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JAPAN INTERNATIONAL COOPERATION AGENCY

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〔VOLUME Ⅲ〕

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JAPAN INTERNATIONAL COOPERATION AGENCY

VOLUME III

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XVII ANNEX

[II] ENGINEERING STANDARD FOR RADIO FREQUENCY ASSIGNMENT — Page

[II]-1 Introduction

Radio frequency assignment is very important for system design in this project too and this engineering standard sets out technical requirements to be met or based on upon assigning new frequencies. The applications of radio frequencies can be classified by their operating modes into fixed service, mobile service, and satellite communication service. This project is intended mainly for fixed service, so that description given hereunder is limited to fixed service.

[II]-2 Frequency Assignment

It is recommended to effect frequency assignment on the basis of the frequency allocation requirements set out in Appended Articles 3 and 5 of the Radio Regulation and the allocation principles of BOC and in consideration of the following items.

(1) Channel capacity below 300 MHz band be less than 6 channels
Channel capacity in 300 ∿ 470 MHz bands be less than 24 channels.

(2) In order to realize effective use of frequencies, as much alternative use of frequencies as practicable is requisite.

[II]-3 Site Location

1. Propagation Course

In view of the radio route and profile, clearance should be secured between the first Fresnel zone and obstructions no building should be constructed so high to become obstruction in the propagation course, and no such disturbance to propagation as close reflection should be involved even if the coefficient of effective radius of the earth, K, varies to 0.8. Use of the over-the-horizon system (OH) should be limited to cases when the number of channels is small even in remote places so as to prevent influence of the OH system to be other systems.

2. Interference

The presence/absence of the influence of interference to and from other stations should be examined. Details are described later in this brochure.

3. Location of Transmitter

Beam feeding, reflector, etc., should not be used in urban areas or areas congested with radio stations as much as possible. Station building should be selected in consideration of environmental conditions such as power supply, roads, and mutual interference, ease of maintenance, etc.

4. Future Expansion

Extension of radio routes, branching, expansion of facilities, etc., should be planned in consideration of future expansion.

5. Repeating Methods

(1) Baseband repeating

This repeating method features ease of drop out and insertion but is inferior to the heterodyne method in distortion character and line quality since modulation and demodulation are repeated every time the signal is repeatered.

(2) Heterodyne repeating

This repeating method requires no modulation nor demodulation upon repeating and is suitable for long-distance transmission. In this repeating method, the signal level variation, distortion, and transmission quality characteristics are excellent.

(3) Direct repeating (RF amplification)

Although this repeating method allows no such drop out nor insertion as those of heterodyne repeating mentioned in (2), it allows, unlike heterodyne repeating, microwave frequencies to be sent and received without frequency conversion.
Accordingly, a reflector incorporating active elements may be employed.

[II]-4 Circuit Design
(Fixed service)

1. Design Procedure

System design of fixed service is usually made by the following procedure, and design data are filed in form of a system design manual.

	Item
Line demand	1 Location (Section) 2 Application 3 Channel capacity 4 Line quality 5 Reliability
System design	1 Line quality 2 Site location 3 Modulation method 4 Repeating method 5 Power system 6 Equipment in use 7 Antenna 8 Feeder 9 Reflector 10 Measures against fading 11 Measures against interference 12 Overall examination of system design
Installation design	1 Station building and steel tower 2 Equipment arrangement 3 Wiring 4 Auxiliary facilities

Is designed to meet requirements?

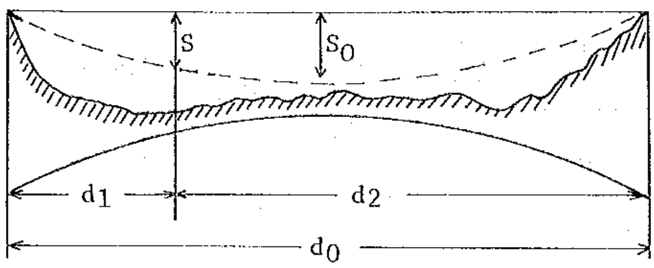
1 Frequency
2 Polarization
3 Power rating

2. Preparation of Course Profile

In order to roughly grasp the topographical condition of the route for proper site location, it is advantageous and efficient to employ a 1/200,000 map or a like. When preparing a profile of a course after general examination, draw a straight line on the proposed route on a map with an accuracy of more than 1/50,000, pick up heights respective points on the straight line above the sea level, and enter these heights on the profile chart to be prepared.

(1) Examination of first Fresnel zone

After preparing the profile, find the dip of the Fresnel zone by figure shown below.

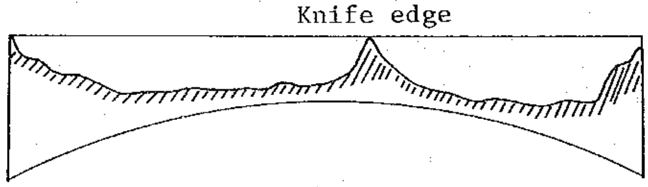


The dip of the first Fresnel zone, S, is given by

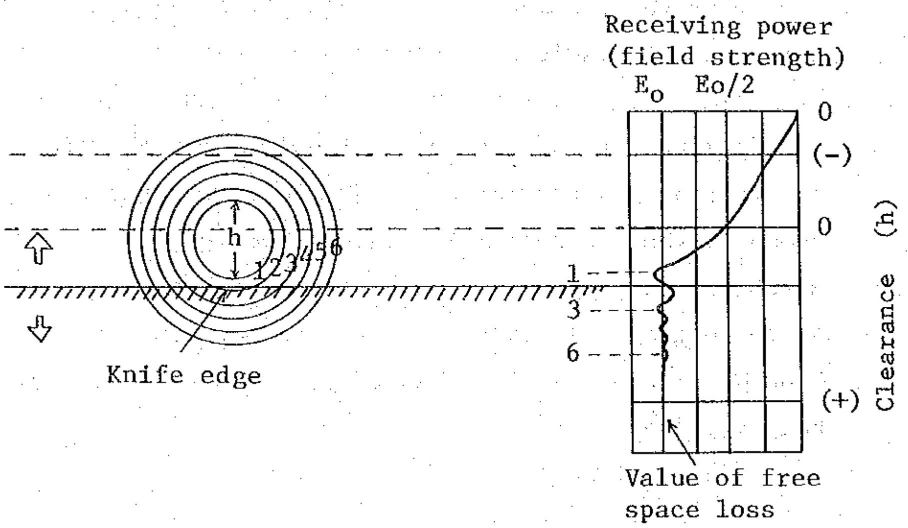
$$S_0(\mu) = 1.58\sqrt{\lambda(cm)\cdot d0(km)} \quad \text{.......... Center of the span}$$

$$S\ (\mu) = 3.16\sqrt{\lambda\ (cm)\ \frac{d1(km)\cdot d2(km)}{d0(km)}} \quad \text{..... At a desired point}$$

When there is an obstruction (knife edge) on the line of sight as shown below, a loss of Eo/2 or 6dB relative to the free space loss is applied.



The influence of obstructions within the Fresnel zone is shown in the following figure which presents receiving power variation for clearance variation. Here, the zone of receiving power deviation in the line-of-sight is called the Fresnel zone and zones in which the second and third deviations occur are called the second and third Fresnel zones.



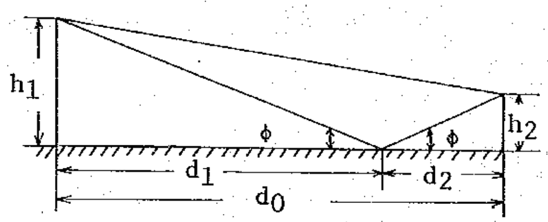
Note 1: Eo(v/m) is receiving power in free space.

Note 2: Numbers show the orders of Fresnel zone.

(2) Location of reflection point

The location of the reflection point is given by the following equations.

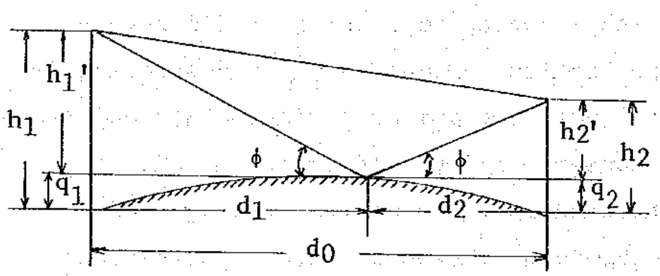
1) Flat ground



$$d_1 = \frac{h_1}{h_1 + h_2} d_0$$

$$d_2 = \frac{h_2}{h_1 + h_2} d_0$$

2) Spherical ground



$$d_1 = \frac{h_1'}{h_1' + h_2'} d_0$$

$$d_2 = \frac{h_2'}{h_1' + h_2'} d_0$$

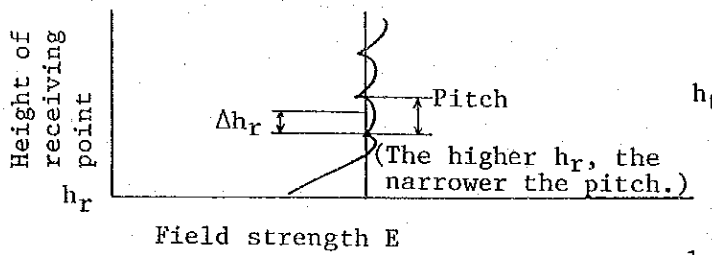
where

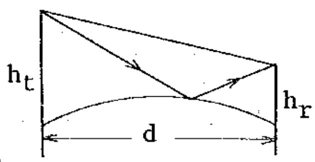
$$h_1' = h_1 - q_1$$

$$h_2' = h_2 - q_2$$

(3) Examine of height pattern

When the antenna height at the transmitting or receiving point is varied with the distance between the transmitting and receiving points being constant, the course difference between the direct wave and the wave reflected from the ground varies and receiving power variation due to interference appears as shown in the figure below. This is called height pattern and is used for determining the presence/absence of any reflecting wave from the ground and measuring the reflection coefficient of the ground. The higher the frequency, the smaller the pitch of the height pattern.





$$\Delta hr = \frac{1}{4} \frac{\lambda(m) \cdot d(m)}{ht(m)} \ (m)$$

Δhr: Half pitch of height pattern

3. Propagation Loss

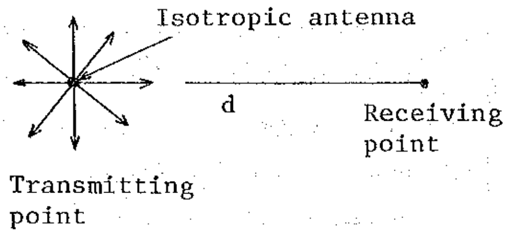
In radio wave propagation, it is necessary to consider, in addition to the free space loss, the influence of ground reflection, mountain diffraction loss, fading due to radio wave condition in the sky, radio wave absorption by rain and snow, etc.

(1) Free space loss

Free space is such a space where any of radio wave refraction, diffraction, reflection, absorption, and scattering is caused between the transmitting and receiving points and radio wave attenuation is caused only by diffusion.

In general, propagation design of radio waves above the VHF range is made on the basis of the use of isotropic antennas. The propagation loss to be caused when isotropic antennas are used as the transmitting and receiving antennas, is designated free space loss.

When the free space loss is Lo, the transmitting output P, and the receiving power W, then we have Lo = P/W, and since isotropic antennas are omnidirectional, power flux density at all points on the sphere of d in radius and the total power flux density is P. Then, power flux density Pu at the receiving point located at a distance of d from the transmitting point is given by



$$Pu = \frac{P}{4\pi d^2}.$$

Also, since the effective aperture area of the isotropic antenna on the receiving side, Ae, is $\lambda^2/4\pi$, receiving power W is given by

$$W = Pu \cdot Ae = \frac{P}{4\pi d^2} \cdot \frac{\lambda^2}{4\pi}$$

Hence, free space loss Lo is obtained by Lo = P/W = $(4\pi d/\lambda)^2$ and can be given in decibels as follows.

$$Lo(dB) = 122 + 20 \log d(km) - 20 \log \lambda(cm)$$

$$= 32.4 + 20 \log d(km) + 20 \log f(MHz)$$

(2) Propagation loss on flat or spherical ground

The antenna height as which the receiving power reaches the free space value in the above-mentioned height pattern figure varies in proportion with the wavelength, so that when a radio wave of below 1000 MHz is employed, consideration should be given to the loss due to the influence of the ground (including the sea surface). This applies when the loss thus obtained is larger than the free space loss. When the loss thus obtained is smaller than the free space loss, the propagation loss becomes equal to the free space loss.

(3) Feeder loss (Lf)

(Unit:dB/m)

Type		7GHz	12GHz
Square		0.1	0.2
	Aluminum	0.07	0.17
	Copper	0.06	0.15
Circle		0.03 (51ϕ)	0.03 (40ϕ)

The losses of major waveguides (standard design values including joints and bents) are given in the table on the left. When there are transmitting/receiving circuit, main/stand-by switching circuit, etc., between the transmission power measuring point of the antenna and the feeder, losses due to these circuits should be added.

5. Absolute gains of transmitting and receiving antennas (Gat and Gar)

The absolute gain of a parabolic antenna using a frequency above 1000 MHz is given by

$$GA = \eta \frac{4\pi}{\lambda^2} A = \eta \left(\frac{\pi d}{\lambda}\right)^2$$

$$GA(dB) = 20 \log D(m) - 20 \log \lambda(cm) + 50 + \log \eta$$

where

η : Aperture coefficient of the parabola (equal 0.6 in the of shallow type and 0.5 in the case of deep type)

A : $\frac{\pi D^2}{4}$ (D: Diameter of the parabola)

4. Passive repeating (reflector element)

Passive repeating is achievable by the following methods.

1) By using antennas back to back.
2) By using a refraction radio wave lens.
3) By varying the direction with a reflector.
4) By using diffraction net (this is advantageous when the change of direction is small).

Methods 1) and 2) provide worse repeating efficiencies and require larger construction expenses, so that the reflector method which employs such a flat reflector that can easily adjust the diffraction and which provides a high efficiency is employed most frequentry.

(1) Gain of reflector element

The gain of a reflector is given by the ratio of the aperture area of the isotropic antenna in use to the equivalent area of the reflector as is the case in obtaining the gain of a parabolic antenna, as follows.

$$GR = (\eta \frac{Ao}{\lambda^2/4\pi})^2 = (\eta \frac{4\pi Ao}{\lambda^2})^2$$

$$GR(dB) = 20\ \log Ao(m^2) - 40\ \log\lambda(cm) + 102 + 20\ \log\eta$$

$$= 20\ \log Ao(m^2) + 40\ \log f(GHz) + 43 + 20\ \log\eta$$

where Ao : Area of the reflector element

η : Efficiency of the reflector element

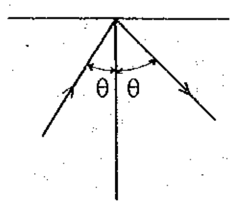
$\lambda^2/4\pi$: Isotropic antenna aperture area

The efficiency is reduced by loss due to rough surface caused by unevenness or irregularity, corrosion, rust, field disturbance due to antenna margin, etc. In general, the efficiency is about 80% and

$$20\ \log\eta = 20\ \log\eta\ 0.8 = -2dB.$$

(2) Angle loss of reflector

The angle loss of a reflector is an additional loss caused by the reduction in the reflector projection area in the direction of propagation when the reflector has a certain angle to the direction of propagation and is given by



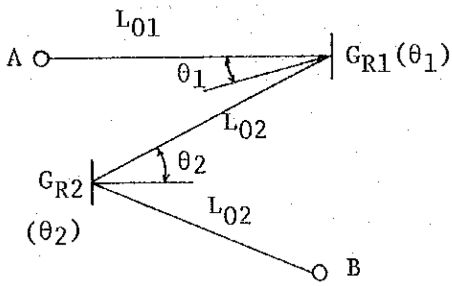
$$L\theta = 20\ \log\ \cos\theta_1,\ \cos\theta_2$$

where θ_1: Incidence angle in the horizontal direction

θ_2: Incidence angle in the vertical direction

(3) When two reflectors element are used:

When $\theta > 60°$, the efficiency to be obtained with one reflector is rather low and repeating should be made by using two reflectors as shown below.



