TECNICAS DE MODULACION PARA COMUNICACIONES DIGITALES: SEÑALIZACION MULTISIMBOLA.

En las secciones anteriores hemos puesto nuestra atención en las formas más simples de sistemas de portadora digitales, esas que involucran modulación binaria en amplitud, fase o frecuencia. En los sistemas PCM vimos que los requerimientos de ancho de banda estaban ligadas con la relación de Nyquist. Se vió que si un conjunto de $M = 2^n$ símbolos, es usado, con n el número de dígitos binarios sucesivos combinados para formar el símbolo apropiado para ser transmitido, 2 n bits/s/Hz pueden ser transmitidos utilizando la banda de Nyquist.

En esta parte, discutiremos especificamente esquemas de señalización de multifase, multiamplitud y multifase/multiamplitud combinadas como ejemplos de sistemas multisímbolos. Estos sistemas no son otra cosa más que una combinación sucesiva de pulsos binarios para formar un pulso más largo que requiere un ancho de banda menor, Como primer ejemplo de un

ESQUEMA MULTISIMBOLO

Considere un sistema en el cual dos pulsos sucesivos binarios se combinan y el conjunto resultante de cuatro pares binarios, 00, 01, 10, 11, se usa para generar una onda senoidal de alta frecuencia de cuatro posibles fases, una por cada par binario. Esto es una extensión obvia para transmisión PSK binaria de cuatro fases. La i-ésima señal, de las cuatro posibles, puede escribirse

$$S_i(t) = \cos(w_c t + \theta_i)$$

$$i = 1, 2, 3, 4 \qquad -\frac{T}{2} \leq t \leq \frac{T}{2} \qquad (8)$$

con forma rectangular considerada hasta este punto por simplicidad. Así, esto extiende la representación binaria de la ecuación (5).

Las posibles elecciones para los cuatro ángulos de fase son

$$\theta_i = 0, \pm \frac{\pi}{2}, \pi \qquad (9)$$

$$\theta_i = \pm \frac{\pi}{2}, \pm \frac{\pi}{4} \qquad (10)$$

En ambos casos las fases son espaciadas $\pi/2$ radianes.

Las señales de este tipo son llamadas PSK cuaternario (QPSK). Estas señales son un caso especial de multi-PSK (MPSK). Las señales PSK son algunas veces clasificadas también como BPSK.

En general, como ya se dijo, n pulsos binarios sucesivos son acumulados y uno de los $M= 2^n$ símbolos es retirado. Si la razón binaria es R bits/s, cada intervalo de pulso binario es $\frac{1}{R}$ segundos.

El símbolo correspondiente de salida es entonces $T= \frac{n}{R}$ segundos. Las señales de la ecuación (8) pueden ser representadas, por expansión trigonométrica, en la forma siguiente:

$$S_i(t) = a_i \cos w_c t + b_i \operatorname{sen} w_c t \; : \; -\frac{T}{2} \leq t \leq \frac{T}{2} \qquad (11)$$

para el caso de la ecuación (9), en que los pares (a_i, b_i) sean dados, correspondiendo respectivamente a los ángulos

$\theta_i = 0, -\frac{\Pi}{2}, \Pi$, y $\Pi/2$, por

$$(a_i, b_i) = (1,0), (0,1), (-1,0), (0,-1) \qquad (12)$$

El correspondiente conjunto de $(a_i, b_i,)$ para (10), esta dado por

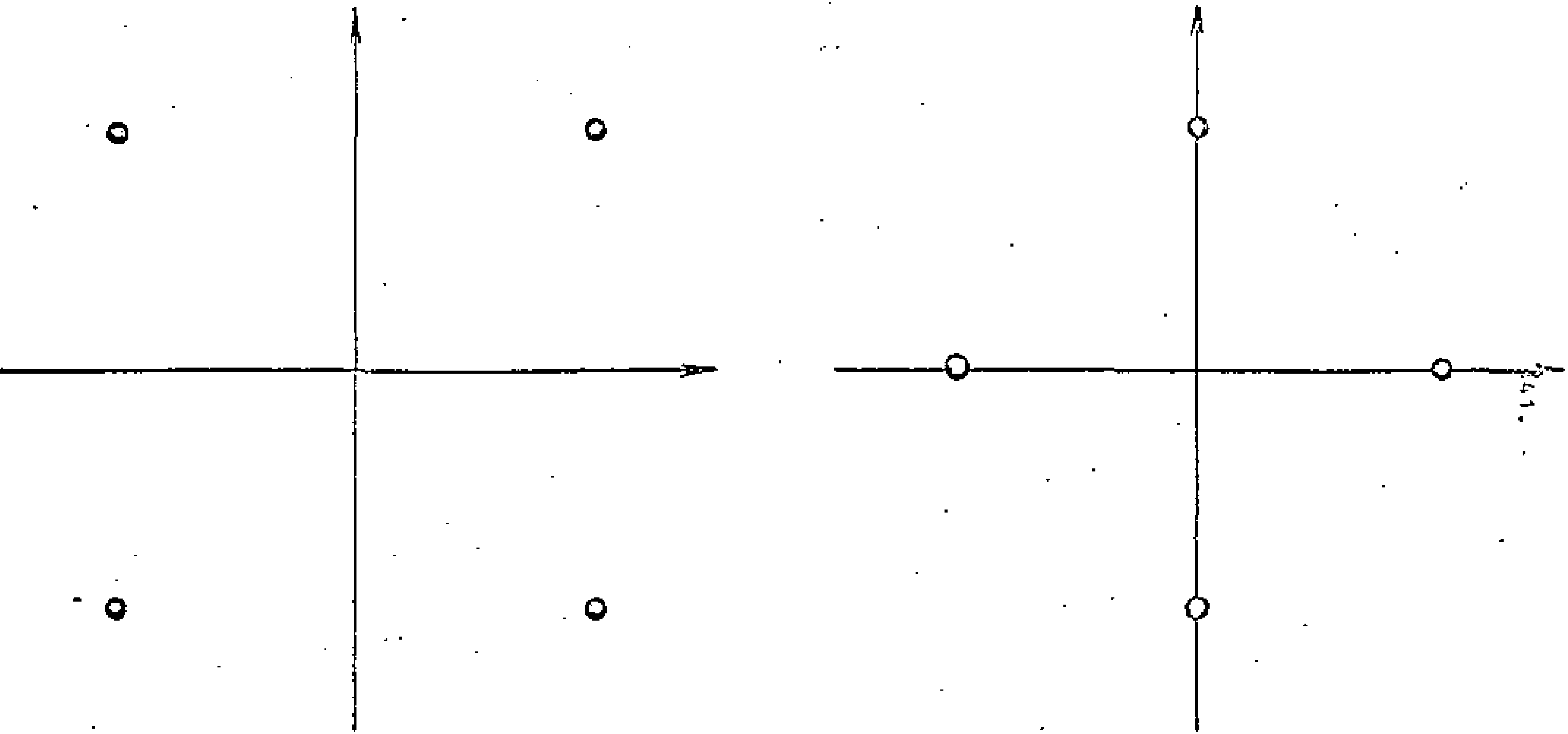
$$(\sqrt{2}\, a_i, \sqrt{2}\, b_i) = (1,1), (-1,1), (-1,-1), (1,-1) \qquad (13)$$

La transmisión de este tipo es frecuentemente llamada transmisión de cuadratura, con dos portadoras en cuadratura de fase una a otra ($\cos w_c t$ y $\operatorname{sen} w_c t$) transmitidas simúltaneamente sobre el mismo canal.

Es útil representar las señales de (11) en un diagrama de dos dimensiones al localizar los diferentes puntos (a_i, b_i). El eje horizontal correspondiente a la localización de a_i es llamado componente en fase y el vertical, en el cual b_i esta localizada se llama componente en cuadratura. Las cuatro señales de (12) se muestran en la fig. 27a., las de la ec. (13) se ilustran en la fig. 27b.

La representación en fase (coseno) y en cuadratura (seno) de las señales QPSK $s_i(t)$ sugiere un posible camino de generar estas señales. Dos pulsos de entrada binarios sucesivos son acumulados y el par de números (a_i, b_i), tomados cada $T = \frac{2}{R}$ segundos, es utilizado para modular dos términos de portadora en cuadratura, cos $w_c t$ y sen $w_c t$, respectivamente, donde uno de los números es cero, esa portadora esta de seguro imposibilitada. Un modulador de este tipo es mostrado en la fig. 28.

Es evidente que la demodulación es llevada a cabo al usar dos detectores síncronos en paralelo, uno en cuadratura con el otro. Un diagrama de bloques de tal demodulador aparece en la fig. 29.



(a) (b)

FIGURA No. 27

CONFIGURACION DE SEÑALES QPSK.

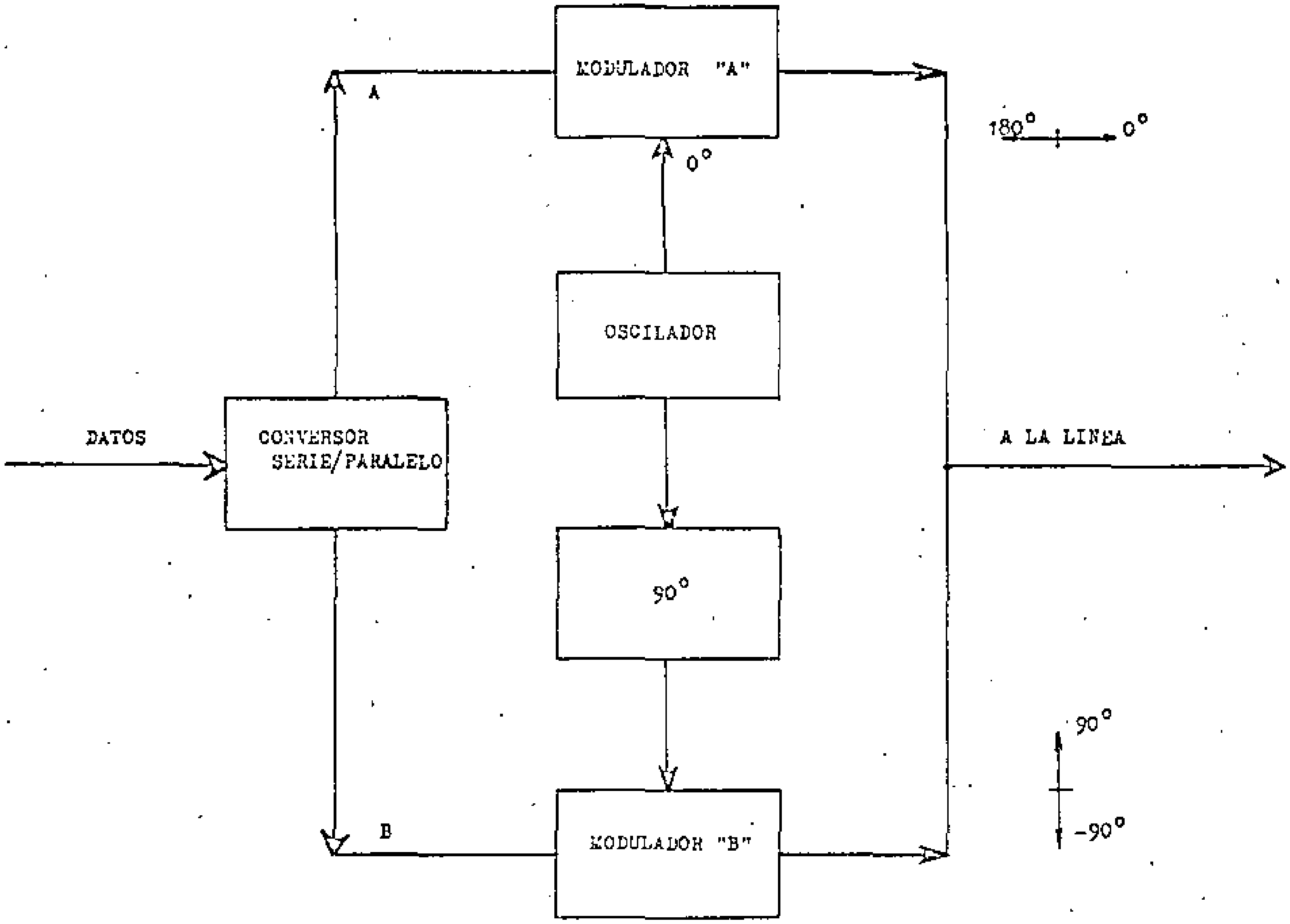


FIGURA No. 28
GENERACION DE SEÑALES QPSK

SEÑALIZACION MULTISIMBOLA

01

00, 01, 10, 11 ⟶ PSK

$$Si(t) = Cos\ (w_c t + \theta_i)$$

$i = 1,2,3,4 \qquad -T/_2 \leq t \leq T/_2$

ó $\theta_i = 0, \pm \pi/_2, \pi$

$\theta_i = \pm\ \pi/4, \pm\ \frac{3\pi}{4}$

$$Si\ (t) = a_i \cos w_c t + bi\ Sen\ W_c t$$

para $\theta_i = 0, -\pi/2, \pi, \pi/2$

$(a_i, b_i) = (1,0), (0,1) (-1,0), (0,-1)$

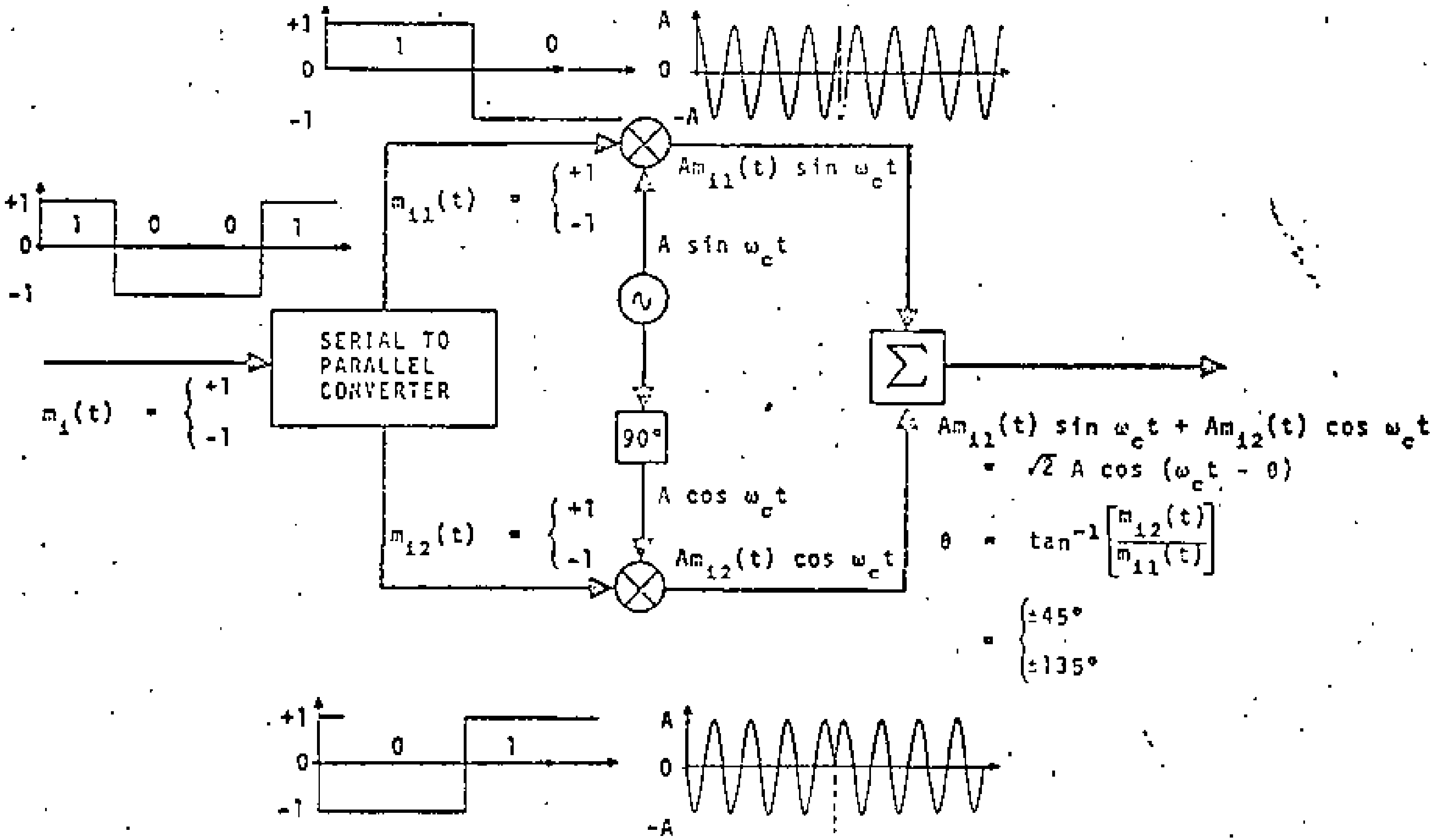
para $\theta_i = \frac{\pi}{4}, -\frac{\pi}{4}, \frac{3\pi}{4}, -\frac{3\pi}{4}$

$(\sqrt{2}\, a_i, \sqrt{2}\, b_i) = (1,1), (-1,1), (-1,-1), (1,-1)$

⟶ PSK Cuaternaria (QPSK)

⟶ Caso especial de PSK Multiple (MPSK)

QUADRIPHASE MODULATION



Mas tipos generales de esquemas de señales de múltiple nivel pueden ser generadas dejando que a_i y b_i en (11) tomen múltiples valores.

Las señales resultantes son llamadas señales de modulación en amplitud en Cuadratura. Estas señales pueden interpretarse como que tienen modulación en amplitud de múltiple nivel aplicada independientemente en cada una de las dos portadoras de cuadratura. El demodulador de la fig. 29 con un detector síncrono, puede entonces usarse para recobrar la información digital deseada.

SISTEMAS 8-PSK

La técnica de modulación 8-PSK puede ser vista como una extensión del sistema QPSK. En el diagrama de bloques del modulador clásico 8-PSK mostrado en la fig. 30., la tasa de datos f_b es dividida en tres flujos paralelos binarios, cada uno teniendo una tasa de transmisión de $f_b/3$. El convertidor de 2 niveles a cuatro produce uno de los cuatro posibles niveles de una señal polar de banda base en a y b. Si el símbolo binario A es un lógico (cero), entonces el nivel de salida a tiene uno de los dos posibles estados (positivo ó negativo). El estado lógico del bit C determina si el nivel más largo ó mas pequeño de la señal debe estar presente en a ó en b. Cuando C = 1, entonces la amplitud de a es mayor que la de b; si C = 0 entonces el proceso inverso es verdadero. Las señales de banda base polares de 4 nive

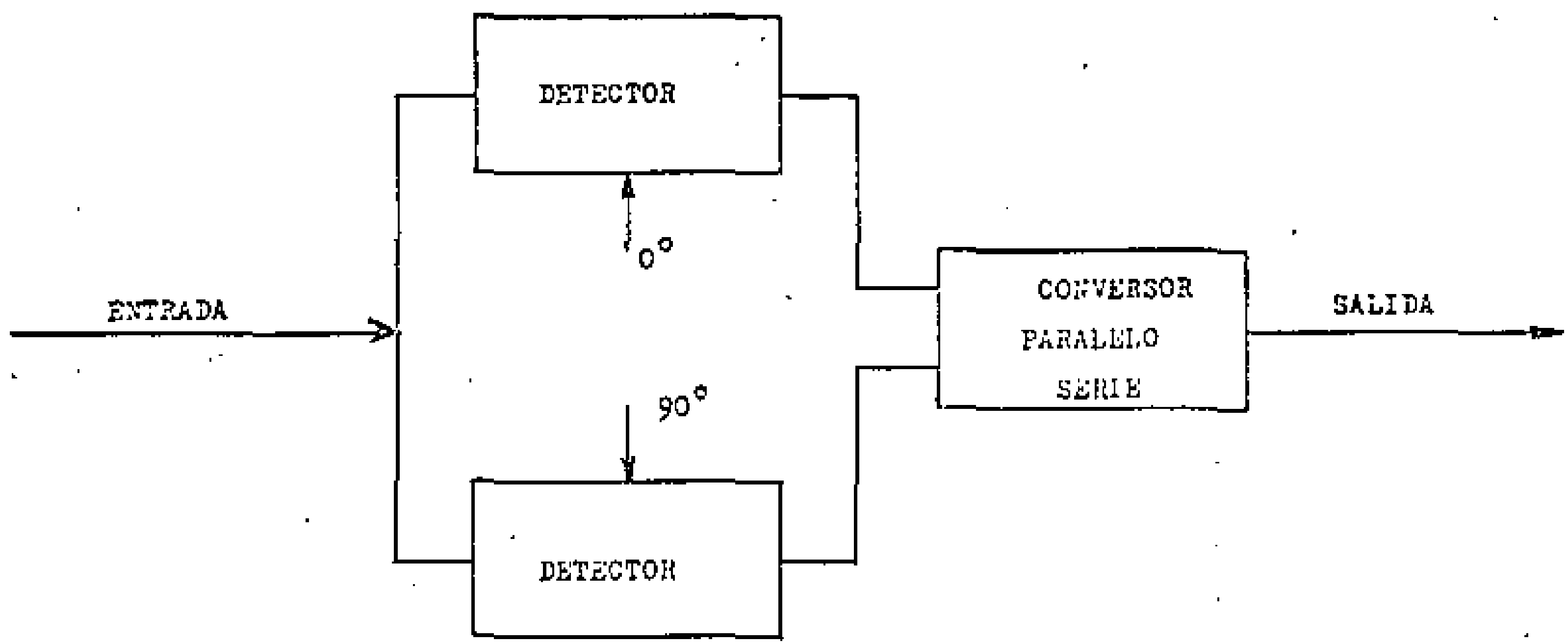


FIGURA No. 29
DEMODULADOR QPSK

les en a y b son utilizadas para modular en amplitud (doble banda lateral con portadora suprimida) las dos portadoras en cuadratura.

Una moderna aproximación en el diseño de un modulador 8-PSK para alta velocidad (90 M_b/s), usando solamente dispositivos digitales ha sido discuitod en referencias. El principio de operación de tal modulador es ilustrado en la fig. 31. La tasa de información binaria de banda base es convertida de serie a paralelo en la unidad distribuidora de datos. Estos flujos paralelos de datos de tasa $f_b/3$ conmutan en encendido o apagado las compuertas lógicas del multicanalizador conmutativo IF de alta velocidad. Dependiendo de los estados lógicos de banda base, uno de los ocho vectores digitales IF es conectado a la salida digital IF.

Esta portadora digital defasada en fase 8 PSK es filtrada por medio de un filtro paso banda convencional; así, una señal 8-PSK limitada en banda es obtenida. La fig. 32 muestra la digitalmente implementada, 90 M_b/s, 8-PSK tarjeta de circuiteria impresa usada por Raythson Data Systems en sus sistemas de microondas de 6 y 11 GHz.

La constelación para una señal QAM de 11 estados aparece en la fig. 33. Note que esta señal puede considerarse como si se generara por dos señales moduladas en amplitud en cuadratura. Ya que cuatro niveles de amplitud son usados en cada una de las portadoras, la señal es algunas veces refe rida como una señal QAM de cuatro niveles. Todos los puntos en la constelación son igualmente espaciados.

MODEMS

Los modems han sido ampliamente adoptados para la transmisión de datos digitales sobre varios medios de transmisión. El ejemplo de un modem PSK de cuatro fases para transmisión digital sobre un canal de 38 KHz en el sistema de satelite SPADE es clásico para mostrar la aplicación de los modems. Un diagrama de bloques simplificado de una combinación transmisor-receptor QAM se ilustra en la fig. 34.

Para una tasa de transmisión de alta velocidad sobre la linea telefónica, señalización de niveles múltiples debe de usarse.

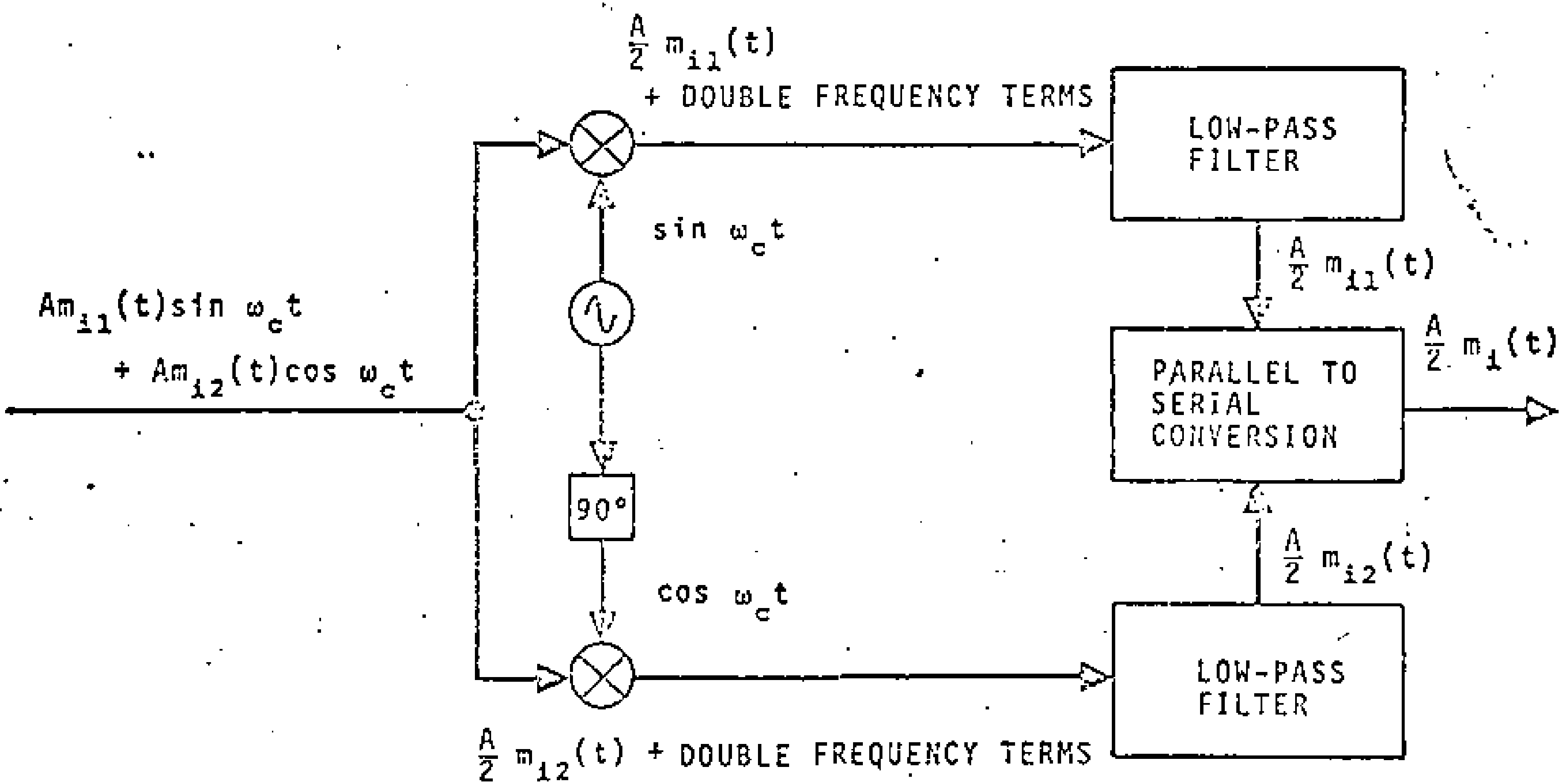
Ejemplos de tres constelaciones y sus correspondientes espectros de transmisión, usados en modems de 2400, 4800 y 9600 bits/s respecitvamente, aparecen en la fig. 35. Los espectros de amplitud mostrados están en la escala de decibeles.

EFECTOS DE RUIDO

Señales de banda base

Un oscilograma típico del voltaje de ruido $n(t)$ se ilustra en la fig. 36. Aunque el ruido es considerado aleatorio tal que no se pueden especificar por adelantado valores particular de voltaje como una función del tiempo, se puede sin em

QUADRIPHASE DEMODULATION (COHERENT)



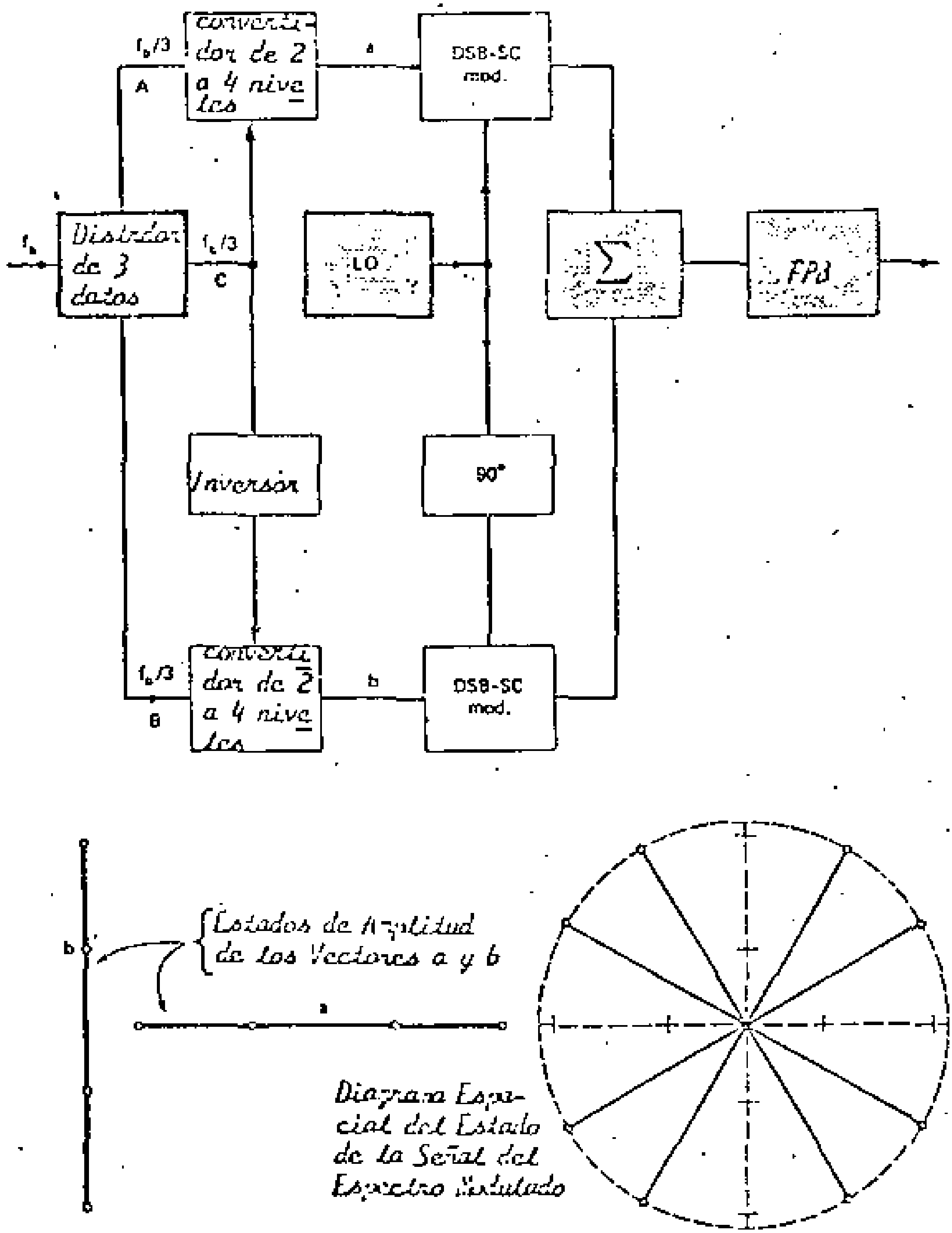


FIGURA N° 30
Modulador PSK Clásico de 8 Fases y Diagrama de Estados de Amplitud.

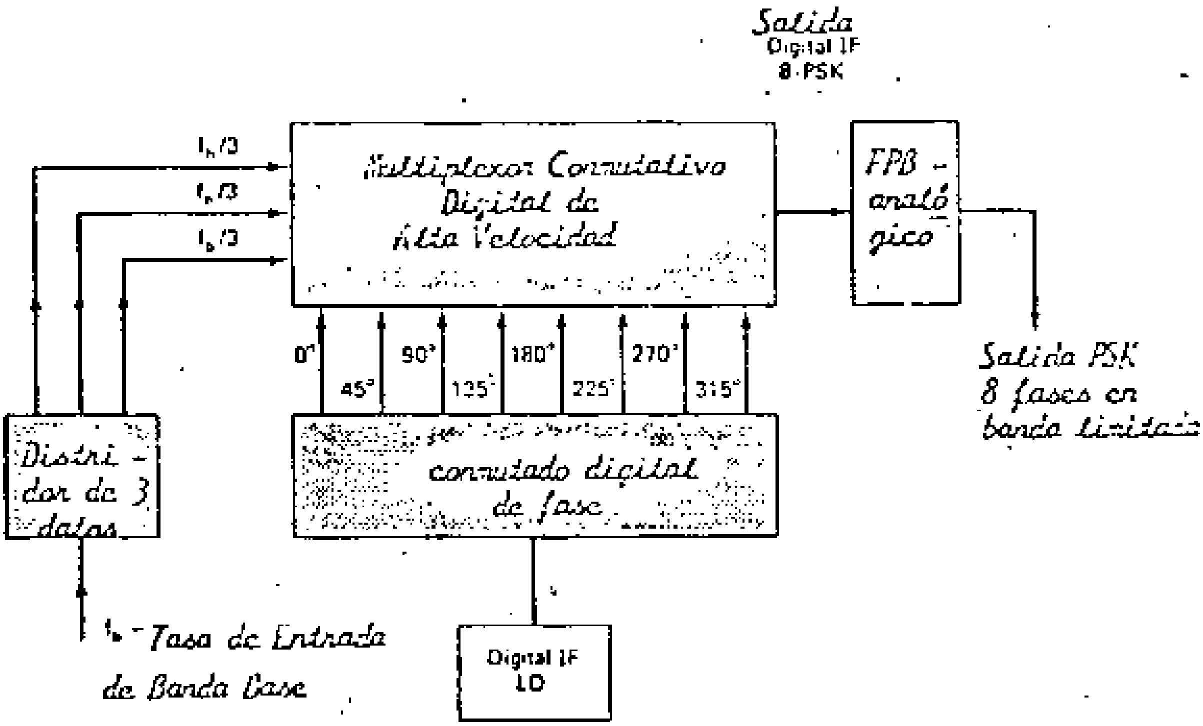


FIGURA N° 31
Modulador PSK de 8 Fases y Alta Velocidad.

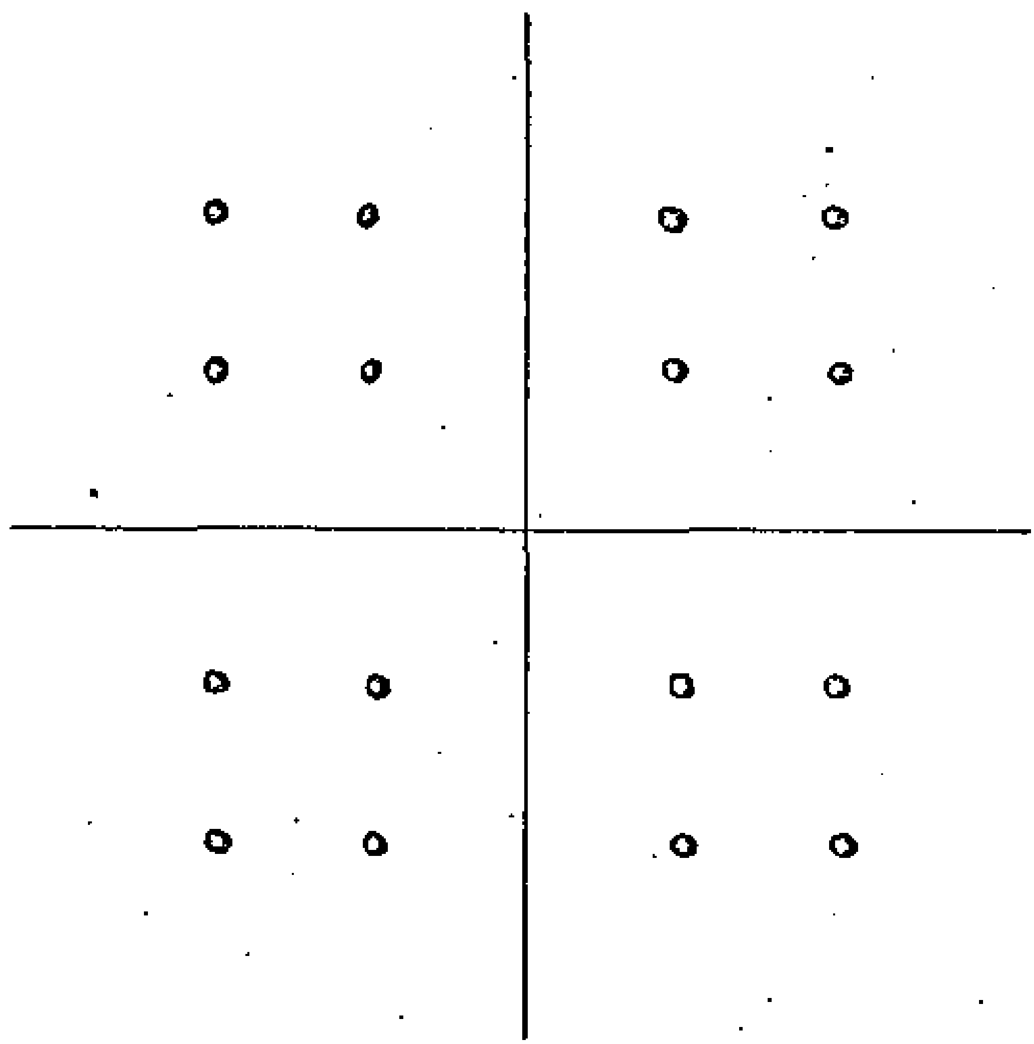


FIGURA No.33

CONFIGURACION QAM DE 4 NIVELES(16-SIMBOLOS)

TECNICA DE MODULACION DIGITAL CONTRA EFICIENCIA ESPECTRAL

TIPO DE MODULACION	NUMERO DE NIVELES LOGICOS	NUMERO DE BITS POR SIMBOLO	ANCHO DE BANDA
A S K	2	1	$B_T = 2B$
F S K	2	1	$B_T = 2B + 2\delta f$
P S K	2	1	$B_T^b = 2B$
4-P S K	4	2	$B_T^{4\phi} = \frac{1}{2} B_T^b$
8- P S K	8	3	$B_T^{8\phi} = \frac{1}{3} B_T^b$
16- P S K	16	4	$B_T^{16\phi} = \frac{1}{4} B_T^b$
Q A M	16	4	$B_T^{QAM} = \frac{1}{4} B_T^b$

MODULACION	VELOCIDAD (bits/s)
F S K	1200
4-P S K	2400
8-P S K	4800
16-P S K	9600

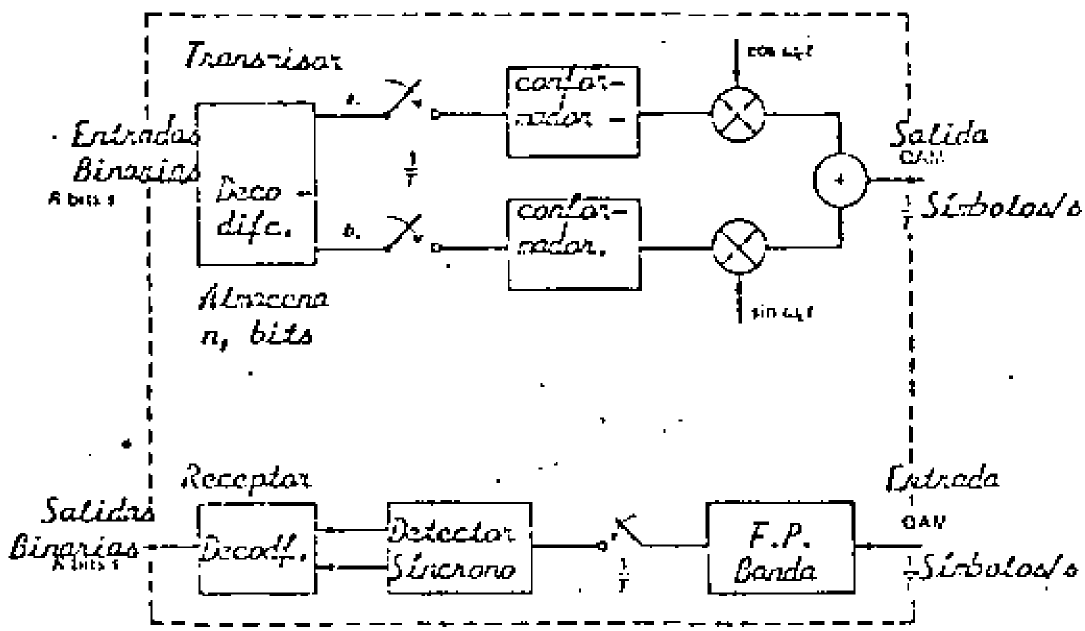


FIGURA Nº 34
Diagrama Simplificado de un Modem QAM.

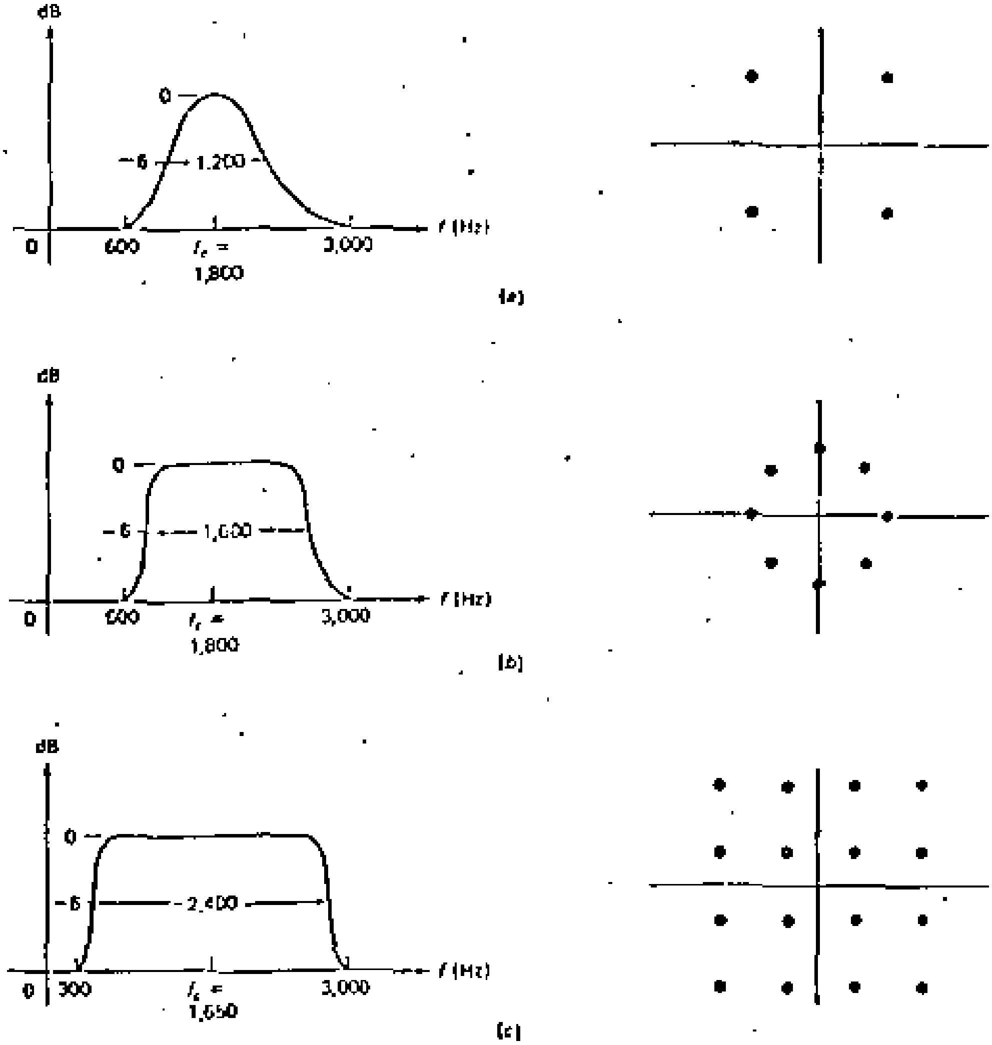


FIGURA N° 35

Espectro y Constelación para Modems de Alta Velocidad: a) 2400 bits/s, PSK de 4 fases, característica cosenoidal; b) 4800 bits/s, PSK de 8 fases, 50 % de factor de conformación; c) 9600 bits/s, QAM, 16 estados, 10 % de factor de conformación.

TASA DE ERROR EN TRANSMISION BINARIA

La probabilidad de que una muestra medida $n(t_1)$ caiga en el rango de n a $n + dn$ está dada por $f(n)\,dn$, con

$$f(n) = \frac{e^{-n^2/2\sigma^2}}{\sqrt{2\pi\sigma^2}}$$

$0 \longrightarrow 1$

error $\quad Pe \longrightarrow P_{Ruido} > A/2$

$1 \longrightarrow 0$.

Si un 0 está presente $v(t) = n(t)$.

Así la función de densidad para v, asumiento que un cero esta presente, es

$$fo(v) = \frac{1}{\sqrt{2\pi\sigma^2}} \; e^{-v^2/2\sigma^2}$$

$$P_{e_0} = \text{Prob } (v > A/2) = \int_{A/2}^{\infty} fo(v)dv$$

Si un 1 se transmite

$$v(t) = A + n\,(t)$$

$$f_1(v) = \frac{1}{\sqrt{2\pi\sigma^2}} \, e^{-(v-A)/2\sigma^2}$$

$$P_{e_1} = \text{Prob}\,(v < A/2) = \int_{-\infty}^{A/2} f_1(v)\,dv$$

Probabilidad total del sistema ?

P_{e_0}, P_{e_1} } Probabilidades condicionales

P_0 y P_1 eventos mutuamente exclusivos

$$\longrightarrow (P_0 + P_1)$$

$$P_e = P_0 P_{e_o} + P_1 P_{e_1}$$

$$P_e = \frac{1}{2}\left(1 - \text{erf} \frac{A}{2\sqrt{2}\,\sigma}\right)$$

donde

$$\text{erf} = \frac{2}{\pi} \int_0^x e^{-y^2}\,dy$$

bargo, asumir que se conocen las características estadísticas del ruido. En particular se considera que el ruido tiene una función de probabilidad gausiana, con $E(n) = 0$. Específicamente, si se muestrea el ruido en cualquier tiempo arbitrario t_1 la probabilidad de que la muestra medida $n(t1)$ caiga en el rango de n a $n+dn$ esta dada por $f(n)\,dn$, con

$$f(n) = \frac{e^{-n^2/2\sigma^2}}{\sqrt{2\Pi\,\sigma^2}} \qquad (1)$$

Este es el modelo estadístico más usado para ruido aditivo en comunicaciones, y es en la mayoria de aplicaciones, una representación válida para el ruido real presente.

Se considera que la varianza del ruido σ^2 es conocida (puede ser medida). La función se muestra en la fig. 37. En este capítulo, analizaremos la probabilidad de error al tomar un nivel de ruido en lugar de señal y viceversa.

Considere que en un sistema binario la amplitud de los pulsos es A volts. La secuencia compuesta de simbolos binarios más ruido es muestreada una vez cada intervalo binario y se hace una decisión si un 1 ó un 0 esta presente. Una simple forma particular de hacer la decisión es decidir un 1 si el voltaje compuesto es mayor que A/2 volts, y 0 si la muestra es menor que A/2 volts.

Ocurriran errores si, con un pulso presente la muestra de voltaje compuesto es menor que A/2, o, con un pulso ausente, si el ruido solo excede a A/2.

Un ejemplo de una posible secuencia de señal, indicando los dos posibles tipos de error, es mostrada en la fig. 38.

Para determinar la probabilidad de error cuantitativamente se consideran los dos posibles tipos de error separadamente. Considerese primero que un cero es enviado, tal que ningún pulso esta presente al tipo de decodificar. La probabilidad de error en este caso es justamente la probabilidad de que el ruido exceda la amplitud A/2 y sea equivocado por un pulso ó un 1 en el código binario. De la misma forma ya que v(t) = = n(t) si un 0 esta presente, el valor muestreado v es una variable aleatoria con la misma característica estadística del ruido. La probabilidad de error es entonces la probabilidad de que v aparecera entre A/2 e ∞. Así la función de densidad para v, asumiendo un cero presente, es justamente

$$f_o(v) = \frac{1}{\sqrt{2 \pi \sigma^2}} e^{-v^2/2\sigma^2} \qquad (2)$$

el índice 0 denota la presencía de un 0 y la probabilidad de error Peo en este caso es el area bajo la curva $f_o(v)$ de A/2 a ∞.

$$Peo = Prob\ (v > \frac{A}{2}) = \int_{A/2}^{\infty} f_o(v)\ dv \qquad (3)$$

la función de densidad $f_o(v)$ se muestra en la fig. 39, con

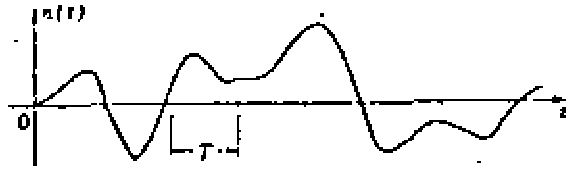


FIGURA N° 36
Típico Oscilograma de Voltaje de Ruido

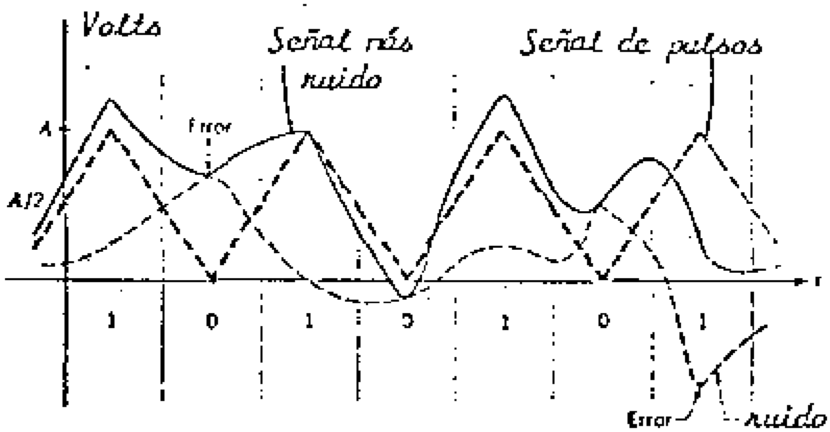


FIGURA N° 38
Efectos del Ruido en la Transmisión de Pulsos Binarios

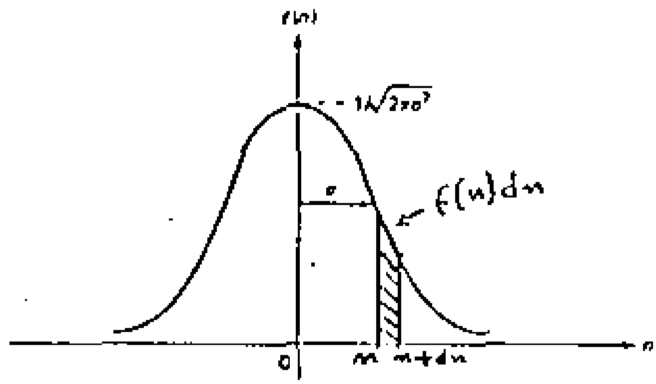


FIGURA N° 37
FUNCION DE DENSIDAD DE PROBABILIDAD
GAUSSIANA.

la probabilidad de error indicada por el area sombreada.

Considerese ahora que un 1 es transmitido. Este aparece en el decodificador como un pulso de amplitud A volts más el ruido superimpuesto. Una muestra $v(t)$ del voltaje compuesto tomado a un tiempo t es ahora una variable aleatoria $A+n(t)$. La cantidad fija A sirve para defasar el nivel del ruido de un promedio de cero volts, a un promedio de A volts. La variable aleatoria v tiene la misma estadística que n, fluctuando respecto a A, y de cualquier modo diferente de cero. Su función de densidad es la misma función gausiana, con la mis varianza, pero con un valor promedio de A. Asi, tenemos

$$f_i(v) = \frac{1}{\sqrt{2\,\Pi\sigma^2}}\, e^{-(v-A)^2/2\sigma^2} \qquad (4)$$

Esta ecuación se muestra en la fig. 39b.

La probabilidad de error corresponde ahora a la posibilidad de que la muestra v de la señal más el ruido caiga abajo de A/2 volts y sea equivocada por ruido solamente (o sea juzgado, incorrectamente, un cero). Este es justamente el area bajo la curva de $f_1(v)$ desde $-\infty$ a $A/2$ y esta dada por

$$P_{e1} = \text{Prob}\left(v < \frac{A}{2}\right) = \int_{-\infty}^{A/2} f_1(v)\, dv \qquad (5)$$

Esta probabilidad de error se indica por el área sombreada de la fig. 39b.

Es interesante preguntar como se definiria la probabilidad de error de todo el sistema. Nótese que los dos posibles tipos de error considerados pertenecen a eventos mutuamente exclusivos; el cero excluye al 1 aparentemente, y viceversa.