There are two types of arrangement depending on the spacing between the two reflectors: for arrangement and close arrangement.

a) Far arrangement

When the two reflectors are located considerably apart from each other, the loss between them is given by

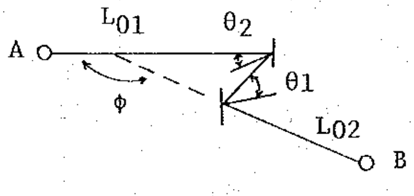
$$\Gamma = L_{01} + L_{02} + L_{03} - GR_1(\theta_1) - GR_2(\theta_2)$$

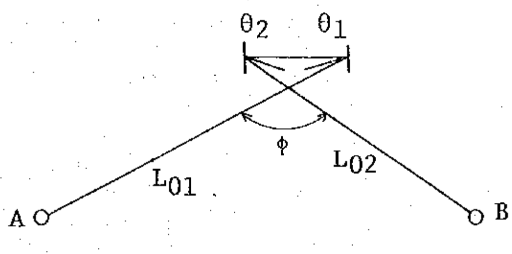
where $L_{01} \sim L_{03}$: Free space losses

$GR(\theta)$: Reflector gain with incidence angle θ

b) Close arrangement

The close arrangement where two reflectors are located close to each other is available in the following two types.





When $\theta = \theta_1 \geqq \theta_2$,

$$\theta_1 - \theta_2 = \frac{180° - \theta}{2}$$

When $\theta = \theta_1 = \theta_2$,

$$\theta = \frac{180° - \phi}{4}$$

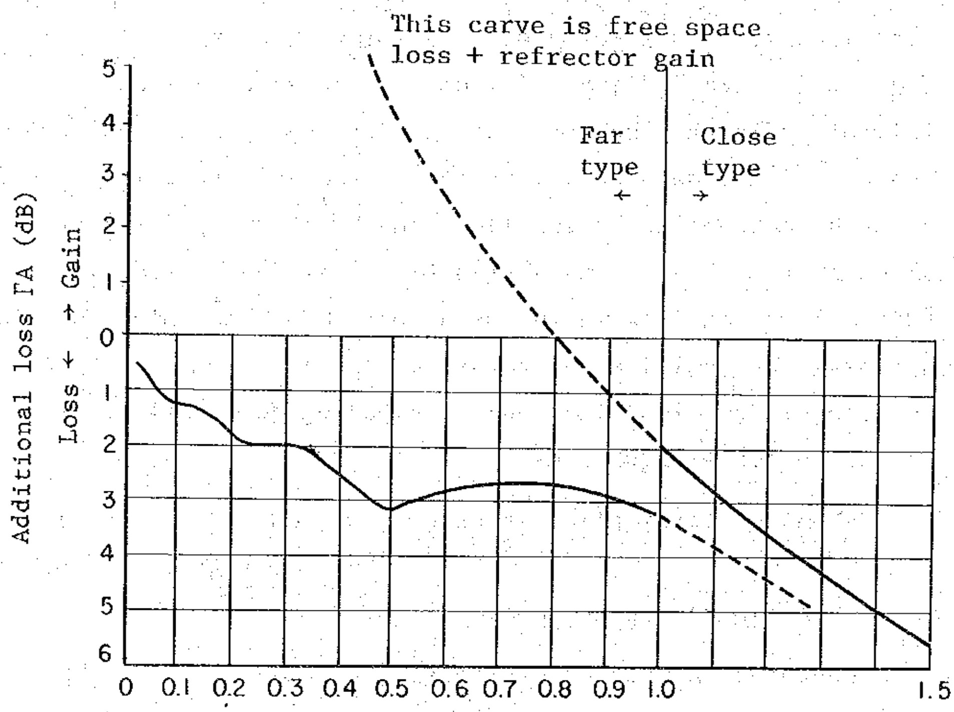
The loss between A and B has given by

$$\Gamma = L_{01} + L_{02} - GR(\theta) + \Gamma A$$

where ΓA : Additional loss due to the two reflectors

It seems as if one of the two reflectors operates the provide the gain and the other reflector is used simply for directional change.

Additional loss for case of double reflector element



$$\sigma = \frac{\lambda d}{A_{cos}\theta}$$

λ : Wave length (m)
d : Refrector separation (m)
A : Refrector area (m^2)
θ : Refrector in shoot angle

The left figure shows the additional loss of two reflectors and whether far or close arrangement depending on the distance between the reflectors.
When σ > 1, calculation should be made for far type.
When σ > 1, calculation should be made for close type.

5. Diffraction Net

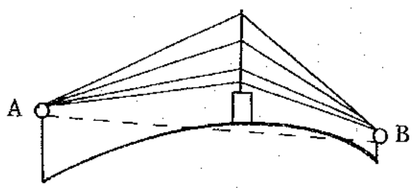
For routes being out of sight the above-mentioned reflectors are employed in general. However when the route approaches a large circle course and there is a hill or a like from which the transmitting and receiving points can be seen within the line of sight, it is often advantageous to employ diffraction.
When diffraction is made by a circular hill, the receiving power is usually somehow reduced from the value obtained by the knife edge diffraction theory due to reflection on irregular or spheric surface. Accordingly, when a knife edge is formed by diffraction and a shielding net is arranged depending on the course difference as shown in the following figure, the influence of the superposition of the positive and negative phases caused by the course difference at receiving point B can be controlled by shielding one polarity by diffraction and, by aligning the phases, the diffraction loss can be reduced.
(Dielectric material may be used instead of the shielding plate for aligning the positive and negative phases.)
In general, the loss improvement by using diffraction, Ls, is given by the following equations.

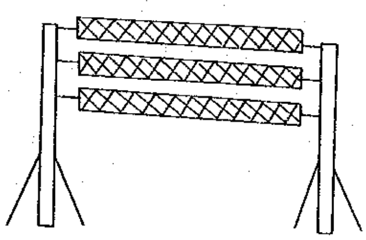
When shielding plate is used: $Ls = 20 \log N + 6$ (dB)

$Ls' = 20 \log N + 12$ (dB)

When dielectric is used:

Where N: Number of diffraction stages





6. Beam Feeding

Beam feeding is employed to applying a sharp beam to the reflector used as shown in the figure below instead of a feeder line which may otherwise become very long and cause a considerably large loss when proper clearance is not achievable but by erecting extremely high steel towers at the transmitting and receiving points for sending the radio wave in the intended direction.

The range of beam feeding, d_1, can be obtained by the following equation.

$$d_1 = \frac{Ae^2}{\lambda} = \frac{\frac{4}{\pi} Ao \cos\theta}{\lambda} \quad (\text{when } \sigma = \frac{\lambda \cdot d}{Ae} \leqq 1)$$

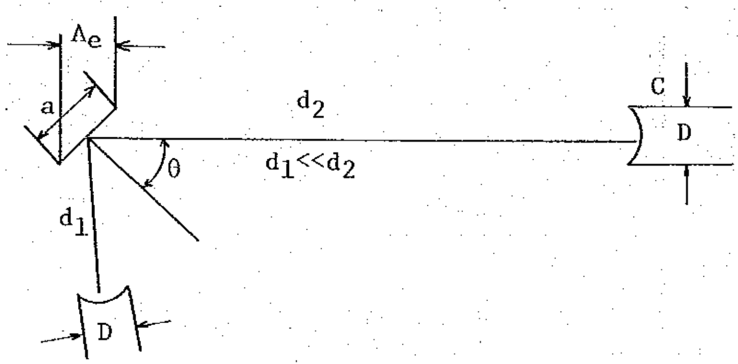
Where λ : Wavelength

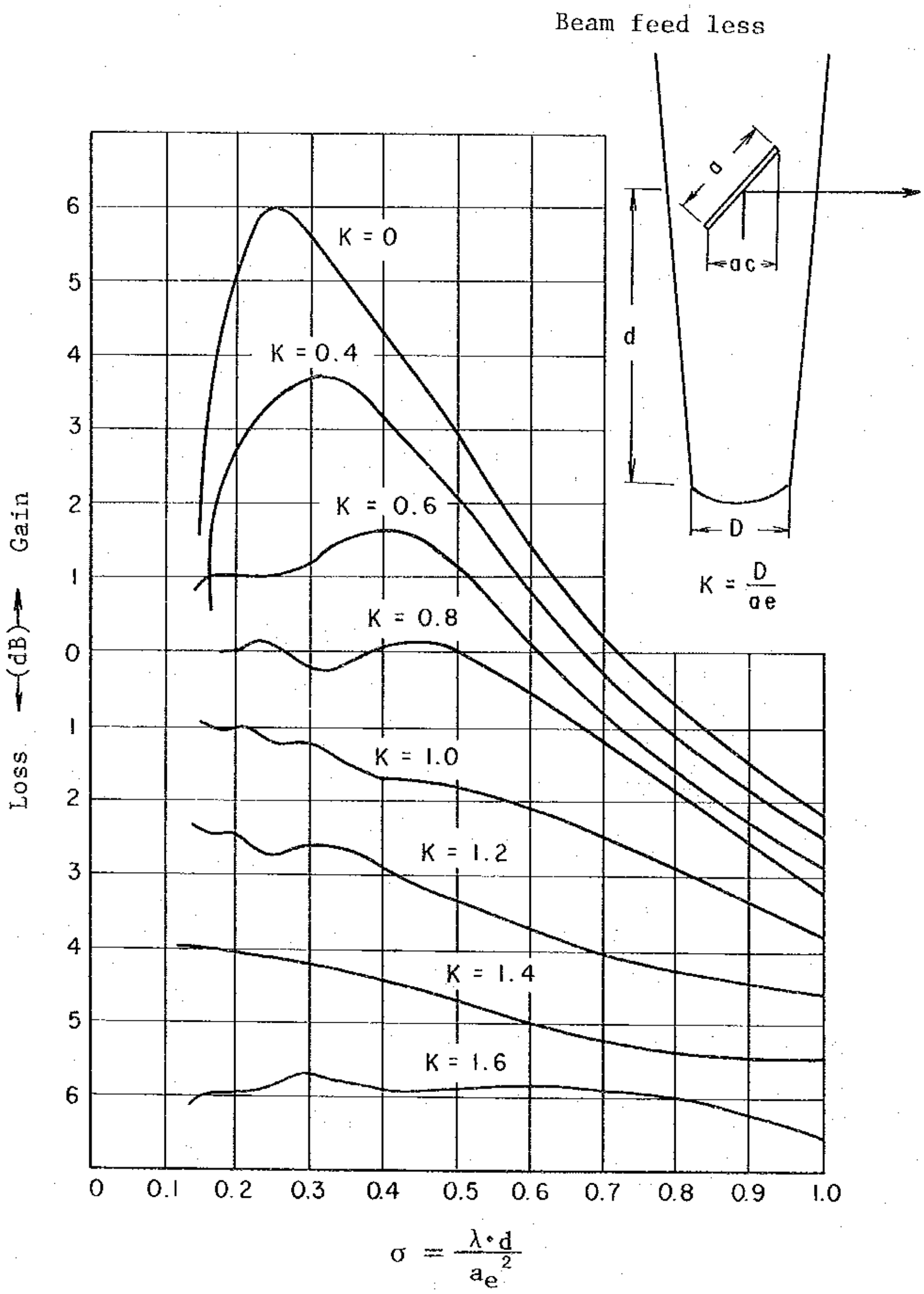
a_e : Diameter of the circle having an area equal to the effective area of the beam reflector (m)

$$a_e = 2\frac{\sqrt{Ao \cos\theta}}{2}$$

d : Distance between the antenna and beam reflector (m)

Ao : Area of the reflector




Beam feed less
K = 0
K = 0.4
K = 0.6
K = 0.8
K = 1.0
K = 1.2
K = 1.4
K = 1.6
Loss ←(dB)→ Gain
a
ac
d
D
K = D/ae
0
0.1
0.2
0.3
0.4
0.5
0.6
0.7
0.8
0.9
1.0
σ = λ·d / ae²

7. Line Reliability

This paragraph describes the method of line reliability calculation. It is desirable for achieving reliability that antenna power P_t (dBm) should fall within the following range.

29.7 MHz ∿ 300 MHz	$A + M\ (dB) > P_t > A$
300 MHz ∿ 10 GHz	$P_t > A$
Above 10 GHz	$P_t > A'$

$$A(dBm) = (Lp + L_f + L_F) - G_{At} - G_{Ar} + P_{th} \quad \text{(below 10 GHz)}$$

$$A'(dBm) = (Lp + L_f + L'_F) - G_{At} - G_{Ar} + P_{th} \quad \text{(above 10 GHz)}$$

These equations are employed for standard cases and A or A' includes losses and gains other than given here (filter loss, reflector gain, etc.).

- Lp : Propagation loss
- L_f : Feeder loss
- M : Usually about 10 dB. 10∿20 dB or so when necessary because of the input level.
- L_F : Fading loss (Standard values are given in the following table.)

Reliability / Frequency range	5×10^{-3} (99.5%)	1×10^{-3} (99.9%)
29.7 MHz ∿ 300 MHz	0.1 dB/km	-
300 MHz ∿ 1000 MHz	0.2 dB/km	Add 3dB to the value corresponding to 5×10^{-3}
1000 MHz ∿ 10 GHz	0.2 ∿ 0.3dB/km	Add 6dB to the value corresponding to 5×10^{-3}

It is to be noted that these values are standard one and ample examination should be made by referring to the fading estimation method, etc., mentioned later so as to meet the required line reliability.

L_F' : Loss due to rain, mist, etc. Attenuation (dB) by rainfall corresponding to a rainfall strength of 0.01%, 10 minutes.

$$L_F' = L_{Fo}' \times kP \times d$$

L_{Fo}' : Attenuation (dB) per km near the propagation route (18°C) at rainfall strength of 0.01%, 10 minutes.

kp : Distance coefficience

d : Distance of the route (km)

P_{th} : Threshold input level

Infrequency modulation and phase modulation, the signal-noise-ratio (S/N) is improved abruptly when the receiving input level becomes above a certain value. This threshold level is give by

$$P_{th}(W) = K \cdot T \cdot B \cdot F \cdot C_f$$

K : Boltzmann's constant (1.37×10^{-23} J/°K)

T : Absolute temperature 273°+°C (In normal temperature: 293°C)

B : Bandwidth of receiver (Hz)

F : Noise figure of receiver

C_f : S/N at threshold level (9dB in the cases of FM and PM)

When the external noise power (P_{rne}) is large, P_{th} is obtained as follows.

$$P_{th}(W) = (P_{rni} + P_{rne})$$

where P_{rni} : Internal noise power

P_{rne} : External noise power

8. Signal-to-Noise Ratio (S/N) of Line

The signal-to-noise ratio (S/N) for thermal noise and external noise per section and standard condition of a transmission line is given by

$$S/N = P_t - (L_p + L_f) + G_{At} + G_{Ar} - P_{rn} + I$$

The S/N for plural sections is given by

$$S/N = \frac{1}{1(S_1/N_1) + (S_2/N_2) + \ldots\ldots 1(Sn/Nn)}$$

where P_t : Antenna power (dBm)

L_p : Propagation loss (dB)

L_f : Feeder loss (dB)

G_{At} : Absolute gain of transmitting antenna (dB)

G_{Ar} : Absolute gain of receiving antenna (dB)

P_{rn} : Receiving noise power (dBm)

I : S/N improvement factor (dB)

(1) Calculation of receiving noise power P_{rn}

a) When a frequency below 470 MHz is employed, receiving noise power P_{rn} is given by the sum of receiver internal noise power P_{rni} (noise power at the receiver input terminal + estimated receiver thermal noise power) and external noise power P_{rne}, which are respectively given by

$$P_{rni}(\text{dBm}) = 10\ \log B + F = 144$$

$$P_{rne}(\text{dBm}) = 10\ \log B/b + E - 20\ \log_f - 77.3 + g - L_f$$

where

B : Receiver equivalent noise bandwidth for noise spectrum (kHz)

F : Receiver noise figure (dB)

E : Mean square value of noise power (dBμ)

b : Instrument equipment bandwidth for noise spectrum in measuring E (kHz)

f : Frequency (MHz)

g : Antenna gain against noise (dB)

In the case of omnidirectional radiation for H polarity: $g = g_s$

In the cases of directional radiation in H polarity and omnidirectional radiation in V polarity: $g \fallingdotseq 0$

In the case of directional radiation in both H and V polarity: $g \fallingdotseq G_s/2$

G_s : Absolute gain for signal

L_f : Feeder loss (dB)

b) When a frequency exceeding 470 MHz is employed, P_{rne} can be neglected for P_{rni}, so that we have $P_{rn} = P_{rni}$.

(2) S/N improvement coefficient I

a) SS-FM:

$$I = 10 \log \left(\frac{fdo^2 \cdot B}{f\nu^2 \cdot f_s} \right)$$

b) SS-PM:

$$I = 10 \log \left(\frac{M^2 c \cdot B}{f_s} \right)$$

f_{d0} : Effective frequency deviation (kHz) at test tone level

B : Receiving bandwidth (kHz)

$f\nu$: Baseband frequency of transmission line

fs : Frequency band (kHz) of transmission line (=3.1KHz)

mc : Effective phase deviation (radians) at test tone level

9. Fading

(1) Types and countermeasures for fading

Fading is depends on all atmospheric conditions (temperature, atmospheric pressure, vapor pressure, etc.) and can be classified into the following types.

1) Fading due to variation of M distribution (Note 1)

i) Fading due to variation of K (Note 2)

ii) Fading due to formation of air duct

iii) Fading due to formation of air reflector

2) Scintillation fading

Electric field variation caused by radio wave scattering due to local atmospheric disturbance and its depth is 2 ∿ 3dB or less. This fading is of no practical problem.

Note 1: The defraction factor of the atmosphere, n, given by

$$n = \sqrt{\varepsilon_s}$$

where ε_s is the dielectric constant of the atmosphere (ε_s = 1.0004 ∿ 1.0009).

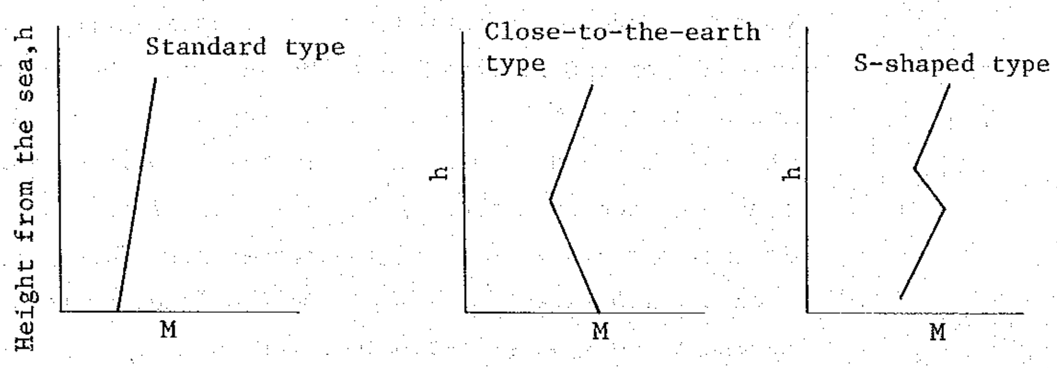
Usually, the higher in altitude, the smaller ε_s and the smaller the refraction factor.

For radio wave propagation, modified refraction factor (M) is used instead of N in consideration of the coverture of the ground surface.

$$M = (n - 1 + \frac{h}{R}) \times 10^6$$

Where h is the height of the propagation course above the sea level and R the radius of the earth.

The relationship between M and h is designated M curve and its examples are shown below.



Note 2: M increases with increased height. As height h becomes larger the propagation course tends to become nonlinear and curve to the direction of larger M, so that the actual line-of-sight distance on the earth becomes larger than the geometrical line-of-sight distance.

In consideration of this, the radius of the coverture under standard atmospheric condition becomes about 4/3 times as large as the radius of the earth, R.

Accordingly, by considering a hypothetical earth is a radius K_R, we can calculate K_R as follows assuming that radio waves go on straight over the hypothetical earth.

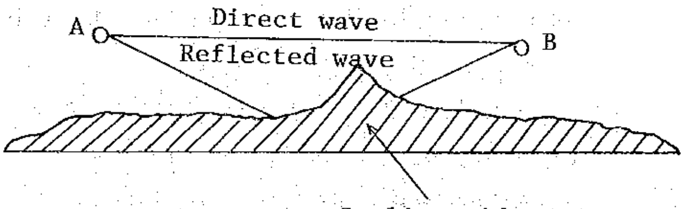
$$K_R = \frac{R}{1 + \frac{R}{n_o} \cdot \frac{dn}{dh}}$$

Where n_o is the refraction factor of the atmosphere on the ground furface. In standard atmosphere, $K_R = \frac{4}{3}R$

a) Fading due to variation of K

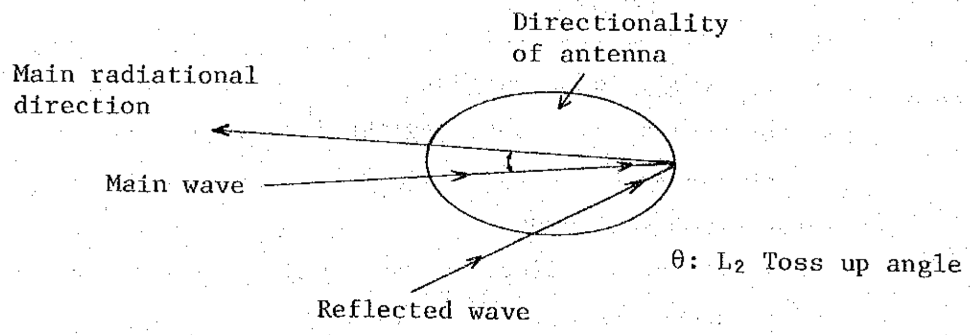
i) In the case of mountain diffraction the depth of diffraction varies with K, varying the diffraction loss.
In systems where sufficient clearance is not secured, the line-of-sight condition may degrade due to variation of K and a large transmission loss may be caused. Accordingly, it is desirable that sufficient clearance should be kept even if K vary to about 0.8 or so.

b) When the line involve reflection from the ground surface (or water surface), the reflection point varies with K and the strength of the resultant wave varies with the course difference (phase difference) between the direct and reflected waves.
The reflection point is smooth in this case, the influence of the reflected wave becomes large and the resultant wave causes intensive fading. For the countermeasure against this, it is necessary to select of large irreguralities in the area expected to be the reflection point or select such a topographical condition that blocks the reflected wave as shown in the following drawing.



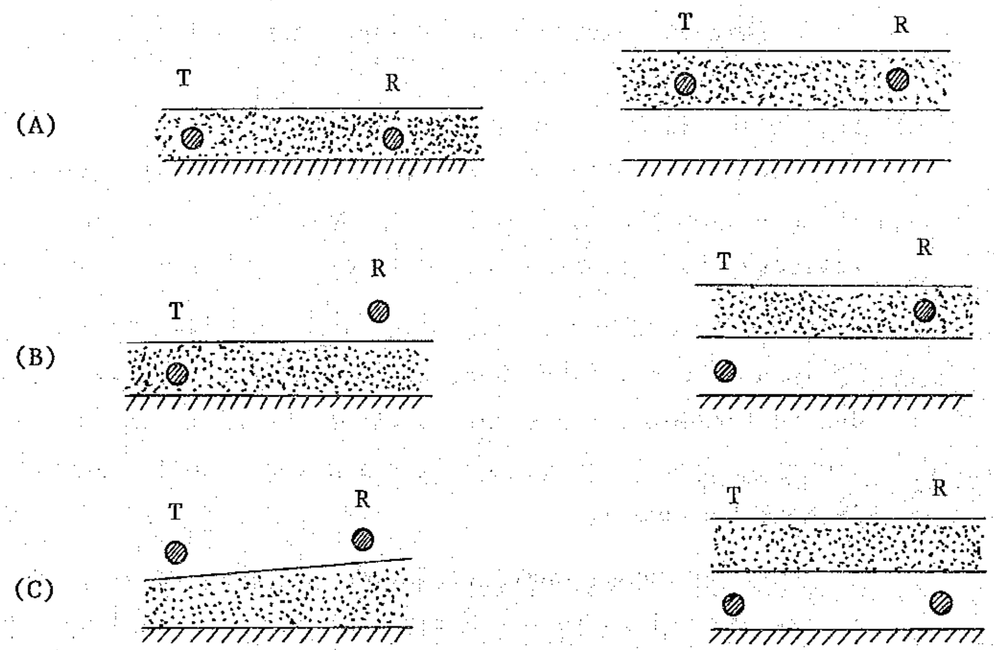
2) When space diversity is employed or in the direction of the deflected wave, such an antenna that provides small gain should

be selected or the antenna should be devised so that the antenna should be raising as shown in the following figure.



b) Fading by formation of air duct

Fading due to air duct presents complex phenomena depending on the relative relationship between the locations of the transmitting and receiving points and the location of the duct and appears in three forms as shown below: (A) where T (transmitting point) and R (receiving point) are located in the duct, (B) where one of the T or R (whichever is located in the duct and the other out of the duct, and (C) where both T and R are out of the duct.



In the case of (A), the receiving power is stronger than the free space receiving power and transmits considerably distant (super-refraction propagation).

In the case of (B), the receiving power is generally weak and its average value lowers gradually over a long time.

In the case of (C), not only the direct wave but also the ground reflection wave and duct reflection wave are combined to cause a large depth of fading.

Countermeasures against fading caused by the formation of air duct in the above-mentioned manners are:

i) The propagation course should not go over the sea, beach, and paddy field as much as possible.

ii) The propagation course should go at an ample height above the ground.

iii) The propagation course should not be too long.

iv) When fading can not be solved by the above-mentioned measures, space diversity, etc., should be employed upon propagation design. When the height of the propagation course above the ground is less than 100 m even if the propagation distance is about 50 km, fading due to air duct may often be caused, so that due care should be taken upon selecting the propagation course.

c) Fading caused by formation of air reflector

When air reflector is formed because of air duct produced in the sky over 100 m, fading may be caused by the influence of the reflected wave from the air reflector.

d) Fading due to other causes

In a transmission line using a reflectors, fading may be caused due to incidence angle variation of the radio wave or angular deviation caused by strong wind if the directionality of the reflector is made too sharp.

The incidence angle variation of radio waves is reported to be almost negligible in the horizontal direction and is ±10 ∿ 20 seconds in the vertical direction due to the atmospheric refraction factor.

(2) Pressure of microwave fading

a) Gist of fading estimation

Fading is extremely complicated and various methods of estimation have been reported.

For representing the magnitude of fading, it is ultimately necessary to grasp such variables varying relative to the center value of the fading distribution as

1) fading range (F_{rf})
2) fading depth (F_d)
3) Standard deviation (σ)
4) following items for large fading
   i) probability of occurrence
   ii) duration
   iii) depth

and for various other fading phenomena, fading estimation is made synthically in consideration of the following general characteristics (excluding scattered waves).

1) In fine, calm weather, fading is stronger than in bad weather. However, when a front or fronts pass even in bad weather, strong fading may be caused.

2) Fading variation is larger in the nighttime than in the daytime, and in particular fading tends to occur early in the morning or late at night on fine, calm days.

3) Less fading is caused in winder seasons and more fading is generally caused during worm seasons.

4) The center value of fading depends less on the time of the day and season and the variation of the center value in a day is within several dB.

5) Interference propagation courses with large equivalent reflection coefficients are unstable.

6) The higher the frequency, the larger the fading ratio (frequency at which the receiving level crosses the center value of the fading per minute in a given period of time).

7) The shape of daily distribution differs from case to case and the distribution for a period of more than 5 ∿ 7 days is approximately dB normal distribution or gamma distribution.

8) Fading of a free space propagation course over the land forms a dB normal distribution.

9) Fading of interference propagation courses and free space propagation courses at large heights above the ground and fading of short periods present their power center values approximately in from of gamma distribution.

10) Interference propagation courses, the distribution of the center value becomes approximately dB normal distribution in winder seasons. In free space propagation courses over the land, the distribution of the center value become approximately gamma distribution in summer seasons.

11) In the distribution shapes of 7) ∿ 10) above, approximation is achieved in a 1% ∿ 99% range of the long-time distribution of receiving power. In deep fading regions which may become the problem in system design, approximation may often be made by Rayleigh distribution.

b) Expression of fading

For expressing fading in accordance with the above-mentioned gist of estimation, the following fading cumulative distribution curve is generally employed and related terms and distribution forms are as follows.

Attenuation ratio:

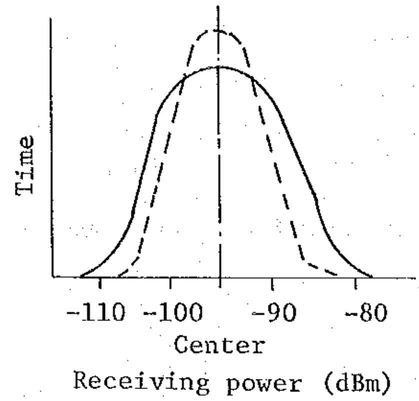
Percent of the time during which the level lowers below the abscissa for the total time

Reliability ratio (reliability):

Percentage of the time during which levels become above the abscissa of the total time for the total time

Relation with standard devistion (σ):

Standard deviation (σ) becomes smaller, the curve changes to dotted one in the following figure.



F_d is the fading width from the center value to the reliability of 99% and F_{rf} is the fading range corresponding to 1% 99% reliability.

Relation between F_d and F_{rf}:

In general, the following relation holds in the case of gamma distribution (including Rayleigh distribution).

$$F_d = (0.65 \pm 0.05) \cdot F_{rf}$$

Relation between F_{rf} and σ:

In general, the following relations hold.

$$F_{rf} = 4.66 \times \sigma \text{ (dB normaldistribution)}$$

$$F_{rf} = 4.8 \times \sigma \text{ (Gamma distribution)}$$

dB normal distribution:

In this distribution, dB value forms the normal distribution (symmetrical to the right and left with reference to the center value) and the cumulative distribution becomes linear on the normal distribution probability chart.)