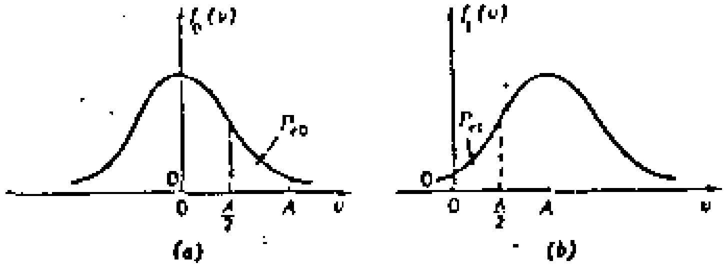


FIGURA N° 39
Densidad de Probabilidad en la Transmisión de Pulsos Binarios: a) Ruido únicamente (se ha transmitido un cero); b) Pulso más ruido (se ha transmitido un uno).

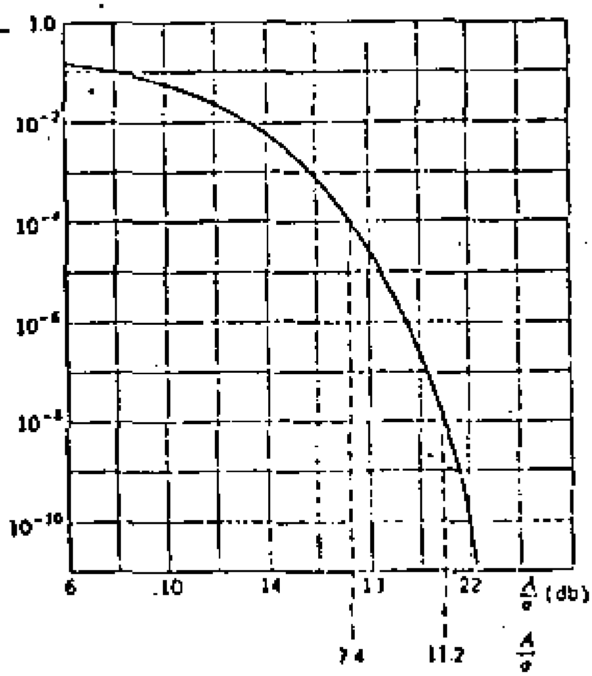


FIGURA N° 40
Probabilidad de Error por Ruido Gaussiano en la Detección binaria.

Por lo que las probabilidades pueden sumarse.

Sin embargo, en este caso, es evidente que P_{eo} y P_{e1} sean ambas probabilidades condicionales, en la primera se asume que esta presente un cero, es la segunda se considera un 1 presente.

Para remover esta condicionalidad se debe multiplicar cada una por su apropiada probabilidad de ocurrencia apriori. Así, considerando que la probabilidad de transmitir un cero es P_o, mientras que la probabilidad de transmitir un 1 es P_1, ambas conocidas tal que $P_o + P_1 = 1$, se tiene que la probabilidad de error total es

$$P_e = P_o P_{eo} + P_1 P_{e1} \qquad (6)$$

Es evidente de la fig. 39 y de la simetría de las curvas gausianas que los dos probabilidades condicionales P_{eo} y P_{e1} son iguales en este ejmplo. Como $P_o = P_1 = \frac{1}{2}$

$$P_e = \frac{1}{2}\left[1 - \text{erf}\ \frac{A}{2\sqrt{2}\ \sigma}\right] \qquad (7)$$

donde

$$\text{erf}\ x = \frac{2}{\sqrt{\pi}}\int_0^x e^{-y^2}\, dy$$

La función de error erf x definida en (1) esta tabulada en libros de estadística ó en varias tablas matemáticas. Con los 1's y 0's considerados con la misma probabilidad de ocurrencia, en un largo mensaje, la ecuación (7) dá la probabilidad de error en la decodificación de cualquier dígito. Note que la probabilidad de error P_e depende únicamente de A/σ , la relación de la amplitud de la señal a la desviación estandard del ruido. Esta cantidad σ es comúnmente referida como el ruido rms. La relación A/σ es entonces la relación señal a ruido rms. La probabilidad de error se muestra graficada contra A/σ en la Fig. 40. Es evidente que $\sigma^2 = N$ (potencia).

Ejemplo

$\frac{A}{\sigma}$ = 7.4(17.4 dB); Pe = 10^{-4}

↓

1 bit en 10^4 es tomado incorrecto

$\frac{A}{\sigma}$ = 11.2 (21 db) ; Pe = 10^{-8}

Si transmitimos 10^5 bits/s

se comete un error cada

1000 s ó 15 min.

Diseñadores usan Pe = 10^{-5} ó 10^{-6}

DETECCION DE SEÑALES BINARIAS Y RUIDO.

Si se recibe señal y ruido en el detector síncrono, tendremos que la entrada en el detector está dada por

$$v(t) = f(t) \cos \omega_o t + n \quad (t)$$

$$= \left[f(t) + x(t)\right] \cos \omega_o t - y(t) \text{ sen } \omega_o t \tag{8}$$

Para PSK $f(t) = \pm A$, para ASK $f(t)$ es $+ A$ ó 0. En el caso FSK ω_c es ω_1 ó ω_2 y $f(t)$ es A si una señal está presente en uno de los dos canales paralelos y 0 si está ausente.

En general, la salida del detector está dada por

$$V_o(t) = f(t) + x(t) \tag{9}$$

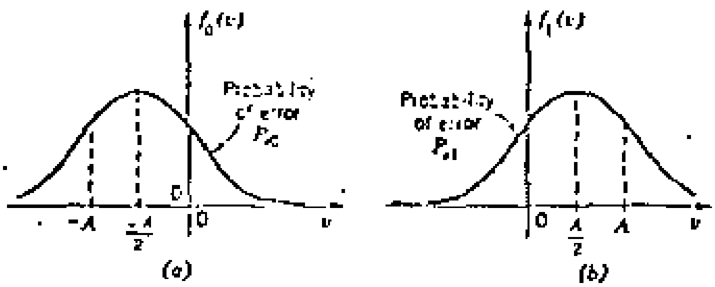
Para señales polares la Fig. 1 muestra las probabilidades de error.

NIVELES DE DECISION OPTIMOS.

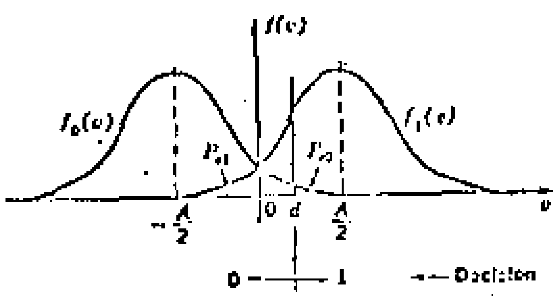
Ya que el decodificador basa su decisión en la amplitud de la señal para poder optimizar v(t), es conveniente variar el nivel de amplitud en el cual la decisión es realizada.

Si 0's ocurren más frecuentemente en el promedio ($P_0 > P_1$) , es conveniente desplazar el nivel de decisión (d) negativamente. Desde luego que el óptimo "d" depende de P_0 y P_1.

Para hacer esta discusión más cuantitativamente debemos regresar a la formulación original de la probabilidad de error.



Probability densities in the transmission of NRZ-polar binary pulses. (a) Negative pulse transmitted. (b) Positive pulse.



Choice of decision level in binary transmission.

$$P_e = P_0 \underbrace{\int_d^{\infty} f_0(v)dv}_{Pe_0} + P_1 \underbrace{\int_{-\infty}^{d} f_1(v)dv}_{Pe_1}$$

$$\frac{\partial P_e}{\partial d} = 0 = - P_0 f_0 (d) + P_1 f_1(d)$$

$$\frac{f_1(d)}{f_0(d)} = \frac{P_0}{P_1}$$

$$\exp \left[\frac{-(d - \frac{A}{2})^2}{2 \sigma^2} + \frac{(d + \frac{A}{2})^2}{2 \sigma^2} \right] = \frac{P_0}{P_1}$$

$$dopt = \frac{\sigma 2}{A} \quad Ln \quad \frac{P_0}{P_1}$$

Para ASK

$$v_o/ASK\ (t) = \begin{matrix} A \\ 0' \\ O \end{matrix} + x(t)$$

ya que la salida es idéntica

$$P_e, \bullet ASK = \frac{1}{2} \left\{1 - \text{erf} \frac{A}{2\sqrt{2}\ N}\right\} = \frac{1}{2} \text{ erfc} \frac{A}{2\sqrt{2}\ N} \quad (10)$$

Para PSK la salida del detector síncrono consiste de una señal polar ± A más ruido. Esto corresponde exactamente a la señal polar analizada anteriormente. Sin embargo, aquí se tiene que la señal es ± A, en lugar de $\pm \frac{A}{2}$. Entonces la probabilidad de error es

$$P_e, PSK = \frac{1}{2} \text{ erfc} \frac{A}{\sqrt{2}\ N} \quad (11)$$

como se puede comparar (11) con (10) el sistema PSK requiere solamente la mitad de la amplitud de la señal que el sistema ASK.

En el caso del sistema FSK las salidas de los dos detectores son comparadas. En cualquier instante un detector tiene señal más ruido, el otro solo tiene ruido. Llamando la salida de ruido de un canal x_1, y la del otro x_2, se tiene al restar las salidas de los dos canales, la salida FSK dada por

$$v_o, FSK = \begin{matrix} +A \\ ó \\ -A \end{matrix} + (x_1 - x_2)$$

La señal de salida es otra vez polar: +A aparece si un 1 ha sido transmitido y -A para un cero, la salida de ruido total es sin embargo $x_1 - x_2$. Si los ruidos en los dos canales son independientes, las varianzas se suman. Se ha, afectivamente, doblado el ruido al sustraer las dos salidas. Sin embargo, ya que la señal de salida es polar, la desviación de la señal efectiva, como en el caso PSK, es dos veces la de ASK. Así, para FSK

$$P_e \ FSK = \frac{1}{2} \operatorname{erfc} \frac{A}{2\sqrt{N}} \qquad (13)$$

Para una probabilidad de error especifico, el sistema FSK requiere 3 dB más de potencia en la señal que el sistema PSK con la misma potencia de ruido, pero es 3 dB mejor que el sistema ASK

La relación señal a ruido de salida de un filtro optimo es: $\frac{A^2}{N} = \frac{2E}{n0}$ para el caso de la detección de un pulso en ruido.

E representa la energía de la señal en el punto donde el ruido blanco gausiano de densidad espectral $\frac{no}{2}$ es agregado.

La fig. 41 ilustra la probabilidad de error para sistemas FSK y PSK en función de la relación señal a ruido $\frac{A^2}{2N}$.

En la práctica de microondas se utilizan los sistemas M-PSK QAM, los cuales serán analizados a continuación en cuanto se refiere a la probabilidad de error.

DETECCION NO COHERENTE

Si la coherencia de fase no se puede mantener, ó si es antieconómico incorporar circuitos de control de fase en el receptor, entonces se usa detección de envolvente.

Es evidente que las señales PSK requieren coherencia de fase para ser demoduladas, de ahí que sólo las señales OOK y FSK utilizan detectores de envolvente.

OOK

$$v(t) = [f(t) + x(t)] \cos wot - y(t) \text{ sen } wot$$

Aquí $f(t) = A$ ó 0

$$v(t) = r(t) \cos [wot + \theta(t)]$$

$$r = \sqrt{(f + x)^2 + y^2}$$

$$\theta = \tan^{-1} \frac{y}{f+x}$$

La probabilidad de error depende de la estadística de r en los dos casos: $f = A$ ó 0.

ESTADISTICAS DE RAYLEIGH Y DE RICIAN

Considérese primero el caso donde la señal está ausente, es decir sólo se tiene ruido, $A = 0$. Con x y y independientes y Gausianas, el problema es determinar la estadística de la envolvente aleatoria r. Hacemos esto primero encontrando la estadística conjuntamente de r y θ y entonces integramos sobre θ para encontrar la función de densidad de r.

$x = r \cos \theta \qquad\qquad y = r \,\text{sen}\, \theta$

$$f_{xy}(x,y)\, dxdy = f_{r\theta}(r,\theta)\, dr\, d\theta$$

$$f_{xy}(x,y) = f_x(x)\, f_y(y) = \frac{e^{-(x^2+y^2)/2\sigma^2}}{2\pi\sigma^2} = \frac{e^{-r^2/2\sigma^2}}{2\pi\sigma^2}$$

$$dxdy = r dr d\theta$$

$$f_{r\theta}(r,\theta)\, drd\theta = \frac{r\; e^{-r^2/2\sigma^2}}{2\pi\sigma^2}\, drd\theta$$

$$f_{r\theta}(r,\theta) = \frac{r\; e^{-r^2/2\sigma^2}}{2\pi\sigma^2}$$

$$f_r(r) = \int_0^{2\pi} f_{r\theta}(r,\theta)\, d\theta$$

$$= \frac{r\; e^{-r^2/2\sigma^2}}{\sigma^2}$$

$f = A$

$$v(t) = \left[A + x(t)\right] \cos w_0 t + y(t) \operatorname{sen} w_0 t$$

$$x' = x + A$$

$$f(x') = \frac{e^{-(x' - A)^2/2\sigma^2}}{2\pi \sigma^2}$$

La envolvente v(t) está dada por

$$r^2 = x'^2 + y^2 = (x + A)^2 + y^2$$

$$\theta = \tan^{-1} \frac{y}{x'} = \tan^{-1} \frac{y}{x+A}$$

con x' y y variables independientes

$$x' = r \cos \theta$$
$$y = r \operatorname{sen} \theta$$

$$f(r,\theta)\, dr d\theta = f(x',y)\, dx'dy$$

$$= \frac{e^{-\left[(x'-A)^2 + y^2\right]/2\sigma^2}}{2\pi\sigma^2}\, dx'dy$$

$$= \frac{e^{-A^2/2\sigma^2} \quad r \quad e^{-(r^2 - 2r A \cos\theta)/2\sigma^2}}{2\pi\sigma^2} \, dr d\theta$$

siendo r y θ variables independientes

$$f(r) = \frac{e^{-A^2/2\sigma^2} \, r \, e^{-r^2/2\sigma^2}}{2\pi\sigma^2} \int_0^{2\pi} e^{r A \cos\theta/\sigma^2} d\theta$$

pero

$$I_o(Z) = \frac{1}{2\pi} \int_0^{2\pi} e^{z\cos\theta} \, d\theta$$

Así que

$$f(r) = \frac{r \quad e^{-r^2/2\sigma^2}}{N} \quad e^{-A^2/2N} \qquad I_o\left(\frac{rA}{N}\right)$$

180.

CALCULO DE PROBABILIDAD DE ERROR

$$Pe = Po \int_{b}^{\infty} fn(r)\, dr + P_1 \int_{0}^{b} fs(r)\, dr$$

181.

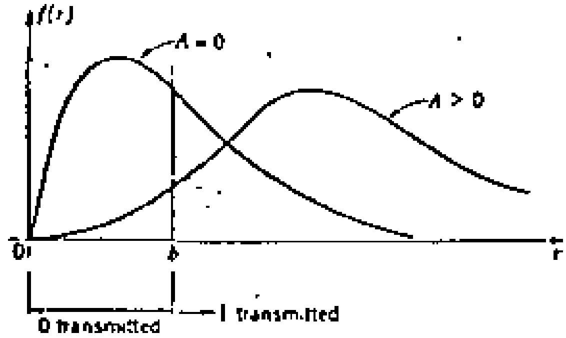
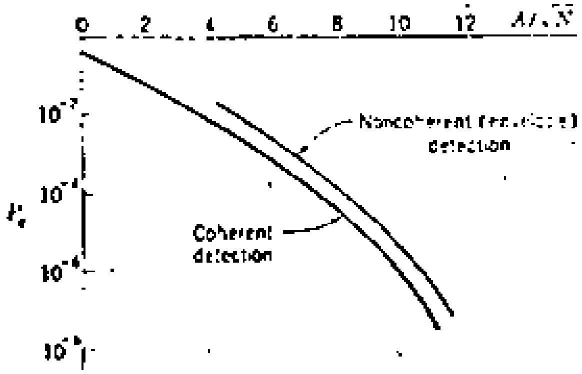


Figure 5-46 Decision regions with envelope-detected OOK signals.



Binary error probabilities, OOK transmission.

FSK

$$P_e = \int_{r_1 = 0}^{\infty} fs(r_1) \left[\int_{r_2 = r_1}^{\infty} fn(r_2) dr_2 \right] dr_1$$

$$P_e = \int_0^{\infty} \frac{r_1}{N} e^{-r_1^2/N} e^{-A^2/2N} I_o \left(\frac{r_1 A}{N}\right) dr_1$$

$$P_e = \frac{1}{2} e^{-A^2/4N}$$

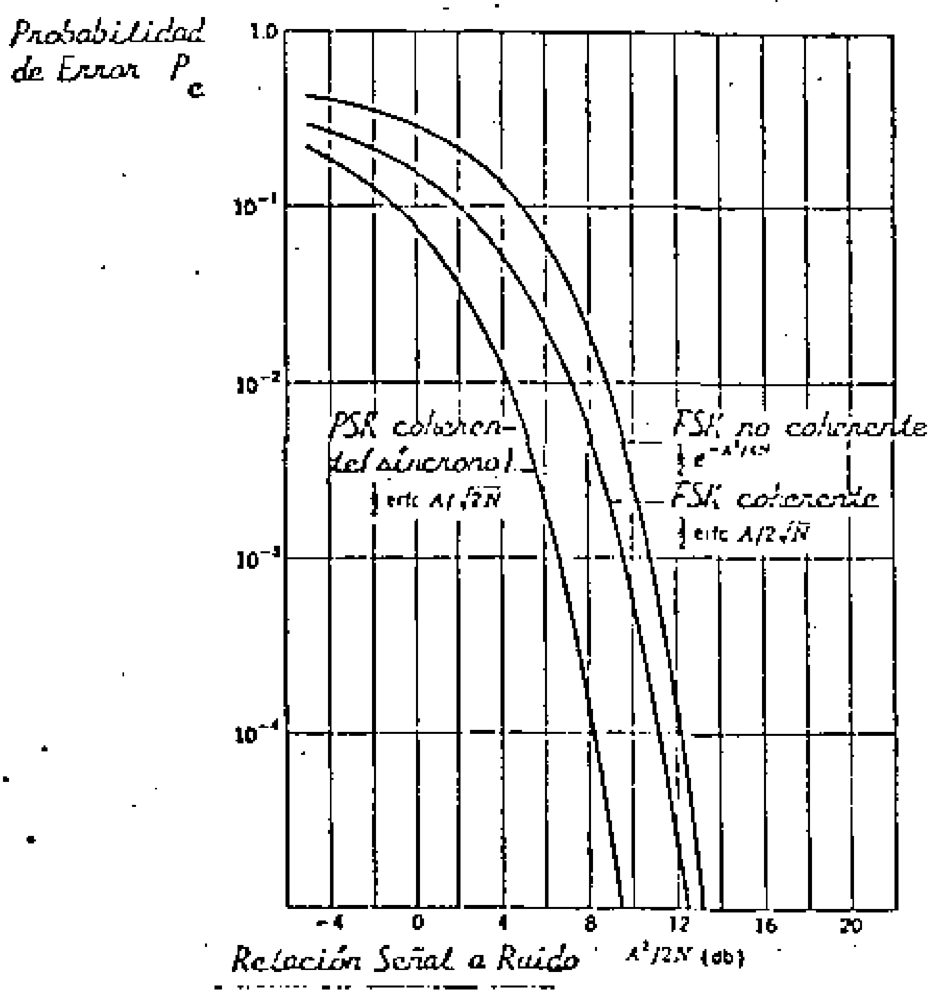


FIGURA N° 41
Transmisión Binaria.

COMPARACION DE DIFERENTES ESQUEMAS DE MODULACION DIGITAL

ESQUEMA	ANCHO DE BANDA	Pe	$\frac{S}{N}$ PARA $P_c=10^{-4}$	COMPLEJIDAD DE EQUIPO
ASK Coherente	2B	$\frac{1}{2}\,\mathrm{erfc}\frac{A}{2\sqrt{2}\,N}$	14.45	Moderado
ASK Incoherente	2B	$\frac{1}{2}\exp(-\frac{A^2 T_b}{16 n_o})$	18.33	Menor
FSK Coherente	>2B	$\frac{1}{2}\,\mathrm{erfc}\,\frac{A}{2\sqrt{N}}$	10.6	Mayor
FSK Incoherente	>2B	$\frac{1}{2}\exp(\frac{-A^2}{4N})$	15.33	Menor
PSK Coherente	2B	$\frac{1}{2}\,\mathrm{erfc}\,\frac{A}{\sqrt{2N}}$	8.45	Mayor
DPSK	2B	$\frac{1}{2}\exp(\frac{-A^2 T_b}{2 n_o})$	9.30	Moderado

La siguiente obtención de la probabilidad de error es ilustrada sobre un diagrama espacial en un sistema QPSK pero que también se aplica al caso binario, y en general, a sistemas M=ary, donde $M = 2, 9, 6, 8, 16, \ldots$, esto es, $M = 2^n$. En el diagrama espacial de la fig. 42 cada estado de fase de igual amplitud representa un símbolo; cada símbolo contiene $n=2$ bits de información. Considerese que el vector $\psi = 0°$ a sido codificado en el transmisor para representar el estado lógico 00, mientras que los vectores de 90°, 180°, y 270° representan los estados lógicos 01, 11, y 10, respectivamente. Consideraremos que cada vector transmitido tiene la misma probabilidad de error; esto es, los datos de entrada en el modulador han sido mezclados y tienen una distribución equiprobable de los estados binarios aleatorios cero y uno.

El diagrama espacial de la señal ilustra que el modem M=ary tiene una simetría circular. Por esta simetría, se puede asumir que en un medio ambiente libre de ruido el vector $\psi = 0°$ que representa el estado 00 ha sido transmitido.

Es también considerado que un modelo de canal de Nyquist esta disponible. Esto es, en el instante del muestreo no hay interferencia entre símbolos. El teoricamente demodulador de fase óptima detectará el estado de fase 00 correctamente si la portadora recibida más el vector de ruido, en el instante de muestreo, esta dentro de la región $-\pi/M$ y π/M. Como un ejemplo ver el vector $v(t) = \bar{V}_{00}$. Si el vector esta dentro la región π/M y π (región de error E_1) ó den tro de la región Π y $-\Pi/M$ (región de error E_2), entonces el vector transmitido que tenga una fase $\phi = 0°$ será erroneamente detectado. En el ejemplo de un vector recibido mostrado en la posición $r(t) = \bar{V}_{01}$, el demodulador decidirá que un vector 01 ha sido transmitido (en lugar de un 00), y así el fasor detectado será un error.

La portadora recibida y la onda de ruido, v(t), de la señal M-ary PSK esta dada por

$$r(t) = A \cos(\omega_c t+\phi) + n_c(t) \cos(\omega_c t+\phi) + n_s(t) \text{ sen } (\omega_c t+\phi) \tag{14}$$

donde A es el valor pico de la portadora recibida, y $n_c(t)$ y $n_s(t)$ representan las componentes de ruido gausiano instantaneas en fase y en cuadratura de fase. Sin pérdidas, puede asumirse que $\phi = 0$.

En la fig. 42 se representa el diagrama vectorial de la portadora y del ruido. Por las figuras (41) y (42) se concluye que un error ocurrira si

$$|\alpha| > \frac{\pi}{M} \tag{15}$$

para derivar la probabilidad de error se tiene que definir, primero, la densidad de probabilidad de α. La función de distribución de probabilidad de α dentro de las regiones

$$r(t) = A \cos (W_c t + \emptyset) + n_c (t) \cos (W_c t + \emptyset)$$

$$+ ns(t) \quad \text{Sen} \ (W_c t + \emptyset)$$

Se tiene error si

$$|\alpha| > \pi/M$$

$$\alpha = \tan^{-1} \frac{ns(t)}{A + nc(t)}$$

Para M - ary PSK

$$P(e) = \int_{\pi/M}^{\pi} P(\alpha) \, d\alpha + \int_{-\pi}^{-\pi/M} P(\alpha) \, d\alpha$$

$$= 2 \int_{\pi/M}^{\pi} P(\alpha) \, d\alpha$$

$$P(\alpha) = \frac{1}{2\pi} e^{-C/N} \left\{ 1 + \sqrt{4\pi \left(\frac{C}{N}\right)} \cos \alpha \, e^{C/N \cos^2 \alpha} \, Q\left(\sqrt{2\left(\frac{C}{N}\right)} \cos \alpha\right) \right\}$$

donde

$$Q(x) = \frac{1}{\sqrt{2\pi}} \int_x^{\infty} e^{-t^2/2} \, dt$$

$$p(e) \cong e^{-C/N} \ \text{Sen}^2 \ \pi/M$$

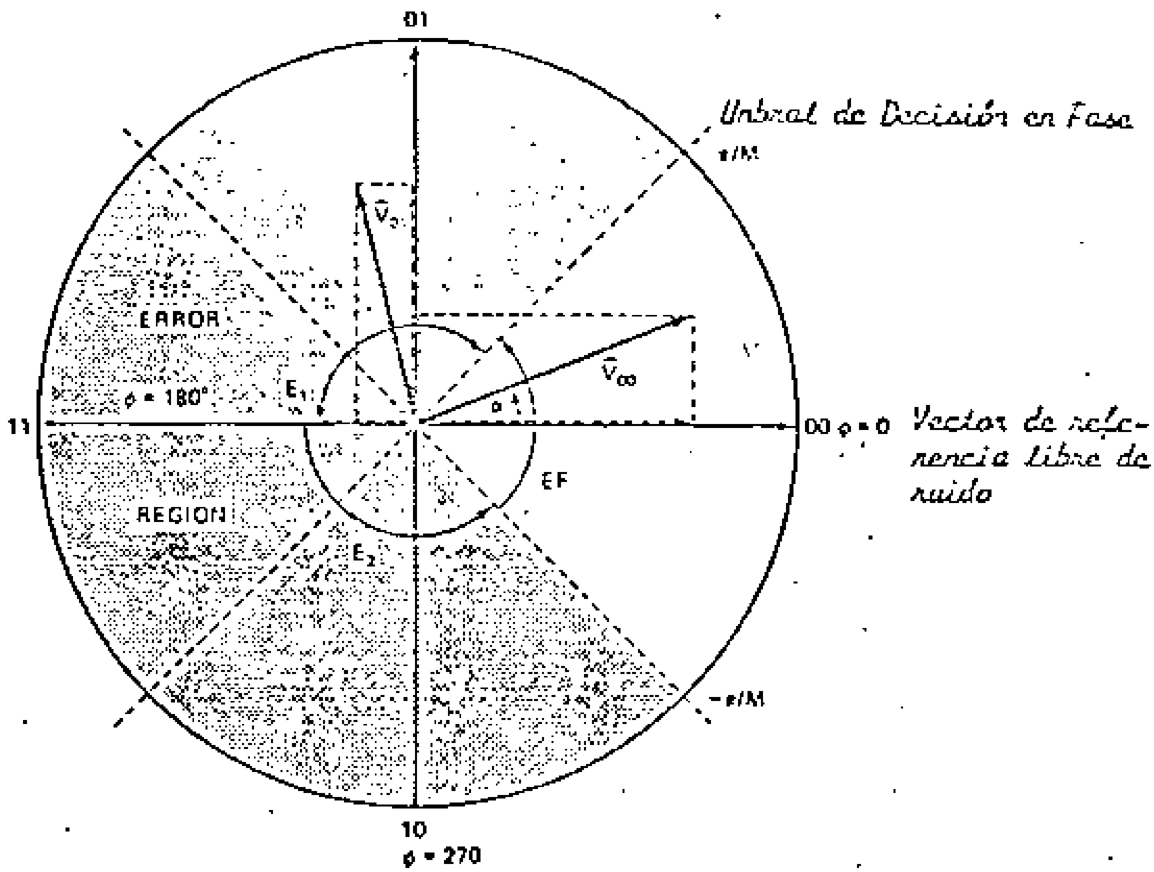


FIGURA N° 42
Región de Error en Demoduladores Coherentes. M 4ary. PSK.

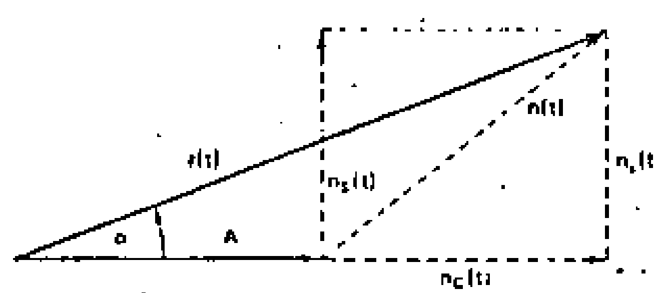


FIGURA N°43
Diagrama Vectorial de una Portadora Recibida y de una - Señal de Ruido.

de error previamente establecidas E_1, Π/M a Π y la región E_2, Π a $-\Pi/M$ esta representada por el area sombreada de la fig. 41 y es la probabilidad de error P_e del sistema. α esta dada por

$$\alpha = \tan^{-1} \frac{n_s(t)}{A + n_c(t)} \qquad (16)$$

La P_e del sistema M-ary PSK es

$$P_e = \int_{\Pi/M}^{\Pi} P(\alpha)\, d\alpha + \int_{-\Pi}^{-\Pi/M} P(\alpha)\, d\alpha = 2 \int_{\Pi/M}^{\Pi} P(\alpha)\, d\alpha \qquad (17)$$

donde $P(\alpha)$ es la función de densidad de probabilidad de α. Esta función para un canal de ruido blanco gaussiano aditivo ha sido obtenida en referenicas y esta dada por

$$P(\alpha) = \frac{1}{2\Pi} e^{-C/n} \left[1 + \sqrt{4\Pi \left(\frac{c}{n}\right)} \cos\alpha \; e^{(C/N)\cos^2\alpha} \; Q\left[\sqrt{2\left(\frac{c}{n}\right)} \;.\; \alpha\right]\right] \qquad (18)$$

donde

$$Q(x) = \frac{1}{\sqrt{2\Pi}} \int_x^{\infty} e^{-t^2/2}\, dt \qquad (19)$$

En la ecuación (18) el término C/N representa la [illegible]ción de la potencia media de la portadora especificada [illegible] cho de banda bilateral de Nyquist el cual es igual [illegible]cho de banda de tasa del símbolo. Como no existe ningu[illegible] ción de forma cerrada que satisfaga las ecuaciones [illegible]

(19), es necesario usar métodos númericos para evaluar la función P(e). La P(e) puede también ser evaluada por la ecuación simple

$$P(e) = e^{-C/N \, sen^2 \, \Pi/M} \qquad (20)$$

Esta aproximación para relaciones C/N altas (C/N > 15 dB) tiene una precisión de 1 dB. Los valores calculados de la curva P(e) = f(C/N), basados en las ecuaciones (17), (18) y (19) han sido gráficados en la fig. 44.

En la mayoría de los sistemas prácticos el ancho de banda de ruido del receptor es mayor que el ancho de banda bilateral de Nyquist. Para proveer una comparación del sistema de ancho de banda mínimo teórico con el sistema práctico de más banda, la ecuación siguiente es frecuentemente usada:

$$\frac{E_b}{N_o} = \left(\frac{C}{N}\right)_{bw} \frac{BW}{fb} \qquad (21)$$

En esta ecuación

E_b = energía promedio de un bit = CTb

f_b = tasa de bit transmitida

T_b = duración de bit unitario

C = Potencia promedio de la portadora

N_o = Densidad espectral de potencia del ruido, esto es, potencia de ruido promedio en un ancho de banda de

1 Hz.

BW = ancho de banda de ruido del receptor.

La probabilidad de error en los sistemas de microondas terrestres esta especificada frecuentemente en términos de la relación C/N, mientras que en sistemas de satelites es especificada en términos de Eb/No.

$Pe = 10^{-6}$

PSK	Detección coherente	PSK	10.5 dB
		4-PSK	10.5 dB
		8-PSK	13.8 dB
PSK	Detección coherente diferencial	DPSK	11.4 dB
		4-DPSK	12.8 dB
		8-DPSK	16.8 dB
FSK	Detección con descriminador	FSK	13.4
		Duobinaria	15.9
		4-FSK	20.1
		8-FSK	25.5

TABLA I

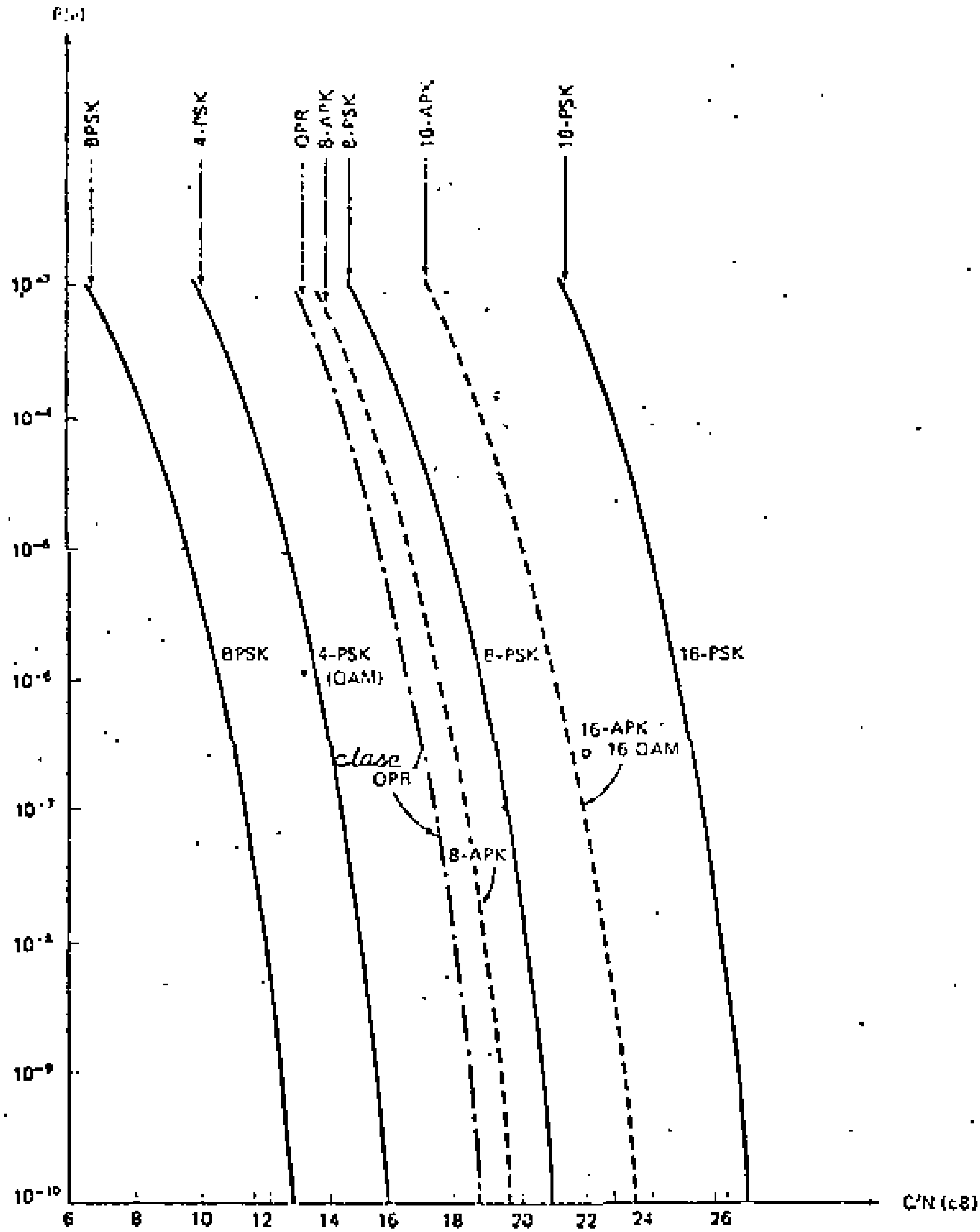


FIGURA N° 44

Representación de la Probabilidad de Error P(e) de Sistemas Coherentes: M-ary PSK, QAM, QPR y M-ary APK. La C/N aquí es especificada en el ancho de banda de Nyquist de doble banda lateral.

2.7 TECNICAS DE ACCESO MULTIPLE POR SATELITE

Existen varias formas para que las estaciones terrenas que se comunican entre sí a través de un satélite determinado puedan utilizar los recursos de potencia y ancho de banda del mismo. A estas formas de utilización se les denomina modos o técnicas de acceso múltiple, dado que son múltiples o varias las estaciones que comparten el mismo satélite. Las técnicas de acceso más comunes, son las siguientes:

- Acceso múltiple por división en frecuencia o FDMA *
- Acceso múltiple por división en tiempo o TDMA **
- Acceso múltiple por expansión de espectro o SSMA +
- Acceso múltiple por desplazamiento de haz o SBMA ++

* FDMA = Frequency Division Multiple Access

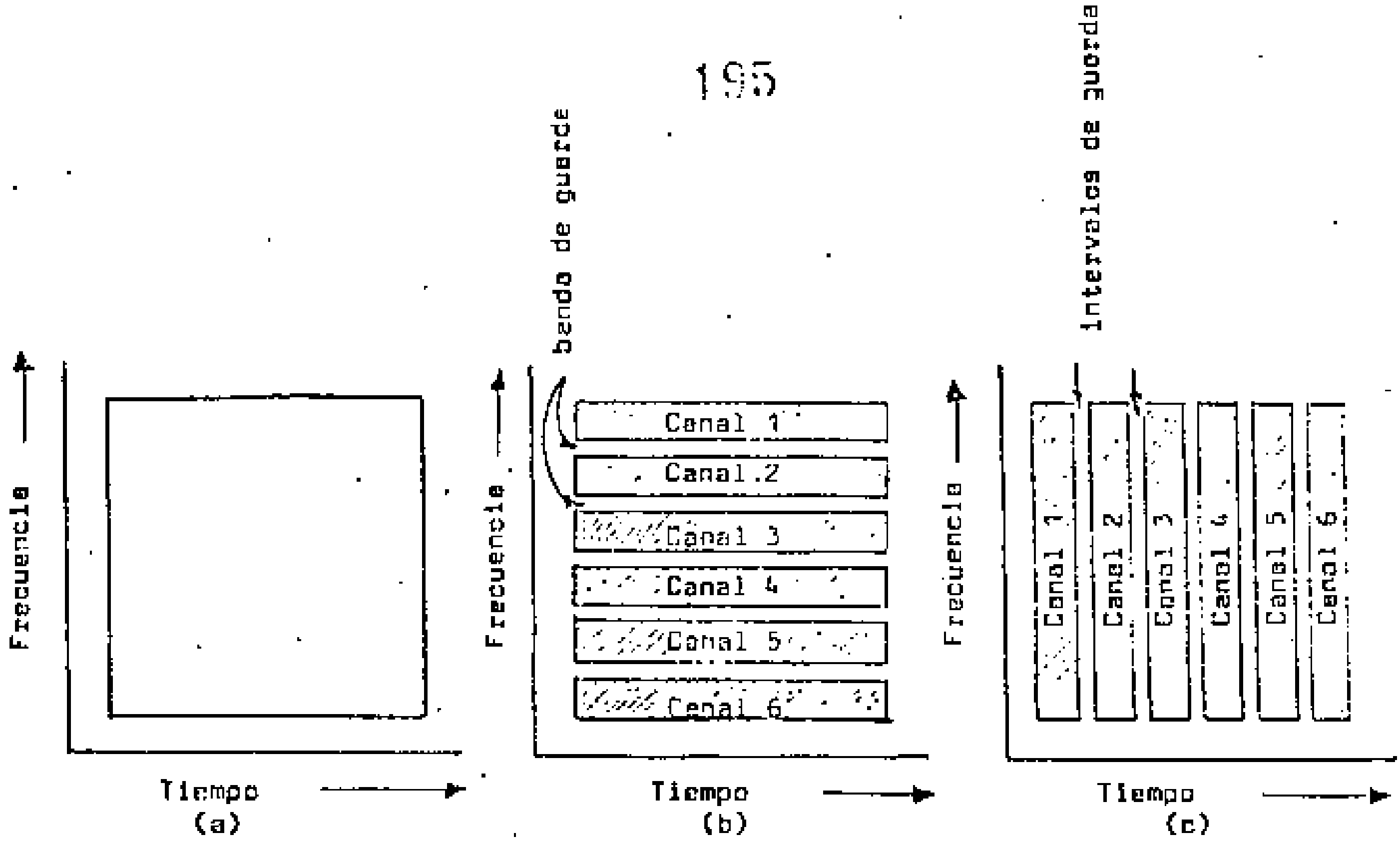
** TDMA = Time Division Multiple Access

\+ SSMA = Spread Spectrum Multiple Access

++ SBMA = Spatial Beam Multiple Access

FDMA es la técnica de mayor uso en la actualidad. Se emplea para transmitir varios grupos de canales de voz multiplexados en frecuencia (FDM) y modulados en frecuencia - (FM). Al conjunto de este esquema de transmisión se le denota como FDM/FM/FDMA, indicando la secuencia en el proceso por el que va pasando la señal.

FDMA y TDMA son analogías de los esquemas de multiplexaje que se utilizan en sistemas terrestres multicanal, - analógicos y digitales, respectivamente. En estas analogías, los elementos que se "multiplexan" en los enlaces vía satélite son las estaciones terrenas participantes en la red. (Ver figura 1).

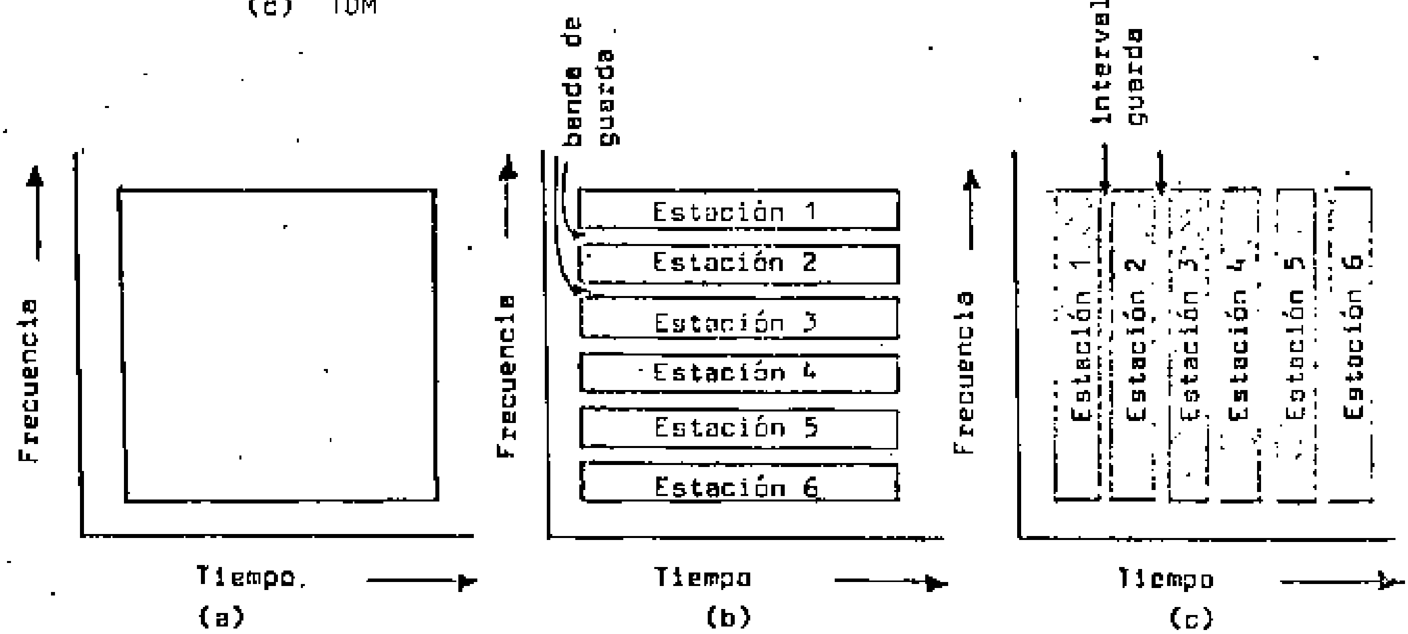


Multiplexaje en sistemas terrestres

(a) espacio disponible para comunicaciones

(b) FDM

(c) TDM



Acceso múltiple en enlaces vía satélite

(a) espacio disponsible para comunicaciones

(b) FDMA

(c) TDMA

Figura 1. Analogía entre FDM y TDM con FDMA y TDMA

El acceso múltiple por expansión de espectro o SSMA, únicamente ha encontrado aplicaciones en sistemas satelitales militares donde es inherente la importancia de la privacía y seguridad. El concepto básico de esta técnica se ilustra en la figura 2.

El acceso múltiple por desplazamiento de haz o división de espacio, comienza a ser popular en satélites de alta capacidad como una forma de re-aprovechar las frecuencias disponibles, dirigiendo haces angostos hacia zonas terrestres específicos (véase la figura 3).

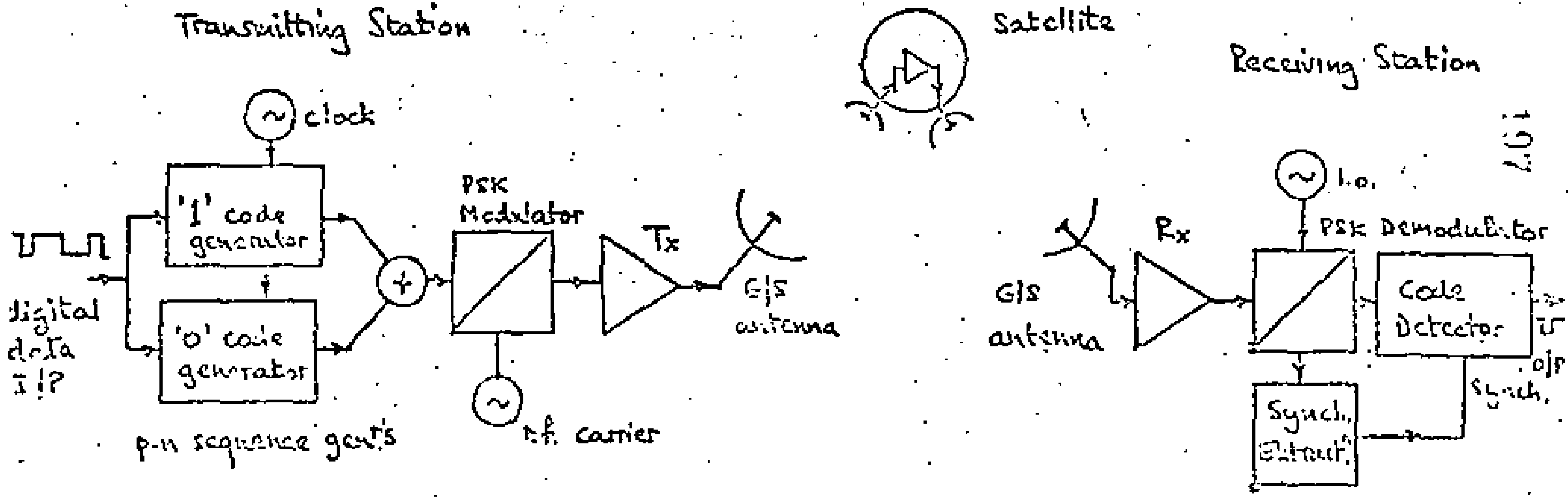
Una desventaja de esta técnica, es que el satélite necesite disponer de una antena muy grande para poder producir haces dirigidos angostos. Por ejemplo, la antena del satélite ATS-6 de la NASA, que puede utilizar esta técnica, mide 10 metros de diámetro y los haces "angostos" cubren varios cientos de kilómetros sobre la tierra. Además, el equipo de conmutación o switcheo de los haces representa un grave riesgo de falla irreparable en el satélite.

En las figuras 4, 5, 6 y 7 pueden verse, respectivamente, el satélite ATS-6, la distribución de alimentadores, los haces producidos por el arreglo norte-sur de alimentadores y algunas de las huellas de iluminación.

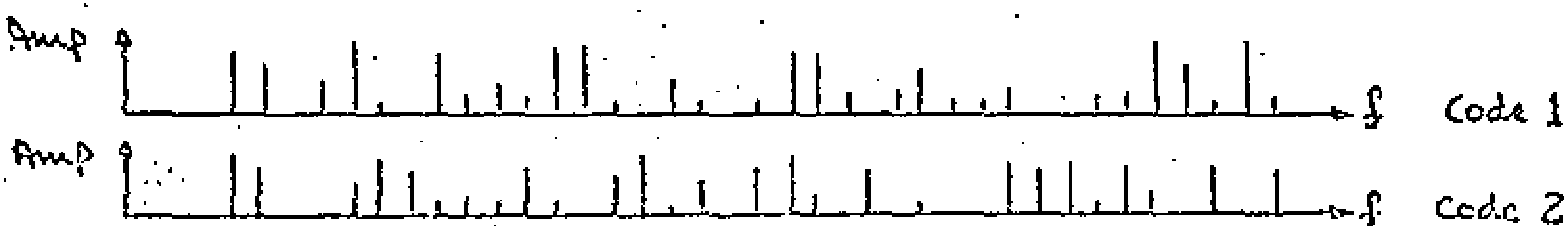
La técnica utilizada universalmente por los satélites comerciales es FDMA y recientemente, aunque en menor escala, TDMA.

Fig. 2

Acceso múltiple por expansión de espectro.



P-N Sequence scheme for SSMA



Part of spectra of two p-n codes

Fig. 3

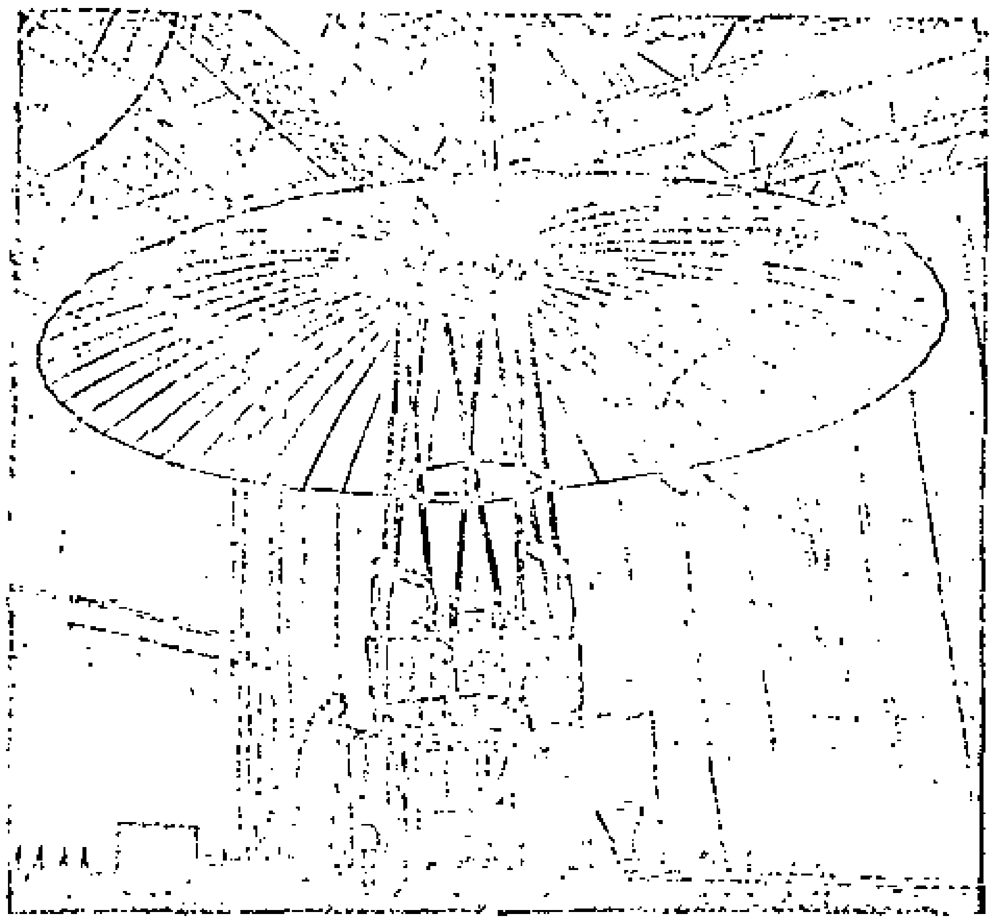


Fig. 4

Antena de 10 metros del satélite ATS-6 de la NASA.