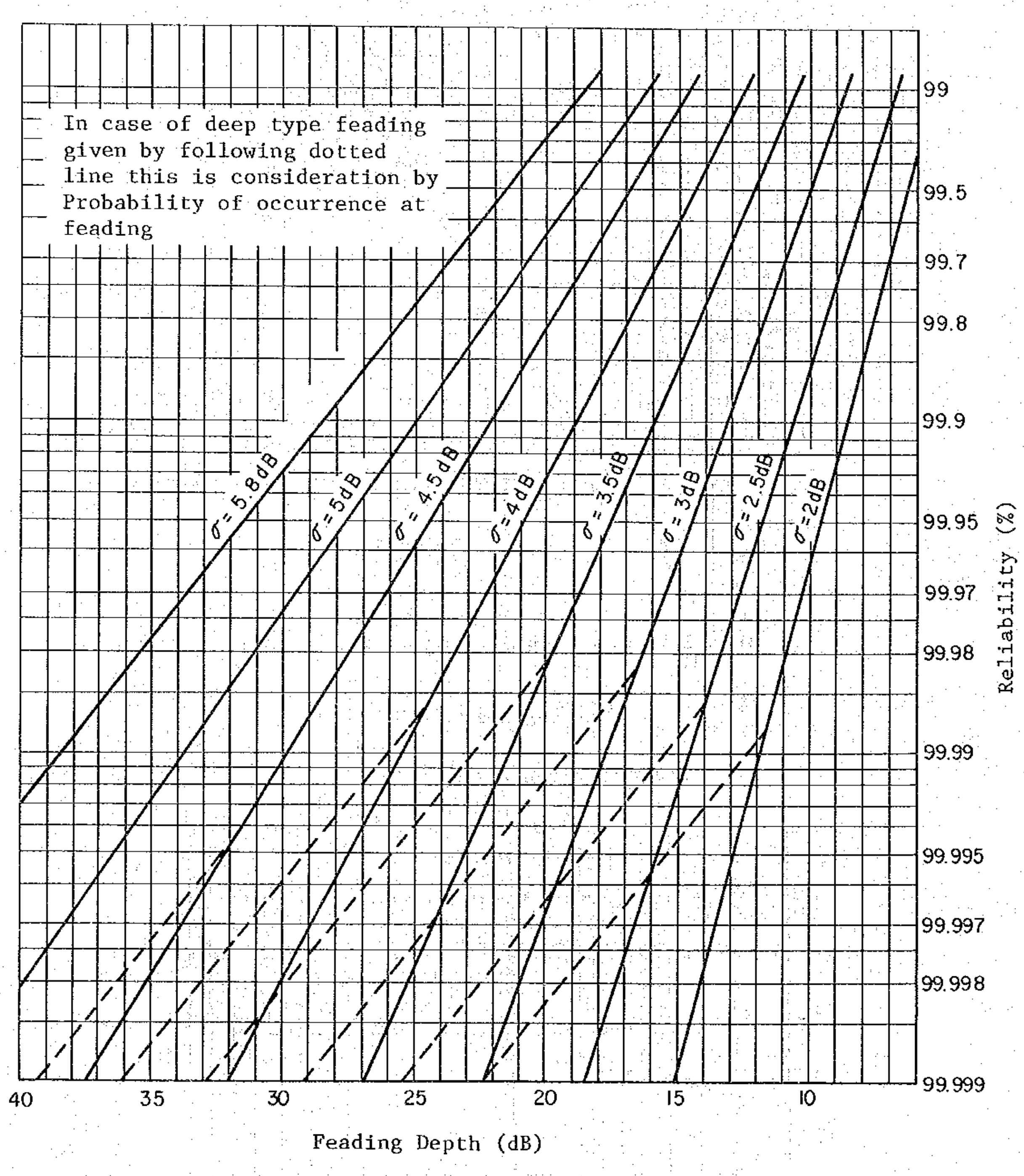
Gamma distribution:

In this distribution, a shape like the foot of a mountain is formed as receiving power lowers and simmetry is not achieved. Accordingly, the cumulative distribution on the normal distribution probability chart does not become linear but is curved toward lower receiving power.

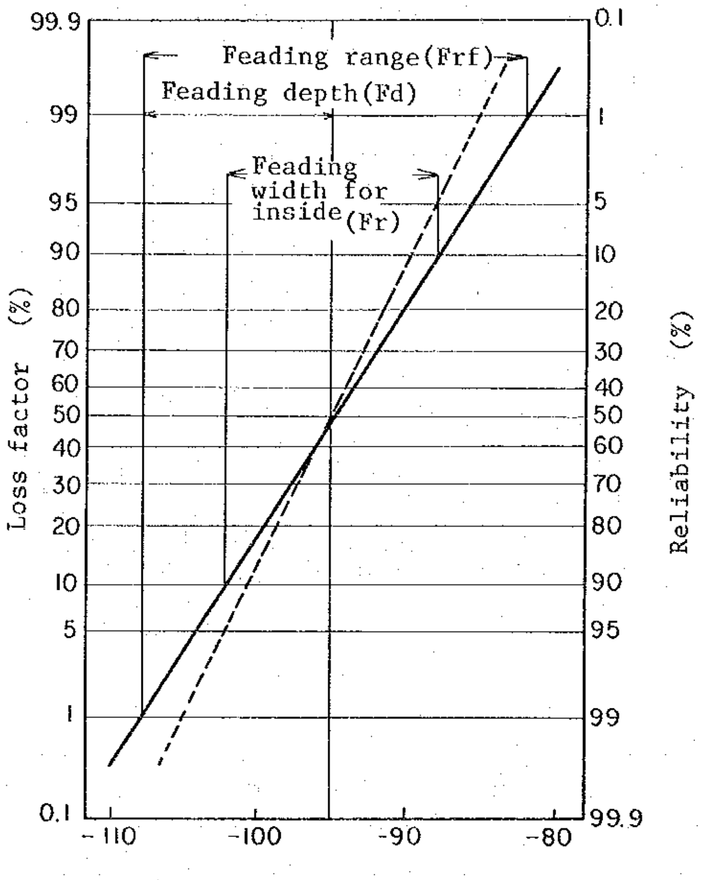
Rayleigh distribution:

This distribution falls when parameter $\gamma = 1$ ($\sigma = 5.8$dB) in gamma distribution and represents the worst theoretical distribution in imbombining multiple frequencies. In this distribution, the % value reduces by 1/10 for every lowering of 10dB relative to the power probability at low levels. (For example, the cumulative propability at which receiving power lowers 20dB relative to the center value is 1% and that at which receiving power lowers 30dB is 0.1%.) In this case, internal fading width F_r = 13.4dB and σ = 5.8dB when the distribution is considered as gamma distribution.

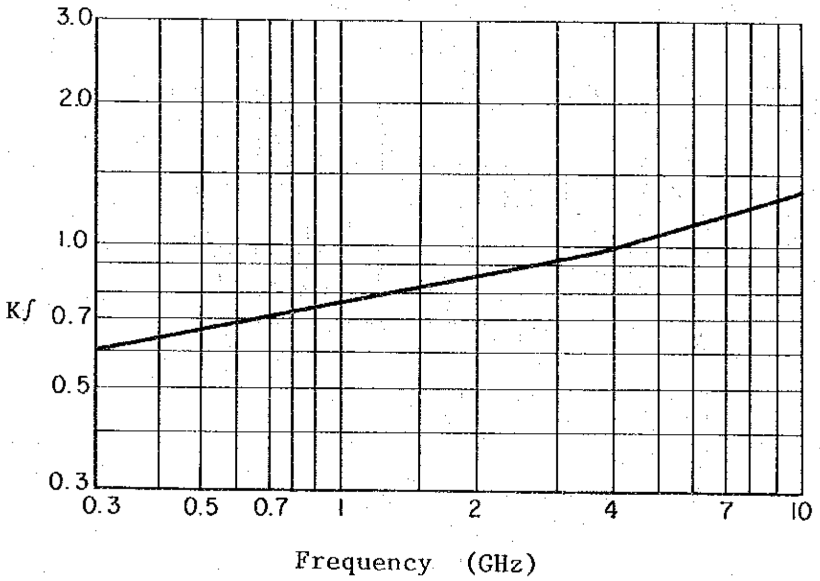
The figure of gamma type feading depth accumulate distribution



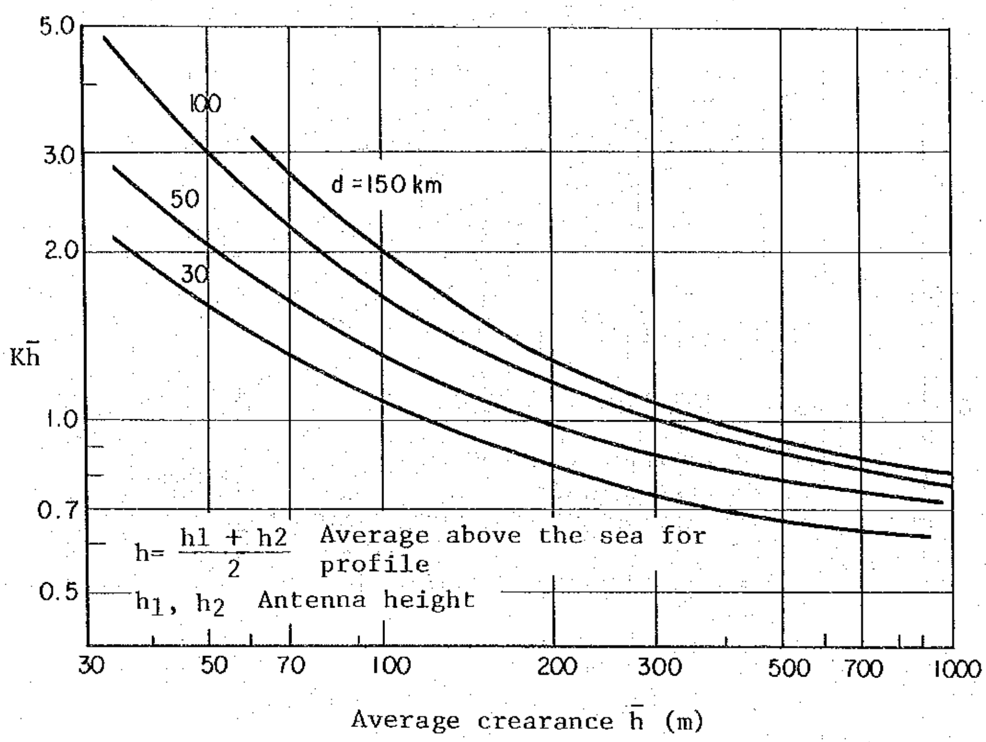
The curve of feading accumulate distribution



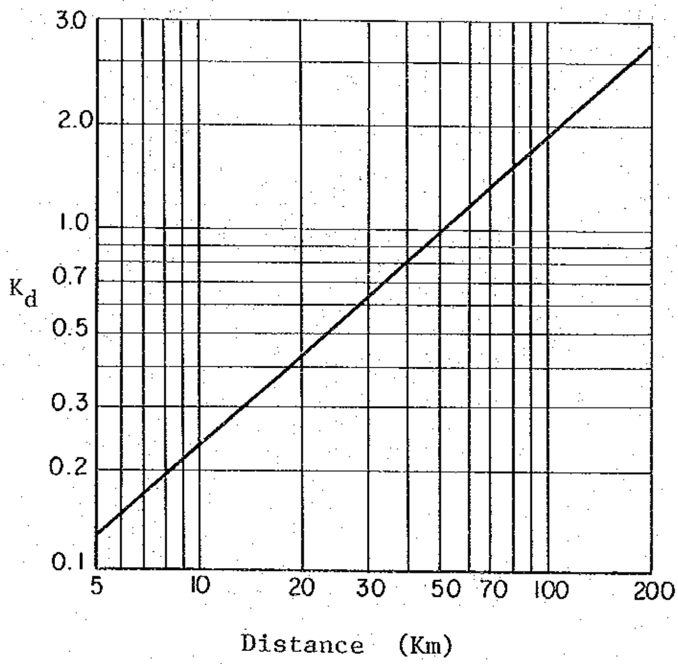
Frequency characteristic



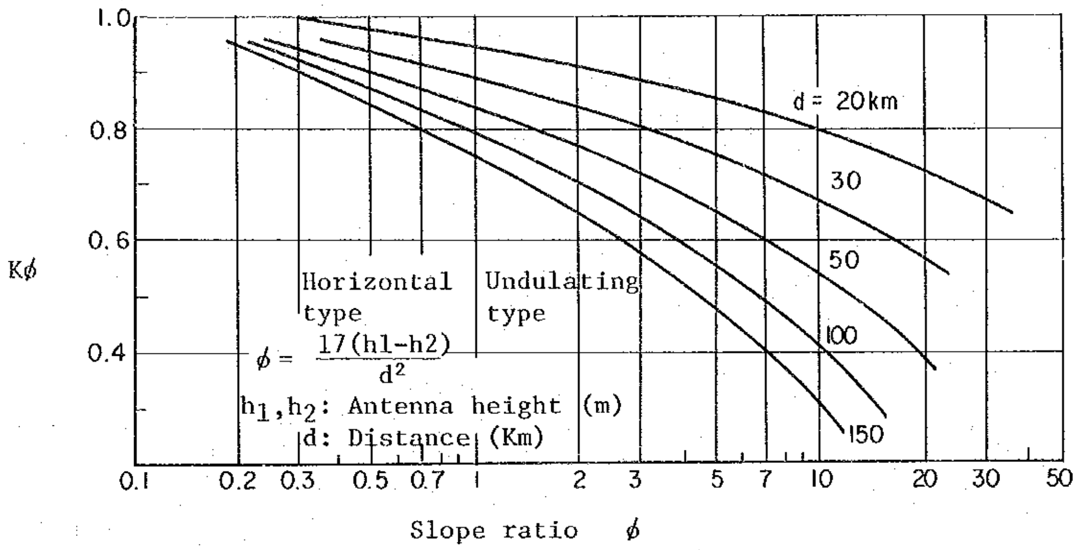
Crearance characteristic



Distance characteristic



Slope characteristic



c) Rayleigh fading estimation equation

i) Rayleigh fading generation probability P

$$P = K\left(\frac{f}{4}\right)^{1.2} \cdot Q \cdot d^{3.5} \qquad P \leqq 0.3 \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots (1)$$

where K : Coefficient (5.1×10^{-9})

f : Frequency (GHz)

Q : Coefficient due to propagation course (Classified in the following table and takes the following values)

Mountain: 0.4

Plain : 1.0

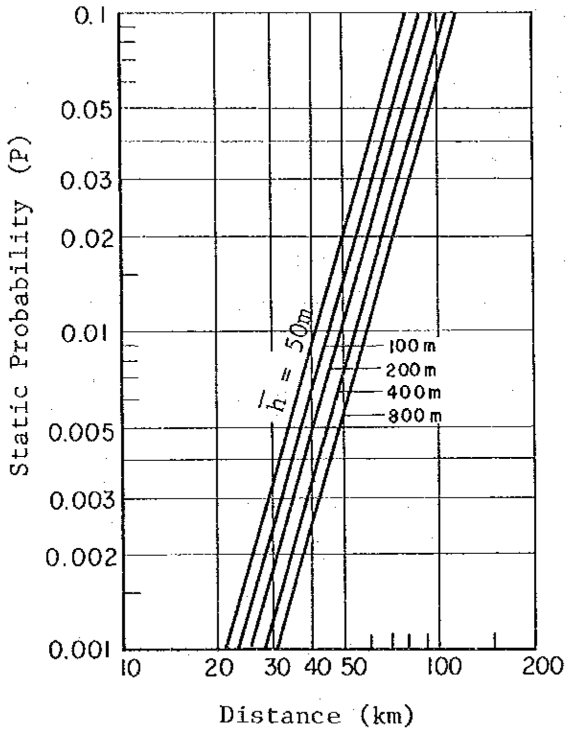
Sea: $3.7 \times \sqrt{\frac{\frac{1}{\bar{h}}}{0.051}}$ where $\bar{h} = \frac{h_1+h_2}{2}$ (m)

h_1 and h_2 are heights above the sea in the case of the sea and those above the ground in the case of beach.