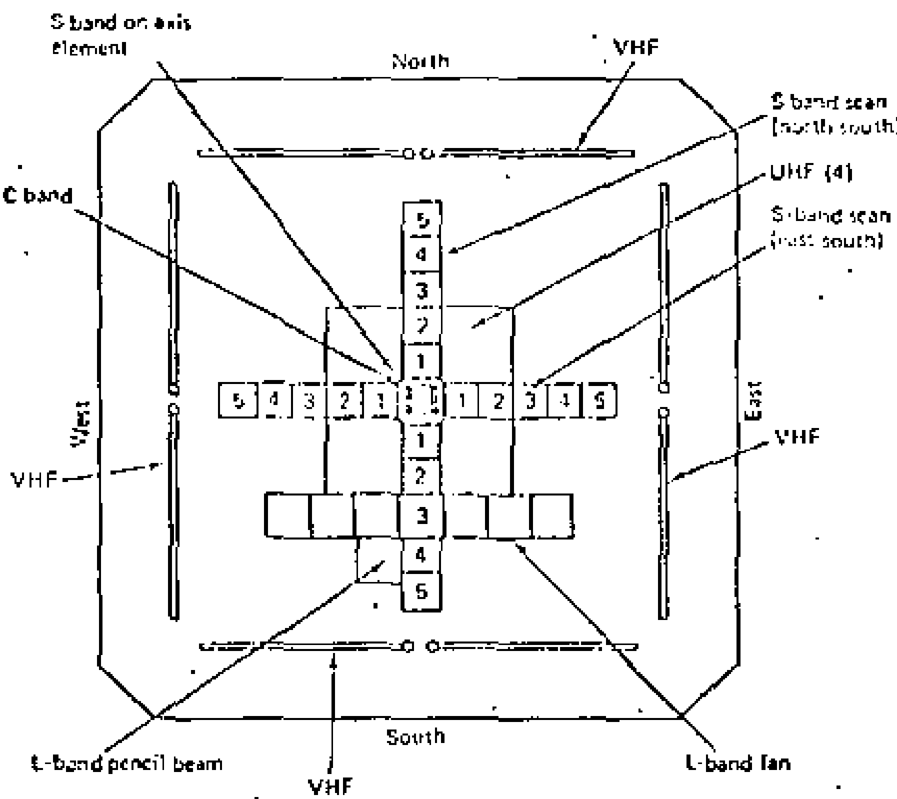
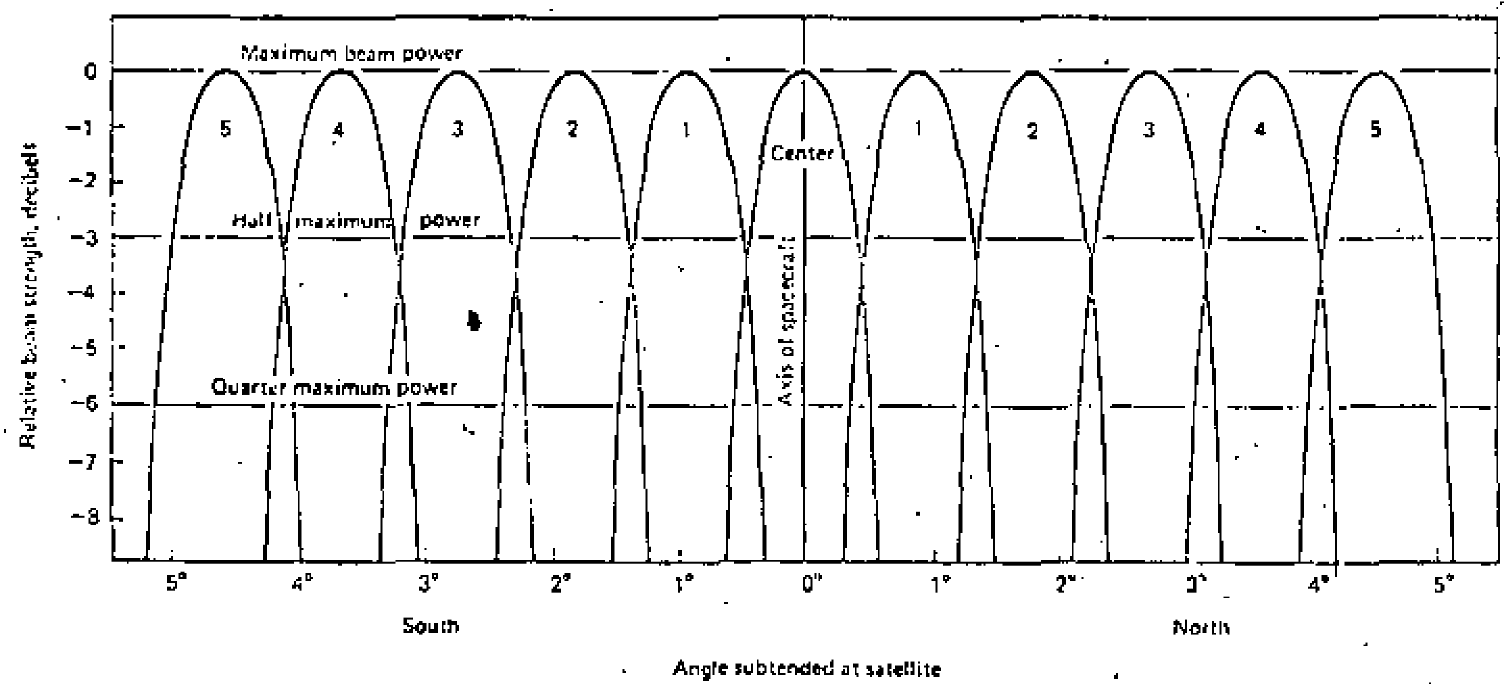

S band on axis element
North
VHF
S band scan (north south)
UHF (4)
C band
S-band scan (east south)
West
East
VHF
VHF
South
L-band pencil beam
VHF
L-band fan

Fig. 5



Maximum beam power
Half maximum power
Quarter maximum power
Center
Axis of spacecraft
Relative beam strength, decibels
0
−1
−2
−3
−4
−5
−6
−7
−8
5 4 3 2 1 1 2 3 4 5
5° 4° 3° 2° 1° 0° 1° 2° 3° 4° 5°
South
North
Angle subtended at satellite

Fig. 6

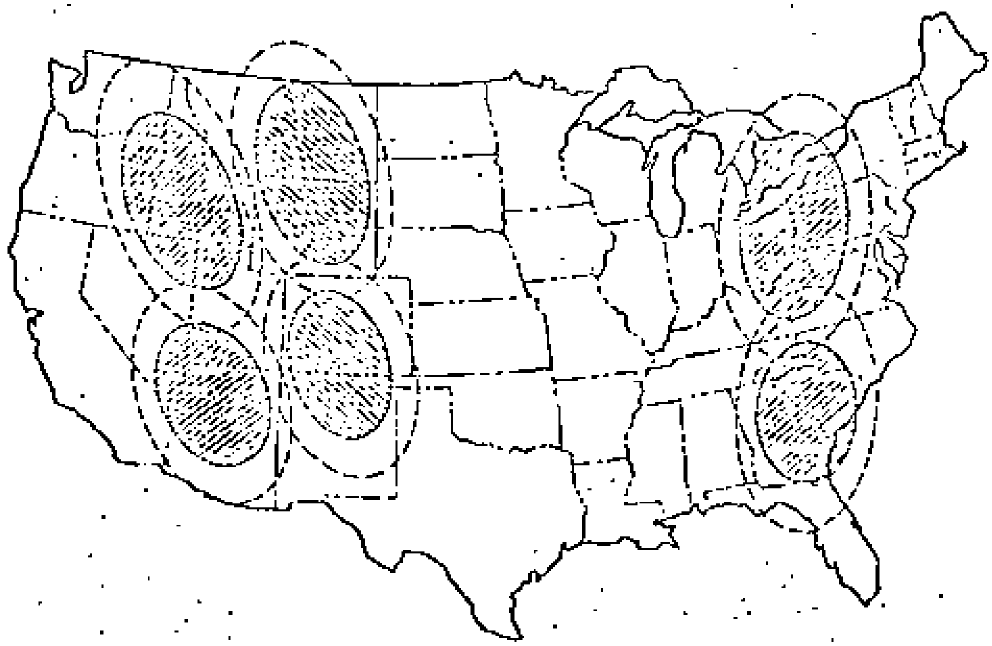


Fig. 7

Acceso múltiple por división en frecuencia (FDMA):

203

- FDM / FM
- SCPC / FM o PSK
  - asignación fija
  - asignación por demanda (DAMA)

En el acceso múltiple por división en frecuencia, la capacidad de ancho de banda de un transpondedor se puede dividir en un número variable de bandas:

(a) se pueden tener pocas bandas de gran capacidad. Cada banda puede manejar un grupo, un supergrupo, o un mastergrupo de voz (FDM)

(b) se pueden tener muchas bandas de poca capacidad. Cada banda puede manejar solamente un canal de voz (SCPC)

(c) se puede tener una mezcla de las dos, FDM y -- SCPC, con señales de voz.

(d) se pueden asignar algunas bandas para transmisión de datos, otras para voz y quizás una para TV. Usualmente, un canal de TV ocupa todo el transpondedor de 36 MHz, pero en algunos - casos se transmiten dos canales de TV en el mismo transpondedor (se tienen 2 bandas o ranuras), o bien un canal de TV y otros tipos de señales como voz y datos.

La figura 8 ejemplifica el uso de un transpondedor por varias estaciones terrenas a través de FDMA.

En la figura 9 se ilustra el ancho de banda de los transpondedores de los satélites Intelsat IV e Intelsat V. Cada - transpondedor puede subdividirse en varias ranuras de frecuencia para su utilización en FDMA, como se ejemplificó en la figura 8.

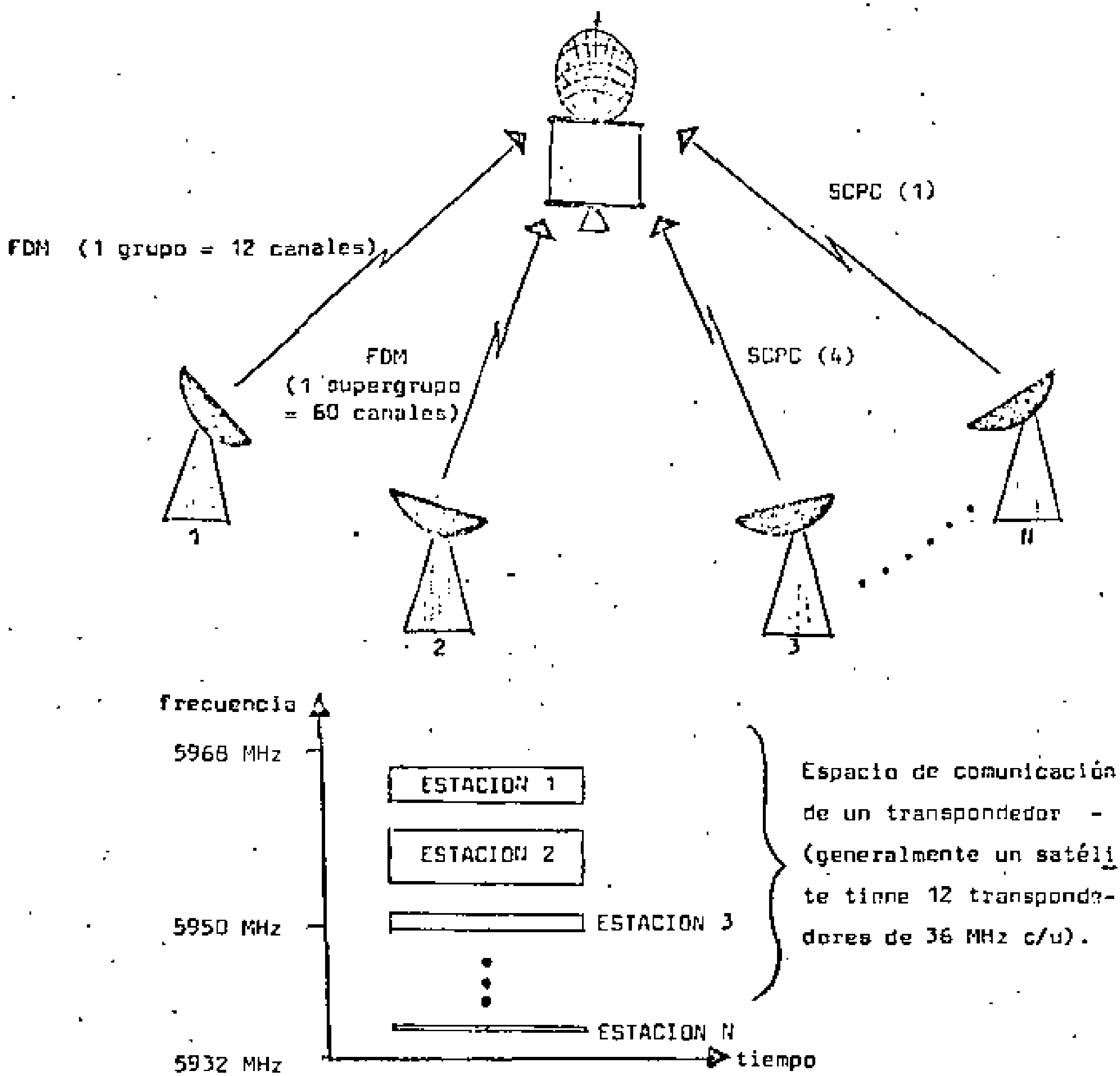


Figura 8. Ejemplificación de FDMA

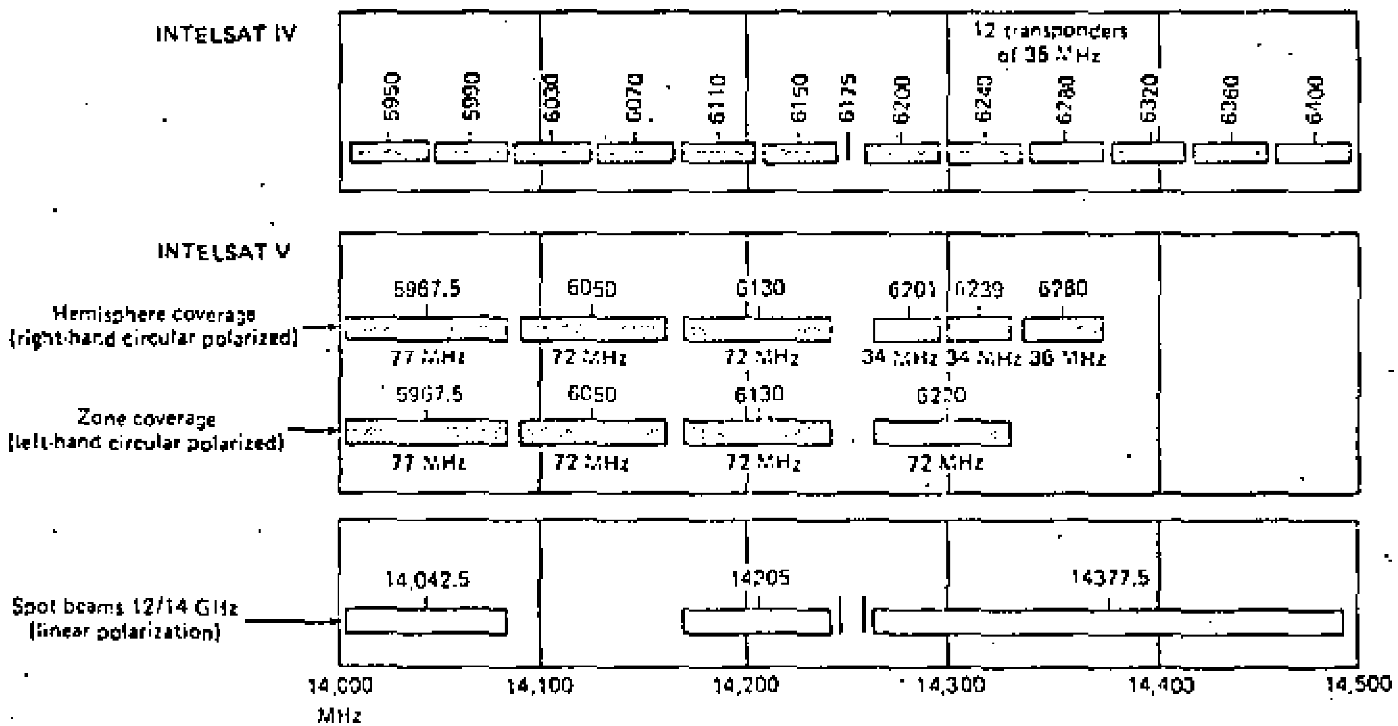
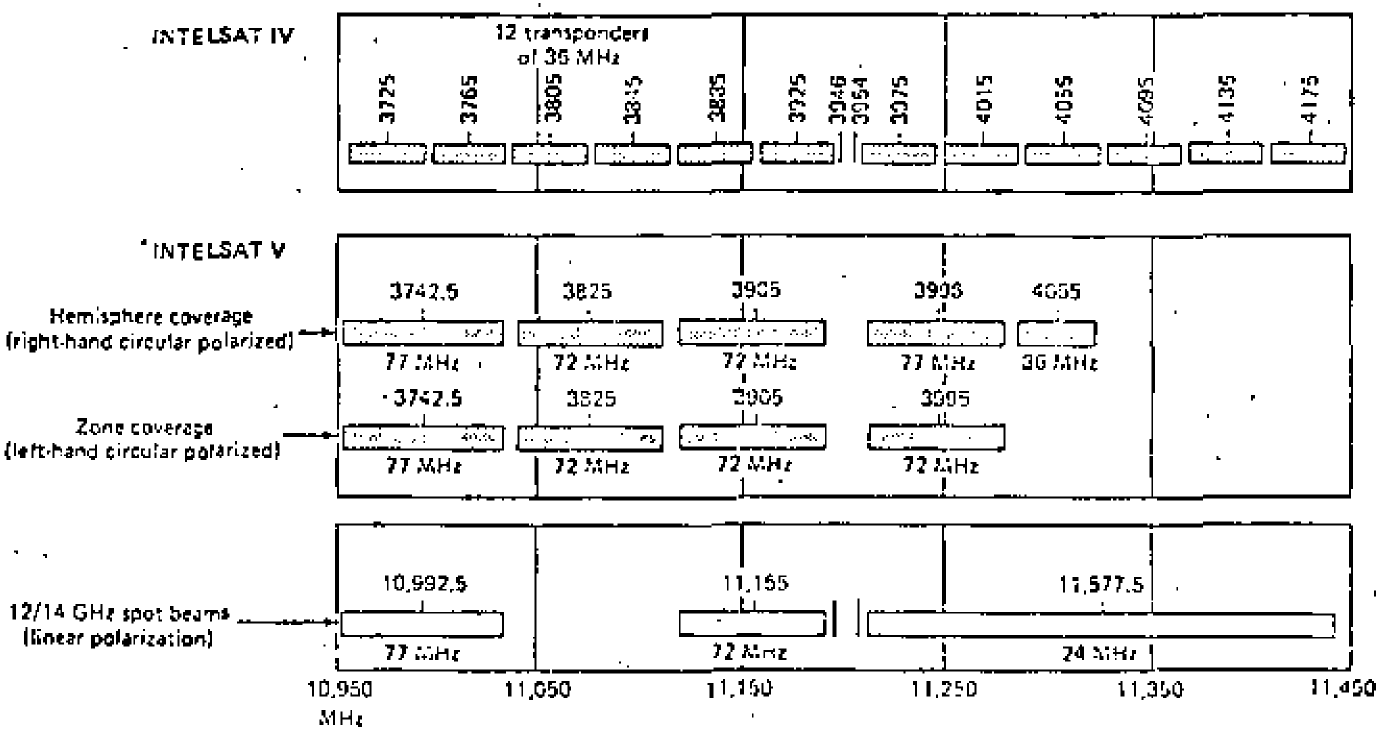


Fig. 9
Frecuencias de subida

Fig. 9 (cont.)
Frecuencias de bajada



- FDM/FM.-

Esta es la técnica utilizada, por ejemplo, por INTELSAT IV y IV-A. Cada estación terrena rearregla los canales y grupos de canales de entrada en supergrupos de 60 canales que ocupan una banda base de 252 KHz, o bien en grupos de 12 canales cuando los requerimientos de tráfico son menores. En las figuras 10 y 11 se ilustra el proceso de FDM.

El supergrupo emitido por una estación A en particular contendrá canales con destinación diferente. Sin embargo, los 60 canales modulan en frecuencia (FM) a una portadora en el rango de los 4 GHz (la frecuencia exacta de esta portadora será la que tenga asignada la estación emisora A). Todas las estaciones terrenas situadas en diferentes partes del mundo y que reciben señales de la estación A demodulan toda la portadora, que tiene un ancho de banda de 5 MHz y extraen los canales que les corresponden.

Al haber varias portadoras presentes en el mismo transpondedor de un satélite, y debido a la característica no lineal del TWT, es necesario operar éste con varios decibeles abajo de su punto de saturación o nivel máximo de potencia de salida. A esta reducción en la potencia aprovechada se le denomina back-off (abreviado BO) se salida. Si el amplificador se operase en una región altamente no lineal, se producirían niveles muy elevados de productos de intermodulación que afectarían significativamente la calidad S/N de las señales amplificadas. En la figura 11* se muestra la característica típica entrada/salida de un amplificador TWT. Notese que el back-off de entrada no es proporcional al back- de salida. En estas mismas notas, en la sección correspondiente a "Estructura básica de un satélite", también se muestra la relación entre el BO de salida de un TWT y la relación portadora / ruido de intermodulación en función del número de portadoras presentes. A manera de ejemplo, los satélites INTELSAT IV y IV-A operan con un BO de 7 dB en haz pincel y 4.2 dB en haz global. El número típico de canales en un transpondedor de 36 MHz en el IV-A con haz pincel es de 900, y de 450 para un haz global con varias portadoras.

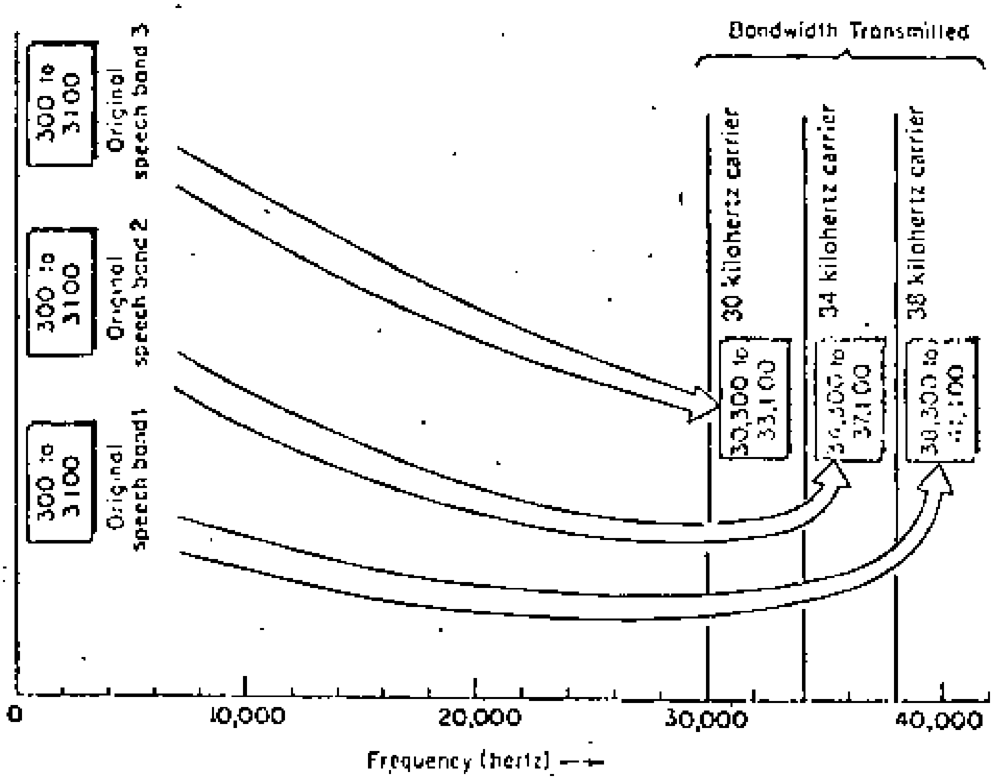
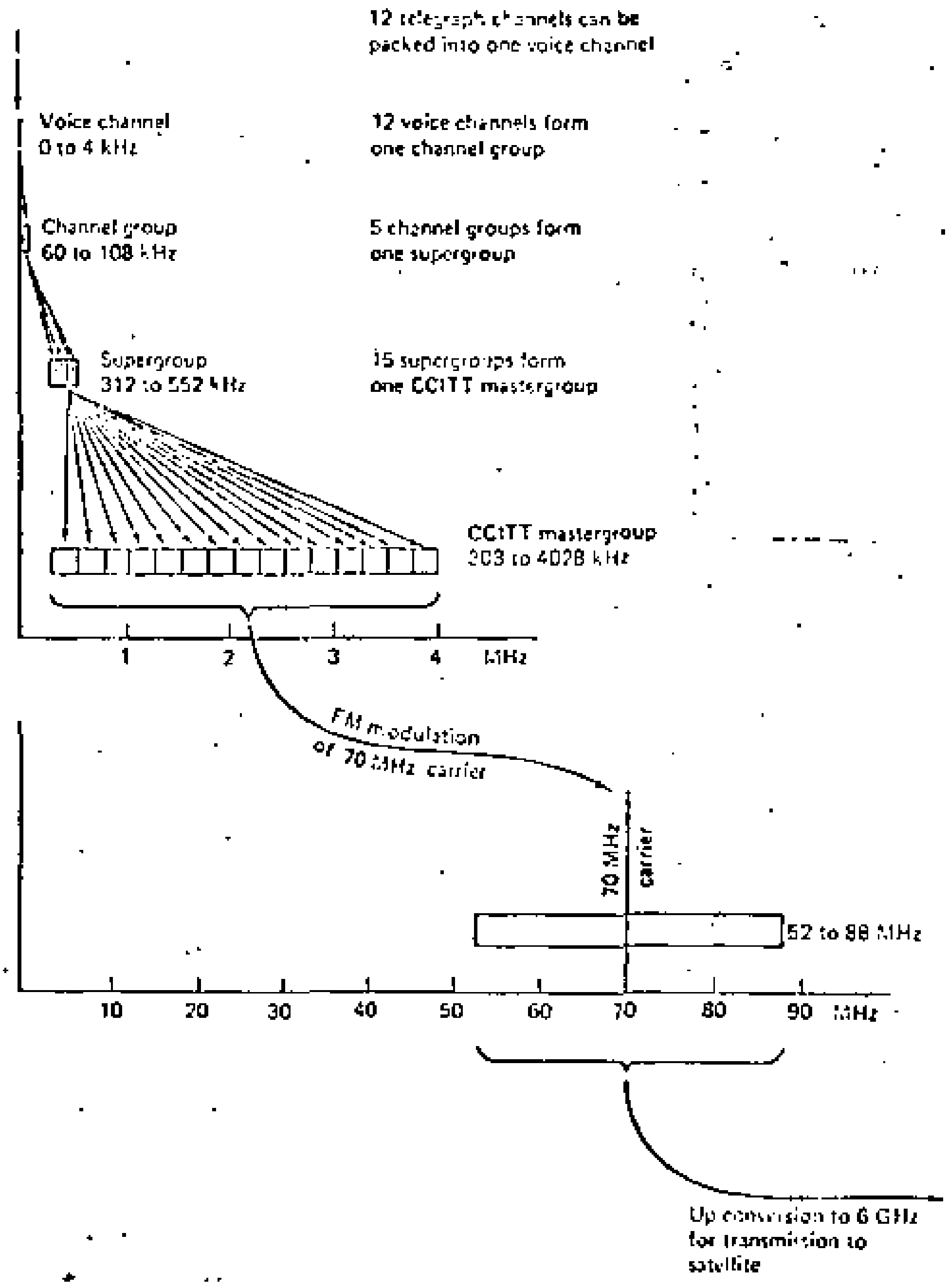
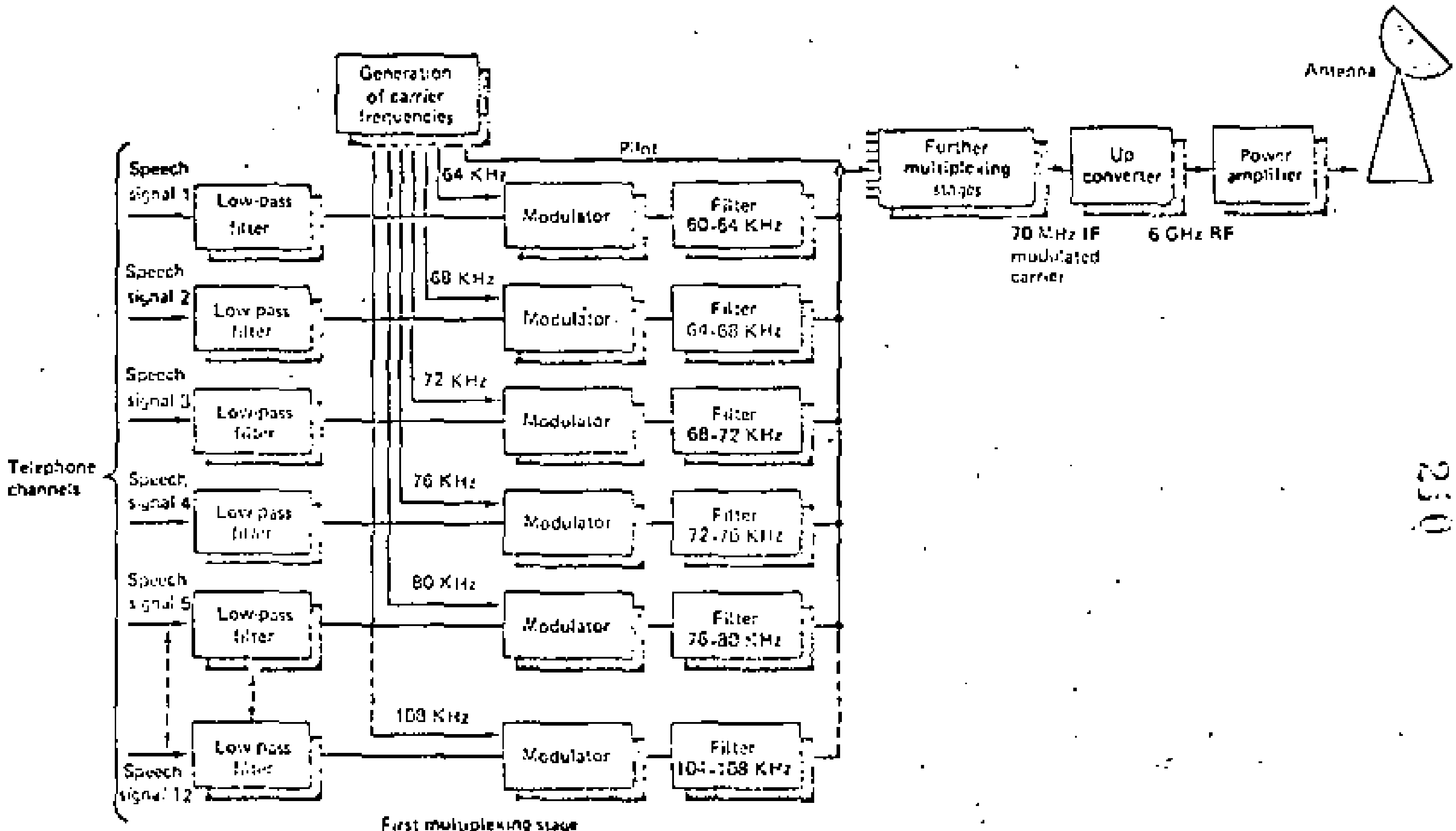

Bandwidth Transmitted
300 to 3100
Original speech band 3
300 to 3100
Original speech band 2
300 to 3100
Original speech band 1
30 kilohertz carrier
34 kilohertz carrier
38 kilohertz carrier
30,300 to 33,100
0
10,000
20,000
30,000
40,000
Frequency (hertz)

Fig. 10

Fig. 10 (cont.)

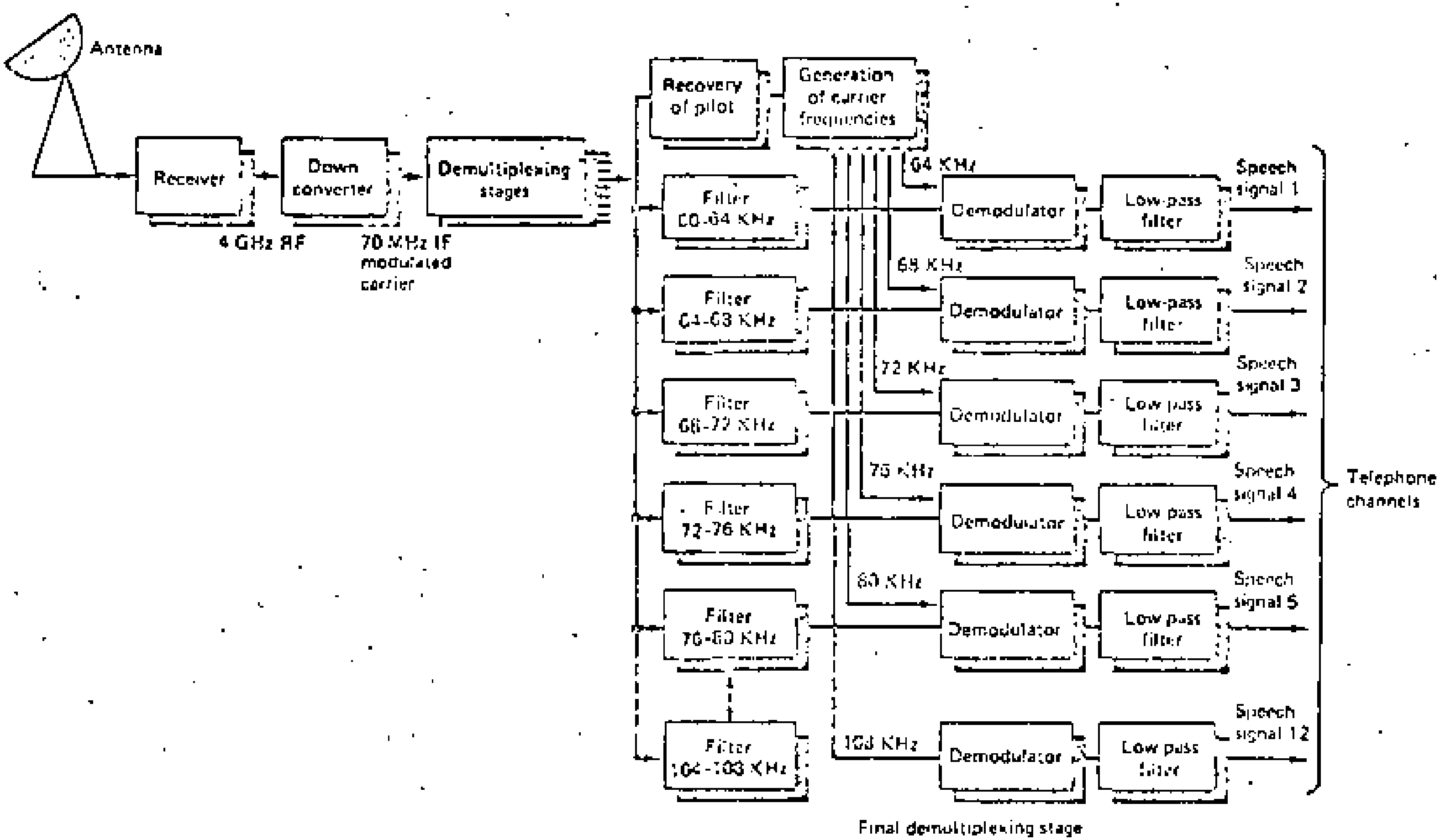

12 telegraph channels can be packed into one voice channel
Voice channel 0 to 4 kHz
12 voice channels form one channel group
Channel group 60 to 108 kHz
5 channel groups form one supergroup
Supergroup 312 to 552 kHz
15 supergroups form one CCITT mastergroup
CCITT mastergroup 303 to 4028 kHz
1
2
3
4
MHz
FM modulation of 70 MHz carrier
70 MHz carrier
52 to 88 MHz
10
20
30
40
50
60
70
80
90
MHz
Up conversion to 6 GHz for transmission to satellite



Generation of carrier frequencies
Pilot
Further multiplexing stages
Up converter
Power amplifier
Antenna
70 MHz IF modulated carrier
6 GHz RF
Telephone channels
Speech signal 1
Speech signal 2
Speech signal 3
Speech signal 4
Speech signal 5
Speech signal 12
Low-pass filter
64 KHz
68 KHz
72 KHz
76 KHz
80 KHz
108 KHz
Modulator
Filter 60-64 KHz
Filter 64-68 KHz
Filter 68-72 KHz
Filter 72-76 KHz
Filter 76-80 KHz
Filter 104-108 KHz
First multiplexing stage

Fig. 11

Fig. 11 (cont.)



Antenna
Receiver
Down converter
Demultiplexing stages
4 GHz RF
70 MHz IF modulated carrier
Recovery of pilot
Generation of carrier frequencies
64 KHz
68 KHz
72 KHz
76 KHz
Filter 60-64 KHz
Filter 64-68 KHz
Filter 68-72 KHz
Filter 72-76 KHz
Filter 76-80 KHz
Filter 104-108 KHz
Demodulator
Low-pass filter
Speech signal 1
Speech signal 2
Speech signal 3
Speech signal 4
Speech signal 5
Speech signal 12
Telephone channels
Final demultiplexing stage

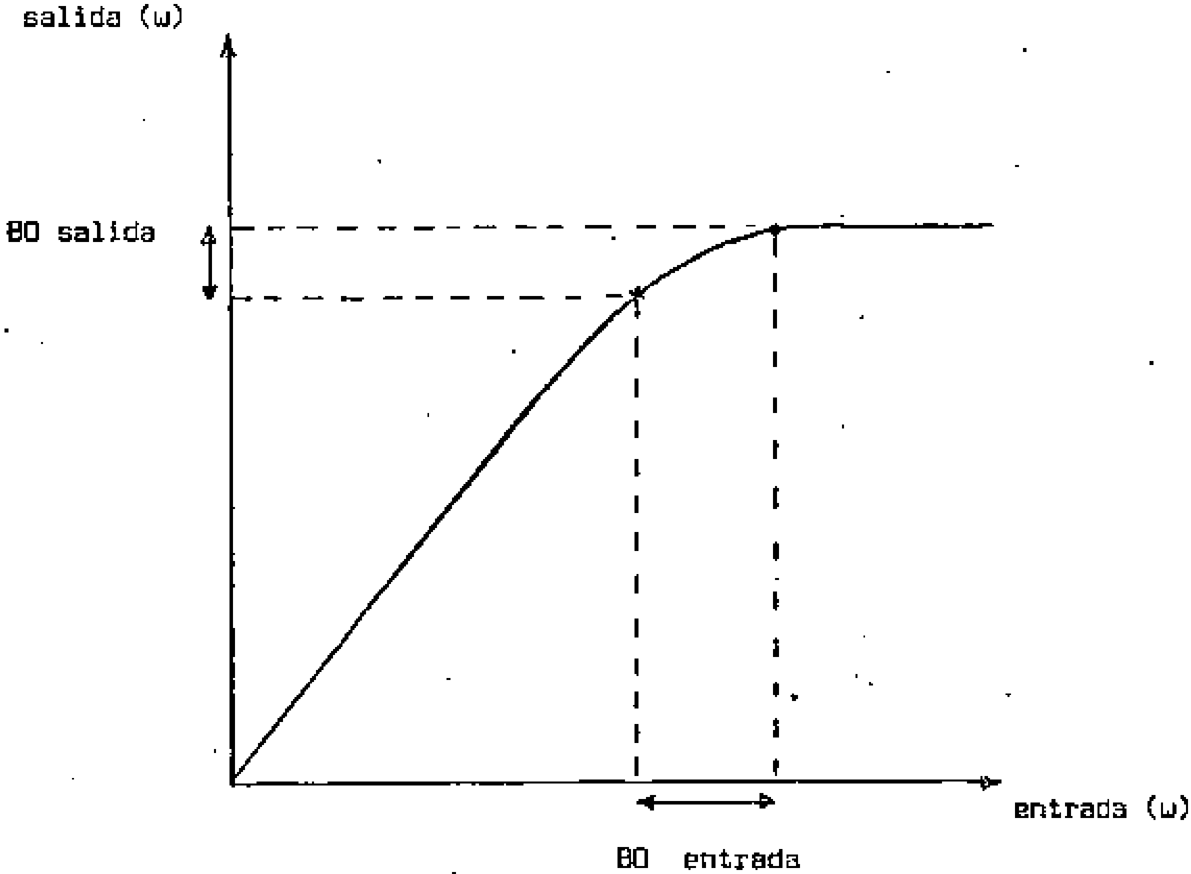


Fig. 11

Caracteristica de amplificación de un TWT.

Como puede notarse, la capacidad de un transpondedor operando en FDM/FM/FDMA varía de acuerdo al número de portadoras, que está íntimamente ligado al número de estaciones accesando el transpondedor. En la tabla I se ilustra esta variación en capacidad.

Tabla 1. Capacidad de un transpondedor de 36 MHz en función del número de portadoras.

No. de portadoras	Ancho de banda por portadora (MHz)	No. de canales por portadora	No. total de canales en el transpondedor
1	36	900	900
4	3 de 10 y 1 de 5	132 60	456
7	5	60	420
14	2.5	24	336

Los transpondedores de 36 MHz normalmente se operan con portadoras moduladas en FM y con ancho de banda de 2.5, 5 ó 10 MHz. Ocasionalmente se usa todo el transpondedor por una sola portadora para telefonía (en estos casos especiales se tiene acceso único y no múltiple). Por lo que respecta a TV, se puede tener un canal en una portadora de 36 MHz, o dos canales en portadoras de 18 MHz c/u usando filtros especiales e igualadores de retraso de grupo. En el caso de México, se transmite TV por los satélites INTELSAT IV y WESTAR IV y en ambos casos se opera en 2 canales de TV de 18 MHz c/u por transpondedor.

FDM/FM/FDMA es muy ineficiente en el aprovechamiento de espectro en el sentido de que cada enlace entre dos estaciones tiene asignada una frecuencia única que no puede ser utilizada por ningún otro enlace en ningún momento, a menos que se emplee re-utilización de espacio (SBMA) o de frecuencia - con otra polarización. Si se consideran, por ejemplo, los enlaces internacionales entre México y Europa a través del --- Atlántico, se observa que las horas de mayor demanda de tráfico están entre las 12 y 17 hrs. en Europa y entre 7 y 12 hrs. en México, cuando hay personal trabajando en las oficinas de ambos lados del océano. Fuera de estas "horas pico" existe muy poca demanda de tráfico para canales de telefonía, pero los canales podrían utilizarse, por ejemplo, entre México y Sudamérica. Esto conduce a considerar otras técnicas de acceso al satélite en FDMA, como son SCPC de asignación - fija y por demanda (DAMA y SPADE).

En la tabla 2 se proporcionan algunos datos relativos al tráfico manejado por la estación internacional de Tulancingo.

Tabla 2. Algunos enlaces vía satélite operados por Tulancingo 1 y 2.

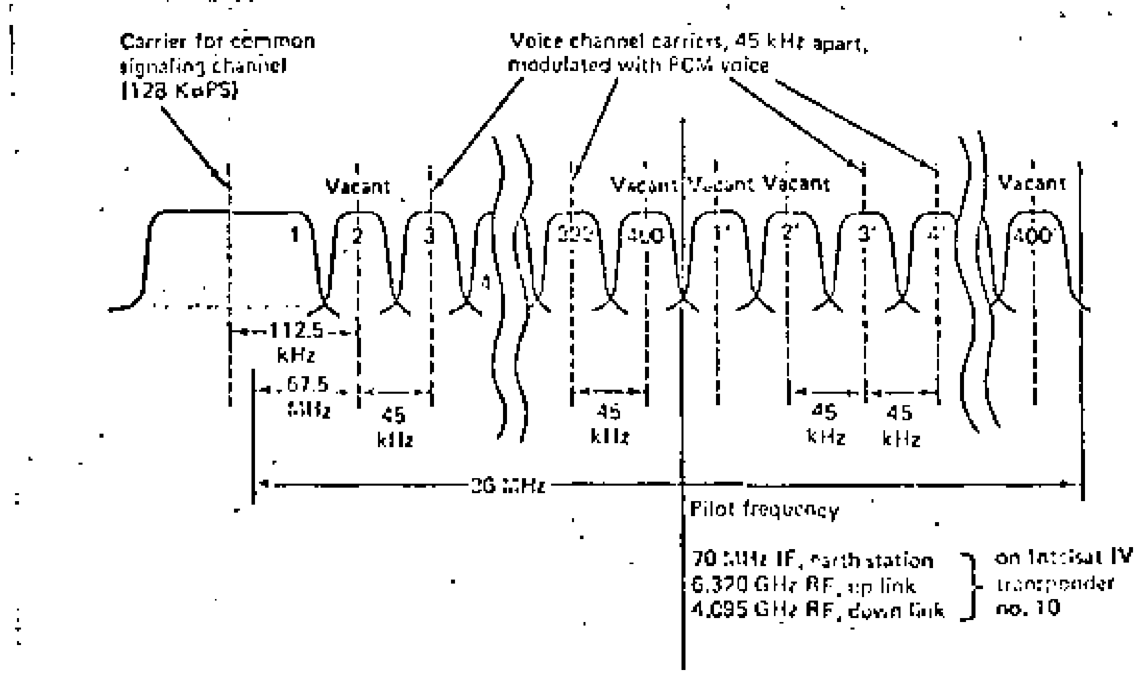
Enlace entre México y	Tx, MHz	Rx, MHz	Tipo de acceso y nº de canales
España	5990.625	3765.625	Asig. fija (1 grupo de 12 c.)
Italia	5984.0	3759.0	" "
Brasil	6038.5	3813.5	" "
Chile	6280.5	4065.5	" "
Venezuela	6077.25	3852.25	" "
Alemania	14060.25	3760.25	" "
Francia	14032.0	3732.0	" "
Inglaterra	14077.25	3777.25	" "
Francia	6042.75	3817.75	" "
Inglaterra	6242.75	4017.5	" "
Cuba	—	—	" 6 canales SCPC
Países participantes de SPADE	—	—	SPADE 12 canales SCPC

- SCPC (Single Channel Per Carrier ó
  Canal Unico por Portadora)

La técnica SCPC tiene gran aplicación cuando se desean - interconectar estaciones terrenas de muy baja capacidad o de-- manda de tráfico. Consiste en que a cada canal de telefonía - se le asigna una frecuencia portadora de RF, misma que es modu lada por la señal de voz en FM o PSK. Dado que las llamadas son aleatorias, el espectro del transpondedor se puede aprovechar eficientemente si las frecuencias portadoras de RF se asignan temporalmente a las estaciones terrenas, es decir, únicamente mientras tengan información que enviar. Cuando una estación A termina de transmitir su información, la frecuencia de portadora que se le había asignado pasa a un "banco de frecuencias" controlado por una computadora central. Si otra estación B desea entonces establecer un enlace, la computadora central le -- asignará una de las frecuencias disponibles en el "banco" y quizás se le otorgue la misma frecuencia que antes había utilizado la estación A. Como el sistema funciona en base a este banco - de frecuencias y al criterio de "servicio a quien pida primero", la técnica recibe el nombre de DAMA (Demand Assignement Multiple Access ó Acceso múltiple de asignación por demanda). Cuando los canales de voz están codificados en PCM (de acuerdo a recomendaciones vigentes de la CCITT), la técnica se conoce como SPADE -- (Single-Channel-per carrier PCM multiple-Access Demand assignment Equipment ó equipo de asignación por demanda en acceso múltiple para canal PCM único por portadora).

INTELSAT IV fue el primer satélite en utilizar SPADE en uno de sus transpondedores. La figura 12 muestra las 800 ranuras de frecuencia en las que se divide un transpondedor de 36 MHz. Se tienen 800 portadoras de RF diferentes, de las cuales 794 se emplean para establecer 397 circuitos telefónicos (un circuito ocupa dos ranuras para el canal de la persona A y el canal de la persona B). El espaciamiento entre cada ranura o canal es de - 45 KHz. Nótese que en el extremo izquierdo se tiene una ranura

Fig. 12


Carrier for common signaling channel (128 KbPS)
Voice channel carriers, 45 kHz apart, modulated with PCM voice
Vacant
Vacant
Vacant
Vacant
Vacant
1
2
3
4
399
400
1'
2'
3'
4'
400'
112.5 kHz
67.5 MHz
45 kHz
45 kHz
45 kHz
45 kHz
36 MHz
Pilot frequency
70 MHz IF, earth station
6.370 GHz RF, up link
4.095 GHz RF, down link
on Intelsat IV transponder no. 10

de mayor ancho de banda; se utiliza para el canal de canalización y en él se transmiten 128 000 bps. Este canal es el que contiene la información variante con eltiempo de qué frecuencias están utilizando unas estaciones y cuales están disponibles para nuevos solicitudes de transmisión.

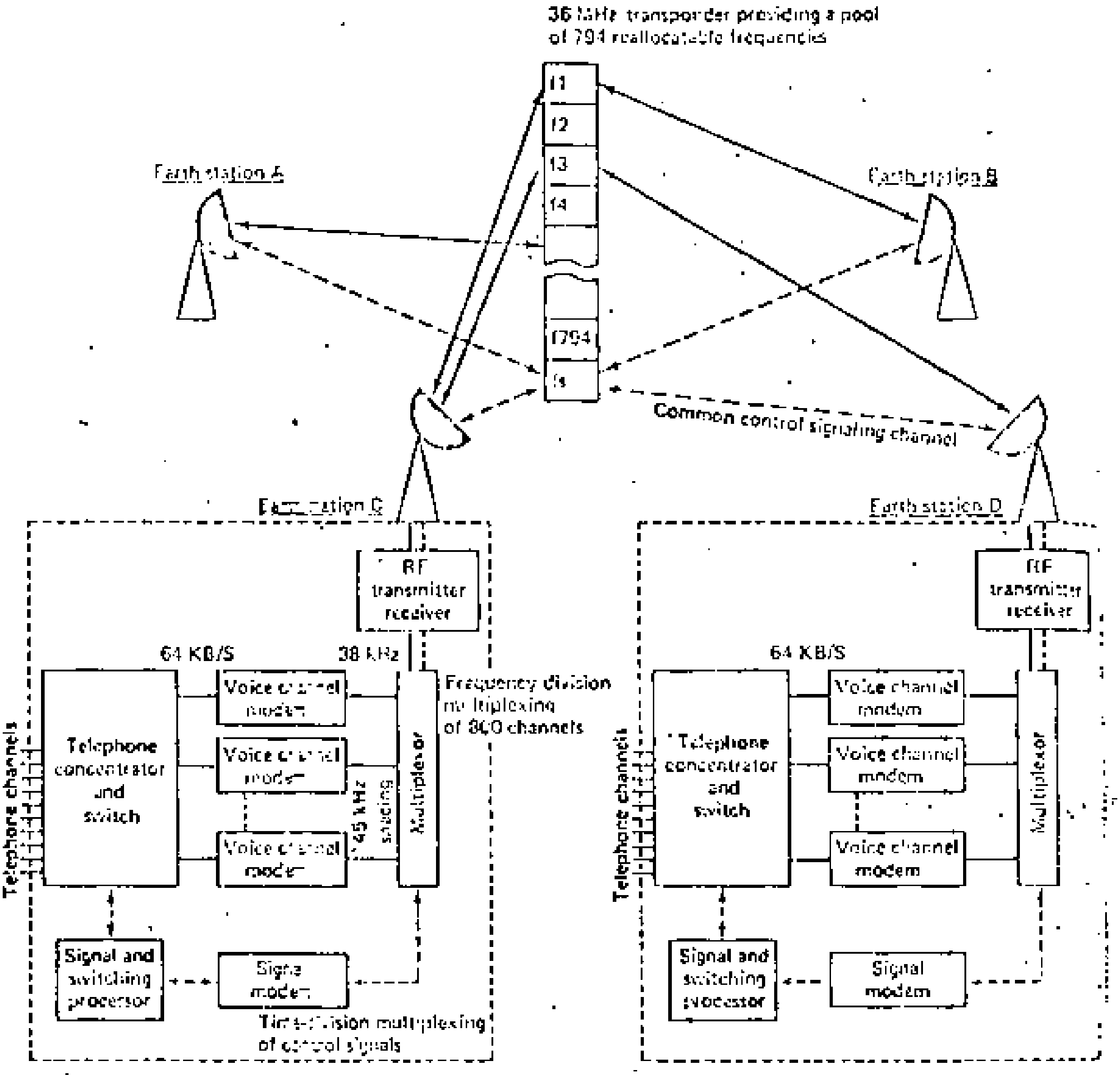
En páginas anteriores, en la Tabla 1, se observó que conforme aumenta el número de portadoras también decrece drásticamente la capacidad en número de canales del transpondedor. De aquí surge la pregunta: ¿cómo es posible tener 800 portadoras en un solo transpondedor? Existen dos razones que justifican la estructura del sistema SPADE:

1a. Al tener un solo canal por portadora, ésta se puede apagar (cero potencia transmitida) cuando no haya voz presente, lo cual sucede cuando menos el 50% del tiempo en que uno establece una conversación, ya que algunas veces uno solamente escucha, y aún hablando se producen pausas entre palabras. Esto provoca que en realidad, se tengan menos de 400 portadoras encendidas al mismo tiempo en un transpondedor.

2a. Los grupos y supergrupos de canales multiplexados en la Tabla 1 son modulados en FM y ocupan más ancho de banda que su equivalente digital. Los canales SPADE son modulados con PSK de cuatro fases; cada canal se codifica en PCM a 64 000 bps y se obtiene una buena calidad subjetiva con espaciamientos de 45 KHz entre canales.

Como puede verse, esta técnica es atractiva, aunque su costo aumenta con respecto al de asignación fija ya que se requiere contar con un complejo controlador DAMA en cada estación.