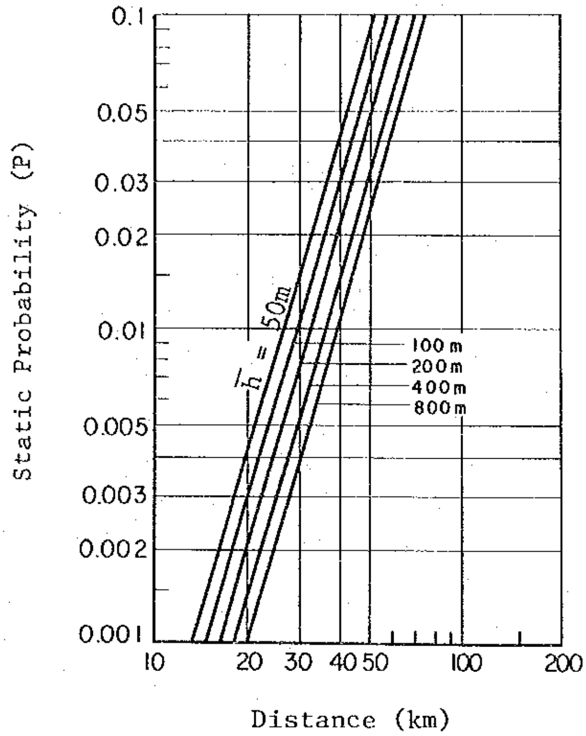
d : Distance

Mountain	When major parts are in mountainous area.
Plain	When major parts are in plain (including bays or the like including beach or the sea and yet including mountainous).
Sea	Above the sea or beach

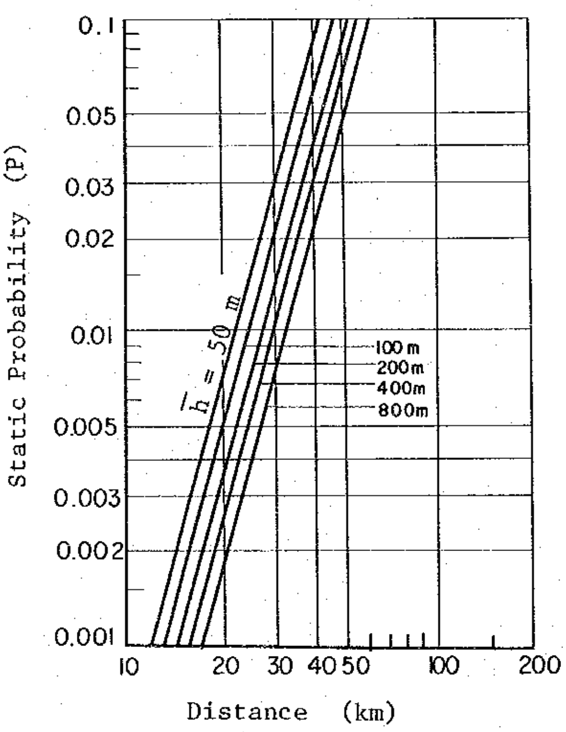
P for Sea Propagate (2,000MHz)



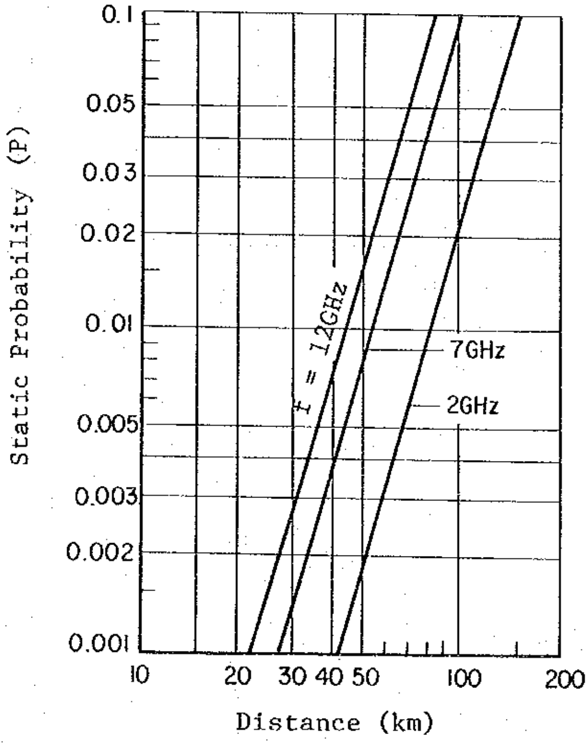
P for Sea Propagate (7 GHz)



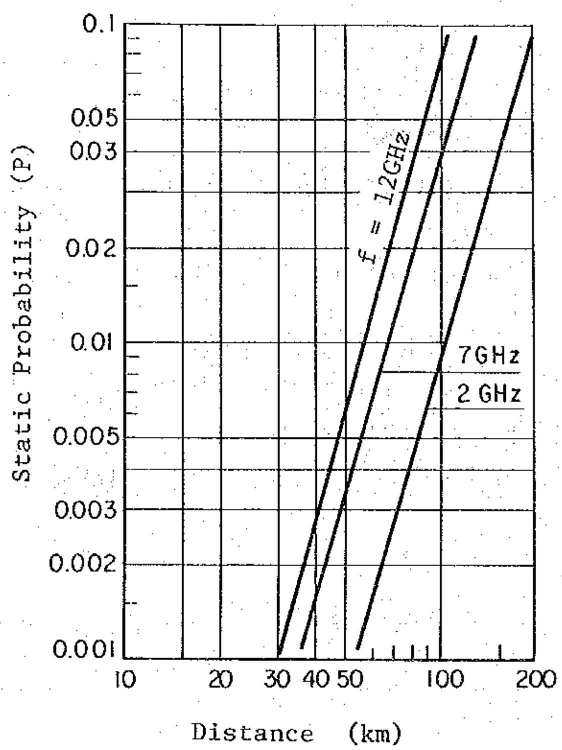
P for Sea Propagate (12 GHz)



P for Plain Propagate



P for mountain Propagate



ii) Standard deviation σ of long-time receiving power distribution in worst season

$$\sigma(dB) = K \cdot (\frac{f}{4})^{0.3} \cdot Q \cdot d^{0.9} \quad \cdots\cdots\cdots\cdots \quad (2)$$

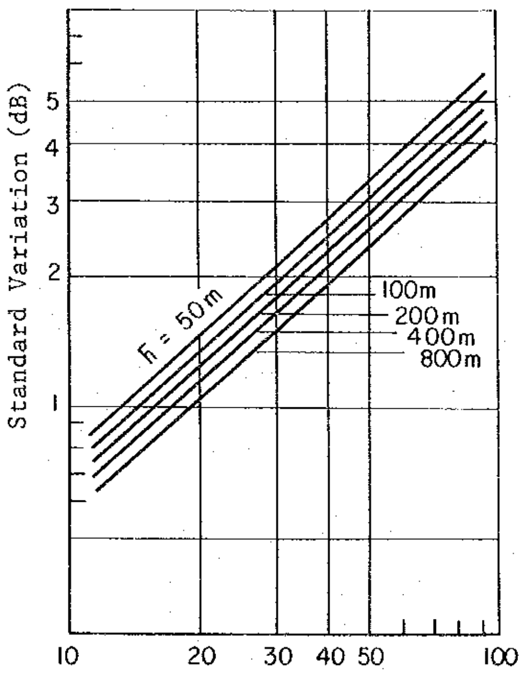
where K : Coefficient (0.068)

Q : Propagation course coefficient

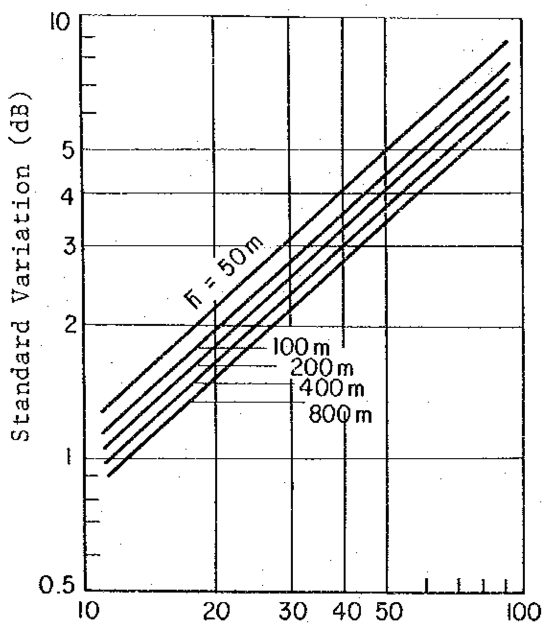
(mountain: 0.8; 1,0; sea: $\frac{(\frac{1}{h})^{0.13}}{0.47}$)

The value gives the standard deviation of the overall receiving power sistribution corresponding to the probability of occurrence and the period of Rayleigh fading.

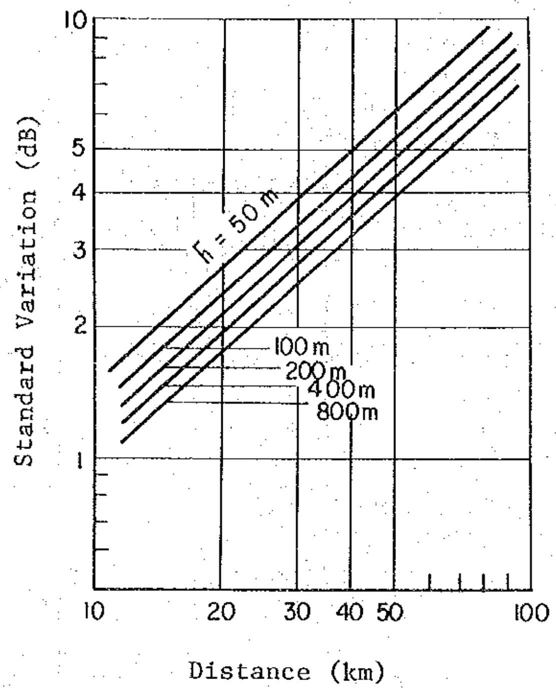
σ for Sea Propagate (2,000MHz)



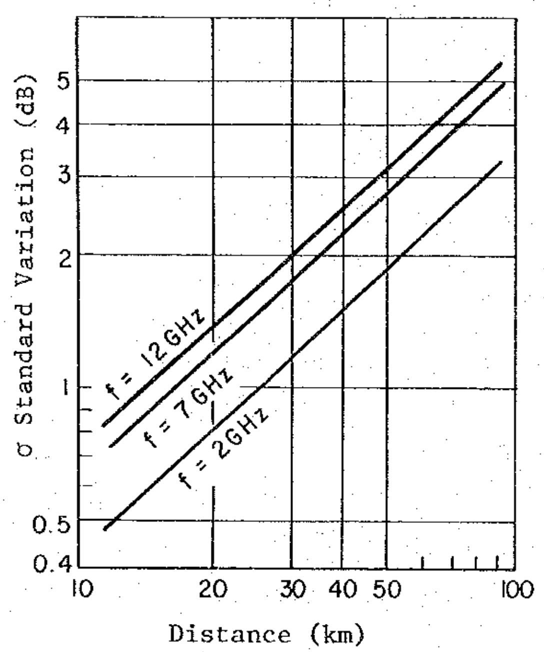
σ for Sea Propagate (7 GHz)



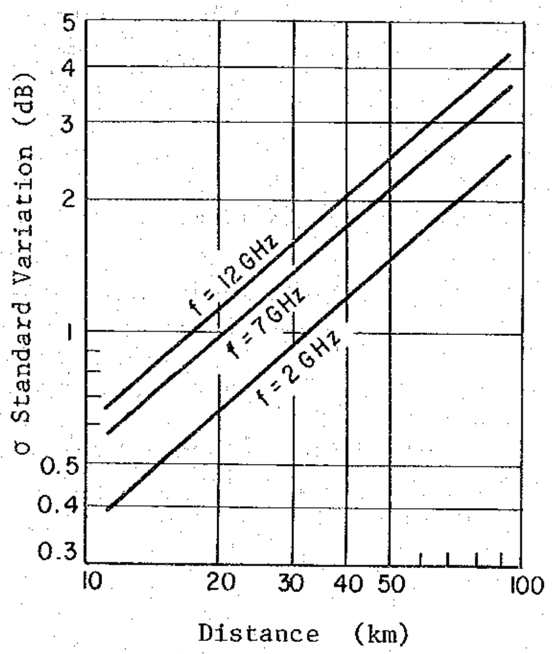
σ for Sea Propagate (12 GHz)



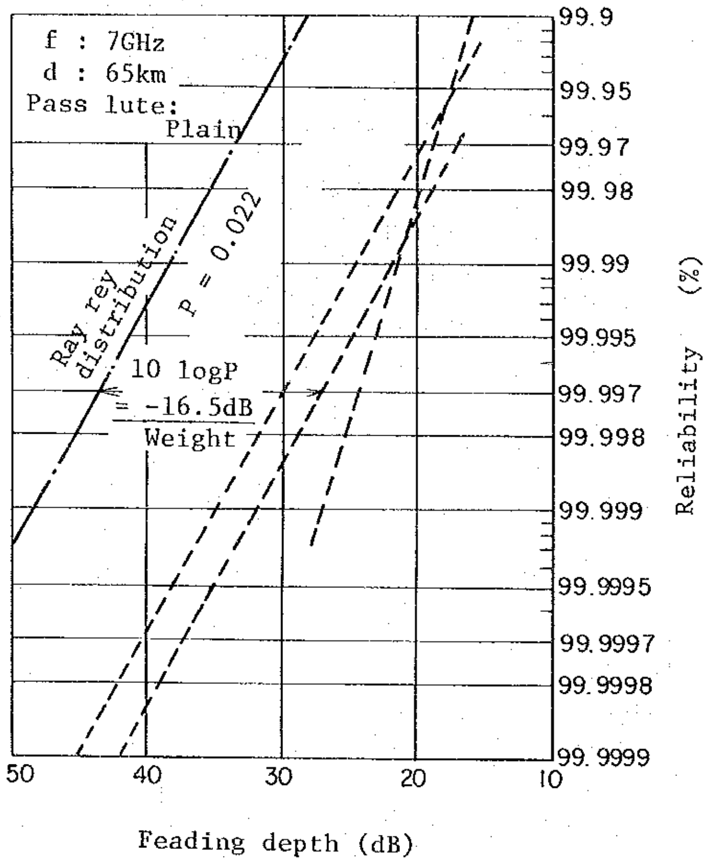
σ for Propagate



σ for mountain Propagate



Conjecture example for power accumulate distribution



[II]-5 Interference and Disturbance

1. Calculation of D/U

The D/U of a system is calculated as follows and the result of calculation should be examined to determine whether or not it meets the required D/U of paragraph 2 above.

$$D/U = D - U \text{ (dB)}$$

Where D : Desired signal level (dBm) at receiver input
U : Interference level (dBm) at receiver input

$$\text{Interference level } U = P_t' - (L_f' + L_p') + G_{At\theta} + G_{Ar\theta} - D_p$$

Where P_t' : Interference radiation power (dBm)

L_f : Feeder loss (dBm) of interfering side (transmitting side) and receiving side

L_p' : Propagation loss (dB) between interfering wave transmitting point and receiving point

$G_{At\theta}$: Gain (dB) of interfering wave transmitting antenna in receiving direction

$G_{Ar\theta}$: Gain (dB) of receiving antenna in the direction of interfering wave transmitting point

D_p : Improvement (dB) cross polarization

At frequencies above 1,000 MHz the D/U is nearly as follows:

$\theta \leq 10°$	$10° < \theta \leq 30°$	$30° < \theta \leq 90°$	$90° < \theta$
approx. 15dB	approx. 10dB	approx. 5dB	0dB

When necessary, differential fading of approx. 5dB should be added to the D/U.

2. Calculation of Required D/U

(1) At same frequency:

Required D/U = Standard S/N − I + 3

where I is S/N improvement coefficient.

(2) At different frequencies

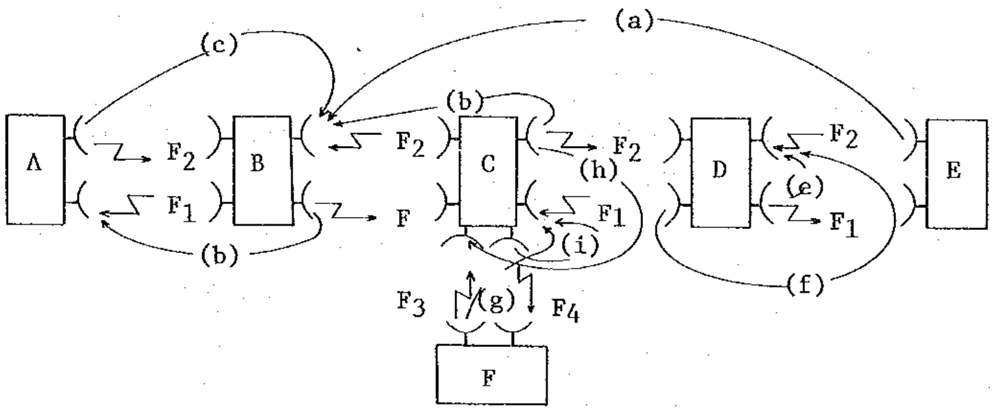
$$\text{Required D/U} = \text{Standard S/N} - I + 3 - L_s$$

L_s: The sum of attenuation due to RF and IF selectivity plus approx. 10dB (dB)

3. Causes of Interference and Countermeasures

In systems and others, interference may be classified into interference for which the carrier frequencies are on the same channel and adjacent channel interference for which the carrier frequencies are on different channels, and the former becomes the problem.

The following figure shows the state of disturbance by interfering waves in a system.



(1) Interference in parallel circuit

1) In the case of the same frequency

i) Overreach (a)

ii) Coupling of front and back of transmitting antenna (b)

iii) Coupling of front and back of receiving antenna (c)

iv) Coupling of IF frequency in repeater station

2) In the case of different frequencies

i) Side-side coupling of antenna (e)

ii) Back to back coupling of antenna (f)

(2) Interference from branch circuit

1) In the same of frequency

i) Front to side coupling of receiving antenna (When F = F) (g)

ii) Front to side coupling of transmitting antenna (When F = F) (k)

2) In the case of difference frequencies

i) Side to side coupling of antenna

ii) Front to side coupling of receiving antenna (When F = F) (g, i)

iii) Front to side coupling of transmitting antenna (When F = F) (k, h)

4. Measures for Reduction of Interference

For reducing interference, due consideration should be given to the following upon system design.

(1) Equalization in receiving power among respective sections

(2) Site location and antenna orientation

(3) Addition of shield plate

Crossing of poralization planes

(4) Adoption of 4-frequency system

5. Interference on Single Channel System

(1) Sensitivity suppression interference

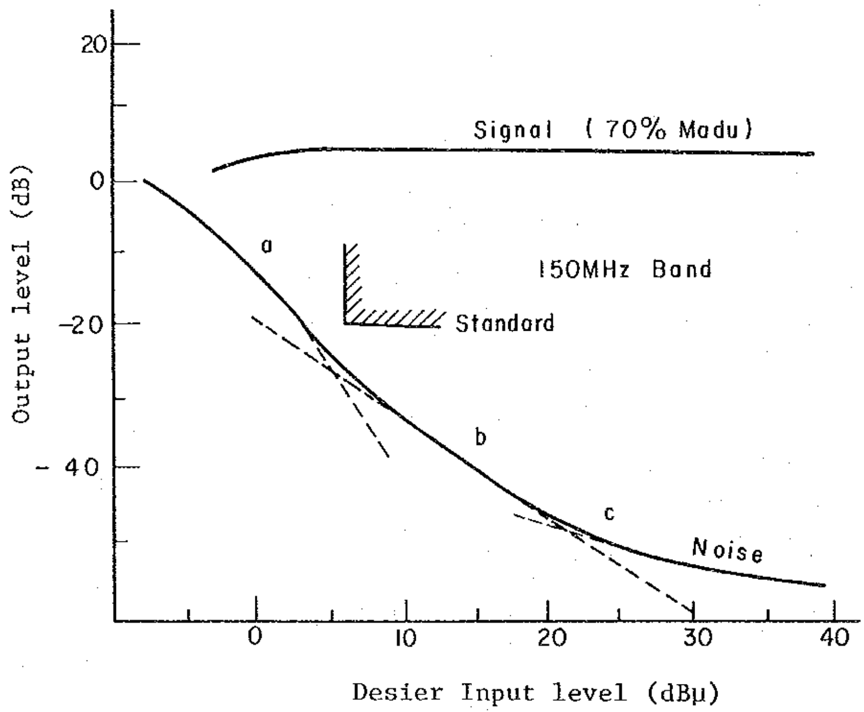
This type of interference is caused when the desired signal is suppressed by an interfering wave of a frequency apart from the receiving band upon receiving the desired frequency by a receiver.

This interference is examined by a interfering level by which, when the desired signal level is incleased 6dB from the 20dB noise suppression sensitivity level*, the state of 20dB noise suppression is re-assumed by the interfering wave.

(* The 20dB noise suppression sensitivity level is the desired carrier signal input (quieting signal) necessary for lowering the level 20dB below the noise level when there is no desired signal input, and the standard value of the 20dB noise suppression level is 2μV (6dBμ).

A standard noise suppression sensitivity characteristic is shown in the right-hand figure where portion a in the noise curve corresponds to a range in which the limiter is not functioning sufficiently and its varies greatly with the input level. Portion b corresponds to the operating range of the limiter. Portion c corresponds to residual noise range scarcely subject to the input level.

Example for Noise Suppression Characteristic



(2) Interference due to intermodulation

Interference due to intermodulation is caused when, in receiving are desired signal, more than 2 strong interfering waves arrive and a frequency equal to the desired signal frequency or IF frequency is generated in the receiver because of the nonlinearity of the receiver to disturb the reception of the desired signal (intermodulation between receivers) and when the interfering wave enters into the receiver through its final stage or from the antenna side to generate undesired signal wave of a new frequency with the desired signal (transmitter intermodulation).

In general, intermodulation is caused by the combination of two interfering waves and is due to the nonlinearity of the third odd order and is generally examined by the interfering input level usually in the 20dB noise suppression condition.

When

$$f_1 \text{ (interfering wave A)} = f_0 + \Delta f$$

$$f_2 \text{ (interfering wave B)} = f_0 + 2\Delta f$$

Where f_0 is the desired frequency, f_1 and f_2 cause nonlinearity distortion by the following relation.

$$2f_1 - f_2 = (2f_0 + 2\Delta f) - (f_0 + 2\Delta f) = f_0.$$

When this interference is caused, due consideration should be given to suppress the interference level as much as possible by a filter, etc., at the receiver input.

(3) Interference due to cross modulation

When the desired signal and interfering wave are both extremely large, the nonlinearity distortion of the receiver is received and the desired signal is modulated with the interfering signal, disturbing the reception. This interference is irrespective of the frequencies of the desired and interfering signals. In general, care should be exercised when the desired input signal level is more than 60dBμ and the interfering signal input level is more than 120 dBμ.

(4) Interference from the same frequency

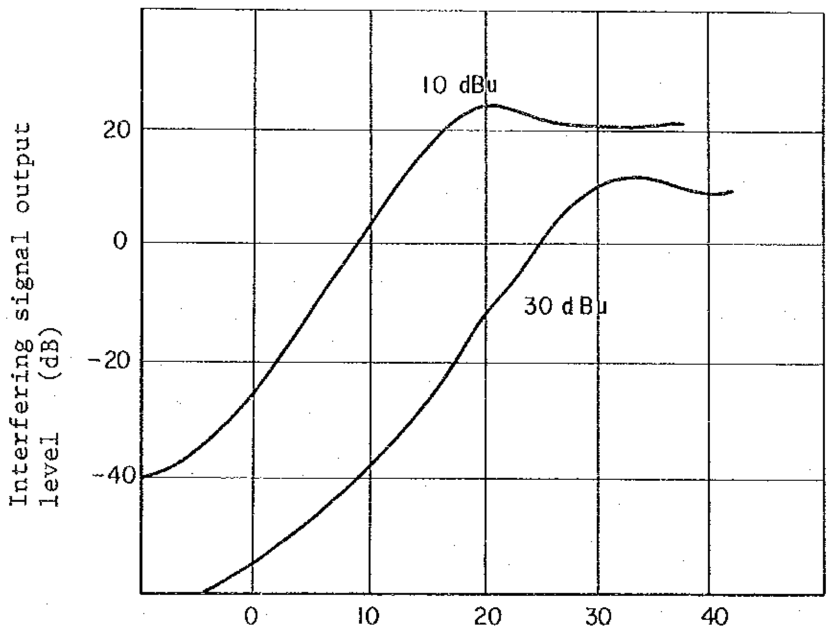
When the desired signal and interfering signal are at the same frequency, a level difference exceeding 10dB is usually required to prevent interferences.

When the levels of the desired and interfering signals approach each other, not only the mixture of communication but also beat due to frequency difference between the two signals are caused. When the interfering signal becomes dominant the desired signal is suppressed by the so-called masking effect.

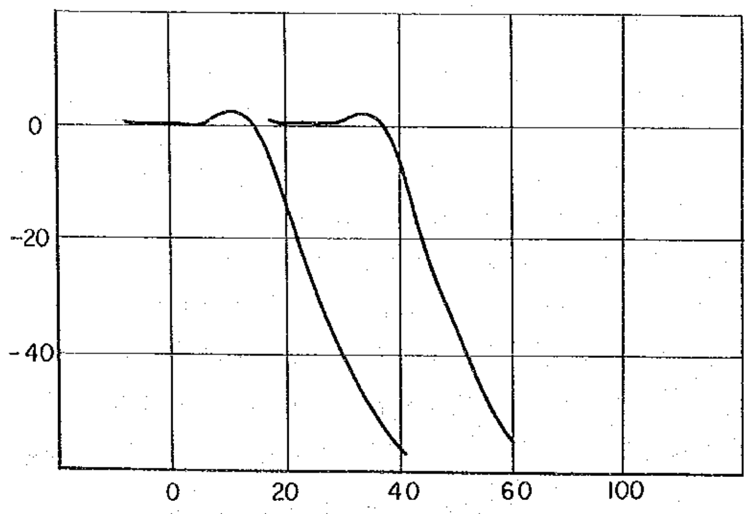
a) The figure on the right shows the influence of an interfering signal of a standard modulation (70%) at F_3 to the desired signal input at the same frequency.

b) The figure on the right shows the influence of an interfering signal at the same frequency (unmodulated) to the desired signal input and output.

10 dB (Desired signal input level)



An example of interference characteristic due to idencical frequency



An example of masking effect

(5) Disturbance due to spurious response

When the receiver incorporates a local oscillator such as super-heterodyne receiver, a number of undesired frequencies appear in the output because of the correlation with the IF frequency and others. The receiver sensitivity against these undesired frequencies is called spurious sensitivity and undesired frequency suppression capability determined by this sensitivity is called spurious response.

In radio communication regulation, 80dB (60 MHz band 150 MHz band) and 70dB (400 MHz band) are generally set out performance requirement for the spurious response, and the value given by

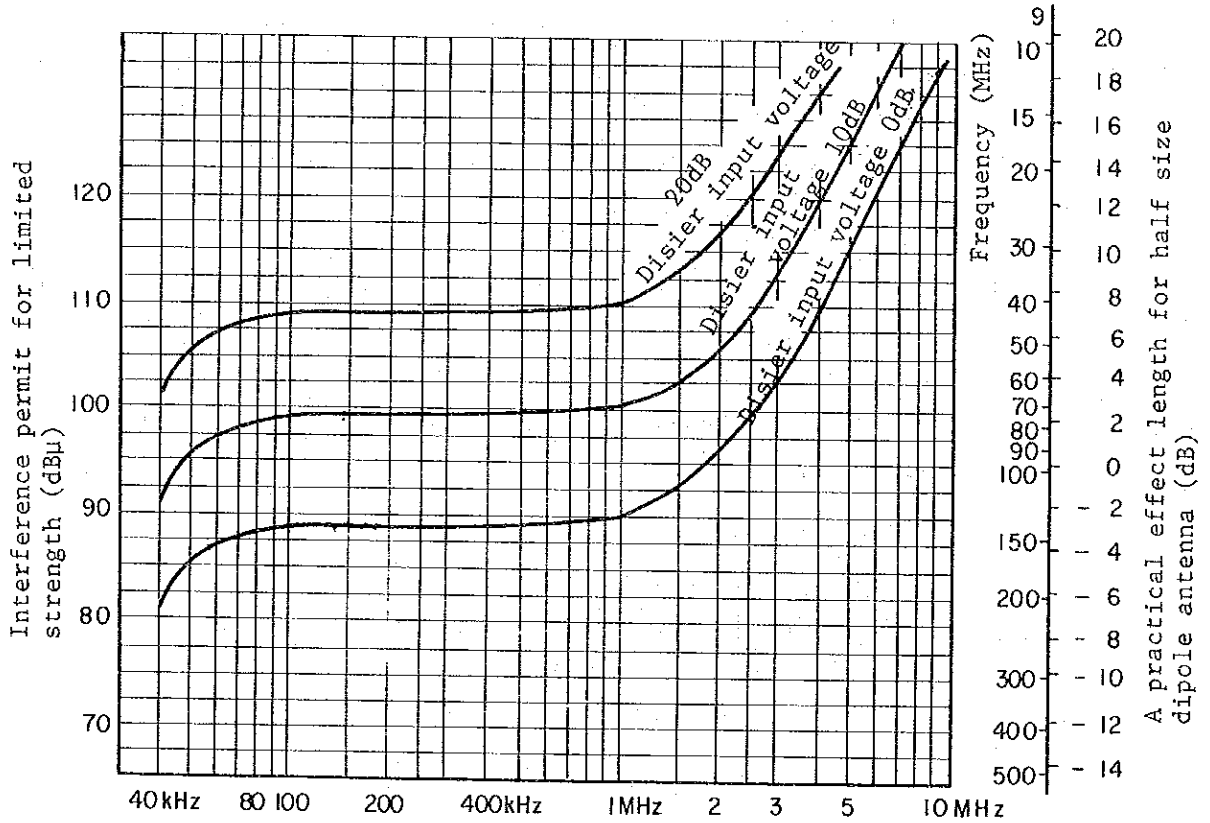
$$U(dB\mu) = Pr + Ps = 10dB$$

where Pr is the minimum standard desired signal input (dBμ) and Rs the spurious response should be examined as the allowable value of the spurious interfering input.

For spurious response, consideration should be given to the following plans.

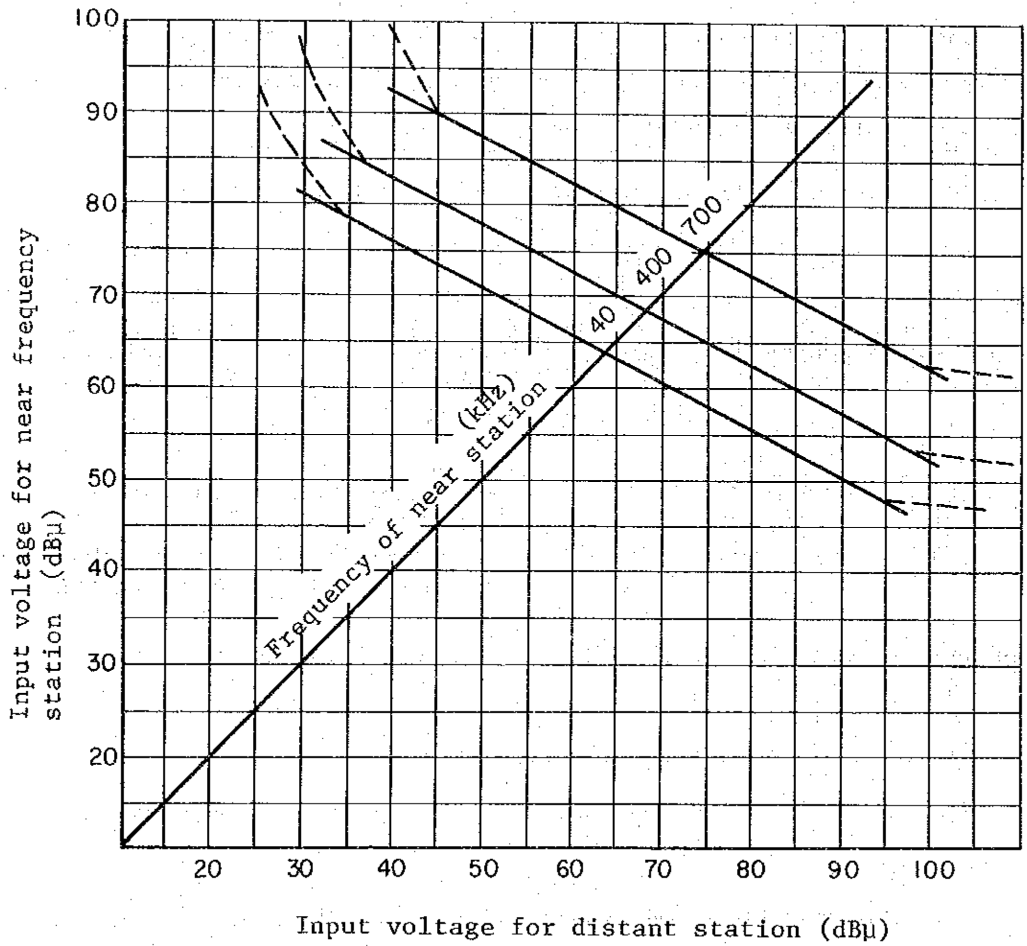
i) Selectivity for image frequencies
ii) Response for higher harmonics
iii) Interference by the same frequency as IF
iv) Multiplication of the above items

The curve of near frequency interference characteristic



Difference in frequency to interference station

The curve of inter moduration characteristic

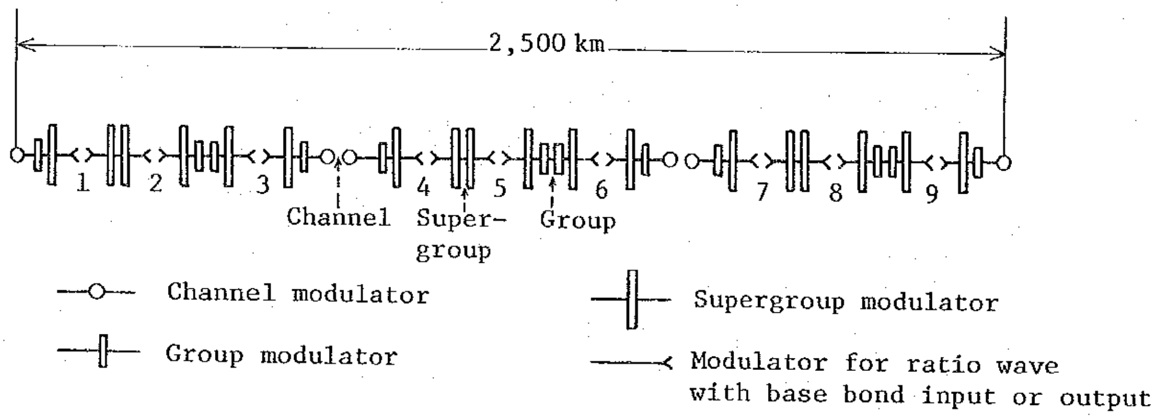


[II]-6 Hypothetical Reference Circuit and International Standards

1. Standard Hypothetical Reference Circuit

The standard hypothetical circuit is provided hypothetically to determine what communication system to use with what configuration or what communication system is used most frequently upon designing communication systems.