En la figura 13 se ilustra el esquema de operación de SPADE. todas las estaciones utilizan secuencialmente todo el canal común de señalización, empleando PSK de dos fases. Para ésto, a cada estación se le asigna un milisegundo para transmitir 128 bits; algunos de éllos son de sincronización, otros de detección de errores y otros de información sobre enlaces en operación y "nuevas



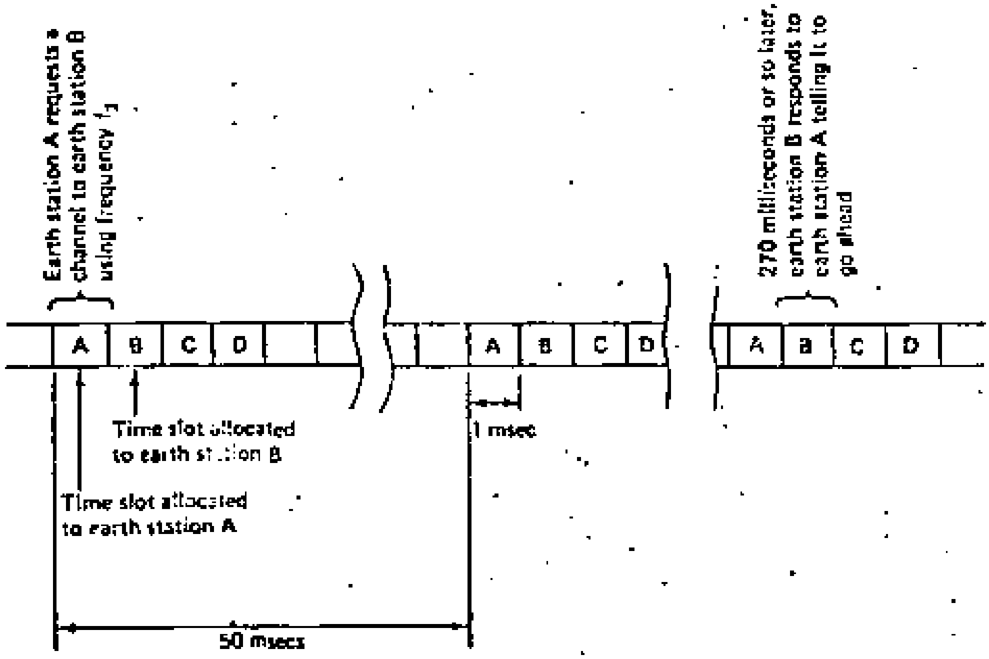
The operation of the SPADE frequency-division demand-assignment system.

Fig. 13

solicitudes". Cada estación dispone de su ranura de tiempo cada 50 milisegundos para actualizar su banco de datos. Por lo tanto, se pueden enlazar hasta 49 estaciones. En la figura 14 se muestra la utilización del canal común de señalización, y en la Tabla 3 se indican las principales características de un canal SPADE.

Tabla 3. Características principales de un canal SPADE

	Canal de comunicación	Canal de señalización
codificación	PCM	—
modulación	PSK-4ϕ coherente	PSK - 2ϕ
velocidad	64 kbps	128 kbps
ancho de banda	38 KHz	—
espaciamiento entre canales	45 KHz	—
estabilidad requerida	$\pm$ 2 KHz	—
tasa de errores	10^{-4}	10^{-7}
acceso	FDMA-SCPC-asignación por demanda	TDMA
longitud del marco	—	50 mseg.
longitud de la ráfaga de datos	—	1 mseg.
número de accesos	397	49 más una estación de referencia para sincronización



Earth station A requests a channel to earth station B using frequency f_3
270 milliseconds or so later, earth station B responds to earth station A telling it to go ahead
A
B
C
D
Time slot allocated to earth station B
Time slot allocated to earth station A
1 msec
50 msecs

Fig. 14

Acceso múltiple por división en tiempo (TDMA)

En páginas anteriores se mencionó que la característica de amplificación de un TWT es no lineal y que si se opera cerca de saturación, se producen niveles elevados de ruido de intermodulación, que afectan la calidad de las señales transmitidas.

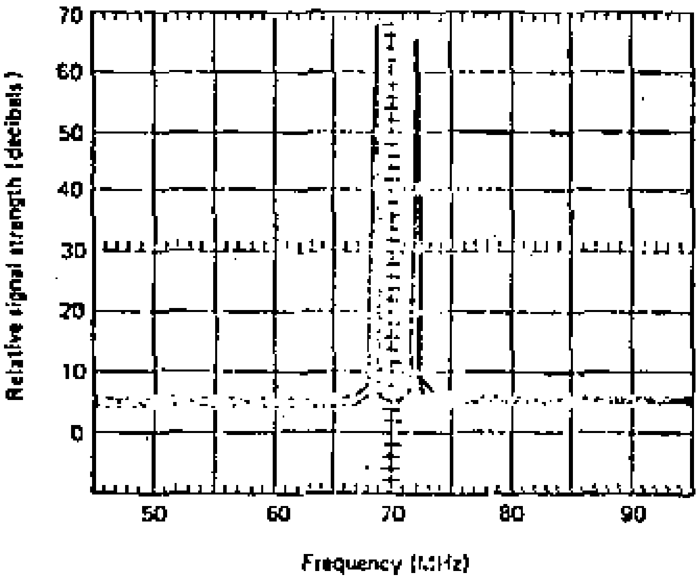
Además, se indicó que el ruido de intermodulación puede reducirse operando al amplificador con un back-off. El posible efecto de estos productos de intermodulación que se salen de la banda de una portadora determinada, puede visualizarse con ayuda de la figura 15. En éila se muestran los resultados experimentales al enviar a través de un transpondedor de un satélite ANIK canadiense a dos tonos de prueba. El experimento consistió en aumentar la intensidad de las dos señales hasta saturar el transpondedor. Al reducir por 3 dB al tono de entrada de mayor frecuencia, se observa una distribución de ruido de intermodulación como la de la figura 15.

El concepto básico de TDMA consiste en usar una sola portadora de banda ancha que ocupe todo el transpondedor. Esto permite operar el satélite a máxima potencia, es decir, en saturación, aún cuando opere en su región no lineal. Se tiene la ventaja de que también en tierra, las estaciones pueden operar en saturación al transmitir. El funcionamiento de un sistema TDMA es el siguiente:

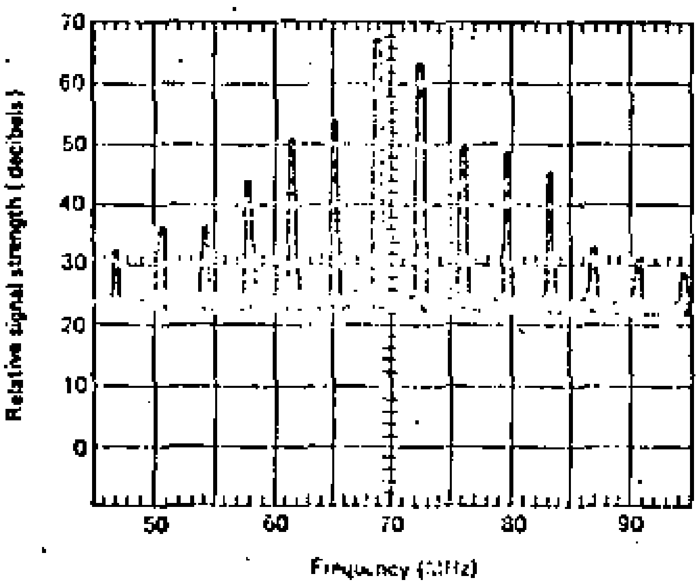
- A cada estación terrena se le asigna secuencialmente un intervalo de tiempo para que utilice todo el transpondedor. La estación puede transmitir dentro de este intervalo una portadora saturada de 36 MHz que contenga información digital mezclada de voz, datos y video.
- En su forma más simple, a todas las estaciones se les asignan secuencialmente intervalos de tiempo de la misma longitud, en forma similar al uso del canal de señalización del sistema SPADE.

Fig. 15

TWO TONES ARE TRANSMITTED TO THE SATELLITE AS SHOWN:



THE FOLLOWING SIGNAL IS RECEIVED AT EARTH STATIONS, SHOWING HOW INTERMODULATION HAS OCCURRED BETWEEN THE TWO TONES:



An example of the intermodulation that occurs when multiple carriers are transmitted via the same transponder. (*Drawn from cathode ray tube photographs in reference 1.*)

- Para operar más eficientemente, cada estación debe tener la flexibilidad de variar su velocidad de transmisión, de modo que las ranuras de tiempo asignadas deben ser de longitud variable (de acuerdo a la demanda o solicitud de cada estación), o bien, las estaciones que así lo necesiten deben tener preferencia y poder transmitir con mayor frecuencia.
- La mayoría de los sistemas operan bajo asignación por demanda. Para éllo se tiene un canal de control que informa a todas las estaciones sobre las asignaciones efectuadas y recibe nuevos solicitudes. A este canal se le denomina algunas veces como canal de servicio (order wire).

El sistema TDMA es muy atractivo, pero requiere de equipo altamente confiable de sincronización. El problema no se limita a asignar intervalos a las estaciones, en forma similar a la de un equipo de conmutación TDM, sino que deben considerarse los desplazamientos del satélite con respecto a su posición normal.

Los satélites están sujetos a varios fenómenos naturales que alteran sus posiciones a largo y corto plazo. Simplemente, en un lapso de 8 días, pueden sufrir oscilaciones de posición como las mostradas en la figura 16, que son efecto de las fuerzas de atracción del sol y la luna. Al cambiar la posición de un satélite, aumenta o disminuye la distancia directa estación X-satélite y, por consiguiente, varía el tiempo de propagación de la señal binaria.

Un transpondedor típico de 36 MHz puede manejar 60 Mbps en QPSK. A estas velocidades, el tiempo entre bits es de 16.67 nanosegundos, y de acuerdo a la figura 16, en donde se muestra la velocidad de alejamiento del satélite con respecto a su posición normal, puede haber variaciones en el tiempo de propagación hasta de 2 nanosegundos, por lo que se requiere mantener una sincronización constante.

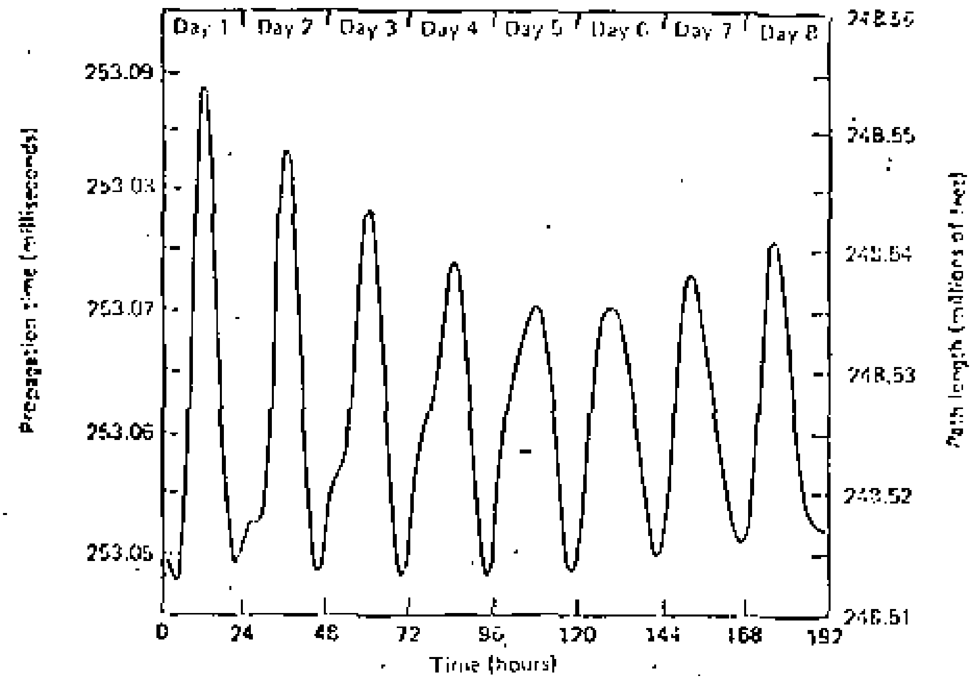
En las figuras 17 y 18 se muestran las estructuras de un marco maestro en TDMA. La duración típica de un marco es de 1 milisegundo, o sea que tiene capacidad para la transmisión de 60 000 bits. La primera ráfaga de bits del marco sombreada en la figura 17, es el canal de supervisión; lo transmite una estación central de control y contiene información de sincronización y de identificación del marco.

La duración de un marco maestro es variable y depende de la demanda de tráfico por cada estación terrena. Un marco maestro está formado por varios marcos, todos iguales, en donde cada estación tiene asignadas ranuras de tiempo fijas. Al variar la demanda, se envían nuevas instrucciones en la ráfaga de la estación central de control, indicando una nueva distribución de tiempos o configuración de marcos, e iniciándose de esta forma un nuevo marco maestro.

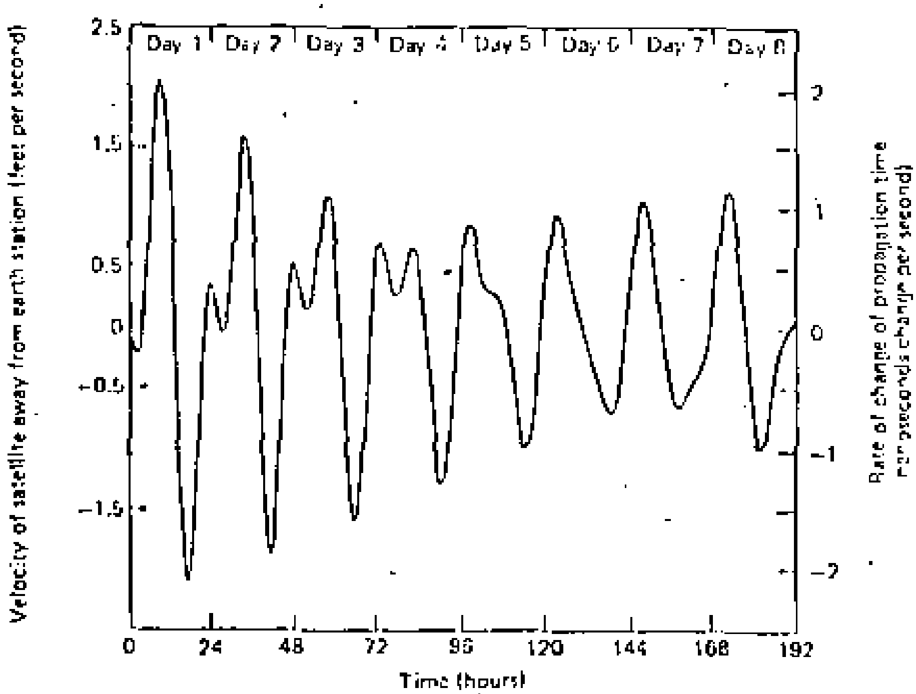
Por lo que respecta a la ráfaga de bits enviada por cada estación, ésta debe contener un préambulo o encabezado para sincronización e identificación, seguido de la información que se desea transmitir. En la figura 19 se ilustra la estructuración de un marco, incluyendo los bits de control, del sistema de COMSAT denominado MATE.

Como complemento final de esta sección, a continuación se anexan algunos diagramas ilustrativos sobre el modo de acceso y el uso de los transpondedores de los satélites canadienses ANIK B y C, en las bandas C y Ku.

Fig. 16

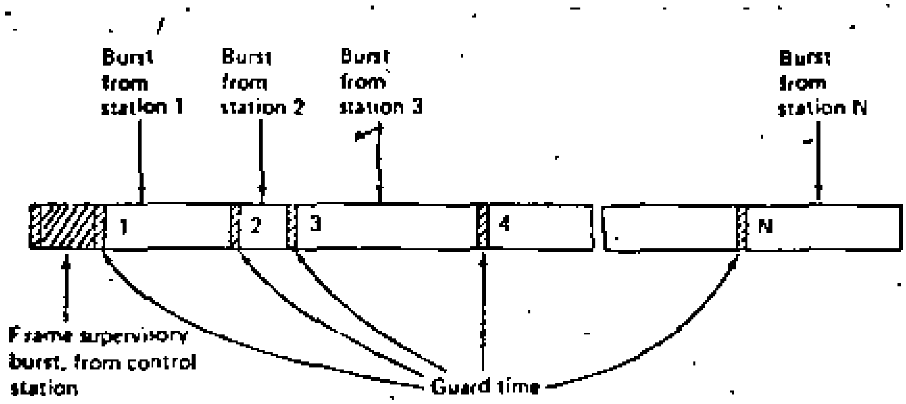


Variations in path length and propagation time via a satellite, measured over an 8 day period



Variations in velocity of the satellite relative to an earth station, for the same 8 days

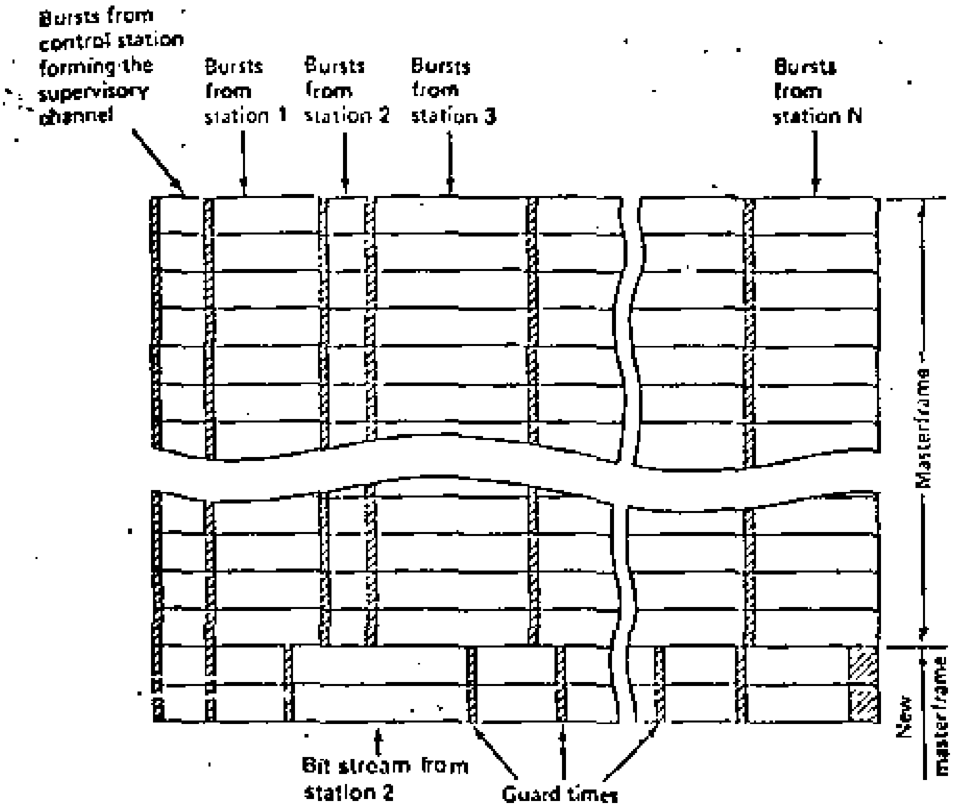
"Tidal" variations in satellite position. A high-speed digital satellite link needs careful synchronization.



A frame in a TDMA system. A typical frame length is on the order of 1 millisecond.

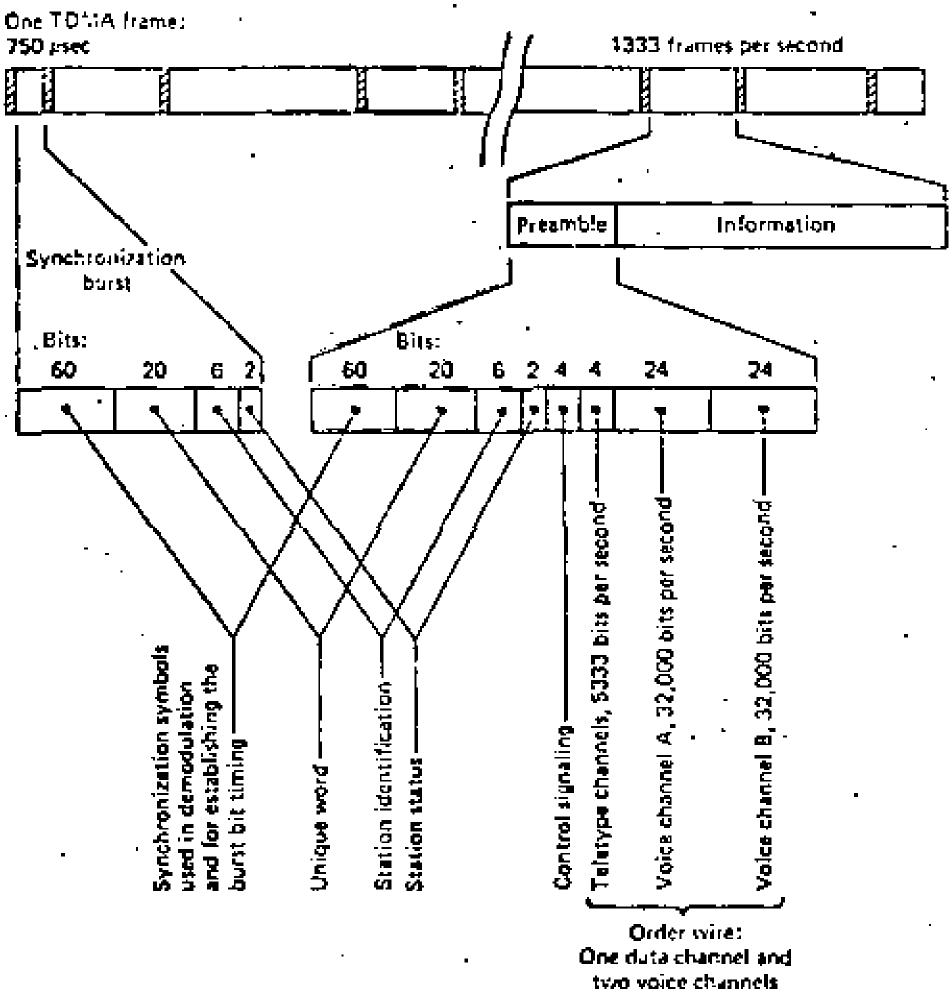
Fig. 17

Fig. 18



A given number of frames follow one another, forming a masterframe. The vertical columns form a continuous bit stream from each earth station containing channels to different earth stations. The burst size allocations will be different in each masterframe if the channel demands vary. Only the supervisory channel will remain the same.

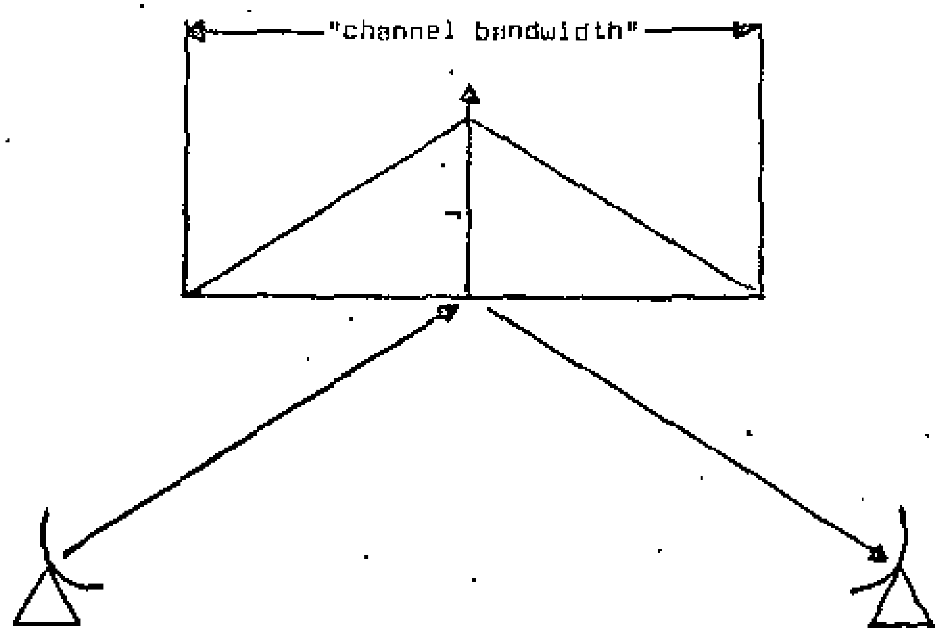
Fig. 19



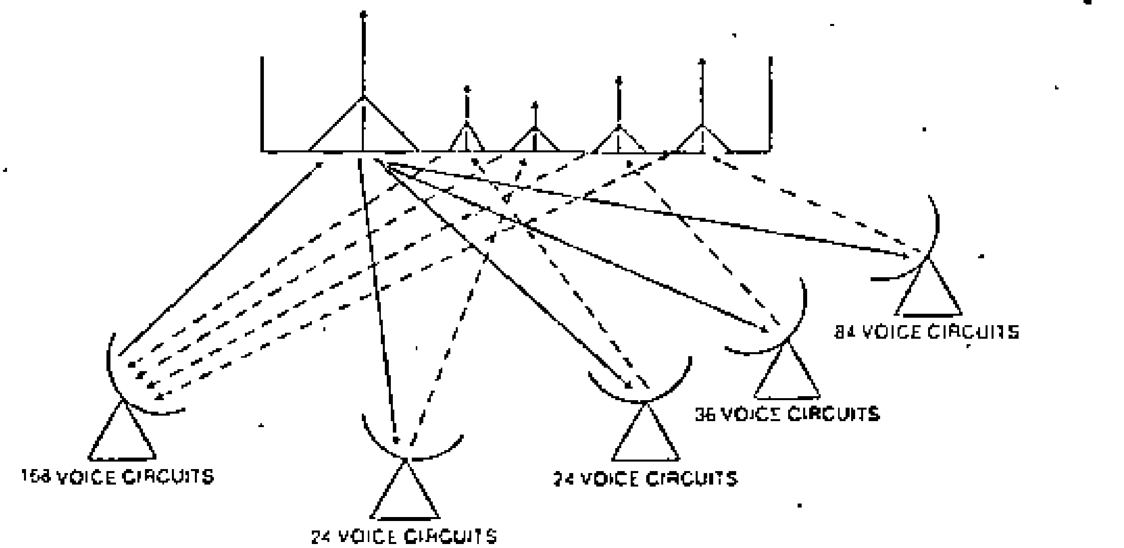
Bits used for synchronization and control of TDMA frames on the Comsat MATE system [2].

Diagramas ilustrativos sobre el modo de acceso y el uso de los transpondedores de los satélites ANIK B y C.

Heavy Route Message Traffic (Frequency Modulation — Single Access)



Medium Density Message Traffic (Frequency Division Multiple Access (FM/FDMA)



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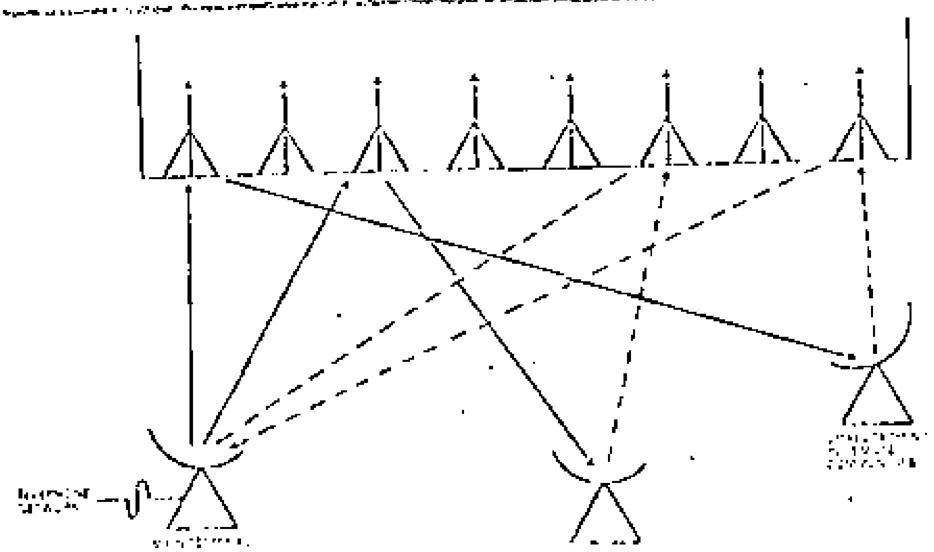
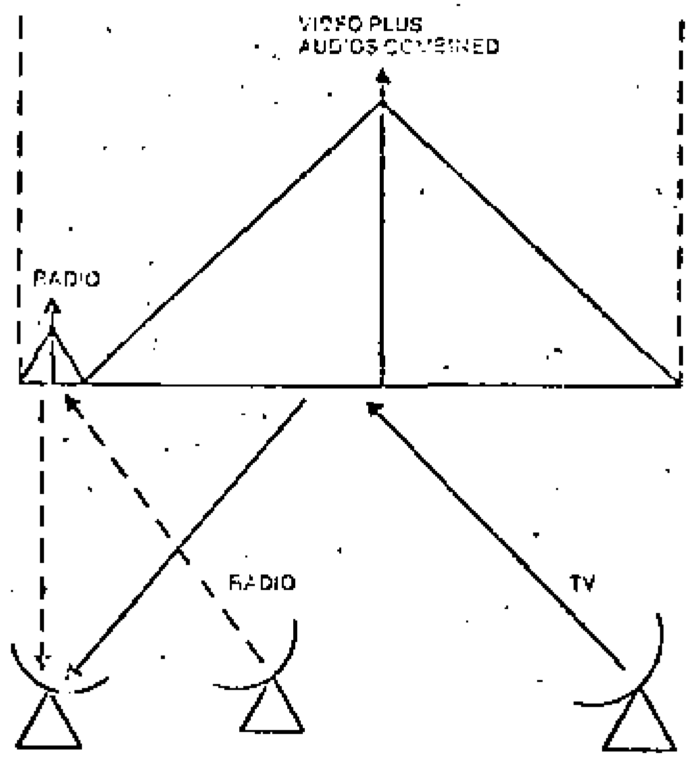
Thin Route Message Traffic (Single Channel Per Carrier SCPC/FDMA)



Television / Radio Channel



Earth Station Characteristics - Message Facilities

TRAFFIC		ANTENNA DIAMETER (m)	G/T (dB/K)	QUALITY	POWER BACK UP	REDUNDANCY
Heavy Route		30	37.5	37.5 dBrncO	Batteries & Generator	Yes
Medium Density FM/FDMA	a	30	37.5	37.5 dBrncO	Batteries & Generator	Yes
	b	10	28	44.0 dBrncO	Batteries & Generator	Yes
	c	8	27.5	44.0 dBrncO	Batteries & Generator	Yes
PCM/TDMA	a	30	37.5	34 dBrncO	Batteries & Generator	Yes
	b	10	31	34 dBrncO	Batteries & Generator	Yes
Thin Route Message	a	30	37.5	Idle Noise 37.5 dBrncO max.	Batteries & Generator	Yes
	b	10	28		Customer Provided	Yes
	c	8	26		Batteries	Yes
	d	4	22		Batteries	Yes
	e	4.5	21		Batteries	Yes
Transportable	a	3.6	19		Batteries	Can be provided
	b	3.6	19		No	No
Light Route TDMA		4.5	22	Idle Noise 35 dBrncO max.	No	No

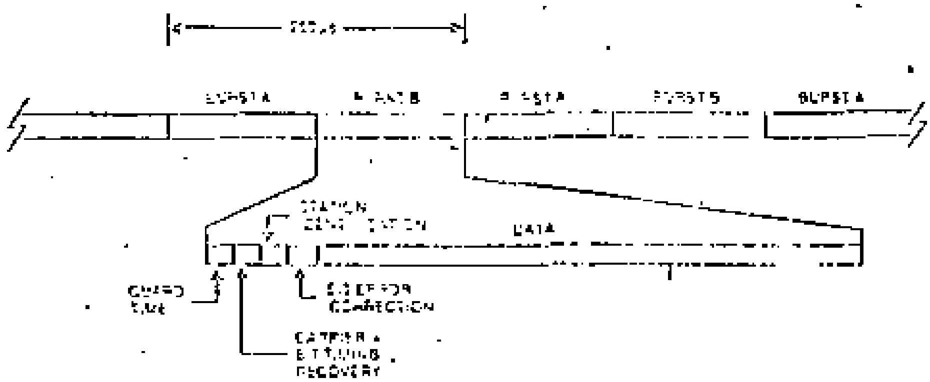
Earth Station Characteristics - Television & Radio 234

TRAFFIC	ANTENNA DIAMETER (m)	G/T (dB/K)	QUALITY	POWER BACK UP	REDUNDANCY
Network TV (Transmit Receive)	a 30 b 10	37.5 28	54 dB Video S/N	Batteries & Generator	Yes Yes
Northern TV (Transmit only)	a 11 b 8	NA NA	Min 45 dB Video S/N	Customer Provided	No No
Remote TV (Receive Only)	a 8 b 8 c 4.5	26 22 21.5	48 dB Video S/N	Some Have Batteries	Some Have
Frontier TV (Receive Only)	a 4.5 B 3.6	18.5 18.5	45 dB Video S/N	No No	No No
Transportable TV (Transmit Receive)	a 4.5 b 4.5	NA NA	43 dB Video S/N	Generator No	Yes No
Facsimile Transmit Receive	8.0 4.5	NA 21.5	Better than 85 line screen	No No	No No

Earth Station Characteristics - 14/12 GHz Facilities 235

TRAFFIC	ANTENNA DIAMETER (m)	G/T (dB/K)	QUALITY	POWER BACK UP	REDUNDANCY
Digital Message	8	35	BER 10^{-7}	Customer Provided	Yes
Television	a 8 b 8	35 29.5	48-54 dB Video S/N	Customer Provided	Yes
	c 4.5	26.5	46-52 dB Video S/N	As Required	As Required
Television Receive Only	a 1.8	16.5	40-42 dB Video S/N	No	No
	b 1.2	13	42 dB Video S/N (Typical)	No	No
Transportable TV Transmit Transmit/Receive	1.8 3.7	N/A 25	43 dB Video S/N	Generator	No

Medium Density Message Traffic (Time Division Multiple Access — TDMA)



2.8 SUPRESORES Y CANCELADORES DE ECO

En toda conexión telefónica ocurren ecos debido al desacoplamiento de impedancias en el circuito híbrido de conversión de 2 a 4 hilos. El eco consiste en que parte de la señal de entrada al híbrido se refleja por la trayectoria de salida hacia el extremo desde donde se originó dicha señal de entrada. La magnitud y característica espectral del eco depende del circuito que se establece para cada conversación telefónica en particular (diferentes alternativas de enlace y combinaciones conversador 1/conversador 2), de tal forma que resulta imposible agregar en el híbrido una impedancia fija de compensación para todas las llamadas.

El eco es particularmente severo en enlaces vía satélite, debido a que el viaje redondo (salto de ida + salto de regreso) puede tomar hasta 0.6 segundos (0.25 X 2 seg. + retraso en el enlace terrestre).

Para enlaces vía satélite existen dos alternativas para reducir ó eliminar los efectos del eco, usando:

- supresores de eco

ó - canceladores de eco.

En un supresor de eco, se introduce un dispositivo que determina qué persona está hablando, y se inserta una pérdida muy grande en la trayectoria de regreso. La desventaja de los supresores de eco es que tienden a "cortar" la parte inicial de las palabras y provocan interrupciones molestas en la conversación (chopping).

Un cancelador de eco no causa el molesto "chopping" de los supresores de eco, ya que consiste en un circuito que genera una réplica del eco, incluyendo retrasos, y la resta de la señal en la trayectoría de regreso. A continuación se anexan detalles sobre el modelo 4B ECHO CANCELER de AT & T International, desarrollado en los laboratorios Bell. Consiste básicamente de un circuito integrado a muy grande escala (VLSI) que contiene 50,000 transistores. Se adapta a cualquier híbrido y cancela el eco en el punto donde se origina.

Pruebas subjetivas han revelado que el usuario prefiere la cancelación del eco, a su supresión, especialmente en enlaces vía satélite. Simplemente en Estados Unidos existen en operación más de 60,000 canceladores fabricados por Western Electric en circuitos terrestres y vía satélite.

Otro ejemplo de cancelador de eco, es el modelo EC-5000 de TeleSystems, subsidiaria de la compañía COMSAT (se anexa copia del catálogo). Este es un cancelador digital multicanal, a diferencia de los canceladores tradicionales de canal único. Opera en configuración full-duplex con circuitos de stándar americano T1 de 24 canales de voz, a 1.544 Mbps. A cada canal de voz de 64 Kbps (codificado en PCM con 8 bits) se le asigna un procesador individual de cancelación de eco, que inserta una cancelación del orden de 45 dB.

A continuación se anexa información relevante respecto a pruebas subjetivas realizadas por INTELSAT, un estudio comparativo de la RCA sobre supresores y canceladores de eco, pruebas subjetivas realizadas por la AT & T, y estandarización de la CCITT.

CANCELADOR DE ECO 239

4 B - A T T

ECHOES ECHOES

AT&T

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ONCE UNAVOIDABLE, NOW ECHOES CAN BE ELIMINATED

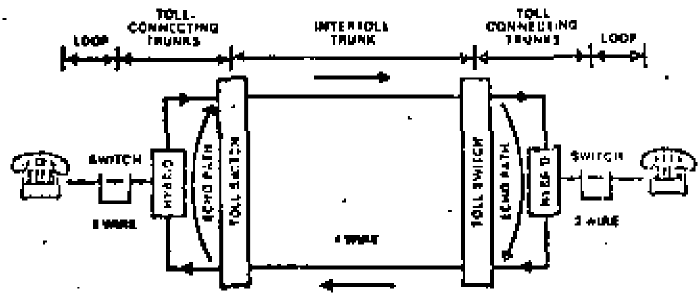
All telephone connections exhibit echoes to some extent. They arise at the hybrids that connect the four-wire circuits used in toll links to the two-wire circuits used in exchange - subscriber loops (see Figure 1 below).

Echoes are particularly troublesome in long-haul circuits; the long delay before hearing the returned echo makes it more noticeable and can influence the customer's ability to converse. With satellite links, the problem is particularly severe. A satellite in geostationary orbit introduces a transit delay of about 0.3 second from sender to receiver. If the satellite is used for both directions of transmission (the normal full-hop mode), the round trip delay is 0.6 second. Tests show delay alone is not especially troublesome to subscribers, but the occurence of echoes with delay causes annoyance.

In the past, telephone engineers have sought to reduce echoes with two principle strategies: loss and suppression. For trunks less than 3,000 kilometers (1,850 miles) long, intentionally adding a bit of loss–up to 3dB–altenuates the returning echo enough to make it less objectionable. Alternatively, an echo suppressor may be used, which operates by determining which party is talking and inserting a high loss in the return path. However, echo suppressors have two serious failings: they tend to clip the initial parts of words, and they lead to a very distracting "chopping" of the conversation, particularly on satellite circuits, when both parties speak at the same time.

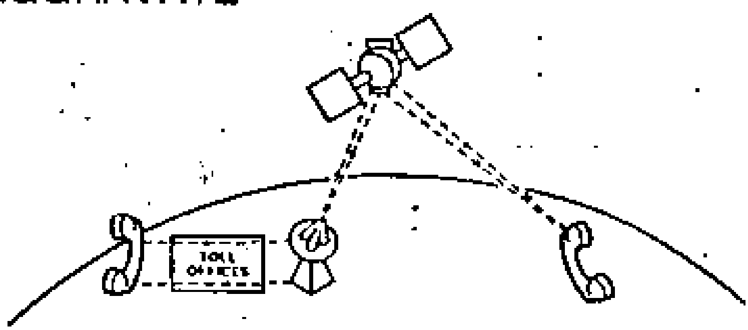
HOW ECHOES ARISE

FIG. 1.
Echoes arise because of impedance mismatches at the two-wire to four-wire hybrid junction of telephone circuits. Some of the incoming signal is reflected as echo back to the sender. Exact shape and magnitude of the echo depends on the circuit established for that particular call, so no fixed impedance can compensate for all calls.



SATELLITE CIRCUITS AGGRAVATE ECHO PERCEPTION

FIG. 2.
Satellite links are much longer than terrestrial links. The resultant delay is not necessarily annoying in itself, but it makes echoes much more noticeable and disturbing.



HOW THE NEW ECHO CANCELERS WORK TO IMPROVE CIRCUIT PERFORMANCE

Canceling the echo before it returns

FIG. 3. VLSI Chip inside brown dotted lines.

The echo canceler, installed at toll offices, samples the received signal (shown in dark purple) and creates a replica of the echo that will be generated in the hybrid and two-wire connection to the local subscriber. At the summing junction, the replica echo (shown in light purple) generated by the echo path model is subtracted from the real echo (also shown in light purple) and the signal from the near end (shown in green), effectively canceling only the echo.

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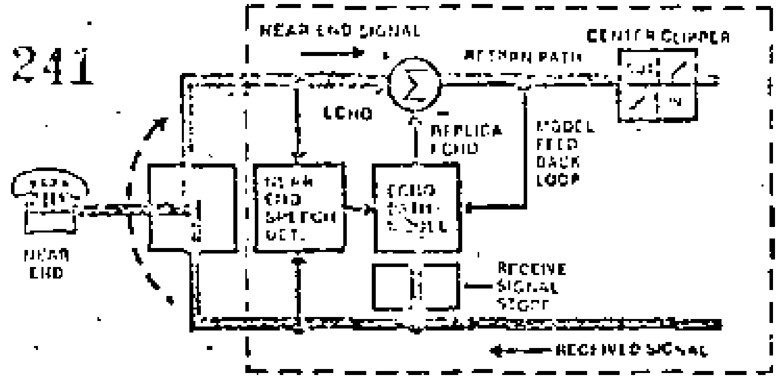


Echo cancelers are a vast improvement over echo suppressors of the past. The principle of echo cancellation which was discovered at Bell Laboratories is entirely different from echo suppression. The echo canceler is an adaptive circuit that generates a close replica of the echo, complete with appropriate delays, and subtracts it from the signal in the return path. This substantially cancels the echo, improving communication, but does not introduce any chopping or clipping.

How is the replica of the echo created? Through the use of special adaptive circuits developed by Bell Laboratories and manufactured by Western Electric. At the heart of the echo canceler is a Very Large-Scale Integrated (VLSI) circuit chip (see photo) containing 50 thousand transistors to perform the following functions:

- Convert A or μ-Law encoded speech to linear and floating point
- Store incoming speech signals in a shift register delay
- Approximate the echo path impulse response adaptively
- Automatically weight each of the delayed samples to produce the optimum match to the returned echo
- Sum the outputs of all weighted, delayed samples and subtract that sum from the returning echo
- Detect near end speech and center clip residual echo as necessary

Thus the echo canceler adapts itself to the particular hybrid and two-wire loop used for each conversation. The echo is canceled at the end where it is created. Ideally, echo cancelers should be installed at both ends of the circuit; however, they are compatible to work with echo suppressors which may be in operation at the other end of the circuit.

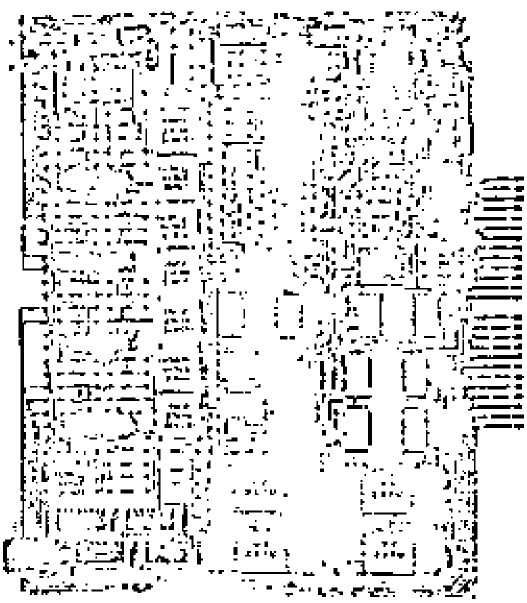
USERS PREFER ECHO CANCELLATION

Subjective tests confirm that users prefer echo cancellation to echo suppression--especially on full hop satellite links. More than 60,000 echo cancelers, manufactured by Western Electric, are at work on terrestrial and satellite circuits in toll offices throughout the United States.

COMMUNICATION SYSTEMS WORLDWIDE

The Echo Canceler Product Family is one of the leading edge technological advancements designed by the Bell System. It represents a commitment to providing communication systems worldwide, solving customer problems, and supplying service excellence. For more information on our products and resources, contact your local AT & T International Representative.

THE 4B ECHO CANCELER SYSTEM CREATED FOR WORLDWIDE USAGE



THE ECH[illegible] CIRCUIT BO[illegible]

The VLSI canceler used in the 4B Echo Canceler System is a digital device designed to cancel echoes on a single voice circuit. Each echo canceler plug-in board contains the necessary electronics to provide simultaneous echo control for two separate telephone circuits.

Board size is 215 mm deep by 283 mm high. The face plate is equipped with jacks to allow testing of the echo canceler while mounted in the shelf.

Transmission level points may be changed in the field by adjusting the attenuation pads mounted on the circuit board.

A tone-detector circuit is available as an option. The tone disabler is designed to meet CCITT Recommendation G. 164 Section 4.

The convergence speed may be selected to be 25, 50, 100 or 200 dB per second.

External pin-outs to the edge connector are provided for remote enabling/disabling or resetting of the echo canceler by an external ground.

The 4B Echo Canceler System has been designed to meet the needs of communication systems worldwide. It is particularly useful on trunk circuits with long propagation delay, especially satellite circuits. To cancel echoes, it uses principles that were discovered at Bell Laboratories and a custom Very Large-Scale Integrated (VLSI) circuit chip that was developed by Bell Laboratories and manufactured by Western Electric. Echo canceler performance is superior to performance obtained with echo suppressors, as demonstrated in user tests and through continuing customer satisfaction.

The 4B Echo Canceler System can be used worldwide without modification.

The 4B Echo Canceler meets the performance requirements specified in CCITT Recommendation G. 165 - Echo Cancelers.

The 4B Echo Canceler System provides a self-contained voice-frequency analog interface and digital echo canceler with 32 msec (nominal) echo tail capability.

A complete 4B Echo Canceler Shelf Assembly consists of a shelf containing echo canceling capability for 24 individual voice-frequency (300 to 3400 Hz) circuits and a power supply working off -48V office power. The shelves are designed to mount in the proposed CEPT 600 mm × 260 mm × 2100 mm bay and feature front-access for easy cabling and maintenance.

[illegible]

While the echo canceler is mounted in the shelf, analog access to both the transmit and receive ports on both the facility and drop side are provided by four splitting jacks. Thus, if the circuit is busied out for maintenance, commonly available equipment, such as a Return Loss Measuring Set or simply a noise source and a meter, together with an attenuator pad, can be used to test the canceler.

4B ANALOG ECHO CANCELER

FEATURES

1. 32-msec echo tail capability
2. Selectable Echo Path Return Loss
3. Adjustable convergence rate of canceler
4. External control of echo canceler enabling and resetting
5. Provision for adjusting the analog Interface TLPs (Transmission Level Point)
6. 2100-Hz tone disabler
7. Center clipper to remove low level residual echo, equipped with noise matching to eliminate noise modulation
8. Front jack access for local testing

4B ECHO CANCELER SPECIFICATIONS

Echo Path Return Loss (ERL):	0 or 6 dB (selectable)
ERL Enhancement:	> 30 dB
Residual Echo with Center Clipping	< −65 dBm0
Convergence Time:	100 to 800 msec (selectable)
Maximum Impulse Response Length	> 32 msec (nominal)
Transmission Levels:	−8 to +7 TLP receive (selectable) −16 to 0 TLP transmit (selectable)
Insertion Loss:	0 ± 0.3 dB
Frequency Response:	± 0.5 dB (400 to 3000 Hz)
Input/Output Impedances:	600 ohms, balanced
Idle Channel Noise:	< 19 dBrnc0
Harmonic Distortion:	< 1%
Dynamic Range:	+3 to −60 dBm0
Tone Disablers:	Meets CCITT Recommendation G. 164 Section 4

COMMUNICATION SYSTEMS WORLDWIDE

The 4B Analog Echo Canceler has been designed by the Bell System to meet the needs of communication systems worldwide. It represents the Bell System's commitment to perform at the leading edge of technological advancement. For more information on the 4B Analog Echo Canceler and our other products, contact your local AT & T International Representative.

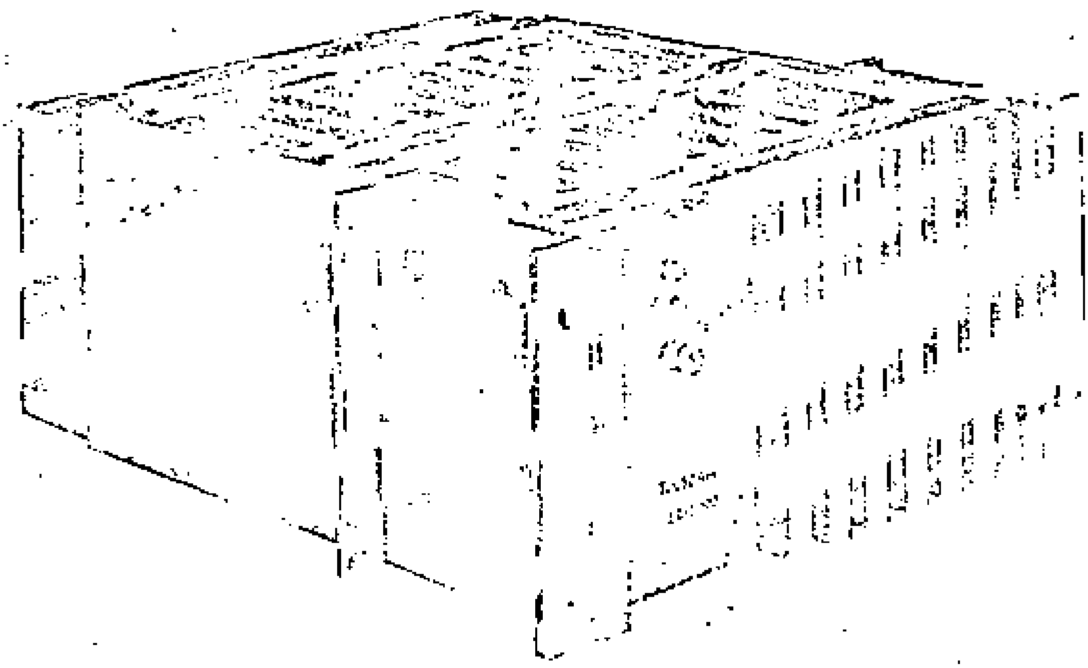
AT&T International

1B - 211A 4/82 Printed In U.S.A.

CANCELADOR DE ECO DIGITAL MULTICANAL

EC-5000, COMSAT GENERAL TELESYSTEMS

EC-5000 MULTICHANNEL DIGITAL ECHO CANCELLER



The EC-5000 is the latest advancement in echo canceller technology for long distance telephone communications over multichannel digital transmission facilities. Complies with CCITT Recommendation G.165.

A QUANTUM LEAP

MSAT General TeleSystems' contribution to the development of echo canceller technology, established by our family of single-channel echo cancellers, takes a quantum leap forward with the introduction of the EC-5000 Series Multichannel Digital Echo Canceller.

Designed to be cost-effective, and efficient, the EC-5000 is a stand-alone device that operates in a full duplex 24-voice channel 1.544 Mbps T1 circuit. Each 64 Kbps 8-bit PCM mu-law companded voice channel is assigned an individual echo cancelling processor, which adapts to the tail circuit on a dynamic basis and provides up to 45 db of echo cancellation on any tail circuit with up to 32 msec (1600 miles) delay.

Advanced Digital Design Complies with CCITT Recommendation G.165

The EC-5000 is a *totally digital device* requiring no adjustments or alignment during installation or operation. *Modular construction* permits easy subassembly replacement for maintenance and [illegible]tional changes.

The design is based on *low power C-mos* technology and a TeleSystems' proprietary design which includes a *VLSI Chip processor* developed specifically for echo canceller functions.

Full Compatibility

The EC-5000 is compatible with alternate *Voice/Data* service on each of the 24 voice channels (when equipped with TD—tone disabler—option or when the user provides canceller disable signal for individual channels).

The EC-5000 conforms to *North American T1 standards* and is directly compatible with D type channel banks such as WECO D1D, D2, D3 or D4; M1C, M12, and M13 Multiplex; T1 Transmux Equipment; and switching equipment which meet DS-1 interface specifications (AT&T Technical Advisory No. 32).

The Unit is *transparent* to T1 and is installed directly into T1 service on the longhaul (four wire) side of a channel bank or digital switch. A/B signaling bit integrity is assured. The EC-5000 inserts 3 PCM frames (375 microseconds) delay into the transmit side and no delay into the receive side of the T1 circuit.

Dynamic Test Routines

TeleSystems' EC-5000 is equipped with dynamic *Self-Test* features which can provide diagnostic tests for each echo canceller automatically when a channel is idle or when a channel is removed from service.

The unit is also equipped with a *Manual, Automatic and Remote Control* and *Test features* that can disable any echo canceller and/or subject it to tests without interfering with normal operation of other channels. The control and test functions are generated on the basis of information derived from *Inband Signaling, Front Panel Controls,* or via a signal interface from a local switch or channel bank.