CCIR Recommendation 392 sets out the following hypothetical reference circuit for such long-distance microwave systems handling more than 60 channels as international systems.



As seen in the figure, this hypothetical reference circuit incorporates:

3 channel modulators,
6 group modulators, and
9 supergroup modulators

in a length of 2,500 km.

In the case of an OH system, the propagation course depends on the topographical condition and division into sections of given lengths is not employed as in a line-of-sight system and Recommendation 396 sets out: "When the propagation distance is L(km), as many sections as 2,500/L (integer) are connected in series, forming a 2,500 km system incorporating 3 channel modulators, 6 group modulators, and 9 supergroup modulators."

2. Allowable Noise Power

The following three recommendations are given for the system noise of frequency division multiplex telephone systems.

For the design of microwave systems, the noise standard for the hypothetical reference circuit (Recommendation 393) is employed. That is, Recommendation 393 specifies the following values for the 2,500km hypothetical reference circuit described in paragraph 1 above.

(1) One hour mean noise power for more than 5% of any month should not exceed a psophometrically weighted noise power of 7,500 pW.

(2) One minute mean power for more than 20% of any month should not exceed a psophometrically weighted noise power of 7,500pW.

(3) One minute mean power for more than 0.1% of any month should not exceed a psophometrically weighted noise power of 47,500pW.

(4) 5ms mean noise power for more than 0.01% of any month should not exceed unweighted noise power of 1,000,000pW.

This recommendation presents objectives for system design as stated before and actual circuits do not necessarily meet the hypothetical reference circuit but usually have different configuration, so that Recommendation 935 sets out noise standard for such actual circuits.

Recommendations on Allowable Noise

Recommendation	Title	Contents of Recommendation
393	Allowable noise in hypothetical reference circuit of frequency division multiplex telephone systems	See the test.
395	Allowable noise in actual circuits of frequency division multiplex telephone system	A. For actual circuits being nearly identical as the hypothetical reference circuit and having length L of 280 ∿ 2,500 km. 1. One hour mean noise power (weighted): less than 3LpW 2. One minute mean noise power (weighted for 20% of any month):

(Continued)

Recommen- dation	Title	Contents of Recommendation
395		less than 3LpW 3. One minute mean noise power (weighted for $\frac{1}{2500}$ x 0.1% of any month): less than 47,500pW B. Actual circuits with configuration different from hypothetical reference circuit 1. One hour mean noise power (weighted): less than (3L + X)pW 2. One minute mean noise power (weighted for 20% of any month): less than (3L + X)pW 3. One minute mean noise power (weighted for Y/2500 x 0.1% of any month): less than 47,500pW where X = 200pW (50km ≤ L ≤ 840km) 400pW (840km < L ≤ 2500km) 600pW (1670km < L ≤ 2500km) Y = 280km (L ≤ 280 km) L km (L > 280 km)
397	Allowable noise power in hypothetical reference circuit in transmitting frequency division multiplex telephone signal by OH system	a) One minute mean noise power weighted for 20% of any month: less than 25,000pW b) One minute mean noise power weighted for 0.5% of any month: less than 63,000pW c) 5ms mean noise power unweighted for 0.05% of any month (with large fading): less than 1,000,000pW

3. Other Standards

Recommendations are also presented for preemphasis characteristic, pilot frequency, intermediate frequency, frequency deviation per channel, baseband width (including pilot) for respective numbers of channels, and interface conditions at connecting points such as level and impedance, as given in Table 1.

For the interface between radio and wired systems, Recommendation 381 specifies that 308 ± 3 kHz should be used for the line regulating pilot, that the outband signal including this pilot should be less than -50dBm0 at the interface point, that the total level of outband signals at the radio equipment output should be less than -17dBm0, etc.

For maintenance requirements, recommendations given in Table 3 are set out for noise measuring test to be performed in routine maintenance for actual traffic and service channel.

Table 1. Recommendation for Transmission System

Recommendation	Title	Contents of Recommendation
275	Preemphasis characteristic	Relative deviation by test tone $= 5 = 10 \log_{10} \left[1 + \dfrac{\frac{6.90}{5.25}}{1 + (\frac{fr}{f} - \frac{f}{f_r})^2} \right]$ where f : Baseband frequency; f_r: Preemphasis resonance frequency
403	Intermediate frequency	1. For radio frequencies below 1 GHz: 35 MHz When the number of channels is large, 70 MHz may be used. 2. For radio frequencies above 1 GHz and up to 1800 channels: 70 MHz 3. For 2700 channels: 100 or 140 MHz
404	Frequency deviation for channel for respective channel capacity	Frequency deviation per channel corresponding to maximum number of channels of the system (see Table 2). It is recommended that the effective frequency deviation to be employed when preemphasis is employed should become equal to when not.

Table 2. Recommendations for Interface, etc.

(Recommendations 380 and 401)

No. of Channels	Frequency of Baseband (kHz)*	Impedance at Interface (Ω)	Relative level per channel at interface point		Frequency Deviation per Channel (rms/CH)	Continuity Pilot	
			Output (dB)	Input (dB)		Frequency (kHz)	Frequency Deviation (rms, kHz)
120	12 ∿ 552 60 ∿ 552	150, bal. 75, unbal.	-15	-45	50, 100, 200	607	25, 50, 100
300	60 ∿ 1,300 64 ∿ 1,296 (60 ∿ 1,364)	75, unbal.	-18	-42	200	1,499, 3,200 or 8,500	100 or 140
600	60 ∿ 2,540 64 ∿ 2,660 (60 ∿ 2,792)	75, unbal.	-20	-45	200	3,200 or 8,500	140
960	60 ∿ 4,028 316 ∿ 4,188 (60 ∿ 4,287)	75, unbal.	-20	-45	200	4,715 or 8,500	140
1800	312 ∿ 8,204 316 ∿ 8,204 312 ∿ 8,120 (300 ∿ 8,284)	75, unbal.	-28	-37	140	9,023	100
2700	312 ∿ 12,388 316 ∿ 12,388 312 ∿ 12,388 (308 ∿ 12,435)	75, unbal.	-28	-37	140	13,627	100

Table 3. Recommendation for Maintenance Conditions

Recommendation No.	Title	Commendts of Recommendation
290	Routine maintenance necessary frequency multiplex system	1. Check of level variation of the line by line regulating pilot 2. Measurement and check of line noise by noise measuring channel and distortion test
398	Noise measurement during operation	Measurement of line thermal noise and distortion noise during operation by providing a measuring channel outside the telephone signal band (at 10% apart from the upper and lower limits).
399	Measurement of line quality by using flat noise	Measurement of thermal noise and distortion noise by using a measuring channel spacified in Recommendation 398 while loading flat noise depending on the channel capacity.
400	Service channels to be provided	1. Channels Channels for orderwire between attended stations, channels for orderwire between attended and unattended stations, switching pilot channels, supervisory channels 2. Transmission methods Independent wires or radio channels (with the same frequency band as the main system or other systems)

[II]-7 Noise Design

1. Noise in Radio Systems

Types and causes of noise produced in multiplex telephone systems can be classified as follows.

- Thermal noise
  - Noise determined by receiving power
  - Noise proper to modulator/demodulator, video amplifier, carrier terminal, etc.
- Distortion noise
  - Nonlinear distortion
    - Noise due to FM modulator/demodulator
    - Noise due to video amplifier (including group amplifier at carrier terminal)
  - Linear distortion
    - Delay distortion due to phase characteristic of IF and RF stages, etc.
    - Amplitude distortion due to amplitude response characteristics of IF and RF stages
    - Echo distortion due to standing wave of antenna system
    - Transmission distortion due to close reflection and diffraction in transmission line
- Interference
  - Interference from channels with identical frequency (Antenna front-back coupling, overreach, etc.)
  - Interference from channels with different frequencies (Antenna side-side coupling, back-back coupling, etc.)
- External noise
  - Mainly city noise

2. Thermal Noise

Thermal noise in its narrow meaning is thermally distorbing noise caused by resistors but in general it is used in comparison with crosstalk noise and interference.

In this latter meaning, all noise produced within a receiver and external noise applied through the antenna are generally termed thermal noise.

Thermal noise caused in a resistor is ascribable to the thermal movement of electrons within the resistor and the noise power N_0 is given by

$$N_0 = KTB$$

where K: Boltzmman's constant (1.38×10^{-23} J/°K)

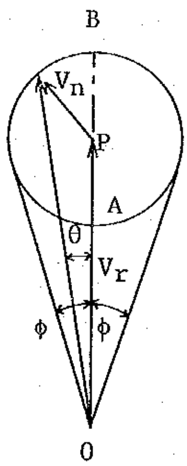
T: absolute temperature

B: Frequency band

This thermal noise is the minimum noise caused in the ideal case and in order to reduce this noise, it is inevitable to lower the temperature.

In general receivers, noise is caused in vacuum tubes and transistors incorporated in the receiver in addition to the above-mentioned thermal noise.

So, noise figure is defined so as to express how many times the thermal noise of the above-mentioned ideal resistor the noise of an actual receiver is. When $\vec{V_r}$ and $\vec{V_n}$ are the vectors of the carrier and noise, respectively, and their amplitudes V_r and V_n, respectively, superposition of the carrier vector is shown in the left-hand figure.



That is, carrier vector $\vec{V_r}$ rotates arround origin 0 at angular velocity $\omega_c = 2\pi f_c$ and noise noise fector $\vec{V_n}$ rotates arround the tip of the carrier vector, P, at angular velocity $\omega_n = 2\pi f_n$.

Accordingly, if the carrier vector is considered stop, only the noise vector $\vec{V_n}$ rotates arround point P at an angular velocity of $\omega_n - \omega_c = 2\pi(f_n - f_c)$.

It is clear from this vector representation that the amplitude of the resultant vector of the carrier and noise vectors varies in an $\overline{OA}$ to $\overline{OB}$ range and its phase in a $-\phi$ to ϕ range.

In other word, the carrier wave undergoes amplitude modulation with modulating wave V_n/V_r at a frequency of $(f_n - f_c)$ and phase modulation at modulation degree ϕ, at a time. As the amplitude of the noise, V_n, can be considered negligibly small when compared with the amplitude of the carrier, V_r, we have

$$\phi = \tan^{-1}\frac{V_n}{V_r} = \frac{V_n}{V_r}.$$

The instantaneous phase value of the resultant vector, θ, is then given by

$$\theta = \frac{V_n}{V_r}\sin 2\pi(f_n - f_c)t.$$

In an FM demodulator, the variation of the phase, that is, the output proportional to the frequency deviation, is taken out and the noise output corresponding to the above-mentioned phase variation is proportional to

$$f_d = \frac{1}{2\pi}\frac{d\theta}{d_t} = \frac{V_n}{V_r}(f_n - f_c)\cos 2\pi(f_n - f_c)t.$$

Most thermal noise in ordinary microwave systems is caused in the receiver and its power per unit frequency band, P_n, is given by

$$P_n = KTF$$

where F denotes the noise figure.

Accordingly, the noise power to be contained in the frequency band of Δf at frequency f is proportional to

$$\frac{DTF\Delta f}{P_r} f^2.$$

If the frequency deviation of the signal (effective value) is S_0, the signal output is proportional $S_0{}^2$ with the same coefficient (demodulator sensitivity) and the signal-to-noise ratio (S/N) is given by

$$[S/N]\ dB = -10\ \log_{10} \frac{KTF\Delta f}{P_r} \cdot \frac{f^2}{S_0{}^2} \cdot$$

If the bandwidth of a telephonechannel is taken for Δf and the effective frequency deviation against the test tone level for S_0, the S/N per channel can directly be obtained.

Some other factors that can be considers to give influence to thermal noise are the sum of thermal noises in multi-stage repeatering and additional thermal noise due to fading, and due consideration should also be given to these items of thermal noise upon actual system design.

3. Distortion Noise

(1) Distortion noise due to nonlinearity distortion

Nonlinearity distortion is such type of distortion that is caused when the signal passes vacuum tubes and transistors and occurs only when the input-to-output relation is not linear or not proportional. Output voltage E can be expressed as a series of input voltage e as follows.

$$F = a_0 + a_1 e + a2e^3 + a3e^3 + \ldots\ldots\ldots$$

Simple sinusoidal wave of e = $A\cos P_t$ is applied as the input voltage to the circuit that can be expressed as above, the output voltage obtain is given by

$$E = a_0 + a_1 A\ \cos\ P_t + a_2 (A\ \cos\ P_t)^2 + a_3 (A\ \cos\ P_t)^3 + \ldots\ldots\ldots$$

$$= a_0 + a_1 A\ \cos\ P_t + a_2 A^2 (\frac{1+\cos\ 2P_t}{2}) + a_2 A^2 (\frac{\cos\ 3P_t + 3\cos P_t}{4}) + \ldots$$

That is, a sinusoidal wave with the same frequency as the input (fundamental wave) is obtained from the term containing A, sinusoidal wave with a frequency twice as large as the input frequency (second harmonic) from the term containing A^2, and a sinusoidal wave with a frequency thrice as large as the input frequency (third harmonic) from the term containing A^3. Accordingly, harmonic distortion factors can be defined by the ratios of the amplitude of the fundamental wave to those of higher harmonics, as follows.

$$\text{Second harmonic distortion: } K_2 = \frac{a2A}{2a1}$$

$$\text{Third harmonic distortion : } K_3 = \frac{a3A^2}{4a_1}$$

It can be understood from the above equations that the second harmonic distortion factor degrades in proportion with the input and the third distortion factor in proportion to the square of the input. This is a fundamental property of distortion in which when the input is raised, the desired output also rises but the distortion voltage increases much rapidly.

When two sinusoidal waves are applied to the input at a time, similar harmonic and resultant waves are produced, as is clear by calculation. Thus, a great numbers of resultant waves are produced because of a number of frequency components contained in the signal, producing noise with a spectrum with respect to frequency.

This noise is called intermodulation noise.

When a noise having a continuous spectrum W(f) in a $0 \sim B$ frequency range is applied at the modulator input, the intermodulation noises of arithmetic summation form per unit bandwidth at frequency f are:

$(\omega_1 + \omega_2)$ type: $\quad a_2^2 \int_0^f \omega(f-x)dx$

$(\omega_1 - \omega_2)$ type: $\quad 2a_2^2 \int_0^{B-f} \omega(x) \cdot \omega(f+x)dx$

$(\omega_1 + \omega_2 + \omega_3)$ type: $\quad \frac{3}{2}a_3^2 \int_0^f \int_0^{f-y} \omega(x) \cdot \omega(y) \cdot \omega(f-x-y)dx \cdot dy$

$(\omega_1 + \omega_2 + \omega_3)$ type: $\quad \frac{9}{2}a_3^2 \int_0^{B-f} \int_0^{f+y} \omega(x) \cdot \omega(y) \cdot \omega(f-x+y)dx \cdot dy$

In this case, the desired signal is the transmission output proportional to input signal $\omega(f)$ and can be represented by a $a_1^2\omega(f)$, so that the signal to intermodulation noise ratio can be obtained by obtaining the ratio of the intermodulation noises given by the above equations.