Complete Flexibility

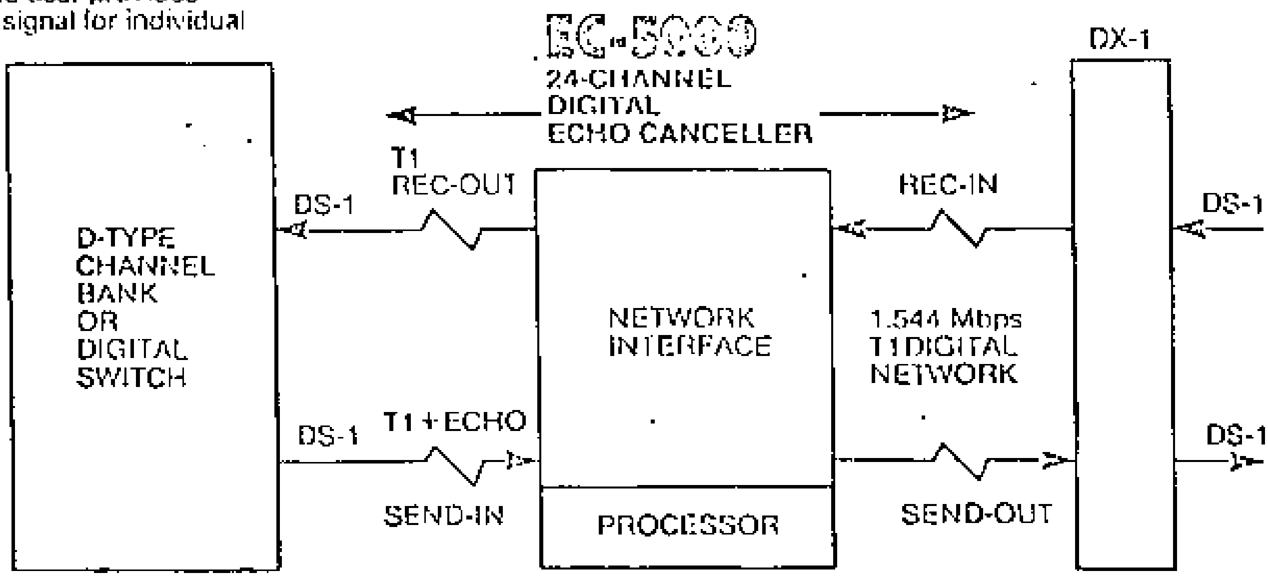
The EC-5000 is designed to operate in a *telephone plant* environment. It mounts into an EIA standard 19" rack, operates from -48 VDC Power Supply, consumes 80 watts at -48 VDC, and requires no forced air-cooling. It complies with general telephone plant engineering practices.

GENERAL DESCRIPTION

The EC-5000 consists of two major functions subsystems: The *Network Interface* and *Processor* (see diagram).

Network Interface functions: receive and send 1.544 Mbps digital bit stream from the longhaul and local circuits; provide synchronization and timing for the entire unit; multiplex and demultiplex T1; signaling bits—extraction and insertion; idle channel code detection; test functions and other controls; front panel control interface; service control interface (optional). Provides an elastic buffer allowing for relative clock drifts between transmit and receive paths and frame deletion or repetition to accommodate buffer overflow/underflow caused by cumulative effects of the drift.

Processor functions: digital echo canceller for each of 24 voice channels consisting of tail circuit modeling; echo estimate generation; echo subtraction; double talk detection; center clipper for removal of residual echo; tone disabler (TD) (optional).



SUMMARY SYSTEM SPECIFICATIONS

Message Format

T1 signal format operating at 1.544 Mbps for 24 64-kbps voice channels with 8-bit mu-law (mu-255) companded PCM encoding.

Network Interface Specifications

Input Signals (*)

Line rate	1.544 Mbps, ±200 bps
Line code	AMI, bipolar, return-to-zero
Line impedance	100 Ohms, nominal
Base-to-peak amplitude	1.5V to 3V, ±10%
Minimum pulse density	One binary 1 in 16 bits (no more than 15 consecutive zeroes)
Average pulse density	Not less than one binary 1 in 8 bits

Output Signals (*)

Line rate	1.544 Mbps, ±200 bps
Line code	AMI, bipolar, return-to-zero
Line impedance	100 Ohms, nominal
Base-to-peak amplitude	3V to 6V, ±10%
Amplitude unbalance between positive & negative pulses	±5% of base-to-peak amplitude
Half-amplitude pulse width	324 ns ±30 ns
Width unbalance between positive & negative pulses	±15 ns
Rise and fall time (10% to 90% base-to-peak amplitude)	80 ns, maximum
Minimum pulse density	One binary 1 in 16 bits (no more than 15 consecutive zeroes)
Average pulse density	Not less than one binary 1 in 8 bits

* (AT&T Technical Advisory No. 32 and No. 24)

Processor Specifications
(Specification applies to each voice channel)

Echo Return Loss Enhancement (w/ERL of 6 db)

Nominal	45 db
W/Center Clipper disabled	24 db
Rate of convergence	90% in 250 msec
Tail circuit delay capacity	32 msec
Tail circuit length (nominal)	1600 miles
Minimum Echo Return Loss (ERL) requirements	6 db
Tone Disabler (option)	A single 2050 to 2240 Hz tone (Complies with CCITT recommendation G.161.)
Power Requirements	-44 to -56 VDC
Power Consumption	80 watts at -48 VDC (w/TD) 70 watts at -48 VDC (w/o TD)

Physical Specifications

Size	19" rack mount chassis 12.2" (31.72 cm) high 18" (46.80 cm) deep

SUMMARY SYSTEM SPECIFICATIONS

Environmental Specifications	
Temperature (non-operating)	-60°C to +70°C
Temperature (operating)	-10°C to +50°C
Humidity	to 95% non-condensing
Altitude (operating)	Sea level up to 3,000 meters
Altitude (non-operating)	Sea level to 10,000 meters
Options	
	117 VAC Power Supply
	Redundant Power Supply
	TD (Tone Disabler) Option
	30 channel CEPT-32 compatible service at 2.048 Mbps

A Word About COMSAT General TeleSystems

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PRUEBAS SUBJETIVAS DE CANCELACION DE ECO REALIZADAS

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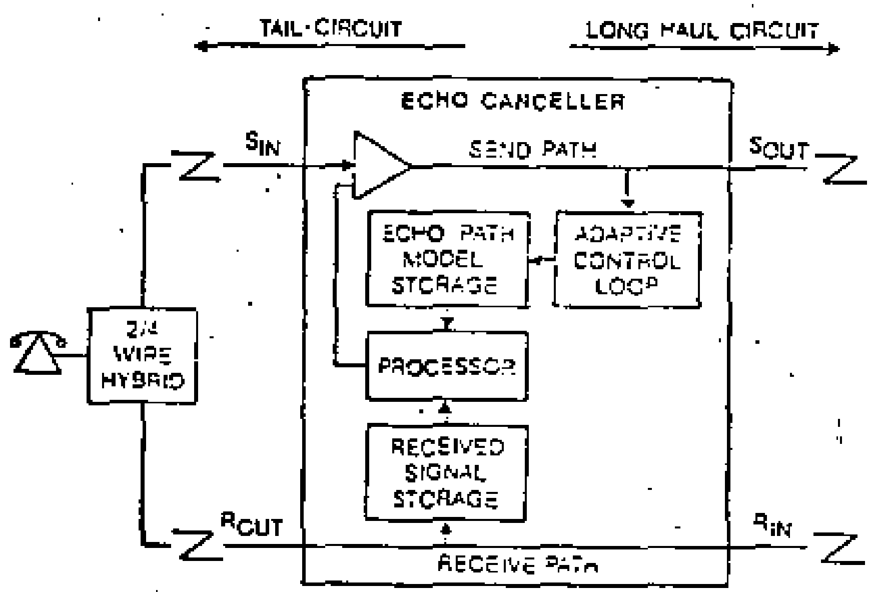
ECHO CONTROL

The following packet of information concerning echo cancellation has been assembled as an aid to those interested in the subject.

INTELSAT has sponsored various programs to study methods of improving echo control. The other articles include customer opinions, CCITT Standards and Comparative Evaluations conducted in Canada and within AT & T.

For further information please feel free to contact our Marketing Department.



(INTELSAT RESEARCH AND DEVELOPMENT PROGRAM -- TEN YEARS OF PROGRESS - 1980)

"The increased propagation time involved in satellite transmissions precipitated a reexamination of the well-known telephone echo problem. Since echo control is a crucial factor in satellite circuits, over the past 10 years INTELSAT has sponsored an intensive program to study methods of improving echo control on satellite circuits."

* * *

"Until recently, echo suppressors have been exclusively employed to control echo. However, the use of echo suppressors, even those of modern design, creates some annoying side effects, which decrease with increasing echo return loss."

* * *

"Recognizing the inherent limitations of echo suppressors, INTELSAT sponsored a program for studying the feasibility of echo cancellation, a technique which promises to eliminate echo-related problems from satellite communications...Echo cancellers were successfully tested between four pairs of INTELSAT countries on satellite circuits spanning the Atlantic and Pacific Oceans.

* * *

"Because of echo cancellation, satellite transmissions equal high-quality terrestrial communications... as illustrated by the data shown in Figure 41, based on an American Telephone & Telegraph (U.S.) report. The present effort under INTELSAT...is receiving wide interest from manufacturers.

* * *

"The echo control project has...brought about the realization of echo cancellers for satellite communications, eliminating the difference in quality between satellite and terrestrial communications which the echo suppressor is unable to achieve."

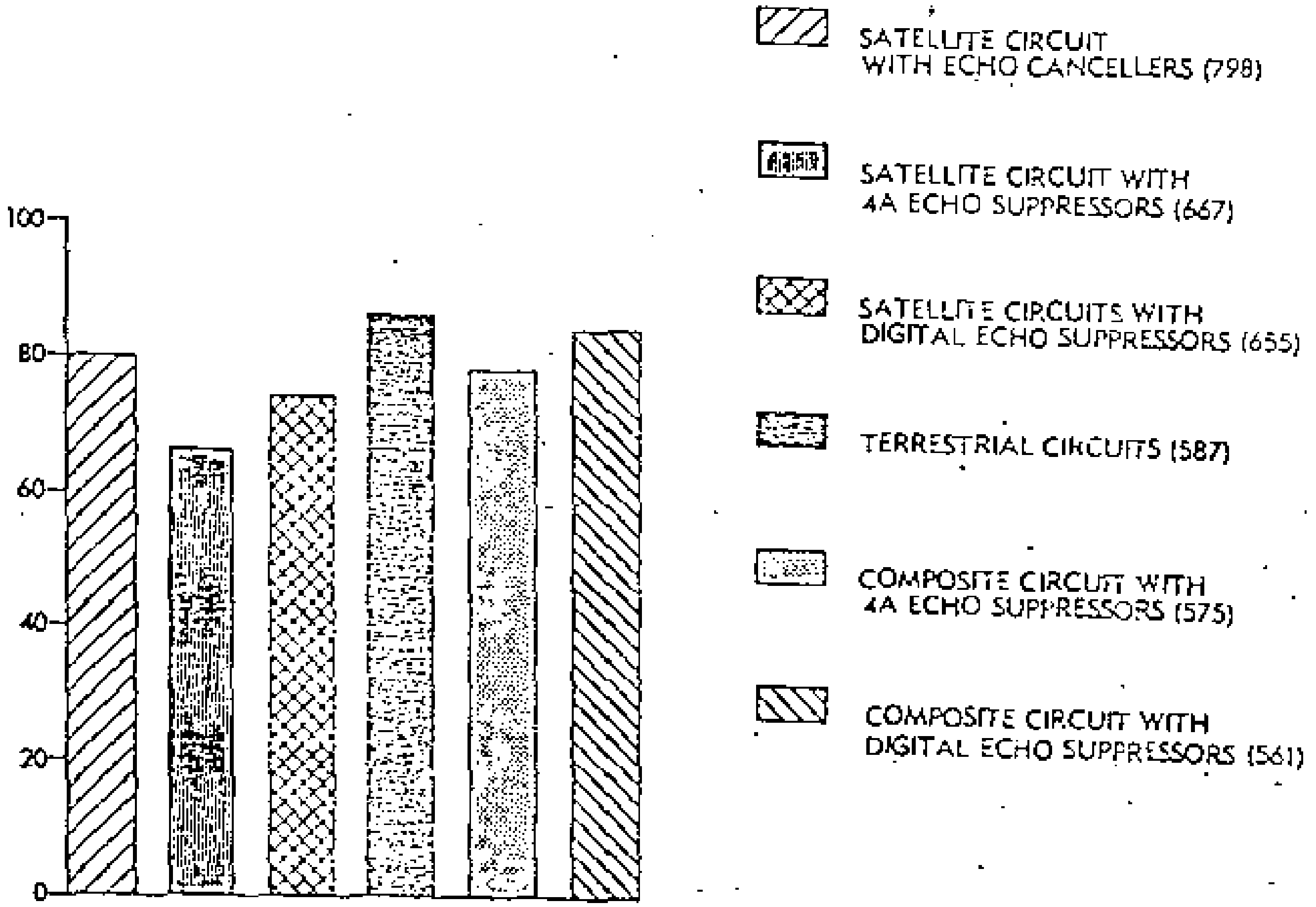


Figure 41. Results of Customer Rating Calls "Good" or "Excellent"

INTELSAT Field Test Trials

ESTUDIO COMPARATIVO DE 253

SUPRESORES Y CANCELADORES

DE ECO

(RCA)

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R. Setzer

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Echo control for RCA Americom satellite channels

Satellite telephone service has been plagued by the echo problem since its inception. The echo canceller has come close to eliminating the problem once and for all — and RCA Americom is leading in the application of this device to its private line circuits.

Abstract: *The fundamental problems caused by echo on telephone plant circuits and the circuit parameters are outlined. How these problems are solved and the efficiency of the solution is of major importance. The need for echo control on satellite channels is defined and two methods of control are discussed and compared: echo suppressors and echo cancellers. The basic principles of echo suppressors are outlined. The conditions that cause the need for echo cancellers and how echo cancellers successfully solve the echo problem are also described.*

Introduction

The introduction of satellites as a transmission facility for voice communications has added a new dimension to the problem of echo control in telephone circuits. Formerly, long line terrestrial transmission links used devices which were designed to provide an acceptable level of telephone circuit noise and echo for 99 percent of the population. However, the satellite link has rendered these devices unsatisfactory for a high percentage of the population, even when appropriate changes were made for the long delay of the satellite link. Therefore, objectionable results were obtained from early satellite telephone circuits because of the application of an outdated echo control technology. In fact, it has been determined from surveys that over 85 percent of potential satellite voice circuit users have been reluctant to use them because of the echo control problem. In the early 1970s, technological programs were introduced to solve the problem.

RCA Americom has been working jointly with industry to advance the state of the art, and currently is leading in the application of the latest device, called an echo canceller, to its private line circuits. The echo canceller will finally come very close to eliminating the echo problem once and for all.

The following article reviews the circuit parameters that cause echo and discusses the operation of the echo suppressor and echo canceller in satellite telephony.

Fundamentals

An echo can be defined as a reflection of electric or acoustic energy. In many applications, reflection of electromagnetic waves is desirable, such as in radar or troposcopic communications, but reflection of speech is generally undesirable. Reflections occur at irregularities of transmission mediums and this is also true of telephone communication systems.

To determine the historical cause of communication circuit irregularities, it is interesting to reconstruct the development of the equipment leading to the present telephone plant. Consider first the telephone instrument itself. Naturally, a microphone is required to convert the speech to electrical energy and a speaker converts the electrical energy to acoustic energy. These devices each normally require a pair of leads.

Next, considering the objective of interconnecting subscribers great distances apart, it would be desirable to carry both the transmitted and received signals on the same pair of wires. This is indeed possible by using a simple transformer incorporated into the telephone set as shown in Fig. 1. Transmitter currents flow in opposite directions through windings A and B and cancel each other's effect in winding C to the same degree that impedances Z_{BAL} and Z_o match each other. These impedances are intentionally different so that a voltage is induced in winding C (called sidetone) and makes the instrument sound "alive". Also, received signals from the central office flow through windings A and B in series and provide the receive audio. The components providing this 2-wire to 4-wire signal separating function have come to be called the terminating set, or "term set."

Next, consider the interconnection of telephone sets which are carried on a 2-wire

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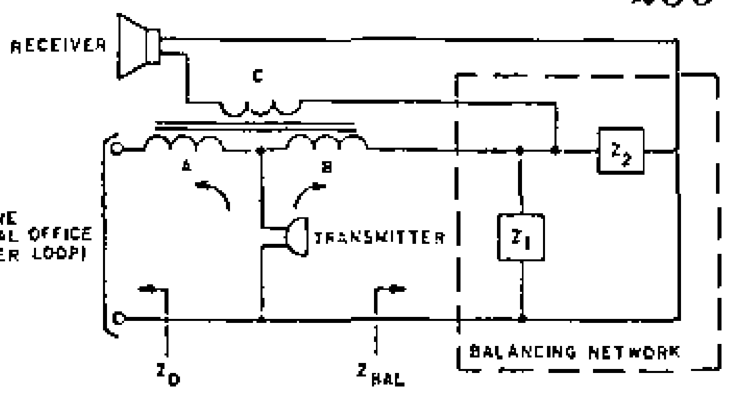

Fig. 1. The transformer in the typical telephone set permits the same pair of wires to carry both the transmitted and received signals.

basis to a switching office and then transmitted on a carrier facility (4-wire basis by nature) where many channels are multiplexed together. This is shown in Fig. 2. Note that the telephone set is shown as a 2-wire device but actually contains an integral term set. A term set is also required at each central office to separate the directions of transmission; but these term sets should be balanced as much as possible to accept the 4-wire receive (RX) from subscriber A, for example, at central office B and transmit it to the 2-wire facility toward subscriber B. This is shown as the direct speech path in Fig. 2. Leakage at the distant end from the 4-wire receive (RX) to the 4-wire transmit (TX) provides an undesirable echo path. The amount of leakage is called return loss (also echo return loss, or ERL) and can be expressed in decibels as:

$$ERL = 20\, log_{10} \left| \frac{Z_{BAL} + Z_{sub}}{Z_{BAL} - Z_{sub}} \right| \text{ dB}$$

Z_{sub} is determined primarily from the constants of the cable transmission line since the impedance of the telephone set can be controlled and the switching office is designed to provide negligible impedance transformation. However, each subscriber varies in distance from the central office and this causes a tremendous variation in Z_{sub}. Impedance compensation for each subscriber line is almost as economically prohibitive as providing 4-wire transmission for each line.

Z_{BAL} is selectable by circuit design but only a compromise value can be chosen to match the average value of Z_{sub}. The wide statistical variation in Z_{sub} is nevertheless well known and results in a distribution of ERL that has been shown to have a normal probability density function. The distribution of subscriber loop impedance can be approximated by a normal probability density function, as shown in Fig. 3, but the approximation is not generally used. In any event, an average value of Z_{SUB} has been historically chosen as the "average subscriber loop."

Thus,

$$Z_{BAL} = \text{Average} \left[Z_{SUB} \right] = R + jX_c,$$

where $R = 900$ ohms, $C = 2.16\ \mu F$.

Studies of the subjective reaction to echo over the range of expected ERL values have shown that the quality of the connection degrades as the echo is delayed in time. For short delays, the echo appears as additional sidetone and is no problem. For longer delays the same amount of echo is more objectionable and finally becomes intolerable. However, within limitations, it is possible to introduce direct loss in each direction of transmission to reduce the echo to an acceptable level. This loss should be introduced as a function of propagation delay (distance). The direct speech signal suffers the loss once but the echo path includes the loss twice.

This technique of introducing loss as a function of distance has been called the Via Net Loss plan (VNL)[1] and has been the basis of the Bell System terrestrial network design. Figure 4 shows the amount of loss required. Above 45 msec, the loss is excessive and an echo suppressor is required. Since the round trip delay of a satellite channel is approximately 600 msec, all satellite channels are equipped with echo suppressors.

Echo suppressors—how they work

The basic operating principle of an echo suppressor is to provide a direct speech path under some condtions and to interrupt the echo path under others. Note that the direct speech path used when one subscriber is talking becomes the echo path when the other subscriber is talking, as shown in Fig. 5. This is a split-type echo suppressor where one device is provided at each end. When only one person is talking, B, for example, B's TX PAD is set to 0 dB as is A's RX PAD. Therefore, no loss is encountered from B to A. Also A's TX PAD is set to 50 dB to prevent B from hearing his own echo, but this is no problem since A is not talking anyway. The circuit is symmetrical for single talker conversation in the reverse direction.

The problem is what to do when both parties talk at the same time. The design shown in Fig. 5 detects this double-talking condition and each suppressor switches in a 6 dB pad in the receive direction. (Actually the suppressor of the active speaker had switched in its 6 dB pad in anticipation of this state). These pads attenuate the direct speech of each talker by 6 dB and, as shown in the previous section, reduce the talker echo by 12 dB. The values chosen evidently represent a compromise. Also, the ERL distribution of 11σ3 dB (meaning normal

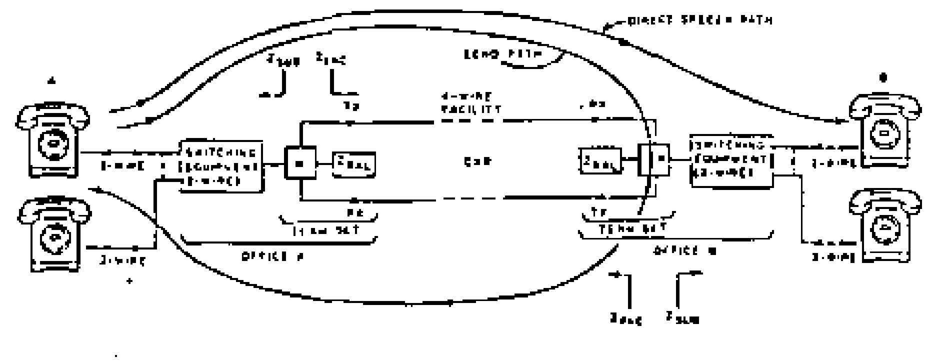

Fig. 2. Illustrating the use of "term sets" at each central office to provide 4-wire capability.

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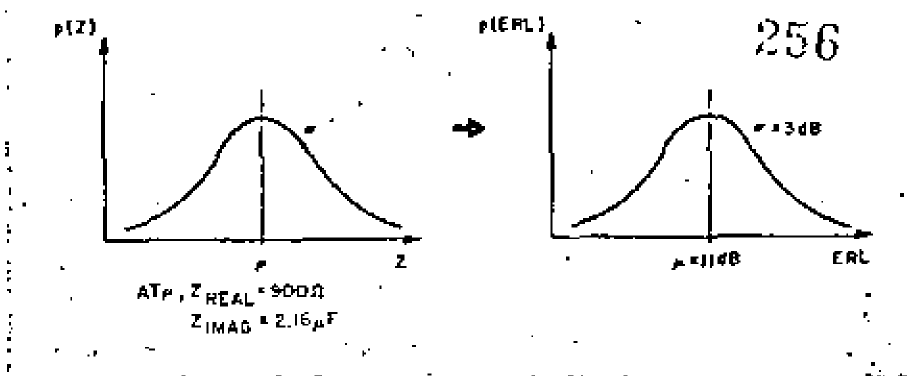

Fig. 3. Probability density functions of subscriber loop impedance and echo return loss.

distribution with mean of 11 dB and standard deviation of 3dB) is added to 12 dB to provide the overall level of echo protection. This approach does not satisfy VNL rules for echo protection but this is only true for the double-talk state.

Suppression loss hangover time

To gain further insight into echo suppressor operation, consider one of the timings involved in switching these pads in and out. The round-trip delay of the terrestrial link from the echo suppressor output through the hybrid and back to the echo suppressor input is called the echo path delay. This delay is a function of distance and, for domestic satellite circuits with both ends in the same country, the terrestrial delay typically ranges 0–30 msec. Now, consider a graceful transition from a single-talker state with A talking to a single-talker state with subscriber B talking, with a quiet interval in between. When A is talking, B's transmit pad is set to 50 dB but when A stops talking, B's pad should remain in the circuit for the echo path delay. The timing involved in holding the transmit pad (suppression pad) in the circuit for the round-trip terrestrial delay is called the suppression loss hangover time and is typically set to 60 msec (for worst case echo path delay). For B's transmit pad to suppress all of A's echo by 50 dB, the quiet interval must be at least:

- one-way terrestrial delay (15 msec) for A, plus
- one-way satellite delay (300 msec), plus
- round-trip terrestrial delay (30 msec) for B.

In other words, all echo can be suppressed by 50 dB if the conversation is not more interactive than 345 msec. However, this is not likely and if B interrupts before this critical interval has expired, the suppression pad must be switched in to carry B's signal, and A's echo is returned (the suppressor loss hangover protection is lost).

The performance becomes even worse when one considers the detection threshold of the double-talker detectors of Fig. 5. Recall that the ERL distribution of the terrestrial link is 11σ3 dB. Using the 2σ low as the minimum ERL to be encountered, 5 dB can be expected. Now, the echo suppressor must be designed so that a constant echo of 5 dB does not create a double-talker or break-in condition, otherwise the unit would always be in double-talk and the performance would be very poor. Therefore, voice levels must be closer than 5 dB to cause break-in, and a comparative difference of +1.5 dB has been selected.

However, the following variations can be listed in the voice levels applied to the echo suppressor:

1. zero to seven decibel variation in subscriber loop losses,
2. zero to three decibel variation in carrier and cable gain stability,[2]
3. talker volume distribution with $\mu = -15$ dBm, $\sigma = 5.8$ dB.[3]

Therefore, variations on the order of 10 dB in average voice levels can be expected, but the echo suppressor cannot break-in until voice levels are within 1.5 dB. As a result, speech bursts can be lost entirely. More commonly, speech chopping results. This problem can be called the "equal-level break-in problem."

These echo suppressor impairments and others have been studied for 40 years[4] and, while substantial improvements have been made, the basic impairments still remain today for channels equipped with echo suppressors. A new technique is desirable and is now available with the echo canceller.

The echo canceller

To take a fresh look at the problem, let us list some of the known conditions.

1. An echo signal will be returned, with expected loss of 11σ3 dB (depends on impedance of 2-wire section).
2. The echo will be delayed in time, with expected delay of 0-30 msec (depends on length of 4-wire terrestrial section).
3. The signal which will become the echo is available in time before the actual echo is generated.

Therefore, if the signal which will become the echo is stored and the terrestrial level and delay are somehow determined, an estimate could be made of the expected echo. This estimate can then be subtracted from the actual circuit echo when it occurs. Furthermore, the estimate can be updated to further reduce the actual echo. A block diagram of such a circuit is shown in Fig. 6 and is called an echo canceller. This concept was basically developed by Comsat Labs.

The echo canceller, under a single talk condition, samples and stores the voice signal present at the satellite receive (SAT. RX) input and uses this as the first estimate of the echo (perhaps reduced by 11 dB). When the actual echo is received at the terrestrial transmit (TERR.TX), the es-

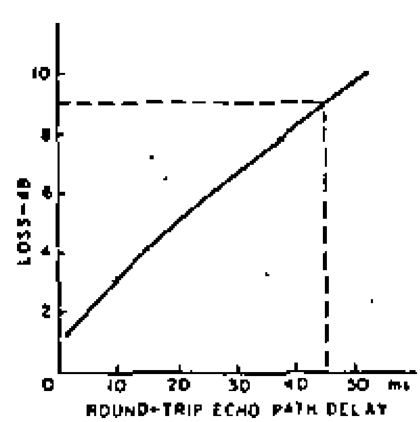

Fig. 4. The amount of loss required to reduce echo level increases with echo path delay.

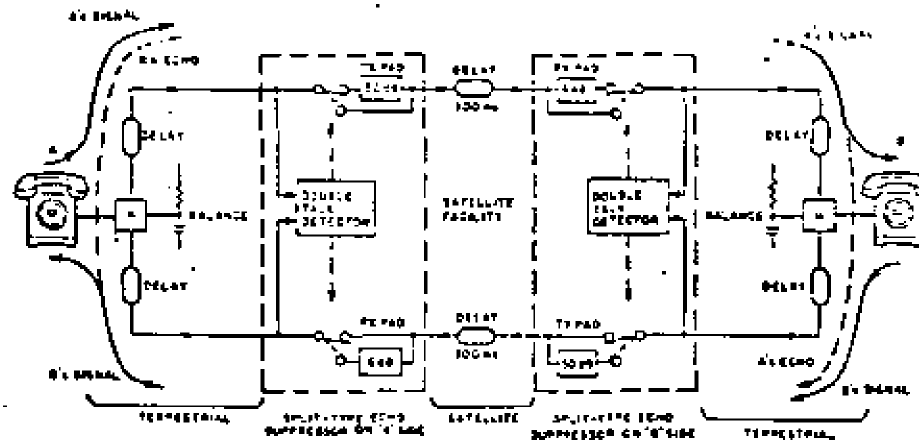

Fig. 5. A split echo suppressor connects the transmit speech path when one subscriber is talking but disconnects it when the other subscriber is talking (since this path is now the undesirable echo path).[7]

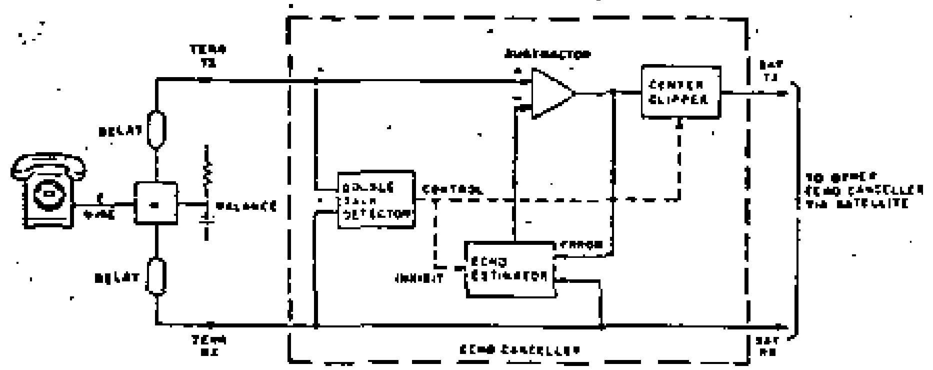

Fig. 6. The echo canceller estimates the expected echo level, then subtracts the estimate from the actual circuit echo when it occurs.

timate is subtracted from it and an error signal is generated. The error signal updates the echo estimate to reduce the error to a minimum. This process is called convergence, and generates a model of the echo path.[8]

Estimating the echo

The accuracy of the echo estimate, and thus the degree of cancellation is determined from several things. First, the actual error signal is contaminated with terrestrial noise. The echo canceller converges on the echo signal and also attempts to converge on the noise which continually changes the echo estimate to a small degree. Also, the structure of sampling, storing and correlating the actual and estimated echo signals limits the amount of resolution. Finally, the step size used in adjusting the echo estimate affects the residual echo left after cancellation, and results in an interesting tradeoft.[5] If the step size is large, the speed of convergence will be fast but the resolution will be limited. If the step size is small, convergence will be slower but the degree of cancellation will be better.

Considering these variables, the raw subtraction process is limited to about 30 dB of echo cancellation, with convergence in approximately 250 msec. It is desirable to achieve even further echo cancellation, for the single talker state, on the order of 45-50 dB. This is accomplished by removing all signals below a certain threshold, in the residual echo path (after subtraction). Signals above the threshold are not appreciably affected. This operation is called center clipping, and is also shown in Fig. 6. The center clipper rejects small signals (i.e., residual echo not cancelled by the subtractor plus terrestrial noise) but passes large signals (i.e., direct speech). The center clipper of an echo canceller is comparable to the suppression pad of an echo suppressor. In fact, an echo suppressor can be considered to be a center clipper with a very high threshold[5]; but it rejects all signals, wanted or unwanted.

The beauty of the echo canceller technique is apparent during the double talk condition. Assuming that a period of single talking has occurred, the echo canceller will have converged on the line characteristics. Approximately 45 dB of echo cancellation is achieved after 250 msec with the center clipper "in" the circuit. Now, when double talk begins, the echo canceller must not be allowed to operate on the local signal and echo at the same time since this will adversely affect the echo estimate. The local direct speech signal would give a tremendous error in the echo estimate since it is not related to the distant signal. Therefore, the echo estimator must be inhibited from updating on the error signal as soon as double talk is detected. However, the echo estimator still maintains an accurate model of the echo path and continues to subtract the echo. The only possible problem occurs if the line characteristics change during the period of double talking but this is not very likely. Also, the center clipper is switched "out" of the circuit since it serves no useful function at this time and would only distort the near-end speech. Therefore, continuous double talking is possible with echo cancellation of 30 dB at each end. No speech bursts are lost since the process is continuous.

The processes involved in echo cancellation lend themselves to microprocessor control. First, the speech signal destined to become the unwanted echo signal must be sampled and stored for the maximum roundtrip terrestrial delay. For sampling at the Nyquist rate of $1/2f_{max}$ (= 125 μsec) and storing for 30 msec, 240 samples are required. Using a 12-bit speech sample and a 9-bit representation of the line impulse response, a minimum mean-square error algorithm used to converge the canceller would require 250 multiplications of 12 by 9 bits and 250 additions of these products during every sampling period. Another Comsat design[7] uses logarithmic encoding to reduce hardware and calculation complexity and saves 28 percent of the memory space required for the previous Comsat design.[4]

Comparison of performance and conclusion

The quality of a satellite channel equipped with an echo suppressor is a function of many variables. In some cases, the quality is excellent and no practical improvement is possible with an echo canceller.[3]

For example, the performance of the

Table I. Performance ratings of circuits using echo suppressors and prototype echo cancellers, all operating under the same conditions (for domestic circuits).

Type of circuit	*Percentage of calls rated fair and poor*
Terrestrial with echo suppressor	10
Composite with echo suppressor	21
Satellite with echo suppressor	34
Satellite with echo cancellor	12

channel equipped with an echo suppressor would no doubt be rated excellent if the following conditions exist:

1. the circuit ERL is 17 dB or greater (2σ high of 11σ3 dB distribution);
2. the terrestrial links exhibit little variation; and
3. the users adjust their talker volumes to overcome the "equal level break-in" problem and subconsciously limit the interactiveness of the conversation (become "polite").

However, as these variables deteriorate, so does the performance. The echo canceller, on the other hand, exhibits a more uniform range of excellent performance over the same range of variables. Some results of a comparison between echo suppressors and prototype echo cancellers operating under the same conditions (for domestic circuits) are given in Table I and include composite circuits (terrestrial channel in one direction and satellite channel in the other direction).[7] Composite circuits have slightly more than one-half the roundtrip delay of a full satellite channel and thus would require less reduction of echo signals than a full satellite channel, for the same level of performance.

Thus the echo canceller makes the satellite channel perform as well as a terrestrial channel. Prototype Comsat Laboratories echo cancellers were used in this test.

RCA Laboratories also conducted comparative tests of echo cancellers and echo suppressors. Their results confirm the excellent performance of echo cancellers under a wide variety of circuit conditions. Then, RCA Americom field-tested several units on demonstration and customer channels and the performance was well received by the customers.

As a result, RCA Americom is intending to apply echo cancellers on a larger scale. This improved channel quality will assist RCA Americom in maintaining a strong market share of the private leased channel business.

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EVALUACION SUBJETIVA DE METODOS DE CONTROL DE ECO

(A T T)

SATELLITE USER REACTION TESTS
A SUBJECTIVE EVALUATION OF ECHO CONTROL METHODS

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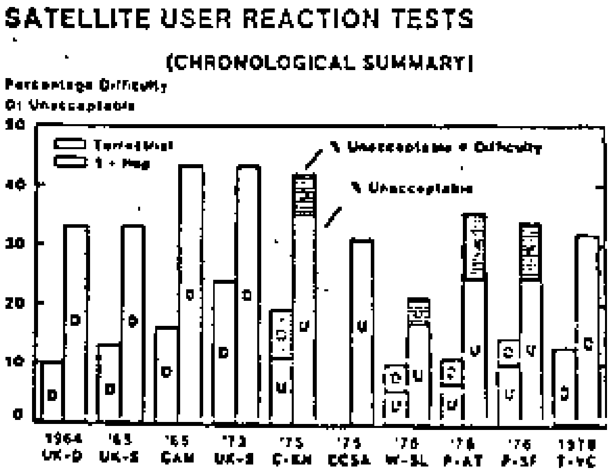
ABSTRACT - This paper provides a composite summary of six Domestic Satellite User Reaction Tests conducted from May 1975 through August 1977. Results from these tests and from earlier experiments consistently show that users experience Difficulty or Unacceptable telephone service two to three times more often on satellite circuits equipped with echo suppressors than they do on conventional terrestrial circuits. Echo cancelers provide sufficient improvement to restore the transmission quality of One-Hop satellite circuits to near-terrestrial performance. The author also demonstrates strong correspondence between users' opinions of transmission quality and their overt actions such as terminating calls early, redialing calls and redialing operators for assistance.

1. INTRODUCTION - Communication satellites have been used for international calls since 1965. More recently, long-distance circuits using satellites as the transmission medium were first used in the continental United States public network in the summer of 1976 when AT&T and GT&E jointly placed the COMSTAR satellite system in service. In 1975, AT&T and Bell Laboratories initiated a continuing series of field tests, using call-back customer interviews, to evaluate user reaction to telephone calls over terrestrial and over domestic satellite (DOMSAT) circuits. Results from these domestic studies and earlier experiments consistently show that users experience a significantly higher level of satisfaction with service via terrestrial circuits as compared with service via satellite circuits equipped with conventional analog echo suppressors.

Figure 1 shows a chronological history of results for ten user reaction tests conducted since 1964. These tests include a range of international and domestic conditions on simulated and on actual satellite facilities. Results are expressed in the percent of calls which are rated unacceptable or in which conversational difficulty is reported. Each test compares the performance of terrestrial circuits with the performance of simulated or actual satellite circuits equipped with analog echo suppressors. The CCSA (Common Control Switching Arrangement) test results are shown for the One-Hop satellite condition only; no terrestrial reference condition was evaluated.

The domestic tests since 1975 also include the percent of calls rated Unacceptable, a more critical category than the rating of Difficulty. In general, customers experienced Difficulty or Unacceptable service on satellite circuits two to three times more often than they did on terrestrial circuits. It is this persistent result which has prompted the Bell System to continue to pursue improved echo control techniques.

Figure 1



CHRONOLOGICAL SUMMARY - KEY

UK-D	US - United Kingdom	Simulated Delay
UK-S	US - United Kingdom	Satellite
CAN	Canadian Domestic	Simulated Delay
UK-S	US - United Kingdom	Satellite
C-KN	Cedar Knolls - Chicago	Simulated Delay
CCSA	CCSA, Pittsburgh - Los Angeles	Satellite
W-SL	Wayne - Salt Lake City	Simulated Delay
P-AT	Pittsburgh - Atlanta	Satellite
P-SF	Pittsburgh - San Francisco	Satellite
T-VC	Toronto - Vancouver	Satellite

2. DOMESTIC SATELLITE TESTS - The six domestic satellite user reaction tests reported in this paper were conducted within the continental United States,[1] including:

(a) A Pilot study using simulated satellite circuits in the public switched network between Cedar Knolls, New Jersey, and Chicago, Illinois, May-June, 1975.

(b) A CCSA network study of Multi-Hop satellite circuits between Pittsburgh, Pennsylvania and Los Angeles, California with access to the public network in Los Angeles, June-October, 1975.[2]

(c) The DOMSAT, Phase 0 tests on simulated, One-Hop satellite circuits and their terrestrial counterparts between Wayne, Pennsylvania, and Salt Lake City, Utah, March-June, 1976.

(d) The DOMSAT, Phase 1A tests on COMSTAR satellite circuits in the public network between Pittsburgh and Atlanta, Georgia, August-October, 1976.[3] These tests include Terrestrial (VNL Design), Half-Hop* with conventional 4A analog suppressors and new digital echo suppressors and One-Hop circuits equipped with 4A analog suppressors, digital echo suppressors and experimental echo cancelers provided by COMSAT Laboratories and Bell Laboratories. Interviews were conducted at the Atlanta end only.

(e) The DOMSAT, Phase 1B tests on COMSTAR satellite circuits in the public network between Pittsburgh and San Francisco, California, October, 1976 - February, 1977.[3] These tests compare user reaction on different conditions including; terrestrial, half-hop and one-hop circuits equipped with various combinations of echo control devices. Interviews were conducted at the San Francisco end only.

(f) The DOMSAT, Phase 1C tests on COMSTAR satellite circuits, also between Pittsburgh and San Francisco, May-August, 1977. Interviews were conducted at the Pittsburgh end.

3. THE CALL-BACK INTERVIEW - The call-back telephone interview, conducted primarily with the called parties,** contained the following questions which provide a basis for subjective transmission quality measures:

(a) Was the quality of the connection acceptable to you?

If no, question (b) was skipped.

(b) Did you have difficulty talking or hearing over that connection?

If the respondent answered "No" to question (a) or "Yes" to question (b), the following question was asked:

(c) Did this cause you or the other person to end your conversation?

If "Yes", questions (d) and (e) were asked.

(d) Did you or the other person call back to complete your conversation?

(e) Did you or the other person contact the operator to inform her of the difficulty or to request assistance?

(f) Which of these four words comes closest to describing the quality of that connection: Excellent, Good, Fair or Poor?

4. COMPOSITE RESULTS - CONVENTIONAL ECHO CONTROL The results obtained in the six domestic tests are given in Table A and have been combined to form a composite of user reaction as shown in Figure 2 for Terrestrial, Half-Hop, One-Hop and Two-Hop satellite configurations. The composite results are for circuit conditions using VNL design and conventional analog echo suppressors. Customers were engaged in conversations on calls they received in the normal conduct of their daily transactions. They were selected for interview, after completing their calls, by virtue of random traffic selection of the intertoll test circuits.

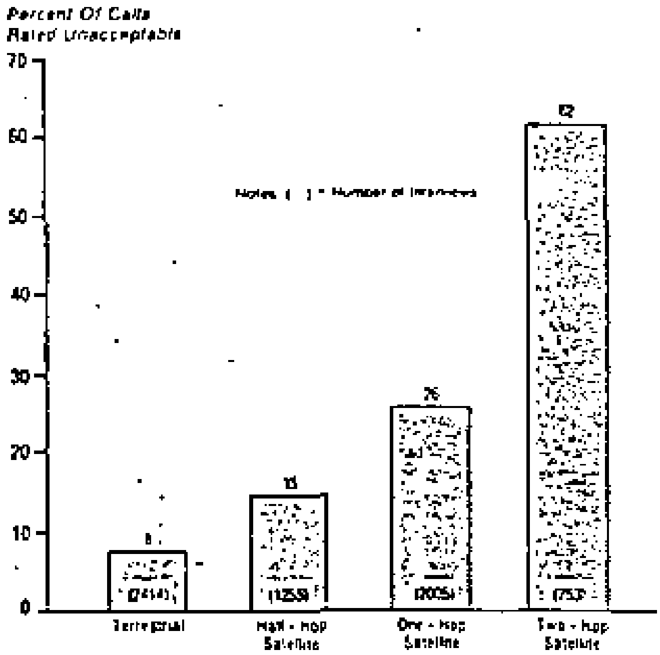
Results are expressed in terms of the percent of connections which were rated Unacceptable (%U). Over 6400 customer interviews are included. Of 2414 interviews conducted for the Terrestrial test circuits, 8.3% of the connections are rated Unacceptable. Of 1255 interviews for Half-Hop circuits, 14.9% are rated Unacceptable. For 2005 interviews for One-Hop satellite circuits, 26.0% are rated Unacceptable and of the 753 interviews for Two-Hop satellite circuits, 62.0% or nearly two-thirds of the connections are rated Unacceptable. In general, customers rated the Half-Hop circuits Unacceptable twice as often compared with the terrestrial reference; One-Hop connections are rated Unacceptable three times and Two-Hop connections nearly eight times more often. Figure 3 shows a breakdown of the individual test results which are used to create Figure 2. In addition, Table A gives the Number of interviews (N), percent Unacceptable or having Difficulty (%U+D), percent Excellent (%E), percent Good (%G), percent Fair (%F), percent Poor (%P), the Mean Opinion Score (MOS), percent calls Terminated Early (%TE), percent calls Redialed (%RE) and percent calls dialed back to Operators for Assistance (%OA).

* Half-Hop circuits are built one way on satellite and one way on terrestrial facilities.

** Nearly all interviews were conducted with the called parties. A small number of interviews were conducted with the calling parties during the Pilot and during the CCSA tests.

Figure 2

Domestic Satellite User Reaction Tests Composite Results (Using Conventional Echo Suppressors)



5. NEW ECHO CONTROL TECHNOLOGY - During the Phase 1 COMSTAR tests, two new echo control devices were evaluated, the #4 ESS digital echo suppressor[4] and experimental echo cancelers.[5], [6] Table B gives the individual test results as well as the composite calculations. Figure 4 reproduces the composite results of Figure 2 along with the composite results obtained from Table B for the test circuits equipped with the new echo control devices. There are three important findings shown in Figure 4:

(a) The digital echo suppressor provides service on Half-Hop satellite circuits comparable to terrestrial service (8.7% vs. 8.3% Unacceptable). The digital echo suppressor improves performance on One-Hop satellite circuits when compared with the conventional echo suppressor results (17.3% vs. 26.0% Unacceptable).

(b) The experimental echo cancelers provide service on One-Hop satellite circuits nearly equivalent to the terrestrial results (9.5% vs. 8.3% Unacceptable).

(c) The echo cancelers provide marked improvement in the quality of Two-Hop satellite circuits (21.3% vs. 62.0% Unacceptable). The absolute round-trip delay (nearly 1.1 seconds) appears to be limiting the quality of Two-Hop satellite circuits which can be achieved with echo cancelers.

Figure 3

Domestic Satellite User Reaction Tests Individual Results

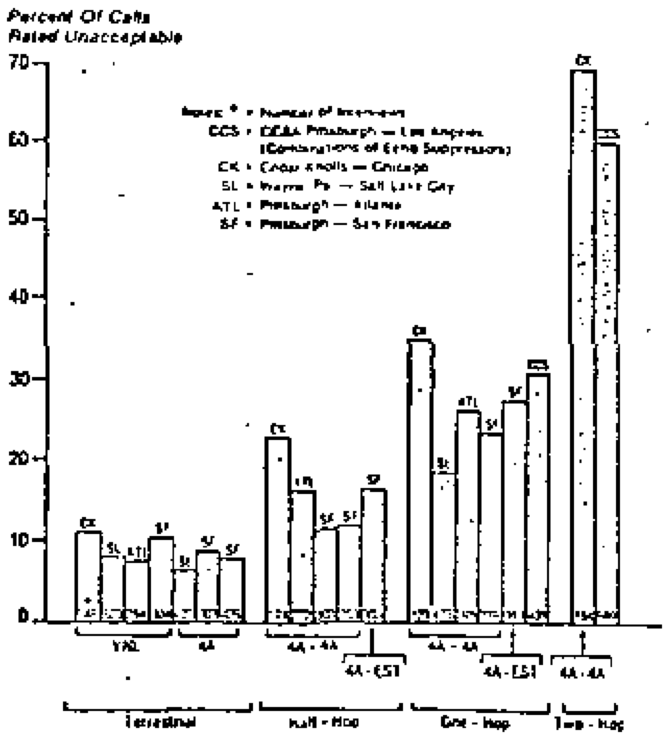
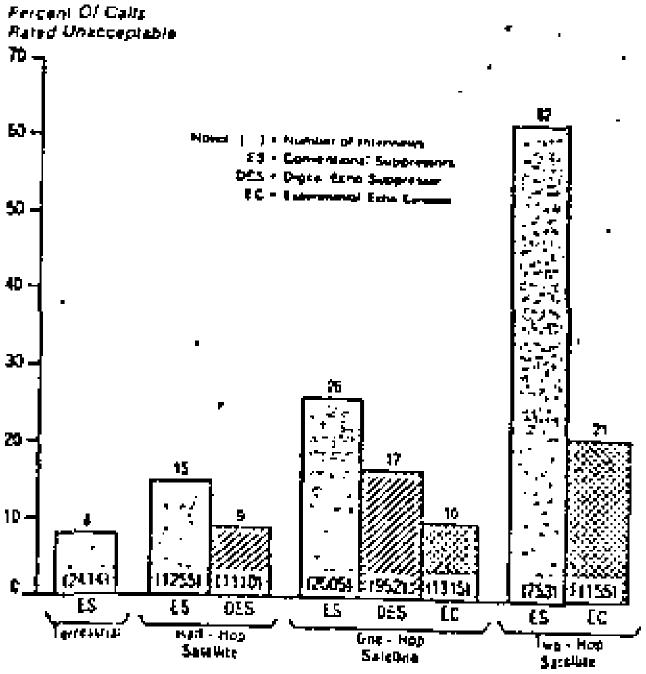


Figure 4

Domestic Satellite User Reaction Test Results Compares Echo Control Methods

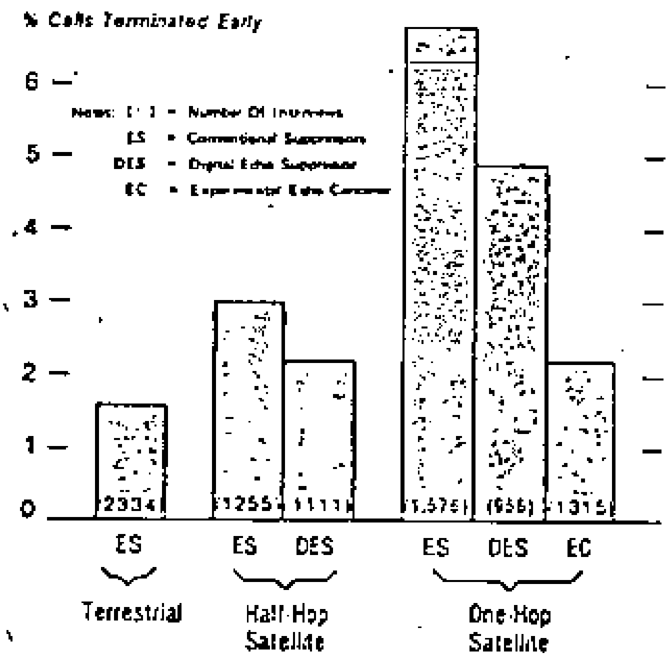


6. CALLS TERMINATED EARLY - Figure 5 displays the composite results for the percent of calls which customers declared were terminated early (Tables A and B). The results for conventional echo suppressors are: 1.6% for the Terrestrial benchmark, 3.1% for Half-Hop Satellite and 6.4% for One-Hop Satellite operation. These results track with the 2:1 and 3:1 ratios of the Unacceptable ratings given in Figure 2. This is a more critical result which implies that customers take some action when they receive poorer quality transmission.

The digital echo suppressor improves performance to 2.1% for Half-Hop and 4.9% for One-Hop satellite operation. The Echo Canceler, with 2.0% of the calls being terminated early, ranks superior to the conventional echo suppressor devices. Due to the small number of occurrences for the terminated early category, the Digital Echo Suppressor on Half-Hop and the Echo Canceler on One-Hop satellite circuits are considered comparable to the Terrestrial performance, consistent with the results in Figure 4 for the Unacceptable ratings.

Figure 5

Domestic Satellite User Reaction Test Results Compares Echo Control Methods

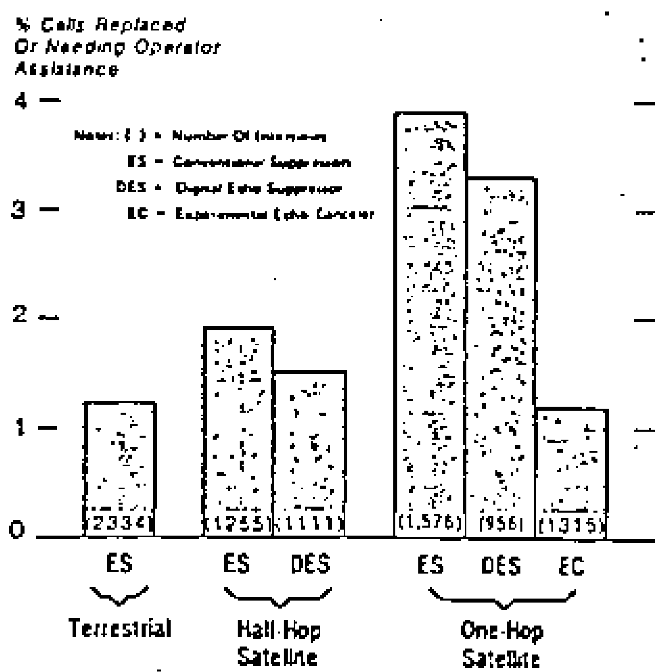


7. CALLS REPLACED OR NEEDING OPERATOR ASSISTANCE During the call-back interviews, customers were asked if the call was replaced and, in a separate question, whether or not they needed operator assistance. The results from these two questions have been combined for two reasons: the combined category represents overt action by customers when they encounter difficulty. The number of occurrences for the separate categories are small and the combined data base gives a better view of the trend. However, it is clear from Figure 6 that the relative performance for Terrestrial, Half-Hop and One-Hop satellite circuits is preserved, consistent with the previous results for Unacceptability and Calls Terminated Early. Half-Hop satellite operation with Digital Echo Suppressors and Full-Hop satellite operation with Echo Cancelers remain comparable to Terrestrial performance.

There are two other important considerations displayed in Table A. First, in the Pilot study, data for the categories of Calls Replaced and Calls Requiring Operator Assistance were gathered from the calling customer. This is a very small data base of 40, 31, 30 and 39 interviews for the TERRESTRIAL, HALF-HOP, ONE-HOP and TWO-HOP conditions, respectively. However, the results for these overt action categories are noticeably higher than in the other experiments. Results in the other tests are predominantly expressed by the called customer, usually not the paying customer. The called customer may not know whether or not the calling customer dialed the operator for assistance. In addition, for the Two-Hop satellite condition in the Pilot test, the User Action is extremely high. For example, of 39 calling party interviews, 10 reported needing operator assistance. Thus, the Pilot results indicate that general user reaction in these more critical categories may be higher than shown, due to the reaction of the paying customer.

Figure 6

Domestic Satellite User Reaction Test Results Compares Echo Control Methods



Secondly, in Tables A and B, the USER ACTION appears highly correlated with the Mean Opinion Score, derived by assigning weights of 4, 3, 2, and 1 to the E, G, F and P ratings, respectively. That is, the USER ACTION steadily increases as the USER OPINION of quality decreases.

8. CONCLUSIONS – The large data base of nearly 10,000 user interviews reported in this contribution demonstrate that:

(a) Telephone users experience Difficulty or Unacceptable service two to three times more often on satellite circuits equipped with Analog echo suppressors than they do on conventional terrestrial circuits.

(b) The difficulties experienced by telephone users are primarily due to the effects of controlling echoes in the presence of the long propagation delay inherent in geo-synchronous satellite systems.

(c) It is possible to reduce the impairment effects of delayed echoes to near-terrestrial performance levels using Echo Cancelers on One-Hop satellite circuits. Similar findings have been suggested in previous contributions to CCITT[(7), (8)].

(d) It is possible to improve the quality of satellite circuit performance to comparable terrestrial circuit performance by engineering Half-Hop satellite circuits, using Digital Echo Suppressors. However, this restricts the application of satellites and may not always be practical.

(e) Two-Hop satellite circuit performance can be improved with the use of echo cancelers, perhaps to a quality level similar to that of One-Hop satellite circuits using analog echo suppressors. Performance comparable to that of terrestrial quality is not achieved.

(f) Telephone users steadily increase their overt actions (Terminate Calls Early, Redial Calls or Redial Operators for Assistance) as they encounter poorer quality connections.

ACKNOWLEDGEMENTS – The author is indebted to many associates at AT&T, Long Lines, Bell Laboratories and the Operating Telephone Companies who contributed to the success of the user reaction tests. Special thanks are given to H. F. McCaffrey for his excellent coordination of the field experiments and particularly to Mr. P. C. Lopiparo of Bell Laboratories for his valuable technical contribution throughout the entire series of experiments. Mr. D. L. Duttweiler is primarily responsible for reducing the echo canceler to practical application.

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(5) D. L. Duttweiler, "A Twelve-Channel Digital Echo Canceler," IEEE Transactions on Communication, Vol. COM-26, No. 5, May, 1978.

(6) See CCITT Study Group XV – Contribution No. 184, (see Annex IV for a discussion of a draft Recommendation for echo cancelers), May, 1976.

(7) CCITT Study Group XVI – Contribution No. 65, Plenary Period 1973-1976.

(8) CCITT Study Group XII – Contribution No. 154 (COM XVI – No. 105), April, 1979.

TABLE A - DOMESTIC SATELLITE USER REACTION TESTS - ANALOG ECHO SUPPRESSORS

I. TERRESTRIAL BENCHMARKS	USER OPINION N	%U	%U+D	%E	%G	%F	%P	MOS	USER ACTION N	%TE	%RE	%OA
Pilot	148	10.8	17.6	60.0	25.0	8.0	7.0	3.38	148	1.90	0.90	2.50
Phase 0-1	523	8.0	11.1	48.2	39.6	6.5	5.6	3.30	441	2.04	1.36	0
Phase 0-2	423	6.4	6.9	50.2	39.1	7.9	2.8	3.37	425	1.18	0.94	0.47
Phase 1A	284	7.8	11.3	47.7	41.7	7.1	3.5	3.34	284	1.06	0.70	0.35
Phase 1B-1	303	9.2	15.7	44.7	36.6	11.2	7.4	3.19	303	2.31	0.99	0.66
Phase 1B-2	304	10.2	13.8	44.6	38.6	11.0	5.8	3.22	304	1.32	0.33	0.33
Phase 1C	429	7.9	13.5	53.2	34.7	8.4	3.7	3.37	429	1.40	0.70	0.47
Composite	2414	8.3	12.1	49.2	37.5	8.4	4.9	3.31	2334	1.58	0.87	0.40

II. HALF-HOP ANALOG ES	USER OPINION N	%U	%U+D	%E	%G	%F	%P	MOS	USER ACTION N	%TE	%RE	%OA
Pilot	124	22.6	28.2	35.0	38.0	14.0	13.0	2.95	124	5.40	5.40	0
Phase 1A	272	16.2	21.0	41.7	36.4	13.4	8.5	3.11	272	4.41	1.47	0.74
Phase 1B-1	303	11.6	18.8	33.8	45.7	14.7	5.8	3.08	303	2.31	0.66	0.33
Phase 1B-2	303	16.5	24.1	28.4	46.8	15.6	9.1	2.95	303	1.32	0.66	0.33
Phase 1C	253	11.9	19.4	39.9	40.9	12.1	7.1	3.14	253	3.56	1.19	0.40
Composite	1255	14.9	21.6	35.6	42.2	14.0	8.2	3.05	1255	3.08	1.41	0.43

III. ONE-HOP ANALOG ES	USER OPINION N	%U	%U+D	%E	%G	%F	%P	MOS	USER ACTION N	%TE	%RE	%OA
Pilot	123	35.0	43.1	32.0	28.0	13.0	27.0	2.62	123	14.00	8.60	10.00
CCSA	431	31.0	-	16.0	47.0	22.0	15.0	2.63	-	-	-	-
Phase 0	473	18.4	21.4	33.4	40.6	13.6	12.4	2.95	473	4.90	1.69	1.27
Phase 1A	348	25.9	32.2	32.0	33.9	16.6	17.6	2.80	348	4.60	1.15	1.44
Phase 1B-1	319	23.5	31.4	26.0	40.6	19.3	14.1	2.79	319	6.90	1.57	0.31
Phase 1B-2	311	27.6	33.8	23.4	39.2	21.8	15.6	2.70	313	7.35	1.92	0.96
Composite	2005	26.0	33.6	26.4	39.3	17.8	15.5	2.77	1576	6.44	2.13	1.22

IV. TWO-HOP ANALOG ES	USER OPINION N	%U	%U+D	%E	%G	%F	%P	MOS	USER ACTION N	%TE	%RE	%OA
Pilot	154	69.5	74.7	14.0	16.0	14.0	56.0	1.88	154	43.50	29.6	25.60
CCSA	599	60.0	-	8.0	27.0	25.0	40.0	2.02	-	-	-	-
Composite	753	62.0	74.7	6.8	24.7	21.9	44.5	1.99	-	-	-	-

TABLE B - DOMESTIC SATELLITE USER REACTION TESTS - NEW ECHO CONTROL METHODS

I. HALF-HOP DIGITAL ES	USER OPINION N	%U	%U+D	%E	%G	%F	%P	MOS	USER ACTION N	%TE	%RE	%OA
Phase 1A	283	10.2	13.4	49.1	37.5	8.6	4.8	3.31	283	1.77	1.06	0.35
Phase 1B-1	299	5.4	10.4	41.4	45.1	10.3	3.2	3.25	299	1.67	0.67	0.33
Phase 1B-2	278	9.4	17.6	42.0	37.9	14.0	6.2	3.16	279	3.23	1.43	0.72
Phase 1C	250	10.0	14.0	51.2	35.8	9.8	3.2	3.35	250	1.60	1.60	0.40
Composite	1110	8.7	13.8	45.7	39.3	10.7	4.4	3.26	1111	2.07	1.17	0.45

II. ONE-HOP DIGITAL ES	USER OPINION N	%U	%U+D	%E	%G	%F	%P	MOS	USER ACTION N	%TE	%RE	%OA
Phase 1A	356	16.0	19.9	39.6	39.4	11.9	9.2	3.09	356	5.62	2.25	0.84
Phase 1B-1	297	19.2	25.9	31.6	41.2	16.0	11.3	2.93	301	4.32	2.66	1.66
Phase 1B-2	299	17.1	24.1	30.8	43.3	16.0	9.9	2.95	299	4.68	1.34	1.00
Composite	952	17.3	23.1	34.3	41.2	14.5	10.1	3.00	956	4.92	2.09	1.15

III. ONE-HOP CANCELER	USER OPINION N	%U	%U+D	%E	%G	%F	%P	MOS	USER ACTION N	%TE	%RE	%OA
Phase 1A	334	8.7	12.6	51.8	34.6	7.7	5.9	3.32	334	1.50	.90	0.60
Phase 1B	464	11.4	17.2	36.5	42.1	15.6	5.8	3.09	464	2.80	.65	0.65
Phase 1C	517	8.3	13.2	49.4	35.5	11.1	4.0	3.30	517	1.74	.58	0.39
Composite	1315	9.5	14.5	45.5	37.6	11.8	5.1	3.23	1315	2.05	.69	0.55

IV. TWO-HOP CANCELER	USER OPINION N	%U	%U+D	%E	%G	%F	%P	MOS	USER ACTION N	%TE	%RE	%OA
Phase 1C	155	21.3	29.7	36.7	35.7	13.3	14.3	2.95	155	8.39	3.87	0.65

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ECHO CANCELLER STANDARDIZATION IN THE CCITT*

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ABSTRACT

Recent technology advances have permitted many innovations in the echo control area. One innovation is echo cancellation. This paper presents the approach taken by the CCITT to produce an echo canceller Recommendation. Highlights of the draft Recommendation are included, specific problem areas are discussed and, for unresolved issues, the points currently being considered are presented.

INTRODUCTION

Advances in solid-state technology have made available a wide range of complex integrated circuit building blocks. These devices have permitted the development of rather sophisticated signal processors which are finding applications in the telecommunications industry. The echo canceller is a recent development which has provided dramatic improvement in echo control in telephone communications [1]-[3]. This paper presents the approach taken in the CCITT to develop a Recommendation for echo cancellers.

The process of standardization within the CCITT consists of the following steps. Near the end of a study period questions are proposed for consideration in the next study period and, if sufficient interest exists, they are incorporated into the scope of work for that period. During the next 4 years member administrations (e.g., British Post Office), operating agencies (e.g., AT&T), and scientific or industrial organizations (e.g., Siemens) submit contributions giving their views in reply to various questions. These contributions are discussed at meetings of Study Groups or smaller working parties, and a consensus of the participants is reflected in a draft Recommendation. If approved at the Plenary meeting of the CCITT, this is published as a numbered Recommendation in the next issue of the CCITT colored books. If all issues cannot be resolved, the Recommendation will contain some provisional items which will remain open for discussion and possible modification in the next study period. Such a process has been used for standardization of echo canceller requirements.

The question proposing the study of echo cancellers was first adopted for study in the 1973-1976 period. However, echo cancellers were still in the research stage, and the question was carried over into the 1977-1980 study period. To date, 14 Contributions have been received and discussed and a draft Recommendation has been prepared.

STUDIES SINCE 1976

Within the CCITT, Study Group XV (Transmission Systems) is responsible for echo control devices. The Echo Suppressor Working Party of Study Group XV is specifically charged with studying questions regarding echo suppressors and echo cancellers. In 1977, the chairman of the Echo Suppressor Working Party (coauthor George Helder) set up a series of meetings with AT&T and COMSAT, two organizations active in echo canceller studies, to discuss the form for a Recommendation on echo cancellers. At that time there existed an echo suppressor Recommendation (G.161) [4] which specified in great detail the operating characteristics of echo suppressors. This form could also be used for echo cancellers. However, it appeared that a more suitable echo canceller Recommendation would address the general functions which a canceller should perform and specify tests which the echo canceller should pass to properly perform these functions. The chairman subsequently prepared such a draft Recommendation as a contribution to Study Group XV (COM XV-No. 96). The contribution was discussed at two working party meetings. Some items requiring additional clarification were identified such as the requirements for transmission performance and compatibility between echo suppressors and echo cancellers used on two ends of the same circuit.

Before the meeting in July 1979, contributions were received on these and other items from Canada (Bell Northern Research) [5], [6], Japan (Kokusai Denshan Denwa) [7] Italy (Italian Administration) [8], and the United States (AT&T and COMSAT) [9]-[11]. As a result of these contributions and the July discussions, extensive revisions were made to the draft Recommendation.

*This paper is based upon the experience of the authors in their roles as delegates to CCITT Study Group XV. Views expressed are not necessarily those of Bell Laboratories or COMSAT Laboratories.

CURRENT FORM OF THE DRAFT RECOMMENDATION

The present draft retains the functional approach to echo canceller specifications. Essentially, on a new telephone connection, an echo canceller should respond to the incoming speech from the far end and its resulting echo, should adapt to the associated echo path in a short time, and should then cancel most of the echo. It should perform this function for essentially any linear, time invariant* (or nearly time invariant) echo path within its memory capacity. In addition, when double talk occurs, the echo canceller must avoid excessive degradation of its echo cancellation performance. Double talk must therefore be detected in the echo canceller and steps taken to prevent further adaption, which would result in a change from the previous converged state.

DRAFT RECOMMENDATION HIGHLIGHTS

The draft Recommendation, currently designated G.YYY, [12], begins by describing echo cancellers, defining compatibility between echo cancellers and echo suppressors, defining terms associated with echo cancellers, and specifying transmission performance requirements. Tests are then prescribed to measure the functional performance of echo cancellers. The test requirements are all provisional and will require further consideration in the next study period.

The draft Recommendation considers two implementation philosophies. The first implementation (type I) provides cancellation so that the residual echo is at a constant low level regardless of the level of the incoming signal or its echo. In such an echo canceller, an additional nonlinear processing device may be necessary to further reduce the residual echo. The echo canceller must not permit any intelligible echo to be returned to the far talker during the single talk operating mode. A center clipper with a clipping level equal to the peaks of the residual echo has been used for that purpose.

A second implementation (type II) provides cancellation which is constant in dB, regardless of the receive input speech level over a specified dynamic range. In such a canceller, the amount of cancellation may be sufficient to produce acceptably low-level echoes for all incoming speech levels and echo return losses in the telephone network. In this case, additional nonlinear processing may be unnecessary.

Both types of echo cancellers have been used in field trials and both have provided substantial subjective quality improvement over echo-suppressor-equipped circuits. The two implementations are based upon different practical economic/performance tradeoff considerations.

In the type I echo canceller, the center clipper is removed during double talk, thus permitting a low-level echo to pass. Subjectively, in the confusion of double talk, this echo may be perceptible but is not objectionable.

In the type II echo canceller, the highest residual echo level (equivalent to the type I residual echo level) occurs for a combination of poor (i.e., low) echo return loss and high receive speech level. This concept relies upon the low joint probability of occurrence of high receive speech level and poor echo return loss given their independent distributions as observed in the network where the echo cancellers are applied.

TEST CONSIDERATIONS

The tests specified in the draft Recommendation are "black box" tests. Figure 1 shows a typical configuration used for a steady-state test. The tests require minimal knowledge of the internal design of the echo canceller and access to the four I/O ports only. In addition, the need for special test equipment was minimized, and certain external controls, including echo canceller adapt/no adapt, memory clear, and center clipper enable/disable, were assumed.

Test 1 in the draft Recommendation addresses the amount of echo cancellation and specifies that the returned echo level (after cancellation) for a fully converged echo canceller must be below certain bounds. This requirement applies to a range of test signal levels, return losses and end delays. For type I echo cancellers the returned echo level must be ≤ 40 dBm0 (when the center clipper is not enabled); for type II the mask in Figure 2 applies.

The second requirement is that the echo canceller must converge rapidly enough to produce a short and unimportant returned echo at the start of a conversation. Thus, for a range of input signal levels, return losses, and end delays, the echo canceller must achieve sufficient cancellation in 500 ms that the combined loss of the echo path, cancellation, and additional nonlinear loss is ≥27 dB. Again, the test signal is noise.

The third requirement is that the echo canceller remain in a converged state during periods of double talk when near-end speech is mixed with the echo. In the test, a simulated double talk noise signal is mixed with the echo at appropriate levels. The requirement specifies that the echo level immediately after the double talking ceases will be no greater than 10 dB above that for the fully converged state.

The fourth requirement is that the echo canceller must converge to an infinite return loss echo path. Echo may be generated as follows. An echo canceller connected to an idle circuit contains the echo path model from the previous connection. If, when a new circuit connection is established, a tandem echo suppressor is encountered in the echo path, the suppression switch will cause the echo canceller to see an infinite echo return loss. If the echo canceller does not converge, it

*The present draft Recommendation assumes that the echo path is linear and time invariant and does not address echo cancellers designed to cancel echos for nonlinear and/or time variant echo paths.

will now process receive speech through the old echo path model, producing an estimated echo which, when subtracted from the new circuit echo, will [illegible] echo canceller generated echo. For this test the canceller initially converges for a defined echo path, the echo path is interrupted, and a noise test signal is applied to the receive input of the echo canceller. The signal produced by the canceller at the send output is monitored and after a defined time it must be below -40 dBm0.

UNRESOLVED ISSUES

Transmission Requirements

An important aspect in the drafting of recommendations is their economic impact. A recommendation should also be flexible enough to permit future innovations, restrictive enough to guarantee satisfactory performance, and compatible with other recommendations so that it may be treated as a subunit of a larger system with its own overriding constraints.

One subject which has caused considerable discussion in formulating the present draft Recommendation is uniformity of transmission performance requirements for analog and/or digital equipment which may be used on analog and/or digital transmission facilities. This problem is not unique to echo cancellers; it is encountered wherever the two technologies are intermixed in the through transmission path. The four most probable echo canceller transmission facility combinations, identified as types A-D, are shown in Figure 3.

The original philosophy concerning the transmission performance characteristics was quite simple. For the type A and type B echo cancellers, the transmission performance characteristics for echo suppressors given in CCITT Recommendation G.161, *Orange Book*, should apply. For type D, the codec must conform to CCITT Recommendations G.711 [13] and G.712 [14], and in both types C and D, the echo canceller must be transparent, providing bit integrity to single talk signals in either transmission path.

Closer inspection of these relevant Recommendations reveals some problem areas. First, the transmission performance characteristics of Recommendation G.161 are quite restrictive in terms of frequency response and delay distortion. This is relatively insignificant in the design of an echo suppressor, since there is no through path processing, and equipment manufacturers are able to meet the requirements of the Recommendation. For a type A or type B echo canceller, the send path may include decorrelation, a sample and hold, and a reconstruction filter. This extra processing makes it very difficult to economically meet the Recommendation G.161 frequency response characteristic. The problem of meeting the delay distortion requirement is even more severe, perhaps requiring the use of third-order compensating networks.

The benefits of these two transmission performance characteristics may not justify the difficulties involved in meeting them. Further, the negative impact on the network of not meeting these performance characteristics may be small compared to the advantage of using echo cancellers [9],[10].

When transmission requirements are established, the intended application of the device should also be considered. An analog echo suppressor is typically one of many equipment units comprising an analog communications channel. Hence, each piece of equipment should meet stringent transmission requirements to maintain their cumulative effects within acceptable limits. The codec requirements on the other hand, need not be so rigid because it is typically used only once in a digital communications channel. Once the signal is in digital form, it is no longer subjected to the same cumulative effects as analog signals.

Based on these considerations and others, a compromise approach is used which allows for some inconsistent performance characteristics in intermixed analog and digital communications equipment applications. This compromise approach includes the following elements:

a. When an inconsistency is discovered the more stringent requirement is not automatically relaxed to the less restrictive requirement, which could result in unnecessary degradation of the existing network.

b. Similarly, the less restrictive requirement is not automatically tightened to the more stringent requirement, which could impose severe economic consequences.

c. Every attempt is made to maintain the highest possible standards consistent with present operational practice and manufacturing capabilities within each technology.

d. When important differences exist, notes and appropriate references to other Recommendations are used to guide the reader in understanding the potential impact of the characteristic under consideration.

These guiding elements were used to derive the provisional values given for some performance characteristics in draft Recommendation G.YYY. One example is the provisional delay distortion specification, which has been relaxed from the Recommendation G.161 specification but is still considerably more restrictive than Recommendation G.712.

Test Signals and Detectors

A second unresolved issue is the definition of the test signal used to stimulate the echo canceller and the instrument used to measure the resulting response to the stimulus. A widely accepted practice in telecommunications is the use of sine wave test signals. They are simple to use, give repeatable results, are well understood, and can be generated and measured with universally available equipment. However, a sine wave is not very suitable for testing the performance characteristics of an adaptive echo canceller. That is, the echo canceller is designed to operate on a speech signal and certain properties not present in a sine wave are required for proper behavior of its adaptation algorithm.

Two different test signals have been proposed. The first is a band-limited white noise test signal; the second is noise shaped in accordance with CCITT Recommendation G.227 [15] to simulate speech loading.

Those advocating white noise are concerned that a more complex signal will require additional specialized test equipment for measuring echo canceller performance characteristics. They believe that echo canceller performance measured with band-limited white noise can be specified so that satisfactory performance for speech will be guaranteed. This approach was used in the draft Recommendation.

The advocates of spectrum shaping given by Recommendation G.227 maintain that it gives a more accurate representation of the signals which will be encountered by the echo canceller. Further, they feel that manufacturers of a device which includes spectrum whitening circuitry will be unduly penalized with respect to manufacturers which do not. This is because the measured performance using white noise will be the same, but for shaped noise, the device having spectrum whitening circuitry will yield a higher figure of merit (such as dB of cancellation).

Another item of concern is the weighting factor applied to the instrument used to measure the residual echo. A contribution describing the relative subjective effect of sub-bands of echo within the voice band [16] shows that center band echo is more noticeable than band edge echo. This factor must be properly considered if the purpose is to establish a high correlation between the objective and subjective performance ratings.

Echo Canceller Performance Boundaries

A third area of continuing deliberations, and perhaps the most difficult one to resolve, is the limits on the echo canceller performance boundaries. Typical areas include convergence time, performance under conditions of high-level interfering signals, terrestrial extension range, dynamic response to step changes in the echo path characteristics, and control parameters for nonlinear devices such as center clippers. In all probability, limits for these items will be derived through a combination of laboratory testing coupled with operating ex- experien as different types of echo cancellers are developed and field tested.

CONCLUSIONS

The Echo Suppressor Working Party has produced a draft Recommendation for echo cancellers, but it is still subject to approval at the Plenary meeting to be held in November 1980. This Recommendation contains many provisional areas which will be the subject of future deliberations within the CCITT. Included are the definition of test signals and possibly the measuring instrument, identification of upper and lower limits on the rate of convergence, and dynamic performance characteristics at the start of double talk and in the presence of high-level interfering signals.

It should be noted that the present draft Recommendation is based upon the results obtained primarily from two echo cancellers which have been successfully tested in field trials. The performance characteristics have been defined using a band-limited white noise test signal. The test specifications, even though provisional, do represent a reasonable set of characteristics. If the spirit and intent of the draft Recommendation is followed, it should, in the opinion of the authors, provide an acceptable echo canceller.

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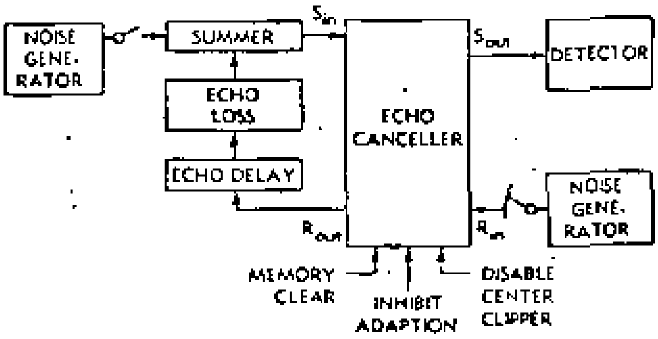

Figure 1. Steady-State Echo Canceller Test Configuration

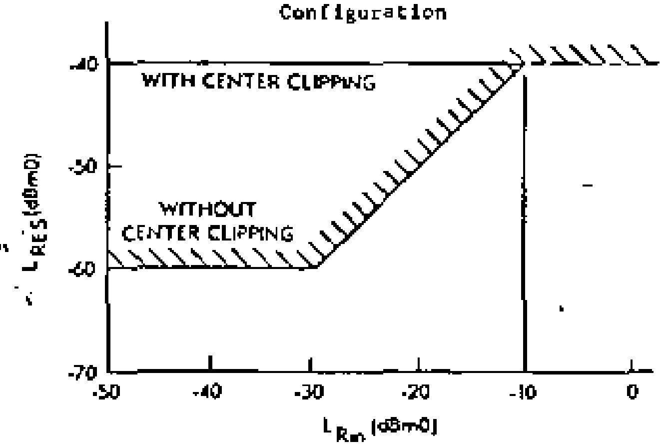

Figure 2. Echo Canceller Steady-State Performance Mask

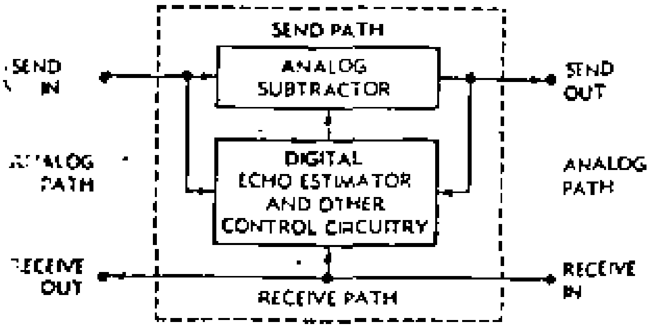

Figure 3a. Type A Echo Canceller

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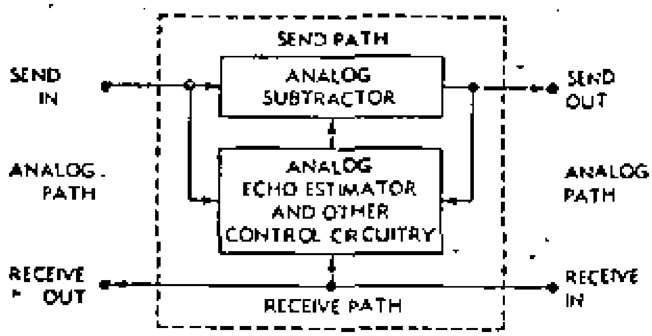


Figure 3b. Type B Echo Canceller

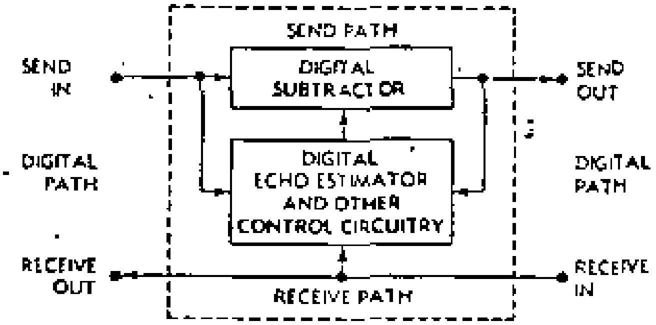


Figure 3c. Type C Echo Canceller

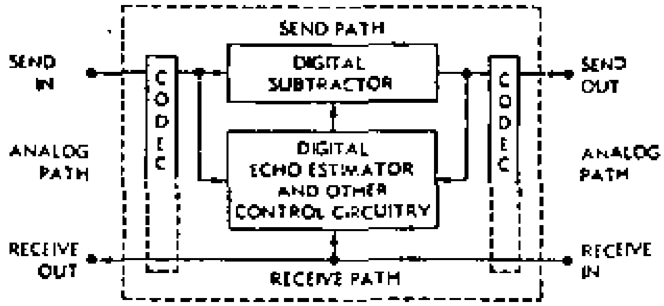


Figure 3d. Type D Echo Canceller

2.9 TECNICAS DE CODIFICACION PARA CANALES VIA SATELITE

En un sistema de transmisión de datos, es deseable incorporar técnicas de codificación dentro de las funciones de los modems, para reducir la razón E_b/N_o requerida para lograr un BER determinado. Esto, sin embargo, reduce la tasa efectiva de la información que puede transmitirse y, por lo tanto, los ahorros en E_b/N_o deben balancearse conjuntamente con la reducción de la tasa de información en un canal limitado en banda ó potencia.

Esencialmente, existen dos variantes de codificación para control de errores:

- Codificación que permite detectar que se produjeron errores en el trayecto de transmisión.
- Codificación que permite detectar y corregir los errores producidos.

En el primer caso, el receptor no es capaz de corregir los errores y tiene que enviar un mensaje al transmisor para que retransmita los paquetes ó bloques detectados con error. Esta técnica se conoce como ARQ (Automatic Repeat Request). Por lo que respecta a enlaces vía satélite, la técnica es poco atractiva debido al retraso que representa el viaje redondo de un cuarto de segundo. Además, se requiere de un sistema de memoria ó buffering para almacenar los bloques ó paquetes temporalmente en caso de que sea necesario retransmitirlos.

En el segundo caso, el receptor utiliza los bits de redundancia para corregir los errores de la transmisión y reconstruir el mensaje original. A la técnica se le conoce como corrección de error por delante ó FEC (Forward Error Correction). Elimina los retrasos en retransmisión y requerimientos de sistemas de memoria involucrados en la técnica ARQ, y el aumento de complejidad en su implementación no es muy grande. Existen dos tipos de códigos para detección y corrección de errores:

1. Códigos de bloques.
2. Códigos convolucionales.

En la codificación por bloques, el codificador añade bits de paridad a la secuencia de información, entregando a la salida bloques más grandes. La notación empleada para describir estos códigos es (n, k, d_{min}), donde n es el número de bits del bloque codificado, k es el número de bits de información, y d_{min} es la distancia mínima de Hamming entre palabras de código.

Algunos ejemplos de códigos por bloques prácticos son los siguientes:

a) Códigos de repetición
$(n,k,d_{min}) = (n,1,N)$, $N \geq 2$

b) Códigos simplex
$(n,k,d_{min}) = (2^k-1,k,2^{k-1})$

c) Códigos de chequeo por paridad sencilla
$(n,k,d_{min}) = (k+1,k,1)$, $k \geq 1$

d) Códigos de Hamming
$(n,k,d_{min}) = (2^m-1,2^m-1-m,3)$

e) Códigos de Hamming extendidos
$(n,k,d_{min}) = (2^m,2^m-1-m,4)$

f) Códigos BCH
$(n,k,d_{min}) = (2^m-1,2^m-1-e \cdot m,\geq 2e+1)$

La principal característica de la codificación por bloques es que los bits de paridad son agrupados y colocados en una sección definida de la palabra codificada.

En los códigos convolucionales, en contraparte, los bits de paridad son contínuamente intercalados dentro de la palabra codificada. A estos códigos también se les conoce como secuenciales ó recurrentes.

Los sistemas que operan con códigos por bloques requieren elementos de almacenamiento ó memoria para el proceso de codificación y decodificación. En los códigos convolucionales, el proceso de codificación y decodificación es contínuo y no se requieren elementos de almacenamiento ó memoria.

Así como en la codificación por bloques hay diferentes variantes, igual sucede en la codificación convolucional. Algunos ejemplos de las técnicas más comunes de decodificación de esta última son los siguientes:

a) Decodificación por umbral.
b) Método secuencial ó probabilístico de Wozencraft.
c) Algoritmo de Viterbi (decodificación por máxima similitud).

El algoritmo de decodificación de Viterbi por máxima similitud (maximum likelihood decoding) es el que se emplea comúnmente en modems para comunicaciones vía satélite. Entre los principales beneficios que ofrece están:

- La relación E_b/N_o para una tasa de error ó BER de 10^{-5}, por ejemplo, se reduce aproximadamente de 4 a 6 dB con respecto a la requerida para BPSK ó QPSK sin codificación.

- La estructura del codificador es relativamente simple, y permite operar con tasas de información hasta de 100 Mbps.

- Los requerimientos de coherencia en la fase de la portadora, aún cuando son más exigentes que para BPSK sin codificación, pueden satisfacerse.

Dado que al codificar por cualquier método a una corriente de información, se gana una reducción en el E_b/N_o requerido a cambio de un mayor ancho de banda ó reducción de la tasa efectiva de la información transmitida, lo mismo sucede con la decodificación Viterbi. En los sistemas de comunicación vía satélite generalmente se tienen limitaciones de potencia disponible y no de ancho de banda, por lo que es deseable reducir el E_b/N_o requerido y emplear una técnica eficiente de codificación/decodificación.

Al dispositivo que efectúa el proceso de codificación y decodificación se le denomina CODEC.

Los parámetros más importantes de un codec son la tasa del código y la ganancia del mismo. La tasa del código es una medida de la expansión necesaria en ancho de banda. Por ejemplo, si se tiene ancho de banda extra disponible, ésta puede balancearse con un esquema determinado de codificación y la potencia disponible, para lograr una cierta tasa de bits en error ó BER. Este balance sigue un comportamiento asintótico, en el sentido de que se tiene cada vez menos ganancia por codificación si se usa dos ó tres veces el ancho de banda original. A manera de ejemplo, considérese un sistema con código convolucional de longitud de restricción igual a 7 y con decodificación Viterbi; la ganancia que se obtiene es de 5.1 dB y 5.6 dB sobre PSK coherente, ideal, sin codificación, usando tasas de código de 1/2 y 1/3, respectivamente, y operando con un BER de 10^{-5}.

En el ejemplo anterior, si la longitud de restricción es 9, se necesita utilizar un 33% más de ancho de banda, y para el mismo BER solamente se obtiene una ganancia de 4.2 dB.

En un enlace vía satélite, la relación total portadora/ruido, C/N, es función de tres relaciones: $(C/N)_{subida}$, $(C/N)_{bajada}$, y $(C/N)_{intermodulación}$. Esta relación total, a su vez, es función de la energía por bit sobre densidad de ruido, E_b/N_o, de acuerdo a la expresión:

$$C/N = (E_b/N_o)\frac{R}{W}$$

en donde R es la velocidad de transmisión, por ejemplo 64 000 bps, y W es el ancho de banda del canal, en Hertz.

No todos los esquemas de modulación son adecuados para la transmisión de datos vía satélite. Los módems más recomendables son BPSK, QPSK, 8PSK y 16PSK. Conforme aumenta el número de fases de la modulación, también aumenta la eficiencia en bits/Hertz, y por consiguiente se reduce el ancho de banda requerido.

Conjuntamente con el tipo de modulación, al haber limitaciones de potencia disponible, habrá que elegir algún esquema de codificación, preferentemente entre codecs con tasa de código de 7/8, 4/5, 3/4, 2/3, ó 1/2.

En la Tabla siguiente se muestra, para un BER de 10^{-6}, la ganancia en E_b/N_o y la expansión en ancho de banda que resultan al usar diferentes tasas de código. Los valores indicados para 7/8 y 4/5 se obtienen con decodificación de umbral, y los de 3/4, 2/3 y 1/2 con decodificación Viterbi.

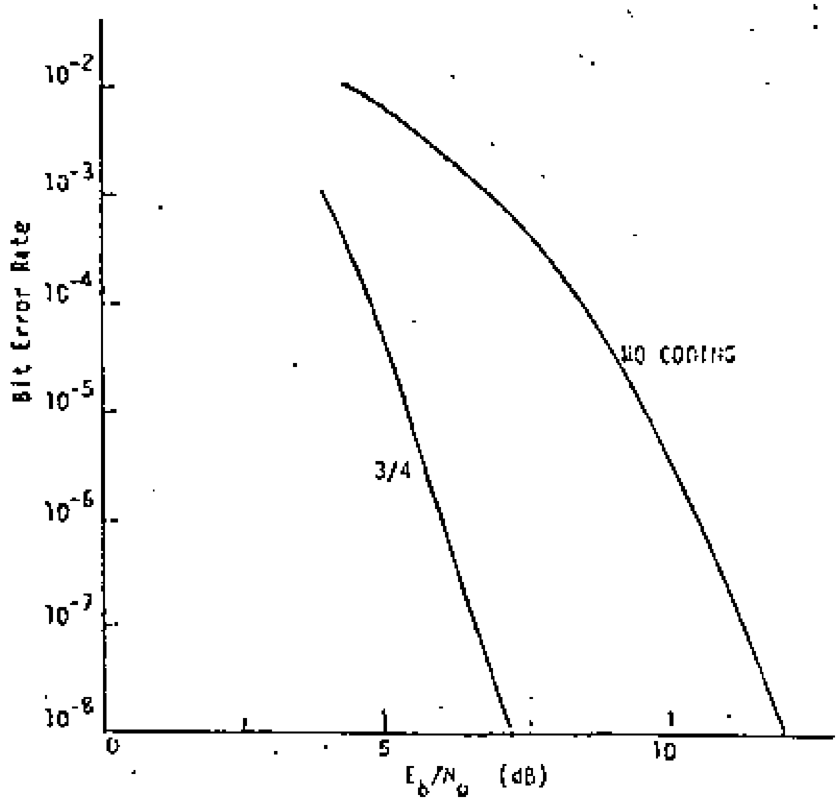
...6

TASA DE CODIGO	1	7/8	4/5	3/4	2/3	1/2
GANANCIA (dB)	0	2.55	3.80	4.39	4.77	5.40
EXPANSION EN ANCHO DE BANDA	1	1.14	1.25	1.33	1.5	2

Se ha demostrado que en sistemas donde se tiene poca potencia disponible, las mejores combinaciones de modulación/codificación son BPSK ó QPSK con tasa de código de 3/4, 2/3, ó 1/2.

En aquellos casos en los que se tengan limitaciones de ancho de banda, y no de potencia, es preferible utilizar por ejemplo 8PSK ó 16PSK con tasas de código de 7/8 ó 4/5.

En la gráfica siguiente se muestra el E_b/N_o requerido sin codificación y con codificación de 3/4 para lograr una tasa de error ó BER determinada.



E_b/N_o VS. P_e FOR CODED AND UNCODED SYSTEM

Ejemplos de sistemas disponibles en el mercado.

a) DIGITAL COMMUNICATIONS CORPORATION.

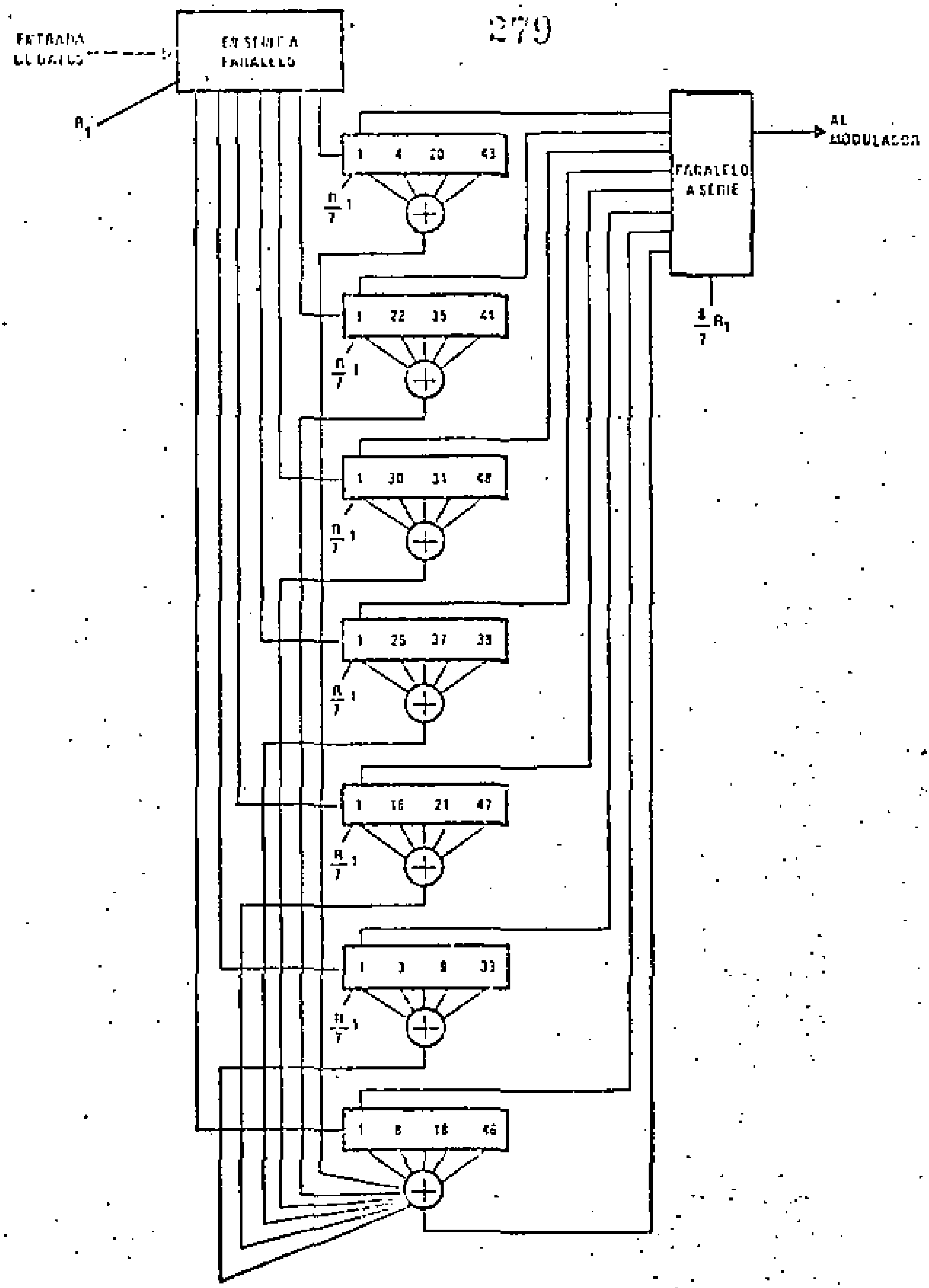
Esta compañía ofrece una serie de estaciones terrenas denominadas DYNAC, para la transmisión de paquetes con datos y voz digitalizada (TDM) por medio de acceso múltiple por división en el tiempo (TDMA). Las estaciones emplean multiplexores estadísticos y las tasas de transmisión pueden seleccionarse desde 268 Kbps hasta 2 Mbps.

La modulación empleada es QPSK y la codificación es del tipo FEC, con tasa de código de 1/2. Ofrecen codecs opcionales con tasas de código de 7/8 y 3/4.

b) HARRIS CORPORATION

Esta compañía ofrece, entre otros componentes, su modem modelo 7005 (LRDM - Low rate data modem) que convierte las señales digitales entrantes a una portadora modulada en BPSK. La tasa del código convolucional empleado es de 7/8 con decodificación de umbral, y la ganancia del codec es de aproximadamente 2.5 dB.

El diagrama de bloques conceptual del codificador de relación 7/8 se muestra en la figura siguiente. El codificador acepta una corriente de datos de entrada en serie y realiza una conversión de serie a paralelo en siete corrientes de datos. Las corrientes de datos se desplazan a través de siete registradores de datos, según el algoritmo de codificación utilizado. Las salidas de las 7 etapas se suman a base de módulo 2 para



Codificador Ortogonal Automático de 7/8

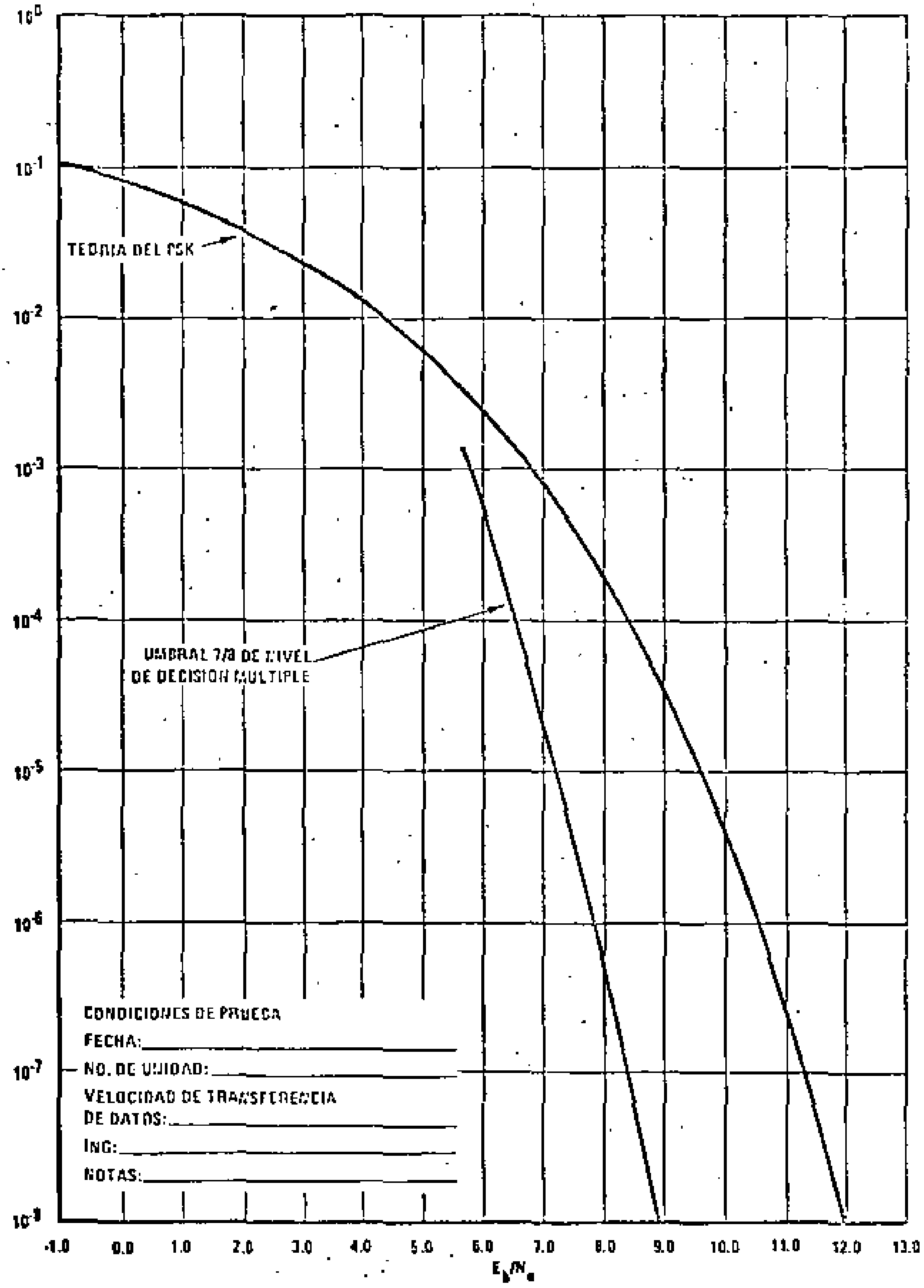
HARRIS

generar el bit de paridad. Los siete canales de datos se recombinan junto con la secuencia de paridad para generar la corriente de datos de salida que entra al modulador.

A continuación se muestra el desempeño del codec de decisión de umbral de 7/8 en comparación con la razón E_b/N_o teórica para PSK sin codificación.

c) TELESYSTEMS

Esta subsidiaria de COMSAT produce una terminal para acceso múltiple por división en el tiempo (TDMA), modelo DST-1000, que puede ser usada en redes privadas y públicas con ráfagas de 15 a 60 Mb/s, con capacidad hasta 100 estaciones por red. El modem de la terminal es QPSK, y emplea codificación por bloques del tipo Hamming con tasa de código 4/5. Se anexa copia del catálogo descriptivo.



"Soft" Codec de Umbral 7/8

HARRIS

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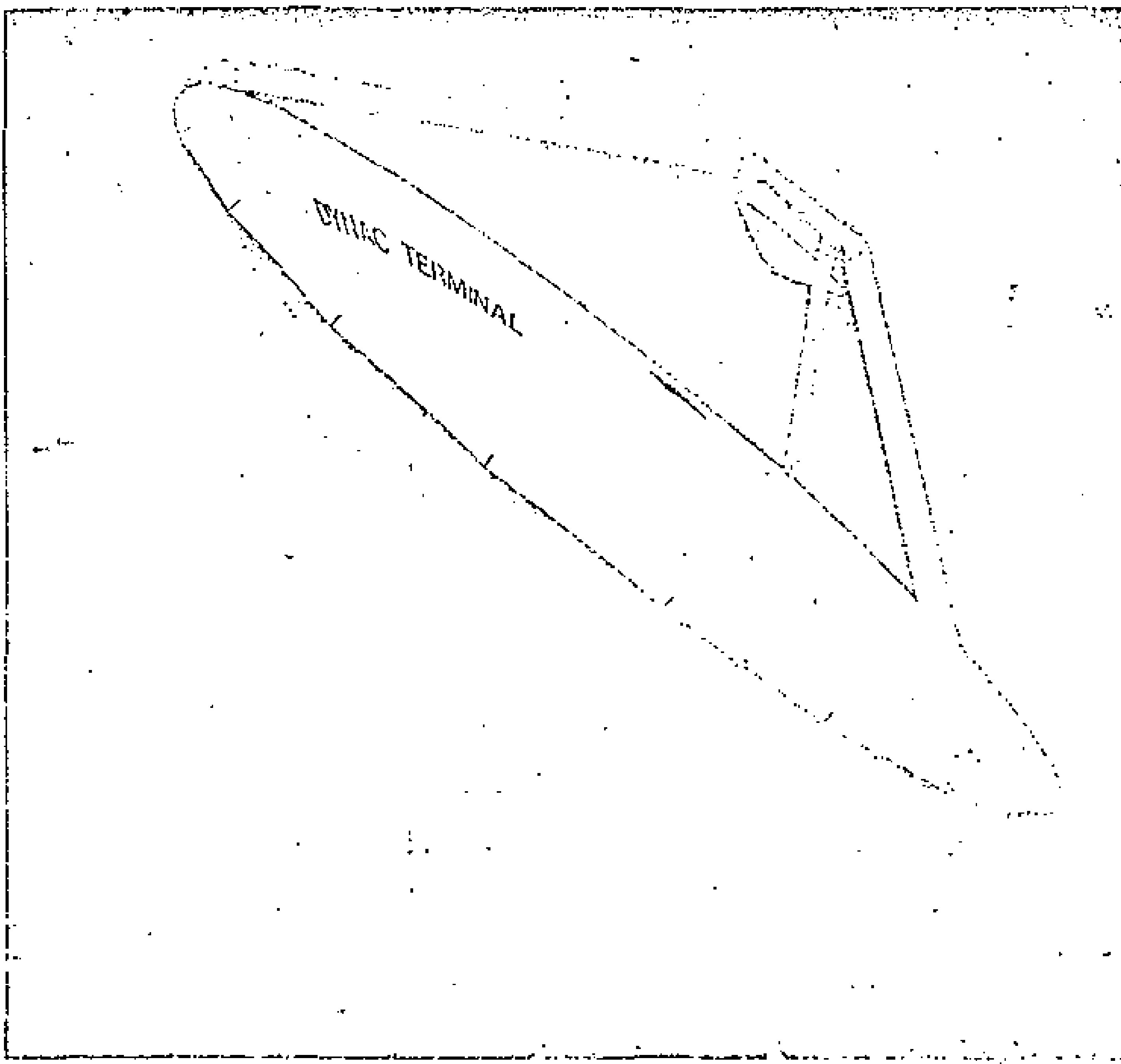
T E R M I N A L D Y N A C

T D M - T D M A

DYNAC™

INTELLIGENT EARTH STATION TERMINALS



Digital Communications Corporation

Dynamic Network Assignment Communications

The DYNAC™ series of satellite earth stations provide unique integrated communications packages designed for multiple access digital data/voice transmission. The earth stations offer maximum flexibility and statistical multiplexing with a selection of transmission rates from 268 kbps to 2 Mbps, a variety of digital and analog interfaces and microprocessor station control.

Implementation of a complete satellite communications network is afforded via the inherent features of the DYNAC™ series earth stations. The network architecture includes mixed data and voice traffic, centralized network control, incremental growth, unattended operation and phased transition from fixed assigned to dynamic assigned operation.

Network Configurations

The DYNAC earth station employs Time Division Multiplex (TDM) and Time Division Multiple Access (TDMA) techniques, and includes real-time dynamic allocation of transmission bandwidth. The combination of these features provides highly efficient space segment utilization and reduced system hardware costs.

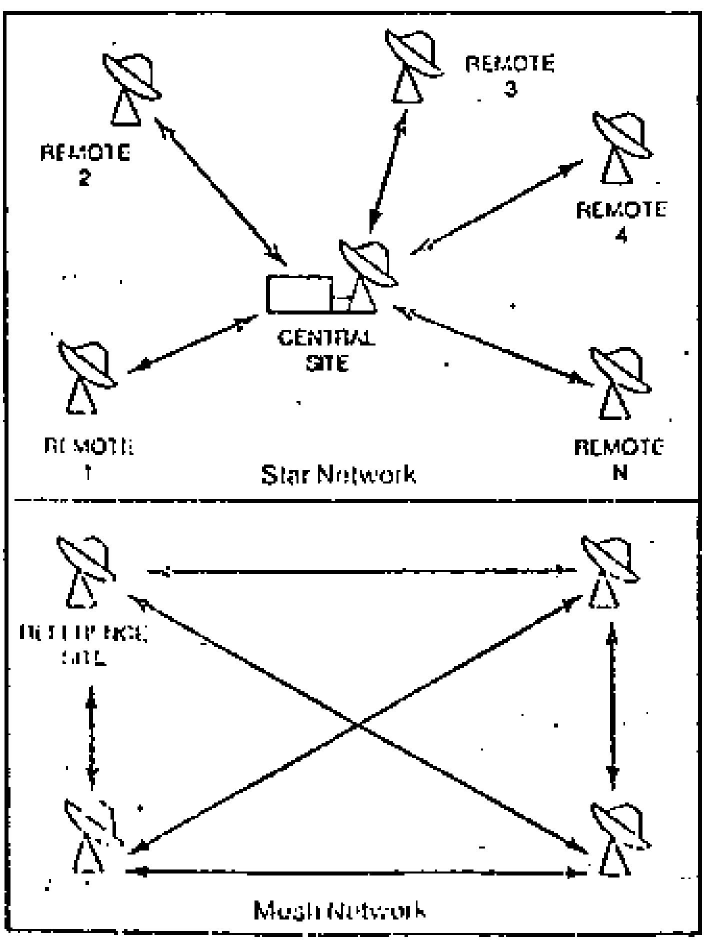


FIGURE 1
Network Configurations

The DYNAC earth station is used in both STAR and MESH network configurations, as shown in *Figure 1*. A STAR network involves a central site that communicates via TDM (continuous) broadcast transmission to all remote sites. Communication in the return path, i.e. from all remote sites to the central, is via TDMA. The STAR network uses a TDM/TDMA carrier pair. Note that remote-to-remote communications is via data routing through the central site. Such an interconnection requires a double satellite hop.

The MESH network uses the same terminal equipment as the STAR, but is a fully-connected network utilizing a single TDMA carrier. No central site exists, but one site is designated as the reference site, which provides network management, burst control and synchronization.

For fixed assigned, low density traffic requirements, SCPC can also be implemented.

Key Features

A. Dynamic Assignment

Based upon real-time requirements, satellite bandwidth is allocated to each station in the network, which allows unused capacity to be reassigned upon demand. Several network management algorithms can be used to determine how capacity shall be assigned. The central (or reference) site communicates signaling information to each station in the network, and reconfigures the network connectivity automatically. A terminal operator may also manually request or assign a capacity change.

B. Overhead Signaling Channel

The signaling channel used for network status and control is part of the transmission frame overhead. This duplex channel provides for remote control commands, network reassignments, requests and assignments, and fault monitoring and reporting. All command/control is handled via a microprocessor controlled unit, the Intelligent Terminal Interface (ITI).

C. Auto-Diagnostics and Fault Reporting

The ITI monitors all station fault indications and performs automatic status reporting to the central site. Remote diagnostics can also be performed, such that unmanned stations can be remotely controlled, various loopback functions performed, and test routines exercised.

In addition, the ITI can support a local maintenance terminal and has auto-dial modem capability, allowing for remote diagnostics to be performed without having a satellite path established.

D. Data Concentration

Two levels of concentration are realizable using the dynamic assignment technique. First, bandwidth may be allocated via time slot assignment on a port-by-port basis, at any of the rates shown in *Table 1*. Channel ports may be set up and disabled either manually, automatically upon terminal request, or by time-of-day control via the Network Management System (NMS). Bandwidth not allocated is returned to a "pool" from which it can be reassigned (similar to a DAMA system in FDMA networks).

Secondly, bandwidth may be varied within an assigned channel port in real-time, based upon an instantaneous requirement. Such a scheme is illustrated in *Figure 2*, which shows a statistical multiplexer supporting a variety of data terminals and supplying a multiplexed, synchronous serial data stream to the DYNAC channel port. The buffer occupancy status of the statistical multiplexer is sent via an asynchronous out-of-band interface to the ITI, which relays this information via the signaling channel to the NMS. The status is interpreted as a request for more (or less) bandwidth, and a new network plan is generated. All sites that may be affected are reassigned a new network plan consisting of channel port rate assignments and TDMA burst position. When the network reassignment is completed, the low speed clock for the affected port is changed to reflect the new rate. Dynamic assignment of bandwidth within a port extends the statistical multiplexer bandwidth advantages through the complete communications link.

TABLE 1

Available Channel Port Rates
300 baud
600 baud
1.2 Kbps
2.4 Kbps
4.8 Kbps
9.6 Kbps
19.2 Kbps
32.0 Kbps (voice)
56.0 Kbps

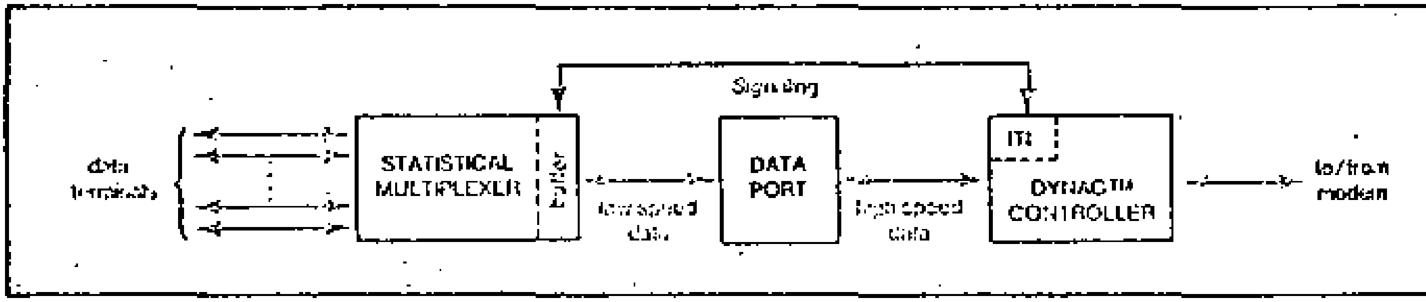


FIGURE 2

Statistical Multiplexer Interface

E. TDMA Burst Synchronization

Burst positions for all earth stations accessing the TDMA frame are assigned by the NMS. Slant path differentials between sites (depending upon the geographical locations of the sites) is known apriori, hence is accounted for in the burst position assignments.

It is necessary to allow for a guard time of sufficient duration between the bursts to preclude the possibility of bursts overlapping at the satellite, due to the doppler shift encountered with the figure-eight motion of the satellite (over one sidereal day). The DYNAC earth station employs open-loop burst synchronization during acquisition, which allows for this minimum guard time (approximately 300 microseconds) in the determination of burst assignments. In steady state (acquired) operation, burst positions are adjusted by the local station to minimize guard times. The scheme is extremely simple and has negligible impact on frame efficiencies at transmission rates below 2 Mbps.

All stations in the network are assigned their transmit burst positions relative to the broadcast reception of the REFERENCE UNIQUE WORD, which is transmitted by the central (or reference) site and denotes the beginning of the TDMA frame.

F. Selectable Frame Lengths

Frame lengths are selectable from 10 to 250 milliseconds duration. Larger frame lengths yield slightly higher bandwidth efficiencies, while shorter frame lengths are used where the transmission delay between interactive terminals is a major consideration.

G. Modularity

All components of the DYNAC earth station baseband equipment are contained on plug-in printed circuit modules, which are accessible from the front of the equipment tray by removing a plexiglass front panel. Test points, indicators and controls are located behind the front panel at the front of the circuit cards, facilitating ease of maintenance and test.

The IF/RF frequency converter, which is housed in a stand-alone chasis, uses stripline technology for high reliability and compact design. A removable plexiglass front panel allows access to test points. All internal assemblies are easily removable providing for ease of maintenance.

Typical Link Performance

Table 2 presents link performance summaries for two cases of a MESH configuration. Case 1 is a 537.6 kbps

Item	Value		Units
	Case 1	Case 2	
[illegible]	[illegible]	5	meters
[illegible]	22.0	22.0	dB/°K
[illegible]	[illegible]	[illegible]	--
[illegible] Output Power	40	125	watts
[illegible] Per Carrier	[illegible]	[illegible]	watts
[illegible]	[illegible]	[illegible]	dBW
Modulation	[illegible]-CPSK	[illegible]-CPSK	
[illegible]	TDMA	TDMA	
[illegible]	[illegible]	[illegible]	kbps
[illegible]	[illegible]	[illegible]	
[illegible] C/N	[illegible]		dB
[illegible] C/N	13.5		dB
[illegible]	[illegible]		dB
[illegible]	[illegible]		dB
[illegible] Margin	2.4		dB

TABLE 2

TDMA network with all stations having a 22.0 dB/°K G/T and transmitters via 40 watt TWT amplifiers. Case 2 is a 1.6128 Mbps TDMA network with the same earth station G/T and operating with 125 watt TWT amplifiers. Rate 1/2 coding is used in both cases with a threshold BER requirement of 10 [illegible] assumed. The LNA noise temperature is 120° K, and the antenna diameter is 5 meter (nominal).

Standard QPSK modem transmission Rates
268.8 Kbps
537.6 Kbps
1.544 Mbps*
1.6128 Mbps
2.048 Mbps*

* These rates do not support 56 Kbps data port operation

TABLE 3

Modulation

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The modulation technique employed is quadrature phase-shift keying (QPSK) with transmission rates to 2.048 Mbps. Standard modem rates are shown in *Table 3* and have been chosen to permit a simple baseband clock regeneration scheme.

Error Coding

The DYNAC controller contains a Rate 1/2 Forward-Error-Correction (FEC) codec that is field selectable.

Rate 7/8 hard decision or Rate 3/4 soft decision FEC codecs are also available. Specific network requirements determine the applicability of these codecs.

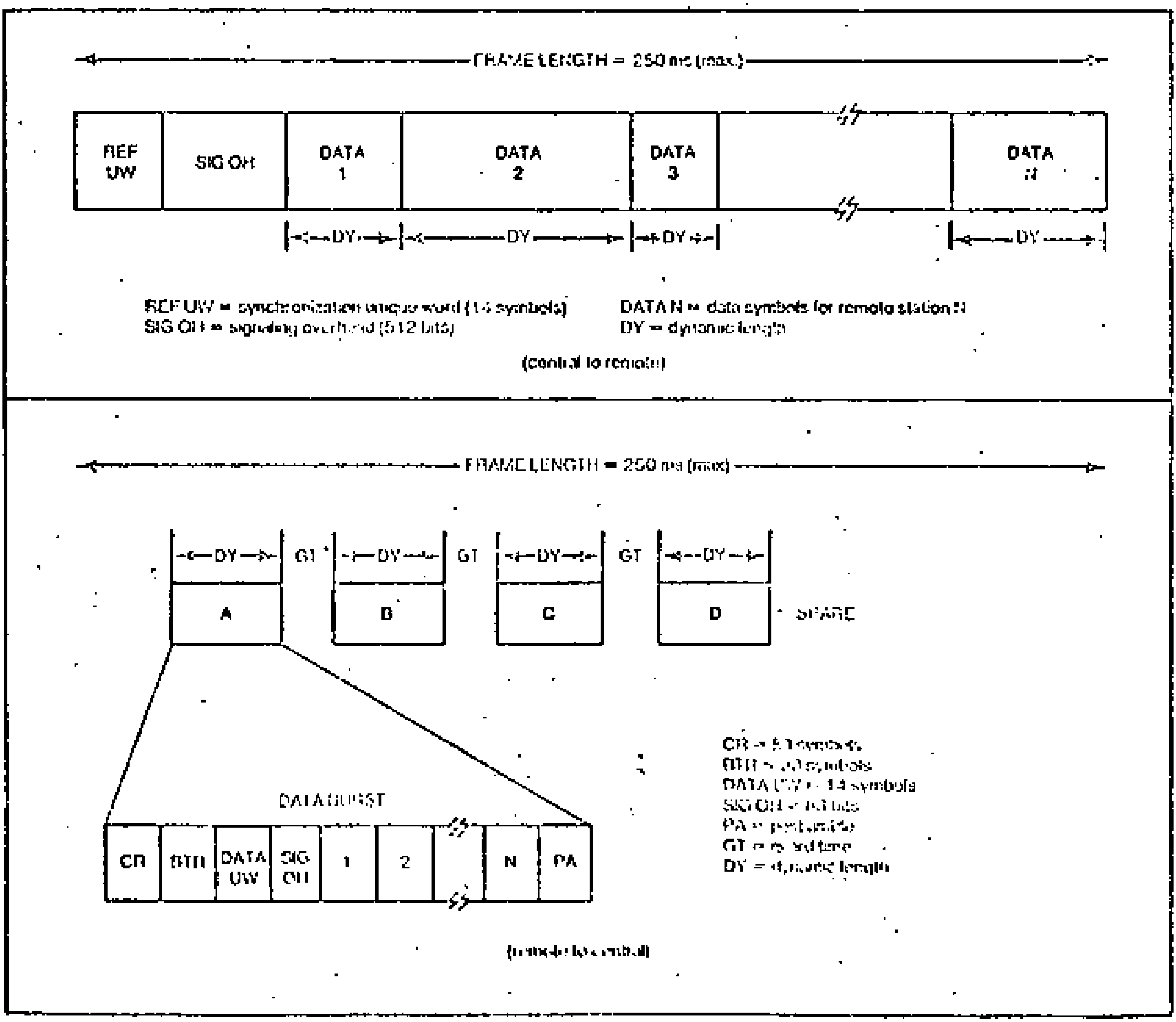

FIGURE 4

Frame Formats (Star Network)

Frame Formats

The frame format selected depends on the network configuration. *Figure 4* presents the formats used for the STAR network. Since transmission from the central station to the remotes is continuous (TDM), the format used is different than that for the remote stations (TDMA). Network synchronization is achieved via the Reference Unique Word in the TDM frame. *Figure 5* illustrates the frame format for the MESH network. Network synchronization is achieved via the REF UW contained in the reference station burst. In both cases, the frame length is dependent on the traffic requirements of the network.

The Carrier Recovery/Bit Timing Recovery (CR/BTR) sequence present at the beginning of each TDMA burst is required to provide for carrier and symbol timing acquisition of the burst demodulator. The CR/BTR sequence is followed by the unique word (UW) which is used to denote the first valid information bit in the burst. The reference unique word is used for frame synchronization as well.

The signaling overhead is the duplex command and control channel used for signaling between DYNAC earth stations and the central site. The outbound (central to remote) channel consists of 64 bytes of information per frame; each inbound (remote to central) channel consists of 11 bytes.

Data blocks for each assigned data port at the terminal are of different lengths, depending upon the rate assignment. Each port is assigned independently, and can operate transmit and receive sides at different rates as well. The time slot (block) for each port is assigned dynamically.

The postamble (PA) is required whenever a FEC codec is used, so that parity bits may be flushed from the error decoder.

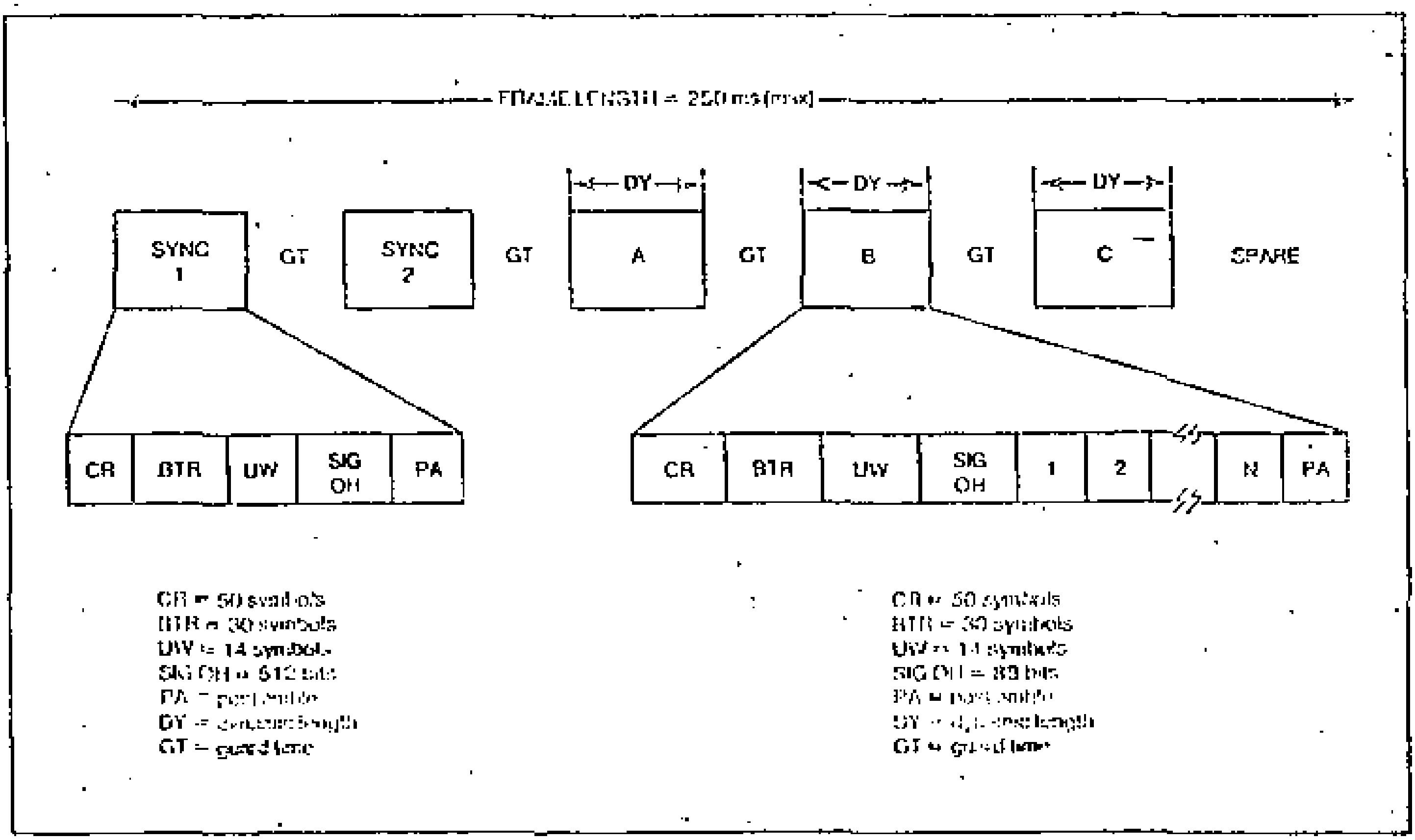

FIGURE 5
Frame Format (Mesh Network)

Terminal Operation

The DYNAC™ earth station is composed of two major subsystems. The first is the IF/RF Subsystem which includes the earth station antenna, low noise amplifier, power amplifier, and frequency converters. The second is the Digital Processing Subsystem which includes the control software and digital baseband hardware. These two major subsystems as well as their individual components are illustrated in *Figure 6* (nonredundant configuration). The operation of the system is described below.

IF/RF Subsystem

The IF/RF subsystem provides the interface between the digital processing subsystem and the satellite. Such interface requirements include the necessary frequency up conversion and power amplification of the outbound QPSK modulated IF carrier as well as the low noise reception and down conversion of the inbound 4 GHz signals from the satellite.

The antenna subsystem consists of a parabolic reflector, a specially shaped subreflector for Cassegrain operation, a dual linearly polarized microwave feed assembly,

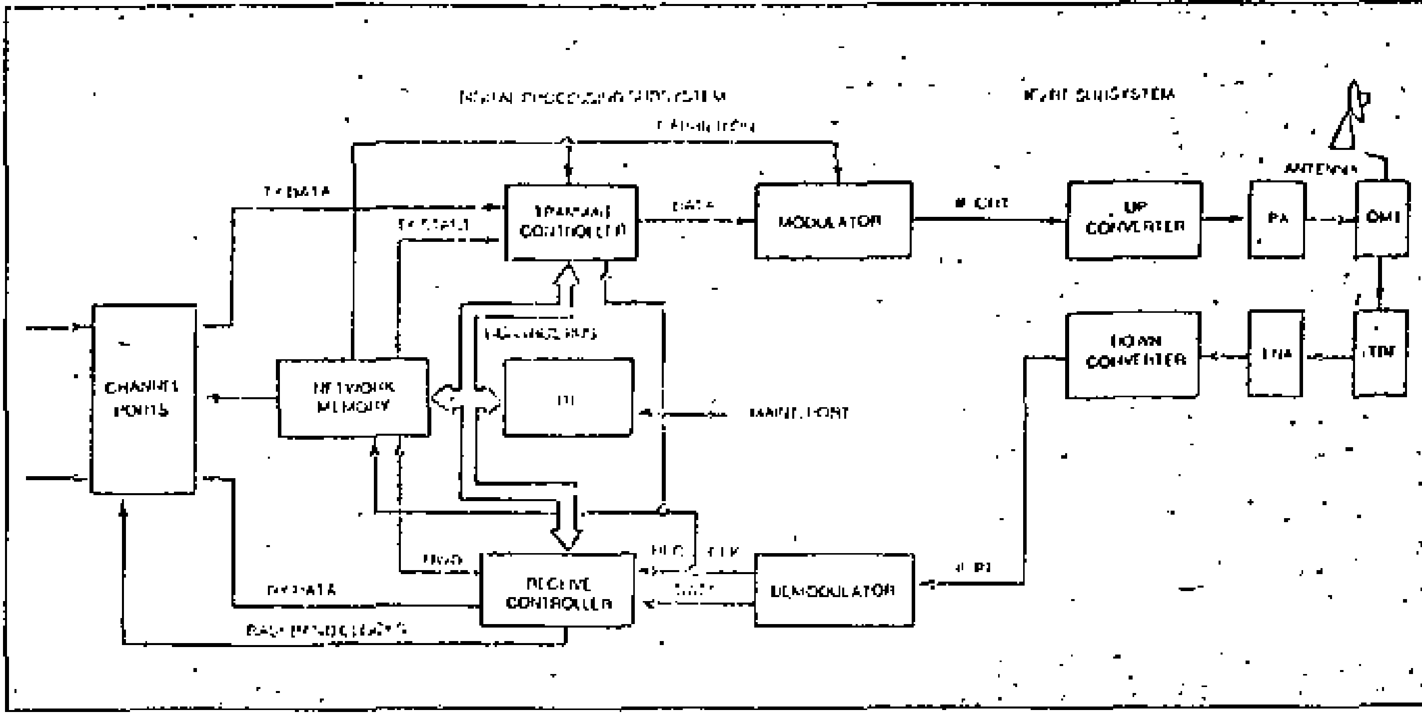

FIGURE 6
DYNAC™ Terminal Block Diagram (non-redundant)

and a manually adjustable mechanical mount. The parabolic reflector and complementary subreflector combined with the dual linearly polarized feed provide high gain and low sidelobe characteristics in the frequency bands of 3.7-4.2 GHz and 5.925-6.425 GHz. The simultaneous transmission and reception of the orthogonally linear polarized signals in the two frequency bands is accomplished via an orthomode transducer (OMT). The OMT is part of the feed assembly and is basically a frequency and polarization selective device.

On the receive side, the 4 GHz signal from the antenna subsystem passes through a transmit reject filter (TRF). The TRF protects the low noise amplifier from the high level 6 GHz signal, being transmitted. The receive signal is then passed through a low noise amplifier (LNA). The low noise properties of this amplifier coupled with its high gain set the noise figure of the receive system. The amplified output of the LNA then passes to the down converter where the 4 GHz signal is translated to a nominal 70 MHz IF frequency. It is this signal which interfaces with the Digital Processing Subsystem.

On the transmit side, the Digital Processing Subsystem supplies a QPSK modulated signal at the nominal 70 MHz IF frequency which is fed into the up converter. This signal is then translated to 6 GHz, amplified by the high power amplifier (HPA) and fed to the antenna. It is important to note that the frequency conversion performed by the up and down converters is done on a transponder basis. Channelization within a transponder is accomplished within the Digital Processing Subsystem.

Digital Processing Subsystem

On the transmit side, synchronous digital data (from any of several baseband data sources or voice digitizers) is written into compression buffers of the channel ports by the assigned low speed clock. The compression buffers serve to rate change the continuous low speed data to the high speed rate. The high speed data is gated from the compression buffers under control of the Transmit Network Memory which has been programmed by information "down-line" loaded from the Network Management System (NMS). The gating information is the network plan that also controls carrier on/off times, preamble generation, and the overhead signaling channel.

Data from all channel ports is multiplexed onto the high speed data bus and fed into the transmit TDMA controller, which adds the modem preamble, signaling channel, and postamble (if applicable), then scrambled with a PN sequence to ensure uniform spectral spreading prior to transmission.

The QPSK modulator accepts the two symbol streams (A and B data) and quadrature-phase modulates an IF carrier with this information. The modulated carrier is gated for transmission under direction of the TDMA controller.

The operation of the receive circuitry is similar to the transmit side. The input QPSK carrier is demodulated and the receive data is synchronized to the appropriate unique word detection (UWD). Descrambling and Rate 1/2 decoding (if applicable) of the data is synchronized to the UWD. Signaling data is demultiplexed from the high speed data stream and sent to the Intelligent Terminal Interface (ITI) for further processing. The remainder of the high speed data is selectively gated into the assigned channel ports under control of the Receive Network Memory. Expansion buffers within the channel port restore the high speed data to the low speed continuous rate. Baseband clocks are provided, as assigned, being generated from the receive symbol timing.

At the heart of the digital processing subsystem is the ITI which performs several key functions. First, all status and alarm information is relayed via the ITI to the NMS location for remote monitoring. Secondly, bandwidth requests and acknowledgements, network plan assignments and update commands, and control information are processed by the ITI. All of the above information is contained in the overhead signaling data. Via the NMS, complete access to the status and control of all remote DYNAC[TM] terminals is provided for unmanned operation of these sites. Automatic control (port assignment and "down-line" loading) of the entire network can be accomplished through the use of the ITI and its associated overhead channel.

Redundancy

The DYNAC earth station equipment is available in redundant or non-redundant configurations. For a redundant site, all IF/RF transmission hardware is provided with 1:1 redundancy and automatic switchover. The DYNAC baseband common equipment is also provided with 1:1 redundancy and auto switchover. Channel port modules are provided with two independent control and data bus interfaces, so that they can be controlled by either on-line system, electrically independent of the other, thus avoiding the possibility of a catastrophic failure. *Figure 7* shows the redundant configuration.

Reference site redundancy is provided by using primary and secondary reference bursts. Either site provides network synchronization, and upon loss of the primary site, the secondary site establishes network control.

Expansion

The baseband equipment may be expanded to provide a variety of channel port interfaces. Up to 40 channel ports may be connected at any earth station site. Five ports may be contained in the start-up tray (containing the DYNAC common equipment), and expansion chassis holds up to ten additional ports. The expansion tray contains provisions for redundant power supplies and control busses.

When the space-segment capacity requirements of the network are exceeded, the baseband equipment can be configured to allow for IF (intermediate frequency) hopping. A Channel Select Oscillator (CSO) module provides independent local oscillator references for the modem, and is controlled by instructions down-line loaded into the Network Memory. The station is then capable of transmitting and/or receiving multiple TDMA bursts per frame on different carrier frequencies, hence the network capacity is increased without the addition of baseband equipment, and most importantly without changing the earth station G/T requirements.

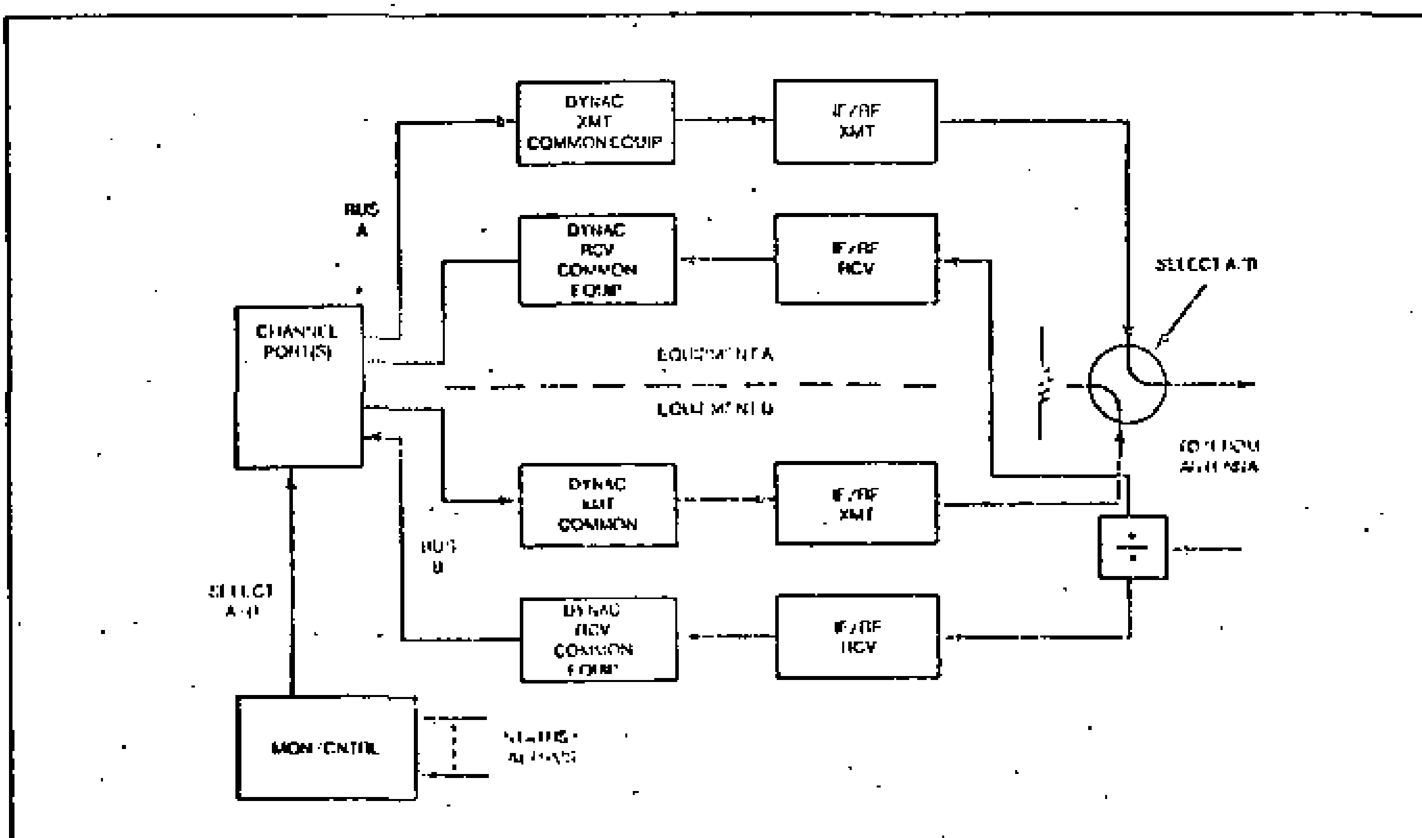

FIGURE 7
Redundancy Scheme

Voice Operation

For applications requiring VF interfaces, the DYNAC equipment can be configured with Continuously-Variable-Slope Delta (CVSD) voice codecs, which convert the analog voice input into a 32 kbps continuous data stream. The voice interface is balanced 600-ohm. The CVSD codec is integrated with a fixed rate 32 kbps data port on one printed circuit module, which is compatible with the data port module slots provided in the equipment tray. Thus, voice and data interfaces can be mixed.

To handle telephony signaling, the DYNAC equipment provides an E&M interface, which is used to detect hook status. For dynamic assignment, the E&M is used as a request message, which is forwarded to the central site processor via the overhead signaling channel.

When the VF circuit is established, multiple-frequency (MF) tone signaling (if required) may be sent in band, or single frequency (SF) or E&M signaling is converted to MF (externally), and then forwarded.

In STAR networks, all VF circuits are connected between the central site and any remote. However, in MESH applications, remote-to-remote connectivity is required. Signaling must first pass in-band (i.e. MF) to the reference site, so that the called earth station can be identified. Once the signaling is received, the network plan is reconfigured, allowing a second VF channel to be established between the reference site and the called earth station. Next, the signaling is forwarded from the reference earth station to the called remote site, upon receipt of which the reference site VF connections are disabled and the remote-to-remote VF is established. This is shown in *Figure 8*.

Timing and end-to-end delays are a consideration in planning this type of network, and interfaces (switches, etc.) frame rates and the network management system must be thoroughly planned. The dynamic assignment of the VF channels, however, allows for a higher space segment utilization versus fixed assigned channels.

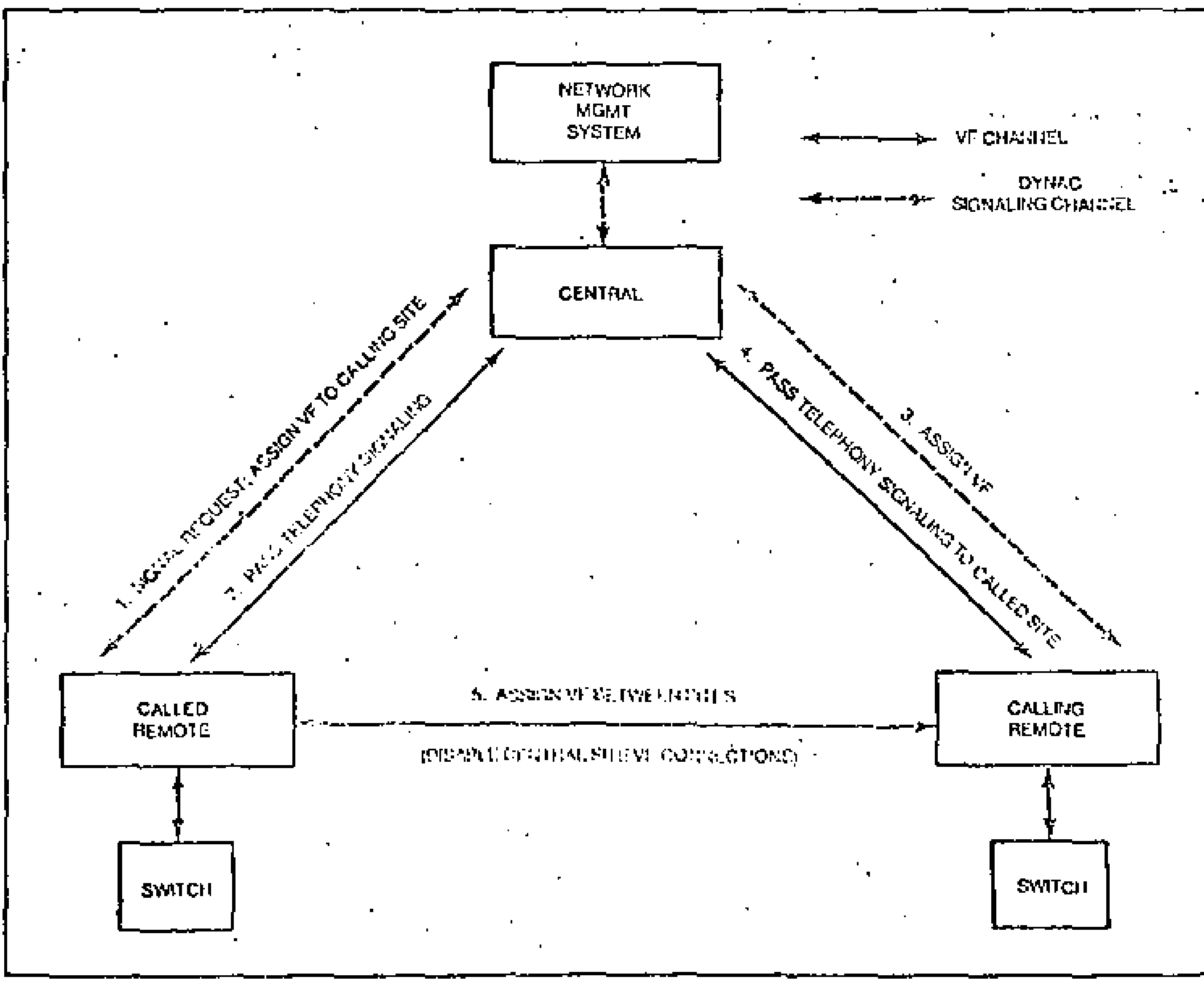

FIGURE 8
Voice Operation

Terminal Specifications

IF/RF Subsystem

Antenna Diameter: Sizes range from 4.6 to 11.0 meters depending on the network configuration and G/T requirements. *Table 4* lists some typical antenna characteristics.

TABLE 4
Typical Antenna Characteristics

PARAMETER	VALUE				
DIAMETER (METERS)	4.6	5.0	7.0	10.0	11.0
TYPE	CASSEGRAIN	CASSEGRAIN	CASSEGRAIN	CASSEGRAIN	CASSEGRAIN
OPERATING FREQUENCY (GHZ)	4 6	4 6	4 6	4 6	4 6
TYPICAL GAIN (dBi)	43.5 46.3	44.5 47.3	47.5 50.5	50.85 53.5	52.0 54.5
MOUNT TYPE	quasi-polar	el-over-az	el-over-az	el-over-az	el-over-az
POINTING ACCURACY (°) IN 30 mph WINDS GUSTING TO 45	0.04	0.03	0.04	0.041	0.041
OPERATING TEMPERATURE (°C)	-20 to +55	-20 TO +60	-20 TO +55	-20 TO +55	-20 TO +55
REC. NOISE TEMP. @ 20° EL.	26°K	23°K	26°K	25°K	23°K

Low Noise Amplifier: Available noise temperatures range from 35°K to 120°K depending on the network configuration and G/T requirements. *Table 5* lists some typical LNA characteristics.

TABLE 5
Typical LNA Characteristics

PARAMETER	Noise Temperature (°K) 35 / 45 / 70 / 90	120
Type	parametric	GaAs FET
operating frequency	3.7 - 4.2 GHZ	3.7 - 4.2 GHZ
min. gain	50 dB	50 dB
max. gain flatness	±0.5 dB	±0.5 dB
Intermodulation (3rd order)	60 dB rejection for two carriers at input levels of -72 dBm	50 dB below each of two carriers at input levels up to -55 dBm
overdrive	0 dBm input with no permanent performance degradation	+10 dBm input with no permanent performance degradation
input connection	CPR-229/G waveguide	CPR-229/G waveguide
output connection	Type N female	Type N female

High Power Amplifier: Saturated output powers range from 5 Watts to 400 Watts depending on the network configuration. *Table 6* lists some typical HPA characteristics.

TABLE 6
Typical HPA Characteristics

PARAMETER	Saturated Output Power (Watts) 5	20 / 75 / 125 / 400
Type	GaAs FET	TWTA
operating frequency	5.925-6.425 GHZ	5.925-6.425GHZ
min. gain	50 dB	50 dB
gain flatness	±0.4dB/40MHZ	±1.0 db across 500 MHZ
noise figure	12 dB max	35 dB max

Subsystem Examples: Typical antenna/LNA/HPA combinations are shown in *Table 7.*

TABLE 7

Typical Antenna/LNA/HPA Combinations

modem bit rate (Kbps)	network configuration: Star central	Star remote	mesh	antenna diameter (M): 4.6	5	7	10	11	HPA sat. output power (watts): 5	20	75	125	400
208.0		x		x					x				
	x							x	x	x (2 carrs)	x (8 carrs)	x (16 carrs)	x (32 carrs)
208.8			x		x					x			
1.0128		x			x						x		
	x						x				x	x (3 carrs)	x (8 carrs)
1.6128			x		x							x	
1.6128			x				x				x		

NOTE: (1) LNA noise temperature is 120°K for above examples

Frequency Converters: Up and down converter characteristics are shown in *Table 8.*

TABLE 8

Frequency Converter Characteristics

PARAMETER	up converter	down converter
General Characteristics		
type	dual conversion stripline	dual conversion stripline
tunnability	mechanical via crystal change	mechanical via crystal change
frequency sense	no spectral inversion	no spectral conversion
Input Characteristics		
operating frequency	52-88 MHZ	3.7-4.2 MHZ
translated bandwidth	40 MHZ	40 MHZ
input level	-20 dBm nominal	up to -30 dBm composite
impedance	50 ohms, unbal.	50 ohms, unbal
local oscillators C band L-band	4762.5-5222.5MHZ 1112.5MHZ	4762.5-5222.5MHZ 1112.5MHZ

PARAMETER	up converter	down converter
Output Characteristics		
output level	0 dBm for single carrier input level of -20 dBm	-10 dBm for desired single carrier input of -55 dBm
1 dB compression point	+10 dBm	0 dBm
impedance	50 ohms, unbal.	50 ohms, unbal.
Transfer Characteristics	center 36MHZ BW	center 36MHZ BW
gain stability (constant temp.)	± 0.25dB	± 0.25dB
gain slope	± 0.1dB/MHZ	± 0.1dB/MHZ
intermodulation (3rd order)	60dBc with two equal RF carriers at an output level of -10dBm	60dBc with two equal IF carriers at an output level of -10dBm

Baseband Interfaces

Data: .... RS-232C or V.35
Format: .... Serial Synchronous
...nnel Port Rates: .... Refer to Table 1
...: .... 4 wire, 600 ohm
Voice Encoding: .... 32 kbps CVSD
Telephony Signaling: .... E&M, SF or MF interfaces available

Modem

Modulation: .... Coherent QPSK, burst or continuous
Operating Bit Rates: .... Refer to Table 2
Ambiguity Resolution: .... Unique Word Detection
Forward Error Correction: .... Rate 1/2 (Selectable)
Rate 7/8 Hard Decision (Optional)
Rate 3/4 Soft Decision (Optional)
AFC: .... Swept Acquisition, ± 40 kHz
Burst-to-burst + 1 kHz
AGC: .... ± 10 dB (Continuous)
± 2 dB (Burst)
Transmit IF Frequency: .. Selectable from 52 to 88 MHz (CSO Option Available)
L.O. Stability: .... Aging < 1 x 10^{-7}/week
BER Threshold .... 10^{-4}@ Eb/No = 10.4 dB (uncoded)

Network Control

Frame Length: .... Selectable from 10 ms to 250 ms
...aling Channels: .... 512 bits (from NMS)
88 bits (to NMS)
Orderwire Channels: ... May be included in OH signaling
Network Configuration: ...... Automatic, in real time, via network controller signaling. Typical reconfiguration times are 2 to 5 seconds.

General Specifications, DYNAC Controller

Fault Indication: ..... Local visual indicators and Form C summary alarms; automatic status reporting to central terminal.
Redundancy: .... 1:1, 1:N configurations available
Power Requirements: .... 115 VAC, + 10%, 60 Hz @ 10 amps
Physical Dimensions: .... 8.75" (222.3 mm) high, 19" (482.6 mm) wide, 28" (711.2 mm) deep
Weight: .... 12 kg (approx.)
Mounting: .... EIA standard
Environment: .... Operating 10°C to 40°C
Storage -25°C to 85°C

CVSD Codec (Option)

Clock Rate .... 32 kbps
Input/Output Impedance ...... 600 Ω balanced ± 10%

Frequency Response: (at -10 dBm0)
300 - 2500 Hz .... ± 1 dB
300 - 3000 Hz .... ± 1 dB, -2 dB
300 - 3400 Hz .... ± 1 dB, -4 dB

Delay Distortion
600 - 2500 Hz .... < 200 μsec
300 - 3000 Hz .... < 500 μsec
300 - 3400 Hz .... < 1000 μsec

Signal to Quantization Noise
(800 Hz, C-message weighting)
Input: + 3 to -20 dBm0 .... 32 dB
-20 to -30 dBm0 .... 30 dB
-30 to -39 dBm0 .... 22 dB

Gain Tracking
(deviation from gain at -20 dBm0)
Input: 800 Hz + 3 to 0 dBm0 .... ± 1.5 dB
0 to -30 dBm0 .... ± 1.0 dB
-30 to -40 dBm0 .... ± 1.5 dB

Idle Noise
(C-message weighting) .... < 65 dBm0

Data Signal Transparency
(2400 bps, bit error rate): .... < 10^{-5}

Support Services

DCC's technical staff is prepared to assist communications managers in the initial definition and planning for a specific network configuration. We are also available to coordinate site surveys and FCC licensing. Installation and maintenance can also be provided.

For further information, contact our satellite communications marketing department.

TERMINAL DST - 1000, TDMA.

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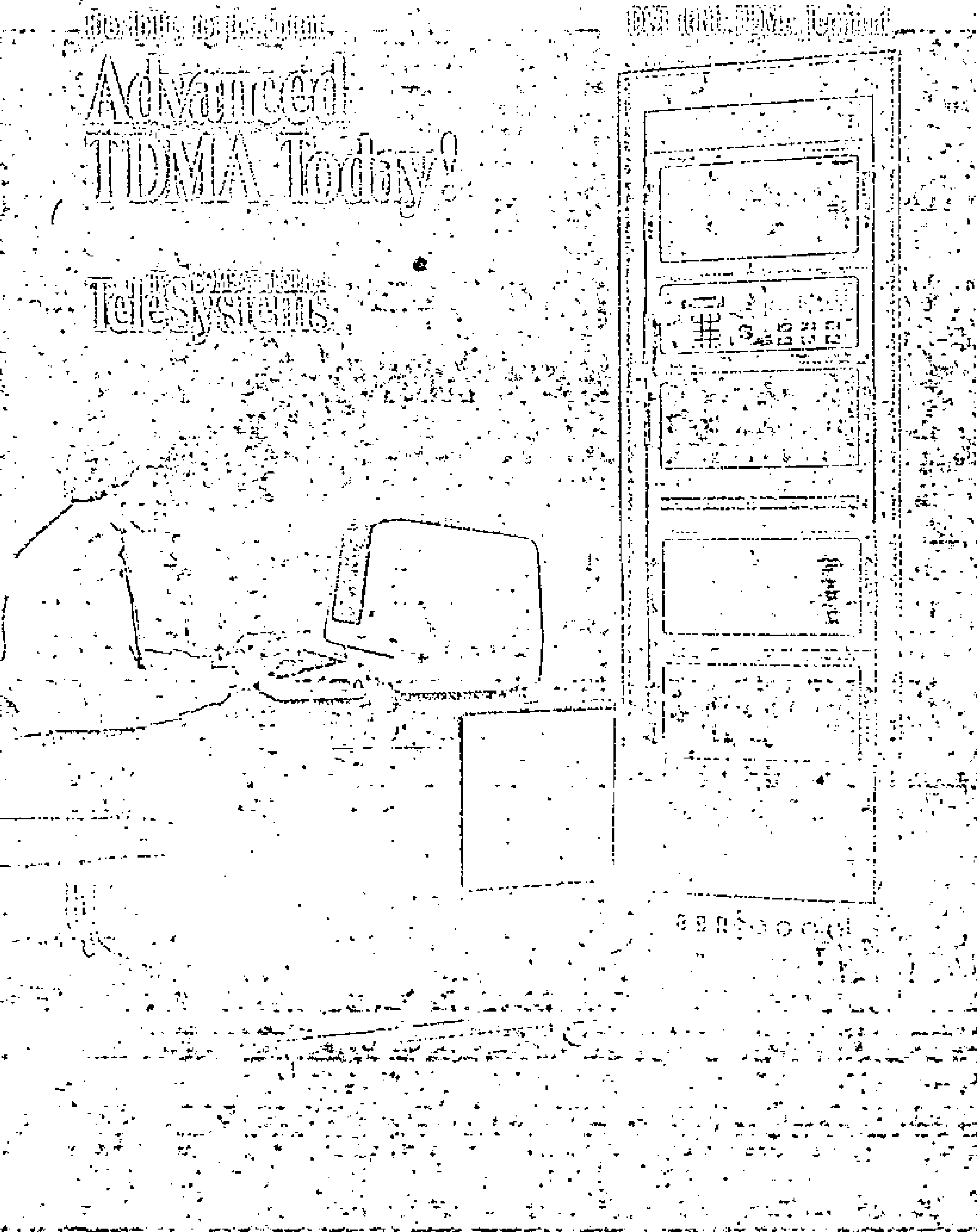
T E L E S Y S T E M S

(COMSAT GENERAL)


Advanced TDMA Today?
TeleSystems

Communications Services

- VOICE: 64 Kb/s PCM embedded in T1 stream with or without E&M, DTMF signaling.
- DATA: 64 Kb/s and N x 64 Kb/s, up to 1.544 Mb/s, embedded in T1 stream.
- VIDEO: encoded into 1.544 Mb/s T1 stream, with or without signaling.
- CCITT, Voice, Data & Video Services at 2.048 Mb/s CEPT-32 Standards.
- OTHER: provision can be made to interface with most conventional digital rates and disciplines.

General Description

The DST-1000 is a medium-rate design for use in private network and common carrier applications requiring burst bit rates of 15 to 60 Mb/s, with capacity for up to 100 stations in the network. Employing the Motorola 68000 family of microprocessors for network management and terminal control, extensive flexibility is achieved in the restructuring of bursts to accommodate changing traffic patterns and user requirements.

Designed for a wide range of applications, the DST-1000 system offers numerous features which lend themselves to evolutionary implementation – as requirements change, the system can be altered to accommodate those needs without making existing elements obsolete. Similarly, plans for future growth and expansion of the network do not significantly impact initial costs – there is no need to purchase extra capability initially in order to accommodate anticipated requirements which may or may not materialize as projected.

Highly effective utilization of space segment capacity is realized by virtue of the fact that burst positions, lengths, and content can readily be adjusted from a central TDMA console. Microprocessor control also achieves considerable reduction in logic components required for the acquisition, synchronization, and traffic management functions, as well as allowing lower rate operation which reduces transponder leasing costs until such time as traffic loads necessitate full transponder operation.

Designed for direct interface with T1 PCM encoded voice channels, any conventional interface operating at a 1.544 Mb/s rate can be accommodated. Video conferencing is a standard feature, as is split T1 operation, permitting individual 64 Kb/s channel streams, or multiples thereof, to be routed individually by source and destination. T1 interfaces are synchronous to the TDMA clock, providing slip free operation through the system. Similarly, CEPT standards of 2.048 Mb/s are equally acceptable by substituting a different family of Terrestrial Interface Modules (TIM's).

A demand assignment (DAMA) feature, which employs a combination of centralized frame management and distributed call processing, is available either as a part of the initial configuration or as an add-on capability. With this feature, the Network Control Center (NCC) at the Reference Terminal assigns bursts to specific stations within the network, and the individual stations (Local Terminals) access and utilize these time slots to transmit traffic channels on a call-by-call basis, eliminating the need for full-time allocation of slots to service traffic peaks or intermittent demands. This also reduces equipment requirements associated with network synchronization and frame management by requiring a maximum of two stations to be so-equipped, at the same time eliminating much of the frame

Frame Architecture

Dual Reference Bursts are incorporated in the frame architecture, as shown in Figure 1, to ensure utmost reliability in network management. The Reference Bursts maintain frame synchronization via the Transmission Instant Channel, adding and deleting frame space and sending self-test instructions to the individual stations via the Frame Management/Command Channel.

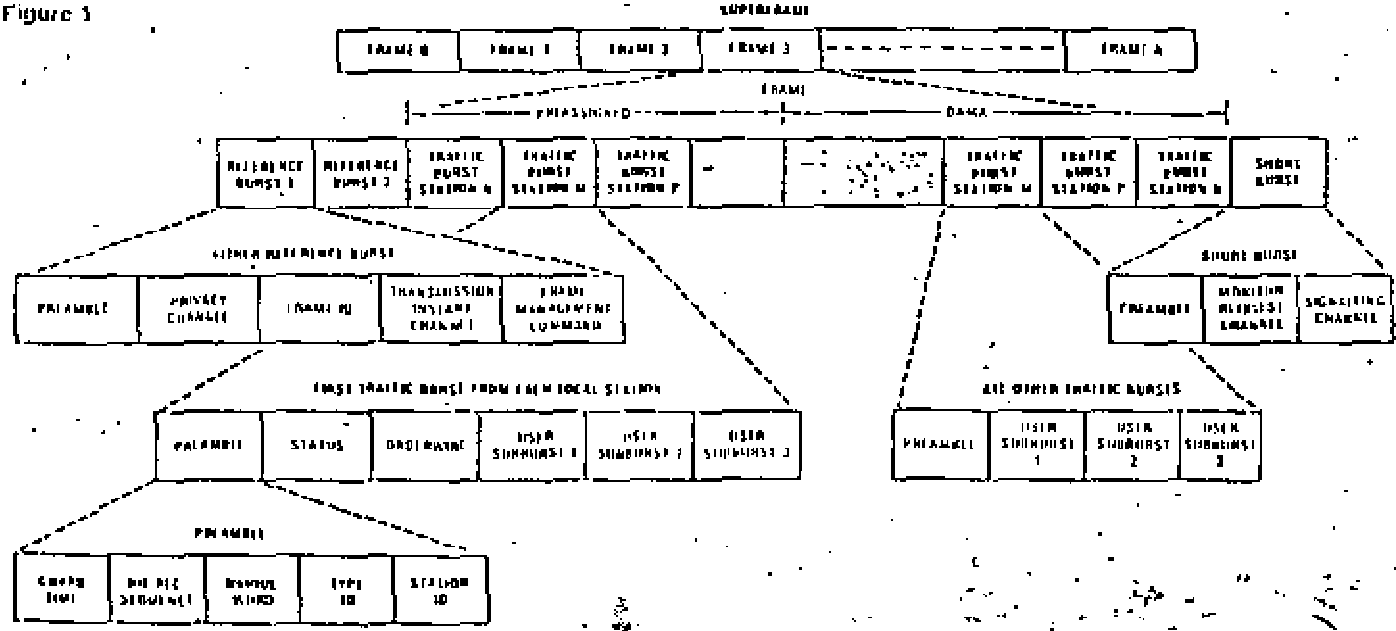
The Privacy Channel of each Reference Burst carries an encrypted scrambler starting key to the Local Terminals; 2?? possible combinations make compromise of network traffic highly improbable.

The future addition of the DAMA feature is accomplished without alteration of the frame architecture by dividing each frame into preassigned and demand assigned segments – the dynamic nature of the frame architecture allows utilization of these slots for preassigned traffic until such time as the DAMA feature is desired.

Status monitoring and request data are carried between the Local Terminals and the Network Control Centers at the Reference Terminals via a Short Burst included in each frame. This burst also carries call-by-call signaling information when the DAMA option is selected.

Traffic Bursts in the preassigned mode have essentially the same composition as Time Slots in the demand assigned mode. However, with DAMA, the Orderwire Channel is eliminated, this function using demand assigned sub-bursts as required.

Figure 1



DST-1000 TDMA Frame Structure

overhead associated with centralized call processing.

Any degree of redundancy can be incorporated. The system can be supplied in a non-redundant configuration where low cost is a predominant consideration and availability [illegible] can tolerate occasional short outages. Where availability is a prime factor, full redundancy with automatic switchover is offered, with 1:1 redundancy of all common equipment and 1:N redundancy of individual channel equipment. Conventionally, N = 3, however, other ratios can be provided as desired. Similarly, the system can be furnished with a single, non-redundant Reference Station, with two fully redundant Reference Stations, or with any desired mix.

Terminal Configuration

A typical Local Terminal consists of Burst Modems, Frame Management Processors, and Terrestrial Interface Units arranged in a redundant configuration to provide automatic back-up in the event of failure of an on-line unit. In a DAMA application, Terrestrial Signaling Units are employed for processing of the individual calls. A Reference Terminal contains this Local Terminal complement plus Frame Supervisory/Network Control Processors and a Reference Clock for network timing. Figure 2 illustrates the Reference Terminal configuration, with the Local Terminal component outlined by a heavy dashed line. The Network Control Center is also a part of the Reference Terminal, and includes interface provision for the Network Monitor and Control System and Centralized Automatic Message Accounting, where required. In a fully redundant network, two such Reference Terminals are employed.

Terminal expansion is greatly facilitated by the bus structure used in interfacing the common equipment with the individual Terrestrial Interface Modules (TIM's). With this arrangement, a variety of terrestrial interfaces can be accommodated without impacting the basic hardware and software structure. Expansion is accomplished by adding TIM's as required. The TIM bus consists of data, address and control lines, with separate buses for transmit and receive paths. During burst transmission, the common equipment presents address and control signals on the transmit TIM bus. Each address identifies a specific TIM and a data block to be read from that TIM, allowing considerable flexibility in multiplexing TIM data and [illegible]. The burst format control network is part of the common equipment. In the receive direction, the opposite functions are performed.

The Modem uses QPSK modulation, employing codeword correlation for resolution of phase ambiguity while avoiding differential encoding loss. Square root Nyquist filters have been specified for the pulse shaping elements, using extensive computer simulation for optimization of BER performance and spectral control.

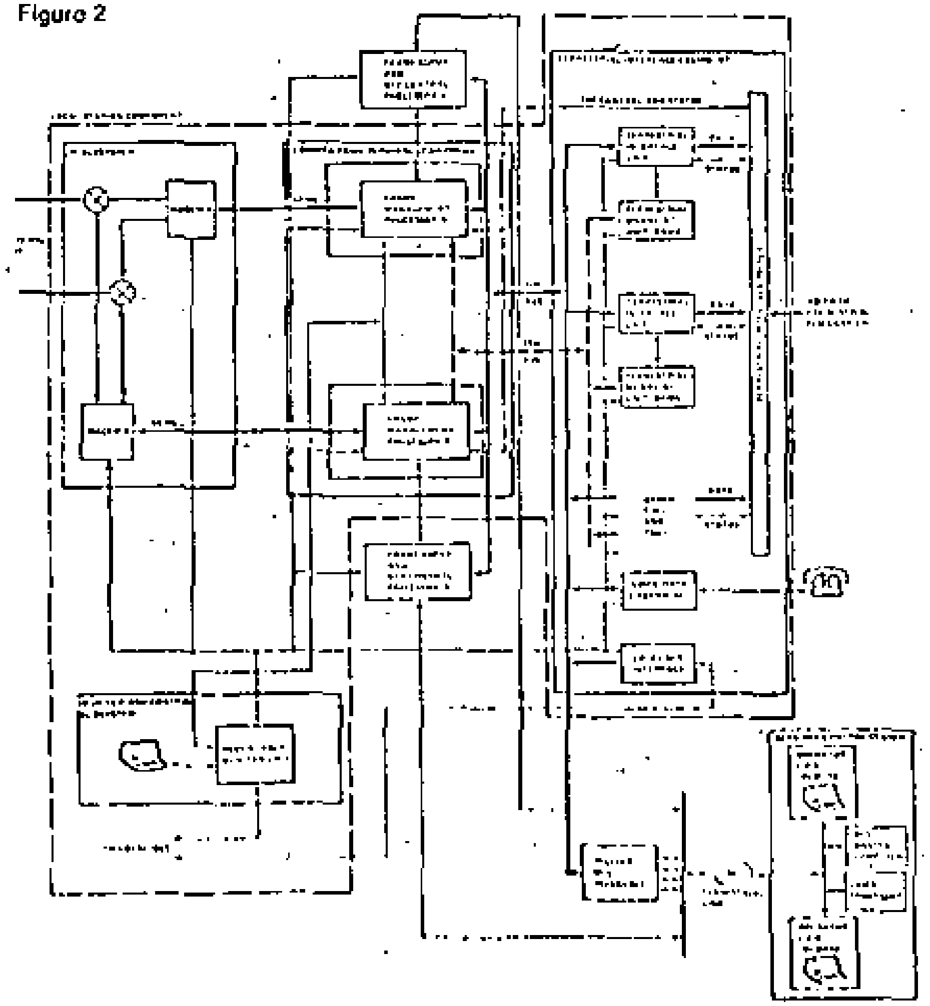
Forward error correction can be employed on any voice, video, data or signaling channel, as required, independent of its use on other channels. A Rate 4/5 Hamming code is utilized, and FEC can be used in DAMA or preassigned operation.

A pseudo random scrambling sequence is introduced to modify all data following the Unique Word, prior to QPSK modulation, dispersing power in the transmitted spectrum per CCIR requirements and permitting clock recovery when long sequences of zeroes are sent. The scrambler starting key is encrypted, using the DES Standard, thus providing a high degree of communications privacy.

Network Maintenance/Operation

- remote monitoring of all local station status by NCC via monitor/request channel
- remote commands by NCC to execute on-line self-test routines at each local station
- loopback testing of interfaces upon command of remote or local operators
- local station monitoring and fault diagnosis performed by redundancy monitor unit
- local or reference station status available for call-up by operator
- faults automatically detected to board level during on-line operation
- detailed fault testing can be performed with the aid of built-in self-test routines
- voice orderwire facilities
- data orderwire facilities can be used by monitor & control system.

Figure 2



DST-1000 Reference Terminal Block Diagram

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Convolutional Codes and Their Performance in Communication Systems

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***Abstract*—This tutorial paper begins with an elementary presentation of the fundamental properties and structure of convolutional codes and proceeds with the development of the maximum likelihood decoder. The powerful tool of generating function analysis is demonstrated to yield for arbitrary codes both the distance properties and upper bounds on the bit error probability for communication over any memoryless channel. Previous results on code ensemble average error probabilities are also derived and extended by these techniques. Finally, practical considerations concerning finite decoding memory, metric representation, and synchronization are discussed.**

I. Introduction

ALTHOUGH convolutional codes, first introduced by Elias [1], have been applied over the past decade to increase the efficiency of numerous communication systems, where they invariably outperform block codes of the same order of complexity, there remains to date a lack of acceptance of convolutional coding and decoding techniques on the part of many communication technologists. In most cases, this is due to an incomplete understanding of convolutional codes, whose cause can be traced primarily to the sizable literature in this field, composed largely of papers which emphasize details of the decoding algorithms rather than the more fundamental unifying concepts, and which, until recently, have been divided into two nearly disjoint subsets. This malady is shared by the block-coding literature, wherein the algebraic decoders and probabilistic decoders have been at odds for a considerably longer period.

The convolutional code dichotomy owes its origins to the development of sequential (probabilistic) decoding by Wozencraft [2] and of threshold (feedback, algebraic) decoding by Massey [3]. Until recently the two disciplines flourished almost independently, each with its own literature, applications, and enthusiasts. The Fano sequential decoding algorithm [4] was soon found to

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greatly outperform earlier versions of sequential decoders both in theory and practice. Meanwhile the feedback decoding advocates were encouraged by the burst-error correcting capabilities of the codes which render them quite useful for channels with memory.

To add to the confusion, yet a third decoding technique emerged with the Viterbi decoding algorithm [9], which was soon thereafter shown to yield maximum likelihood decisions (Forney [12], Omura [17]). Although this approach is probabilistic and emerged primarily from the sequential-decoding oriented discipline, it leads naturally to a more fundamental approach to convolutional code representation and performance analysis. Furthermore, by emphasizing the decoding-invariant properties of convolutional codes, one arrives directly to the maximum likelihood decoding algorithm and from it to the alternate approaches which lead to sequential decoding on the one hand and feedback decoding on the other. This decoding algorithm has recently found numerous applications in communication systems, two of which are covered in this issue (Heller and Jacobs [24], Cohen *et al.* [25]). It is particularly desirable for efficient communication at very high data rates, where very low error rates are not required, or where large decoding delays are intolerable.

Foremost among the recent works which seek to unify these various branches of convolutional coding theory is that of Forney [12], [21], [22], *et seq.*, which includes a three-part contribution devoted, respectively, to algebraic structure, maximum likelihood decoding, and sequential decoding. This paper, which began as an attempt to present the author's original paper [9] to a broader audience,[1] is another such effort at consolidating this discipline.

It begins with an elementary presentation of the fundamental properties and structure of convolutional codes and proceeds to a natural development of the maximum likelihood decoder. The relative distances among codewords are then determined by means of the generating function (or transfer function) of the code state diagram. This in turn leads to the evaluation of coded communication system performance on any memoryless channel. Performance is first evaluated for the specific cases of the binary symmetric channel (BSC) and the additive white Gaussian noise (AWGN) channel with biphase (or quadriphase) modulation, and finally generalized to other memoryless channels. New results are obtained for the evaluation of specific codes (by the generating function technique), rather than the ensemble average of a class of codes, as had been done previously, and for bit error probability, as distinguished from event error probability.

The previous ensemble average results are then extended to bit error probability bounds for the class of time-varying convolutional codes by means of a generalized generating function approach; explicit results are obtained for the limiting case of a very noisy channel and compared with the corresponding results for block codes. Finally, practical considerations concerning finite memory, metric representation, and synchronization are discussed. Further and more explicit details on these problems and detailed results of performance analysis and simulation are given in the paper by Heller and Jacobs [24].

While sequential decoding is not treated explicitly in this paper, the fundamentals and techniques presented here lead naturally to an elegant tutorial presentation of this subject, particularly if, following Jelinek [18], one begins with the recently proposed stack sequential decoding algorithm proposed independently by Jelinek and Zigangirov [7], which is far simpler to describe and understand then the original sequential algorithms. Such a development, which proceeds from maximum likelihood decoding to sequential decoding, exploiting the similarities in performance and analysis has been undertaken by Forney [22]. Similarly, the potentials and limitations of feedback decoders can be better understood with the background of the fundamental decoding-invariant convolutional code properties previously mentioned, as demonstrated, for example, by the recent work of Morrissey [15].

II. Code Representation

A convolutional encoder is a linear finite-state machine consisting of a K-stage shift register and n linear algebraic function generators. The input data, which is usually, though not necessarily, binary, is shifted along the register b bits at a time. An example with $K = 3$, $n = 2$, $b = 1$ is shown in Fig. 1.

The binary input data and output code sequences are indicated on Fig. 1. The first three input bits, 0, 1, and 1, generate the code outputs 00, 11, and 01, respectively. We shall pursue this example to develop various representations of convolutional codes and their properties. The techniques thus developed will then be shown to generalize directly to any convolutional code.

It is traditional and instructive to exhibit a convolutional code by means of a tree diagram as shown in Fig. 2.

If the first input bit is a zero, the code symbols are those shown on the first upper branch, while if it is a one, the output code symbols are those shown on the first lower branch. Similarly, if the second input bit is a zero, we trace the tree diagram to the next upper branch, while if it is a one, we trace the diagram downward. In this manner all 32 possible outputs for the first five inputs may be traced.

From the diagram it also becomes clear that after the first three branches the structure becomes repetitive. In fact, we readily recognize that beyond the third branch the code symbols on branches emanating from the two nodes labeled *a* are identical, and similarly for all the

[1] This material first appeared in unpublished form as the notes for the Linkabit Corp. "Seminar on convolutional codes," Jan. 1970.

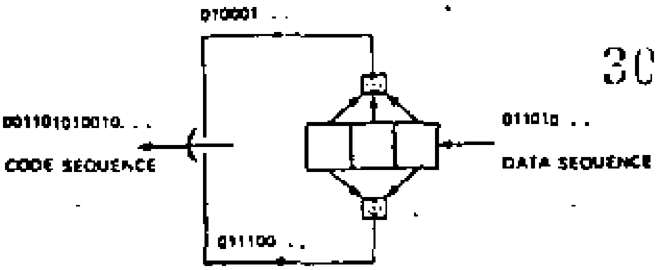

Fig. 1. Convolutional coder for $K = 3$, $n = 2$, $b = 1$.

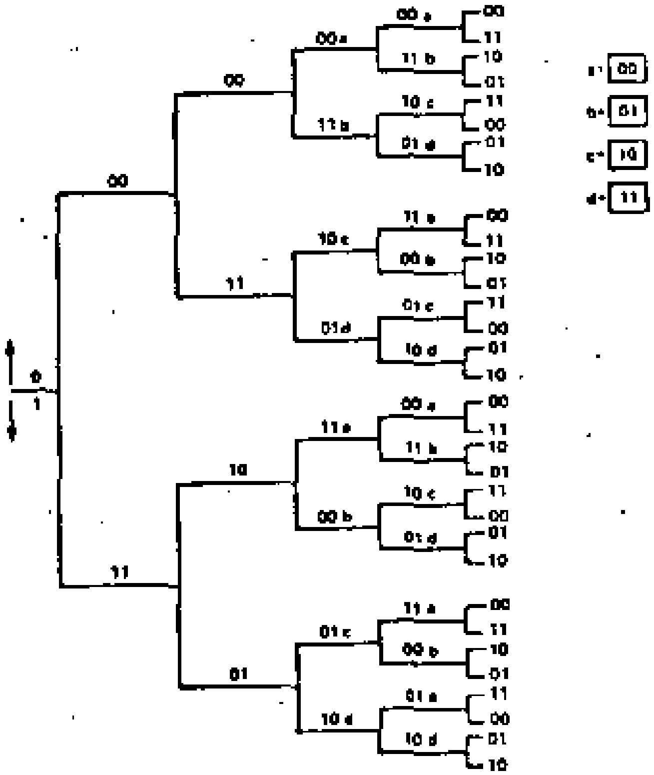

Fig. 2. Tree-code representation for coder of Fig. 1.

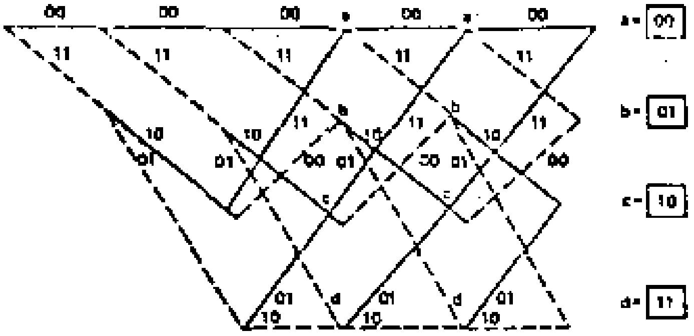

Fig. 3. Trellis-code representation for coder of Fig. 1.

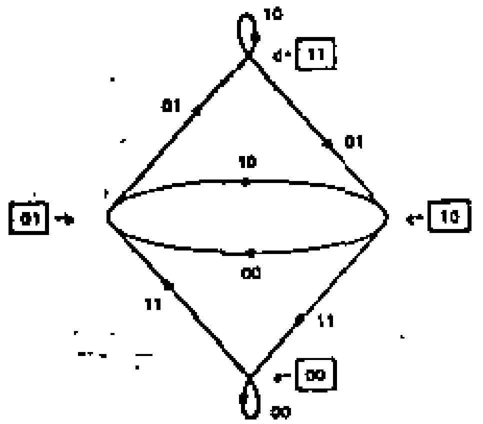

Fig. 4. State-diagram representation for coder of Fig. 1.

identically labeled pairs of nodes. The reason for this is obvious from examination of the encoder. As the fourth input bit enters the coder at the right, the first data bit falls off on the left end and no longer influences the output code symbols. Consequently, the data sequences $100xy\cdots$ and $000xy\cdots$ generate the same code symbols after the third branch and, as is shown in the tree diagram, both nodes labeled a can be joined together.

This leads to redrawing the tree diagram as shown in Fig. 3. This has been called a trellis diagram [12], since a trellis is a tree-like structure with remerging branches. We adopt the convention here that code branches produced by a "zero" input bit are shown as solid lines and code branches produced by a "one" input bit are shown dashed.

The completely repetitive structure of the trellis diagram suggests a further reduction in the representation of the code to the state diagram of Fig. 4. The "states" of the state diagram are labeled according to the nodes of the trellis diagram. However, since the states correspond merely to the last two input bits to the coder we may use these bits to denote the nodes or states of this diagram.

We observe finally that the state diagram can be drawn directly by observing the finite-state machine properties of the encoder and particularly the fact that a four-state directed graph can be used to represent uniquely the input-output relation of the eight-state machine. For the nodes represent the previous two bits while the present bit is indicated by the transition branch; for example, if the encoder (machine) contains 011, this is represented in the diagram by the transition from state $b = 01$ to state $d = 11$ and the corresponding branch indicates the code symbol outputs 01.

III. Minimum Distance Decoder for Binary Symmetric Channel

On a BSC, errors which transform a channel code symbol 0 to 1 or 1 to 0 are assumed to occur independently from symbol to symbol with probability p. If all input (message) sequences are equally likely, the decoder which minimizes the overall error probability for any code, block or convolutional, is one which examines the error-corrupted received sequence $y_1y_2\cdots y_j\cdots$ and chooses the data sequence corresponding to the transmitted code sequence $x_1x_2\cdots x_j\cdots$, which is closest to the received sequence in the sense of Hamming distance; that is, the transmitted sequence which differs from the received sequence in the minimum number of symbols.

Referring first to the tree diagram, this implies that we should choose that path in the tree whose code sequence differs in the minimum number of symbols from the received sequence. However, recognizing that the transmitted code branches remerge continually, we may equally limit our choice to the possible paths in the trellis diagram of Fig. 3. Examination of this diagram indicates that it is unnecessary to consider the entire received sequence (which conceivably could be thousands or millions of symbols in length) at one time in deciding upon the most likely (minimum distance) transmitted sequence. In particular, immediately after the third branch we may determine which of the two paths leading to node or state a is more likely to have been sent. For example, if 010001 is received, it is clear that this is at distance 2 from 000000 while it is at distance 3 from 111011 and consequently we may exclude the lower path into node a. For, no matter what the subsequent received symbols will be, they will effect the distances only over subsequent branches after these two paths have remerged and consequently in exactly the same way. The same can be said for pairs of paths merging at the other three nodes after the third branch. We shall refer to the minimum distance path of the two paths merging at a given node as the "survivor." Thus it is necessary only to remember which was the minimum distance path from the received sequence (or survivor) at each node, as well as the value of that minimum distance. This is necessary because at the next node level we must compare the two branches merging at each node level, which were survivors at the previous level for different nodes; e.g., the comparison at node a after the fourth branch is among the survivors of comparisons at nodes a and c after the third branch. For example, if the received sequence over the first four branches is 01000111, the survivor at the third node level for node a is 000000 with distance 2 and at node c it is 110101, also with distance 2. In going from the third node level to the fourth the received sequence agrees precisely with the survivor from c but has distance 2 from the survivor from a. Hence the survivor at node a of the fourth level is the data sequence 1100 which produced the code sequence 11010111 which is at (minimum) distance 2 from the received sequence.

In this way we may proceed through the received sequence and at each step for each state preserve one surviving path and its distance from the received sequence, which is more generally called *metric*. The only difficulty which may arise is the possibility that in a given comparison between merging paths, the distances or metrics are identical. Then we may simply flip a coin as is done for block codewords at equal distances from the received sequence. For even if we preserved both of the equally valid contenders, further received symbols would affect both metrics in exactly the same way and thus not further influence our choice.

This decoding algorithm was first proposed by Viterbi [9] in the more general context of arbitrary memoryless channels. Another description of the algorithm can be obtained from the state-diagram representation of Fig. 4. Suppose we sought that path around the directed state diagram, arriving at node a after the kth transition, whose code symbols are at a minimum distance from the received sequence. But clearly this minimum distance path to node a at time k can be only one of two candidates: the minimum distance path to node a at time $k-1$ and the minimum distance path to node c at time $k-1$. The comparison is performed by adding the new distance accumulated in the kth transition by each of these paths to their minimum distances (metrics) at time $k-1$.

It appears thus that the state diagram also represents a system diagram for this decoder. With each node or state we associate a storage register which remembers the minimum distance path into the state after each transition as well as a metric register which remembers its (minimum) distance from the received sequence. Furthermore, comparisons are made at each step between the two paths which lead into each node. Thus four comparators must also be provided.

There remains only the question of truncating the algorithm and ultimately deciding on one path rather than four. This is easily done by forcing the last two input bits to the coder to be 00. Then the final state of the code must be $a = 00$ and consequently the ultimate survivor is the survivor at node a, after the insertion into the coder of the two dummy zeros and transmission of the corresponding four code symbols. In terms of the trellis diagram this means that the number of states is reduced from four to two by the insertion of the first zero and to a single state by the insertion of the second. The diagram is thus truncated in the same way as it was begun.

We shall proceed to generalize these code representations and optimal decoding algorithm to general convolutional codes and arbitrary memoryless channels, including the Gaussian channel, in Sections V and VI. However, first we shall exploit the state diagram further to determine the relative distance properties of binary convolutional codes.

IV. Distance Properties of Convolutional Codes

We continue to pursue the example of Fig. 1 for the sake of clarity; in the next section we shall easily generalize results. It is well known that convolutional codes are group codes. Thus there is no loss in generality in computing the distance from the all zeros codeword to all the other codewords, for this set of distances is the same as the set of distances from any specific codeword to all the others.

For this purpose we may again use either the trellis diagram or the state diagram. We first of all redraw the trellis diagram in Fig. 5 labeling the branches according to their distances from the all zeros path. Now consider all the paths that merge with the all zeros for the first time at some arbitrary node j.

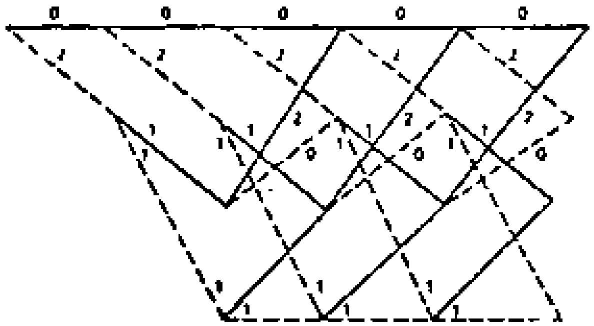

Fig. 5. Trellis diagram labeled with distances from all zeros path.

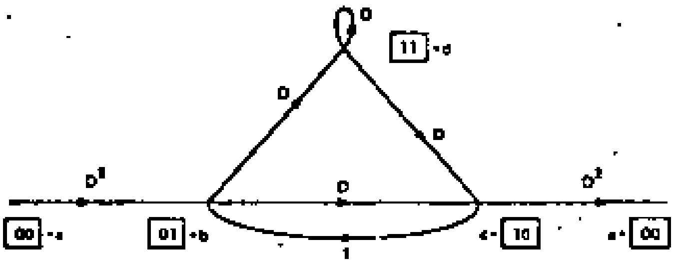

Fig. 6. State diagram labeled according to distance from all zeros path.

It is seen from the diagram that of these paths there will be just one path at distance 5 from the all zeros path and this diverged from it three branches back. Similarly there are two at distance 6 from it, one which diverged 4 branches back and the other which diverged 5 branches back, and so forth. We note also that the input bits for distance 5 path are $00 \cdots 0100$ and thus differ in only one input bit from the all zeros, while the distance 6 paths are $00 \cdots 01100$ and $00 \cdots 010100$ and thus each differs in 2 input bits from the all zeros path. The minimum distance, sometimes called the minimum "free" distance, among all paths is thus seen to be 5. This implies that any pair of channel errors can be corrected, for two errors will cause the received sequence to be at distance 2 from the transmitted (correct) sequence but it will be at least at distance 3 from any other possible code sequence. It appears that with enough patience the distance of all paths from the all zeros (or any arbitrary) path can be so determined from the trellis diagram.

However, by examining instead the state diagram we can readily obtain a closed form expression whose expansion yields directly and effortlessly all the distance information. We begin by labeling the branches of the state diagram of Fig. 4 either D^2, D, or $D^0 = 1$, where the exponent corresponds to the distance of the particular branch from the corresponding branch of the all-zeros path. Also we split open the node $a = 00$, since circulation around this self-loop simply corresponds to branches of the all zeros path whose distance from itself is obviously zero. The result is Fig. 6. Now as is clear from examination of the trellis diagram, every path which arrives at state $a = 00$ at node level j, must have at some previous node level (possibly the first) originated at this same state $a = 00$. All such paths can be traced on the modified state diagram. Adding branch exponents we see that path $a\ b\ c\ a$ is at distance 5 from the correct path, paths $a\ b\ d\ c\ a$ and $a\ b\ c\ b\ c\ a$ are both at distance 6, and so forth, for the *generating functions* of the output sequence weights of these paths are D^5 and D^6, respectively.[3]

Now we may evaluate the generating function of all paths merging with the all zeros at the jth node level simply by evaluating the generating function of all the weights of the output sequences of the finite-state machine.[4] The result in this case is

$$T(D) = \frac{D^5}{1 - 2D}$$

$$= D^5 + 2D^6 + 4D^7 + \cdots + 2^k D^{k+5} + \cdots. \quad (1)$$

This verifies our previous observation and in fact shows that among the paths which merge with the all zeros at a given node there are 2^k paths at distance $k + 5$ from the all zeros.

Of course, (1) holds for an infinitely long code sequence; if we are dealing with the jth node level, we must truncate the series at some point. This is most easily done by considering the additional information indicated in the modified state diagram of Fig. 7.

The L terms will be used to determine the length of a given path; since each branch has an L, the exponent of the L factor will be augmented by one every time a branch is passed through. The N term is included only if that branch transition was caused by an input data "one," corresponding to a dotted branch in the trellis diagram. The generating function of this augmented state diagram is then

$$T(D, L, N) = \frac{D^5 L^3 N}{1 - DL(1 + L)N}$$

$$= D^5L^3N + D^6L^4(1 + L)N^2 + D^7L^5(1 + L)^2N^3 + \cdots + D^{k+5}L^{k+3}(1 + L)^k N^{k+1} + \cdots. \quad (2)$$

Thus we have verified that of the two distance 6 paths one is of length 4 and the other is of length 5 and both differ in 2 input bits from the all zeros.[4] Also, of the distance 7 paths, one is of length 5, two are of length 6, and one is of length 7; all four paths correspond to input sequences with three ones. If we are interested in the jth node level, clearly we should truncate the series such that no terms of power greater than L^j are included.

We have thus fully determined the properties of all paths in the convolutional code. This will be useful later in evaluating error probability performance of codes used over arbitrary memoryless channels.

[3] Alternatively, this can be regarded as the transfer function of the diagram regarded as a signal flow graph.

[4] That is, if the all zeros was the correct path and the noise causes us to choose one of the incorrect paths, two bit errors will be made.

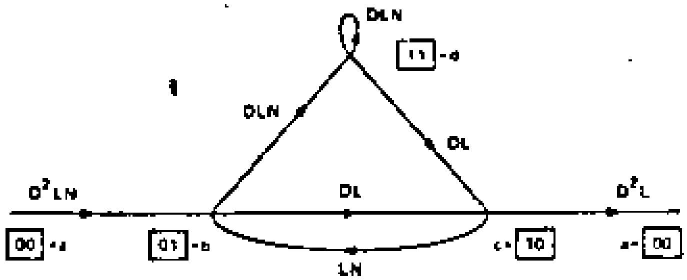

Fig. 7. State diagram labeled according to distance, length, and number of input ones.

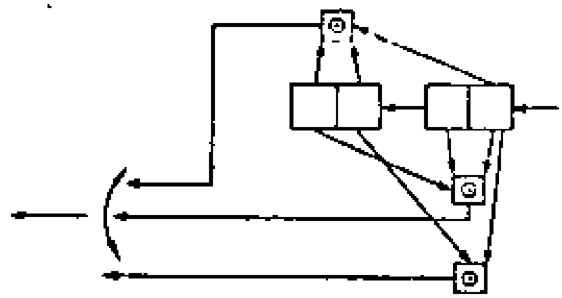

Fig. 8. Coder for $K = 2$, $b = 2$, $n = 3$, and $R = 2/3$.

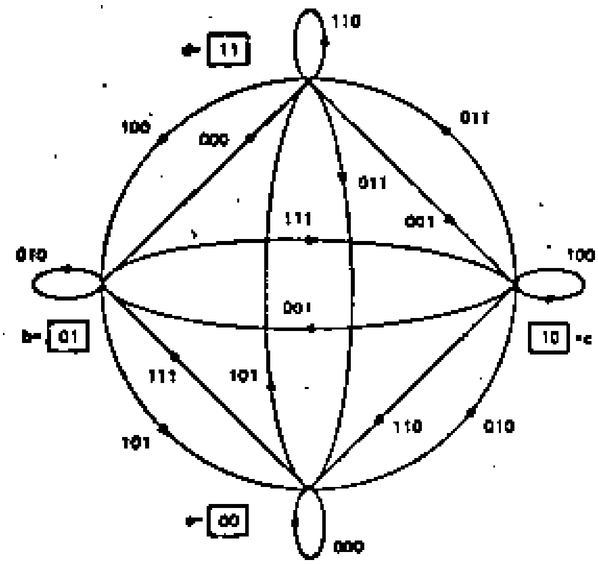

Fig. 9. State diagram for code of Fig. 8.

V. Generalization to Arbitrary Convolutional Codes

The generalization of these techniques to arbitrary binary-tree ($b = 1$) convolutional codes is immediate. That is, a coder with a K-stage shift register and n mod-2 adders will produce a trellis or state diagram with 2^{K-1} nodes or states and each branch will contain n code symbols. The rate of this code is then

$$R = \frac{1}{n} \text{ bits/code symbol.}$$

The example pursued in the previous sections had rate $R = 1/2$. The primary characteristic of the binary-tree codes is that only two branches exit from and enter each node.

If rates other than $1/n$ are desired we must make $b > 1$, where b is the number of bits shifted into the register at one time. An example for $K = 2$, $b = 2$, $n = 3$, and consequently rate $R = 2/3$ is shown in Fig. 8 and its state diagram is shown in Fig. 9. It differs from the binary-tree codes only in that each node is connected to four other nodes, and for general b it will be connected to 2^b nodes. Still all the preceding techniques including the trellis and state-diagram generating function analysis are still applicable. It must be noted, however, that the minimum distance decoder must make comparisons among all the paths entering each node at each level of the trellis and select one survivor out of four (or out of 2^b in general).

VI. Generalization of Optimal Decoder to Arbitrary Memoryless Channels

Fig. 10 exhibits a communication system employing a convolutional code. The convolutional encoder is precisely the device studied in the preceding sections. The data sequence is generally binary ($a_i = 0$ or 1) and the code sequence is divided into subsequences where $\mathbf{x}_j$ represents the n code symbols generated just after the input bit a_j enters the coder: that is, the symbols of the jth branch. In terms of the example of Fig. 1, $a_3 = 1$ and $\mathbf{x}_3 = 01$. The channel output or received sequence is similarly denoted. $\mathbf{y}_j$ represents the n symbols received when the n code symbols of $\mathbf{x}_j$ were transmitted. This model includes the BSC wherein the $\mathbf{y}_j$ are binary n vectors each of whose symbols differs from the corresponding symbol of $\mathbf{x}_j$ with probability p and is identical to it with probability $1 - p$.

For completely general channels it is readily shown [6], [14] that if all input data sequences are equally likely, the decoder which minimizes the error probability is one which compares the conditional probabilities, also called likelihood functions, $P(\mathbf{y} \mid \mathbf{x}^{(m)})$, where $\mathbf{y}$ is the overall received sequence and $\mathbf{x}^{(m)}$ is one of the possible transmitted sequences, and decides in favor of the maximum. This is called a maximum likelihood decoder. The likelihood functions are given or computed from the specifications of the channel. Generally it is more convenient to compare the quantities $\log P(\mathbf{y} \mid \mathbf{x}^{(m)})$ called the log-likelihood functions and the result is unaltered, since the logarithm is a monotonic function of its (always positive) argument.

To illustrate, let us consider again the BSC. Here each transmitted symbol is altered with probability $p < 1/2$. Now suppose we have received a particular N-dimensional binary sequence $\mathbf{y}$ and are considering a possible transmitted N-dimensional code sequence $\mathbf{x}^{(m)}$ which differs in d_m symbols from $\mathbf{y}$ (that is, the Hamming distance between $\mathbf{x}^{(m)}$ and $\mathbf{y}$ is d_m). Then since the channel is memoryless (i.e., it affects each symbol independently of all the others), the probability

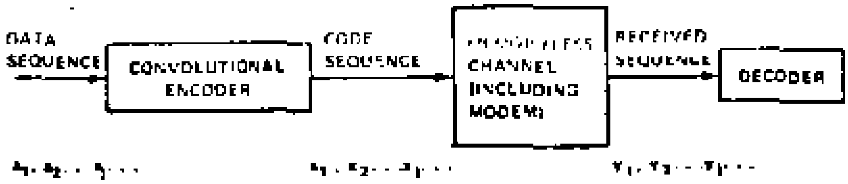

Fig. 10. Communication system employing convolutional codes.

that this $\mathbf{x}^{(m)}$ was transformed to the specific received $\mathbf{y}$ at distance d_m from it is

$$P(\mathbf{y} \mid \mathbf{x}^{(m)}) = p^{d_m}(1-p)^{N-d_m}$$

and the log-likelihood function is thus

$$\log P(\mathbf{y} \mid \mathbf{x}^{(m)}) = -d_m \log\left((1-p)/p\right) + N \log(1-p)$$

Now if we compute this quantity for each possible transmitted sequence, it is clear that the second term is constant in each case. Furthermore, since we may assume $p < 1/2$ (otherwise the role of 0 and 1 is simply interchanged at the receiver), we may express this as

$$\log P(\mathbf{y} \mid \mathbf{x}^{(m)}) = -\alpha d_m - \beta \qquad (3)$$

where α and β are positive constants and d_m is the (positive) distance. Consequently, it is clear that maximizing the log-likelihood function is equivalent to minimizing the Hamming distance d_m. Thus for the BSC to minimize the error probability we should choose that code sequence at minimum distance from the received sequence, as we have indicated and done in preceding sections.

We now consider a more physical practical channel: the AWGN channel with biphase[4] phase-shift keying (PSK) modulation. The modulator and optimum demodulator (correlator or integrate-and dump filter) for this channel are shown in Fig. 11.

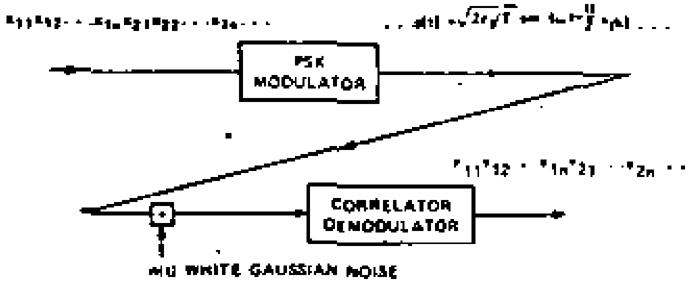

Fig. 11. Modem for additive white Gaussian noise PSK modulated memoryless channel.

We use the notation that x_{jk} is the kth code symbol for the jth branch. Each binary symbol (which we take here for convenience to be ± 1) modulates the carrier by $\pm\Pi/2$ radians for T seconds. The transmission rate is, therefore, $1/T$ symbols/second or $b/nT = R/T$ bit/s. The function ϵ_s is the energy transmitted for each symbol. The energy per bit is, therefore $\epsilon_b = \epsilon_s/R$. The white Gaussian noise is a zero-mean random process of one-sided spectral density N_0 W/Hz, which affects each symbol independently. It then follows directly that the channel output symbol y_{jk} is a Gaussian random variable whose mean is $\sqrt{\epsilon_s}\,x_{jk}$ (i.e., $+\sqrt{\epsilon_s}$ if $x_{jk} = 1$ and $-\sqrt{\epsilon_s}$ if $x_{jk} = -1$) and whose variance is $N_0/2$. Thus the conditional probability density (or likelihood) function of y_{jk} given x_{jk} is

$$p(y_{jk} \mid x_{jk}) = \frac{\exp\left[-(y_{jk} - \sqrt{\epsilon_s}\,x_{jk})^2/N_0\right]}{\sqrt{\Pi N_0}}. \qquad (4)$$

The likelihood function for the jth branch of a particular code path $\mathbf{x}_j^{(m)}$ is

$$p(\mathbf{y}_j \mid \mathbf{x}_j^{(m)}) = \prod_{k=1}^{n} p(y_{jk} \mid x_{jk}^{(m)})$$

since each symbol is affected independently by the white Gaussian noise, and thus the log-likelihood function for the jth branch is

$$\begin{aligned}
\ln p(\mathbf{y}_j \mid \mathbf{x}_j^{(m)}) &= \sum_{k=1}^{n} \ln p(y_{jk} \mid x_{jk}^{(m)}) \\
&= -\frac{1}{N_0}\sum_{k=1}^{n}\left[y_{jk} - \sqrt{\epsilon_s}\,x_{jk}^{(m)}\right]^2 - \frac{1}{2}\ln\frac{\Pi}{N_0} \\
&= \frac{2\sqrt{\epsilon_s}}{N_0}\sum_{k=1}^{n} y_{jk}x_{jk}^{(m)} - \frac{\epsilon_s}{N_0}\sum_{k=1}^{n}\left[x_{jk}^{(m)}\right]^2 \\
&\quad - \frac{1}{N_0}\sum_{k=1}^{n} y_{jk}^2 - \frac{1}{2}\ln\frac{\Pi}{N_0} \\
&= C\sum_{k=1}^{n} y_{jk}x_{jk}^{(m)} - D
\end{aligned} \qquad (5)$$

where C and D are independent of m, and we have used the fact that $|x_{jk}^{(m)}|^2 = 1$. Similarly, the log-likelihood[5] function for any path is the sum of the log-likelihood functions for each of its branches.

We have thus shown that the maximum likelihood decoder for the memoryless AWGN biphase (or quadriphase) modulated channel is one which forms the inner product between the received (real number) sequence and the code sequence (consisting of ± 1) and chooses the path corresponding to the greatest. Thus the *metric* for this channel is the inner product (5) as contrasted with the distance[6] metric used for the BSC.

[4] The results are the same for quadriphase PSK with coherent reception. The analysis proceeds in the same way, if we treat quadriphase PSK as two parallel independent biphase PSK channels.

[5] We have used the natural logarithm here, but obviously a change of base results merely in a scale factor.

[6] Actually it is easily shown that maximizing an inner product is equivalent to minimizing the Euclidean distance between the corresponding vectors.

For convolutional codes the structure of the code paths was described in Sections II–V. In Section III the optimum decoder was derived for the BSC. It now becomes clear that if we substitute the inner product metric $\Sigma y_{hj} x_{hj}^{(m)}$ for the distance metric $\Sigma d_{hj}^{(m)}$, used for the BSC, all the arguments used in Section III for the latter apply equally to this Gaussian channel. In particular the optimum decoder has a block diagram represented by the code state diagram. At step j the stored metric for each state (which is the maximum of the metrics of all the paths leading to this state at this time) is augmented by the branch metrics for branches emanating from this state. The comparisons are performed among all pairs of (or in general sets of 2^b) branches entering each state and the *maxima* are selected as the new most likely paths. The history (input data) of each new survivor must again be stored and the decoder is now ready for step $j + 1$.

Clearly, this argument generalizes to any memoryless channel and we must simply use the appropriate metric $\ln P(\mathbf{y} \mid \mathbf{x}^{(m)})$, which may always be determined from the statistical description of the channel. This includes, among others, AWGN channels employing other forms of modulation.[7]

In the next section, we apply the analysis of convolutional code distance properties of Section IV to determine the error probabilities of specific codes on more general memoryless channels.

VII. Performance of Convolutional Codes on Memoryless Channels

In Section IV we analyzed the distance properties of convolutional codes employing a state-diagram generating function technique. We now extend this approach to obtain tight upper bounds on the error probability of such codes. We shall consider the BSC, the AWGN channel and more general memoryless channels, in that order. We shall obtain both the first-event error probability, which is the probability that the correct path is excluded (not a survivor) for the first time at the jth step, and the bit error probability which is the expected ratio of bit errors to total number of bits transmitted.

A. Binary Symmetric Channel

The first-event error probability is readily obtained from the generating function $T(D)$ [(5) for the code of Fig. 1, which we shall again pursue for demonstrative purposes]. We may assume, without loss of generality, since we are dealing with group codes, that the all zeros path was transmitted. Then a first-event error is made at the jth step if this path is excluded by selecting another path merging with the all zeros at node a at the jth level.

Now suppose that the previous-level survivors were such that the path compared with the all zeros at step j is the path whose data sequence is $00 \cdots 0100$ corresponding to nodes $a \cdots a\ a\ b\ c\ a$ (see Fig. 4). This differs from the correct (all zeros) path in five symbols. Consequently an error will be made in this comparison if the BSC caused three or more errors in these particular five symbols. Hence the probability of an error in this specific comparison is

$$P_5 = \sum_{e=3}^{5} \binom{5}{e} p^e (1-p)^{5-e}. \tag{6}$$

On the other hand, there is no assurance that this particular distance five path will have previously survived so as to be compared with the correct path at the jth step. If either of the distance 6 paths were compared instead, then four or more errors in the six different symbols will definitely cause an error in the survivor decision, while three errors will cause a tie which, if resolved by coin flipping, will result in an error only half the time. Then the probability if this comparison is made is

$$P_6 = \frac{1}{2}\binom{6}{3} p^3 (1-p)^3 + \sum_{e=4}^{6} \binom{6}{e} p^e (1-p)^{6-e}. \tag{7}$$

Similarly, if the previously surviving paths were such that a distance d path is compared with the correct path at the jth step, the resulting error probability is

$$P_k = \begin{cases} \displaystyle\sum_{e=(k+1)/2}^{k} \binom{k}{e} p^e (1-p)^{k-e}, & k \text{ odd} \\ \displaystyle\frac{1}{2}\binom{k}{k/2} p^{k/2} (1-p)^{k/2} + \sum_{e=k/2+1}^{k} \binom{k}{e} p^e (1-p)^{k-e}, & k \text{ even.} \end{cases} \tag{8}$$

Now at step j, since there is no simple way of determining previous survivors, we may overbound the probability of a first-event error by the sum of the error probabilities for all possible paths which merge with the correct path at this point. Note this *union bound* is indeed an upper bound because two or more such paths may both have distance closer to the received sequence than the correct path (even though only one has survived to this point) and thus the events are not disjoint. For the example with generating function (1) it follows that the first-event error probability[8] is bounded by

$$P_E < P_5 + 2P_6 + 4P_7 + \cdots + 2^k P_{k+5} + \cdots \tag{9}$$

where P_k is given by (8).

In Section VII-C it will be shown that (8) can be upper bounded by (see (39)).

$$P_k < 2^k p^{k/2} (1-p)^{k/2}. \tag{10}$$

Using this, the first-event error probability bound (9)

[7] Although more elaborate modulators, such as multiple FSK or multiphase modulators, might be employed, Jacobs [11] has shown that the most effective as well as the simplest system for wideband space and satellite channels is the binary PSK modulator considered in the example of this section. We note again that the performance of quadriphase modulation is the same as for biphase modulation, when both are coherently demodulated.

[8] We are ignoring the finite length of the path, but the expression is still valid since it is an upper bound.

can be more loosely bounded by

$$P_E < \sum_{k=5}^{\infty} 2^{k-5} 2^k [p(1-p)]^{k/2} = \frac{[2\sqrt{p(1-p)}]^5}{1-4\sqrt{p(1-p)}} = T(D)\big|_{D=2\sqrt{p(1-p)}} \tag{11}$$

where $T(D)$ is just the generating function of (1).

It follows easily that for a general binary-tree ($b = 1$) convolutional code with generating function

$$T(D) = \sum_{k=d}^{\infty} a_k D^k \tag{12}$$

the first-event error probability is bounded by the generalization of (9),

$$P_E < \sum_{k=d}^{\infty} a_k P_k \tag{13}$$

where P_k is given by (8) and more loosely upper bounded by the generalization of (11)

$$P_E < T(D)\big|_{D=2\sqrt{p(1-p)}}. \tag{14}$$

Whenever a decision error occurs, one or more bits will be incorrectly decoded. Specifically, those bits in which the path selected differs from the correct path will be incorrect. If only one error were ever made in decoding an arbitrary long code path, the number of bits in error in this incorrect path could easily be obtained from the augmented generating function $T(D, N)$ (such as given by (2) with factors in L deleted). For the exponents of the N factors indicate the number of bit errors for the given incorrect path arriving at node a at the jth level.

After the first error has been made, the incorrect paths no longer will be compared with a path which is overall correct, but rather with a path which has diverged from the correct path over some span of branches (see Fig. 12). If the correct path $\mathbf{x}$ has been excluded by a decision error at step j in favor of path $\mathbf{x}'$, the decision at step $j+1$ will be between $\mathbf{x}'$ and $\mathbf{x}''$. Now the (first-event) error probability of (13) or (14) is for a comparison, at any step, between path $\mathbf{x}$ and any other path merging with it at that step, including path $\mathbf{x}''$ in this case. However, since the metric[8] for path $\mathbf{x}'$ is greater than the metric for $\mathbf{x}$, for on this basis the correct path was excluded at step j, the probability that path $\mathbf{x}''$ metric exceeds path $\mathbf{x}'$ metric at step $j+1$ is less than the probability that path $\mathbf{x}''$ exceeds the (correct) path $\mathbf{x}$ metric at this point. Consequently, the probability of a new incorrect path being selected after a previous error has occurred is upper bounded by the first-event error probability at that step.

Moreover, when a second error follows closely after a first error, it often occurs (as in Fig. 12) that the erroneous bit(s) of path $\mathbf{x}''$ overlap the erroneous bit(s) of path $\mathbf{x}'$. With this in mind, we now show that for a

[8] Negative distance from the received sequence for the BSC, but clearly this argument generalizes to any memoryless channel.

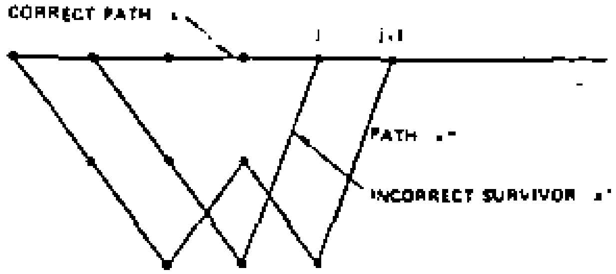

Fig. 12. Example of decoding decision after initial error has occurred.

binary-tree code if we weight each term of the first-event error probability bound at any step by the number of erroneous bits for each possible erroneous path merging with the correct path at that node level, we upper bound the bit error probability. For, a given step decision corresponds to decoder action on one more bit of the transmitted data sequence; the first-event error probability union bound with each term weighted by the corresponding number of bit errors is an upper bound on the expected number of bit errors caused by this action. Summing the expected number of bit errors over L steps, which as was just shown may result in overestimating through double counting, gives an upper bound on the expected number of bit errors in L branches for arbitrary L. But since the upper bound on expected number of bit errors is the same at each step, it follows, upon dividing the sum of L equal terms by L, that this expected number of bit errors per step is just the bit error probability P_B for a binary-tree code ($b = 1$). If $b > 1$, then we must divide this expression by b, the number of bits encoded and decoded per step.

To illustrate the calculation of P_B for a convolutional code, let us consider again the example of Fig. 1. Its transfer function in D and N is obtained from (2), letting $L = 1$, since we are not now interested in the lengths of incorrect paths, to be

$$T(D, N) = \frac{D^5 N}{1-2DN} = D^5 N + 2D^6 N^2 + \cdots + 2^k D^{k+5} N^{k+1} + \cdots. \tag{15}$$

The exponents of the factors in N in each term determine the number of bit errors for the path(s) corresponding to that term. Since $T(D) = T(D, N)|_{N=1}$ yields the first-event error probability P_E, each of whose terms must be weighted by the exponent of N to obtain P_B, it follows that we should first differentiate $T(D, N)$ at $N = 1$ to obtain

$$\left.\frac{dT(D, N)}{dN}\right|_{N=1} = D^5 + 2\cdot 2D^6 + 3\cdot 4D^7 + \cdots + (k+1)2^k D^{k+5} + \cdots = \frac{D^5}{(1-2D)^2}. \tag{16}$$

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Then from this we obtain, as in (9), that for the BSC

$$P_B < P_5 + 2\cdot 2P_6 + 3\cdot 4P_7 + \cdots + (k+1)2^k P_{k+5} + \cdots \tag{17}$$

where P_k is given by (8).

If for P_k we use the upper bound (10) we obtain the weaker but simpler bound

$$\begin{aligned} P_B &< \sum_{k=5}^{\infty} (k-4)2^{k-5}[4p(1-p)]^{k/2} \\ &= \left.\frac{dT(D,N)}{dN}\right|_{N=1,\,D=2\sqrt{p(1-p)}} \\ &= \frac{[2\sqrt{p(1-p)}]^5}{[1-4\sqrt{p(1-p)}]^2}. \end{aligned} \tag{18}$$

More generally for any binary-tree ($b = 1$) code used on the BSC if

$$\left.\frac{dT(D,N)}{dN}\right|_{N=1} = \sum_{k=d}^{\infty} c_k D^k \tag{19}$$

then corresponding to (17)

$$P_B < \sum_{k=d}^{\infty} c_k P_k \tag{20}$$

and corresponding to (18) we have the weaker bound

$$P_B < \left.\frac{dT(D,N)}{dN}\right|_{N=1,\,D=2\sqrt{p(1-p)}}. \tag{21}$$

For a nonbinary-tree code ($b \neq 1$), all these expressions must be divided by b.

The results of (14) and (18) will be extended to more general memoryless channels, but first we shall consider one more specific channel of particular interest.

B. AWGN Biphase-Modulated Channel

As was shown in Section VI the decoder for this channel operates in exactly the same way as for the BSC, except that instead of Hamming distance it uses the metric

$$\sum_i \sum_{j=1}^{n} x_{ij} y_{ij}$$

where $x_{ij} = \pm 1$ are the transmitted code symbols, y_{ij} the corresponding received (demodulated) symbols, and j runs over the n symbols of each branch while i runs over all the branches in a particular path. Hence, to analyze its performance we may proceed exactly as in Section VII-A except that the appropriate pairwise-decision errors P_k must be substituted for those of (6) to (8).

As before we assume, without loss of generality, that the correct (transmitted) path $\mathbf{x}$ has $x_{ij} = +1$ for all i and j (corresponding to the all zeros if the input symbols were 0 and 1). Let us consider an incorrect path $\mathbf{x}'$ merging with the correct path at a particular step, which has k negative symbols ($x_{ij}' = -1$) and the remainder positive. Such a path may be incorrectly chosen only if it has a higher metric than the correct path, i.e.,

$$\sum_i \sum_{j=1}^{n} x_{ij}' y_{ij} \ge \sum_i \sum_{j=1}^{n} x_{ij} y_{ij}$$

or

$$\sum_i \sum_{j=1}^{n} (x_{ij}' - x_{ij}) y_{ij} \ge 0$$

where i runs over all branches in the two paths. But since, as we have assumed, the paths $\mathbf{x}$ and $\mathbf{x}'$ differ in exactly k symbols, wherein $x_{ij} = 1$ and $x_{ij}' = -1$, the pairwise error probability is just

$$\begin{aligned} P_k &= \Pr\left\{\sum_i \sum_{j=1}^{n} (x_{ij}' - x_{ij}) y_{ij} \ge 0\right\} \\ &= \Pr\left\{\sum_{r=1}^{k} (x_r' - x_r) y_r \ge 0\right\} \\ &= \Pr\left\{-2\sum_{r=1}^{k} y_r \ge 0\right\} \\ &= \Pr\left\{\sum_{r=1}^{k} y_r \le 0\right\} \end{aligned} \tag{22}$$

where r runs over the k symbols wherein the two paths differ. Now it was shown in Section VI that the y_{ij} are independent Gaussian random variables of variance $N_0/2$ and mean $\sqrt{\epsilon_s}\,x_{ij}$, where x_{ij} is the actually transmitted code symbol. Since we are assuming that the (correct) transmitted path has $x_{ij} = +1$ for all i and j, it follows that y_{ij} or y_r has mean $\sqrt{\epsilon_s}$ and variance $N_0/2$. Therefore, since the k variables y_r are independent and Gaussian, the sum $Z = \sum_{r=1}^{k} y_r$ is also Gaussian with mean $k\sqrt{\epsilon_s}$ and variance $kN_0/2$.

Consequently,

$$\begin{aligned} P_k &= \Pr(Z < 0) = \int_{-\infty}^{0} \frac{\exp\left(-(Z - k\sqrt{\epsilon_s})^2/kN_0\right)}{\sqrt{\pi k N_0}}\, dZ \\ &= \int_{\sqrt{2k\epsilon_s/N_0}}^{\infty} \left[\frac{\exp(-x^2/2)}{\sqrt{2\pi}}\right] dx \triangleq \operatorname{erfc}\sqrt{\frac{2k\epsilon_s}{N_0}}. \end{aligned} \tag{23}$$

We recall from Section VI that ϵ_s is the symbol energy, which is related to the bit energy by $\epsilon_s = R\epsilon_b$, where $R = b/n$. The bound on P_E then follows exactly as in Section VII-A and we obtain the same general bound as (13)

$$P_E < \sum_{k=d}^{\infty} a_k P_k \tag{24}$$

where a_k are the coefficients of

$$T(D) = \sum_{k=d}^{\infty} a_k D^k \tag{25}$$

and where d is the minimum distance between any two paths in the code. We may simplify this procedure considerably while loosening the bound only slightly for this channel by observing that for $x \ge 0$, $y \ge 0$,

$$\operatorname{erfc}\sqrt{x+y} \le \exp\left(\frac{-y}{2}\right) \operatorname{erfc}\sqrt{x}. \tag{26}$$

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III. The Communication System

3.1 Source Coding

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