Causes for producing distortion which may cause intermodulation noise in microwave systems using frequency modulation are given in paragraph 1.

(2) Linear distortion

When a signal frequency modulated by a multiple signal or a like is sent through a microwave repeater line, the frequency-modulated signal obtained after demodulation becomes different from the initial signal due to the amplitude frequency response of the FM transmission system and phase frequency characteristic of the transmission line.

Linear distortion is such distortion as caused in such a demodulated signal.

Although the circuit which causes nonlinear distortion exhibits nonlinearity in the input-to-output level ratio in the case of nonlinear distortion mentioned in the foregoing paragraph, the circuit which causes linear distortion exhibits linearity in the input-to-output level ratio or the presense of nonlinearity gives no influence to distortion but the amplitude frequency response characteristic or phase frequency response characteristic determines the amount of distortion in the case of linear distortion.

This can be said when only the FM transmission line is picked up singly to be discussed. However, when only the modulator and demodulator are connected to the transmission line and all circuits ranging from the modulator input to the demodulator output are considered, the transmission line involving linear distortion causes nonlinearity in the relationship between the modulator input and demodulator output, so that it can be understood that the distortion which appears in the output wave form is caused by this nonlinearity. From this principle, the differential gain and differential phase characteristics are used to express the non-

nonlinearities of FM transmission lines.

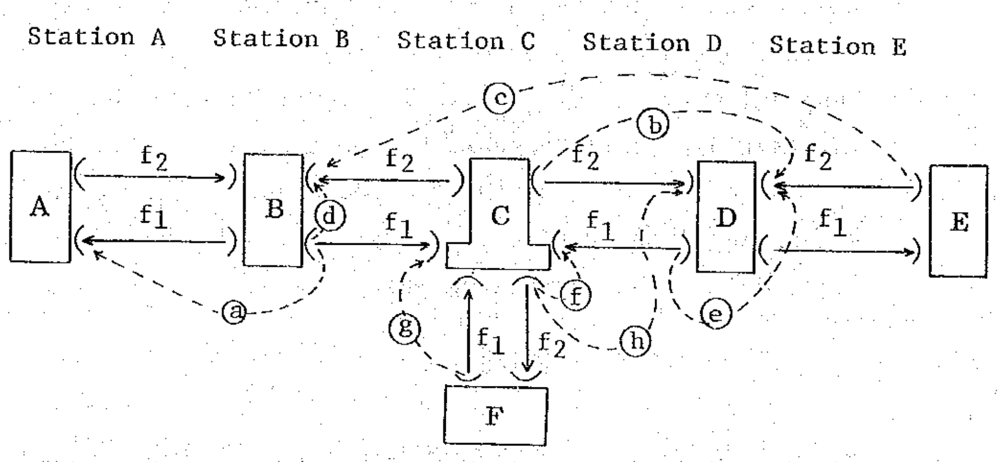
From these characteristics intermodulation noise due to linear distortion can be obtained.

4. Interference Noise

In microwave systems using frequency modulation, two radio frequencies are used in general.

Accordingly, it can be observed at a repeater station that the same frequency is sent from both directions and the same frequency is received in both direction.

Antennas used in microwave systems have sharp directionality but have certain degrees of directionality toward the sides and rear. Accordingly, the following interference is caused by radio frequencies.



(1) Interference within local circuit

Interference within the same frequency channel:

- a) Front-back coupling of transmitting antenna
- b) Front-back coupling of receiving antenna
- c) Overreach

Interference from different frequency channels

- d) Side-side coupling of antenna
- e) Back-back coupling of antenna

(2) Interference from branch circuits

- f) Side-side coupling of antenna
- g) Front-side coupling of receiving antenna
- h) Front-side coupling of transmitting antenna

(Interference from branch circuits exists in form of interference at the same frequency and interference from different frequencies.)

In order to reduce interference coupling between antennas, it is necessary to employ such antennas that sufficiently reduce gains at angles outside the main beam direction.

It is to be noted that in the front-back interference interfering waves transmit on different channels then the desired signal channels and the desired signal may become weak due to fading, etc., or interfering waves may become strong because of free space condition, so that it is necessary to consider this differential fading upon examining interference.

It is also to be noted upon calculation of interference noise that when the desired signal and interfering wave are both unmodulated and of the same frequency, a beat with a frequency corresponding to the frequency difference between the desired and interfering signals appears in the demodulator output.

When the ratio of the desired signal level to the interfering signal level is [D/U](dB), the signal-to-interfering noise ratio [S/I](dB) in the channel on which the beat falls is given by

$$[S/I] = [D/U] - 20 \log \frac{(Fd - Fu)}{\sqrt{2}S}$$

where F_d and F_u respectively denote the carrier frequencies of the desired and interfering signals and S the frequency deviation at the test tone level.

When the desired signal or interfering signal (or both) is modulated and comes to have a spectrum, the interfering component which has been a beat when unmodulated becomes interfering noise with a certain spectrum.

In general, when the power-frequency stectra of the desired and interfering signals, $\omega_d(F)$ and $\omega_u(F)$, are known, the spectrum of the interfering noise after demodulation, $\omega i(f)$, can be obtained by integrating the beat of the frequency components corresponding to $\omega_d(F)$ and $\omega_u(F)$.

That is, in the case of frequency modulation,

$$\omega_i(f) = \frac{f^2}{2P_d^2} \int_{-\infty}^{\infty} \omega_d\ (x)\ \omega_u\ (x+f)\ dx$$

Where P_d is the total power of the desired signal and given by

$$P_d = \int_{-\infty}^{\infty} \omega_d\ (x)\ dx$$

When the modulation degree of the frequency modulation is extremely large, this integration can be obtained comparatively easily.

That is, the power-frequency spectrum of a FM signal modulated with random noise is verified to have normal distribution with the effective frequency deviation being the standard deviation when the maximum frequency of the baseband is sufficiently small when compared with the effective frequency deviation.

Accordingly, we have, by using the above integration.

$$\omega_i(f) = \frac{1}{2} \frac{P_u}{P_u} \frac{f2}{\sqrt{2\pi(\sigma d^2+\sigma u^2)}} e^{-\frac{(f-fo)^2}{2(\sigma_d^2+\sigma_n^2)}}$$

Where P_u, and P_d are respectively the total powers of the desired and interfering signals, σ_d and σ_u are respectively the effective frequency deviations of the desired and interfering signals, and f_0 the frequency difference between the desired and interfering signals.

In the case of interference at the same frequency due to front-back coupling in the same circuit, etc., we have

$$\sigma_d = \sigma_u = \sigma \text{ and } f_0 = 0,$$

so that

$$\omega i(f) = \frac{1}{2} \frac{P_u}{P_d} \frac{f^2}{2\sigma\sqrt{\pi}} e^{-\frac{f^2}{4\sigma^2}}$$

Hence, when the test tone frequency deviation is S, the signal-to-interfering noise ratio is given by

$$[S/I]dB = [D/U] + 10 \log \frac{f2}{S2} \frac{1}{4\sqrt{\pi\sigma}} e^{-\frac{f^2}{4\sigma^2}}$$

5. Noise Assignment

When designing a microwave system for use in certain transmission line, it is necessary to allot standard noise assigned to the transmission line to various noise sources and determine equipment specifications so as to meet respective noise values thus allotted to equipment.

The system design of a microwave system is represented by this noise allotment and the features of the microwave system thus designed can be characterized by the noise allotment.

Furthermore, the system economy is determined by the results of noise allotment.

(1) Basic concept

The standard noise of 7500pW specified for a 2500 hypothetical reference circuit is equal to -51.2dBm relative to 1mW. It is equal to 51.2dB in signal-to-noise ratio for the test tone. This value of 51.2dB is a psophometrically weighted value. This value is equivalent to 48.7dB in unweighted value.

In alloting this value to 9 sections composing the hypothetical reference circuit, conversion of baseband configuration is performed by carrier terminal equipment at the respective connecting points between sections and noise caused between sections is all summed up in form of power summation and noise is allotted by 1/9 equally to all sections. That is, the noise allotment to one modulation section of 280 km is

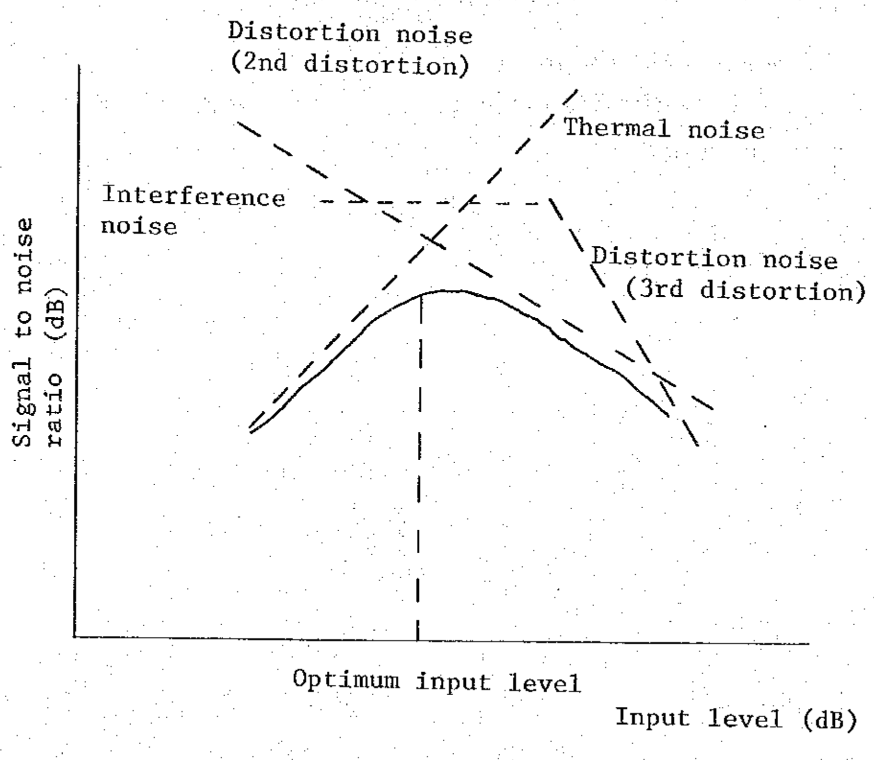
$$48.7 + 10 \log_{10} 9 = 58.2\text{dB}$$

in dividing this noise value into three types of noise, that is, thermal noise, distortion noise (intermodulation noise), and interference noise, the following property of each noise against input level variation must be considered. That is, while the signal-to-ratio due to thermal noise is improved in proportional with the input level (that is, effective frequency deviation) the signal-to-noise ratio due to distortion noise degrades inversely proportional to the input level (in the case of secondary distortion) or in higher ratios (in the case of distortion higher than third order distortion).

On the other hand, the signal-to-noise ratio due to interference noise generally exhibits complecate variation with input level variation.

In the case of interference from the same frequency in a system with a high degree of multiplication, it can be considered that the signal-to-noise ratio does not very greatly with the input level. Accordingly, it is understood, by measuring the signal-to-noise ratio of any microwave system by loaded noise characteristic test, that the signal-to-noise ratio varies as shown in the following figure with the input level.

Signal-to-noise ratio vs. input level



It is now understood that there is such an input level at which the signal-to-noise ratio become maximum and it should be operated at that input level to achieve optimum sufficiency.

In microwave systems, noise caused in the FM section is most dominant and the input level variation in the figure can be considered as the variation of test tone frequency deviation.

The signal-to-noise ratio becomes optimum at a point where the thermal noise and distortion noise are nearly equal to each other. It is also advantageous that at this point the signal-to-noise ratio hardly varies with the input level.

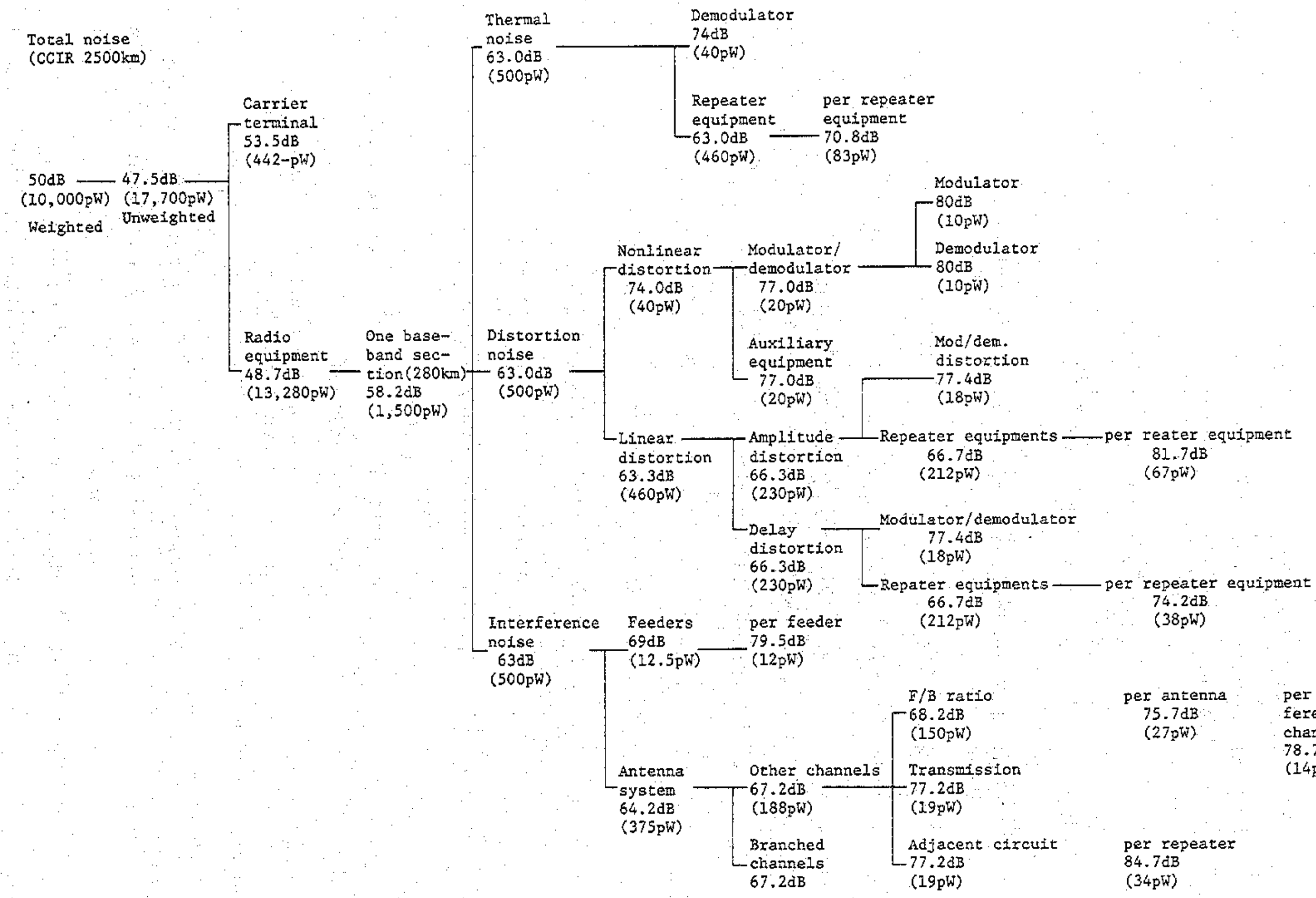
Interference noise need not be directly related with other types of noise but since it can be considered that the predominance of one noise requires rather strict control of other types of noise, nearly the same amount of noise as alloted to thermal noise and distortion noise is usually alloted to interference noise as well. Of the signal-to-noise ratio is 58.2dB for each modulation section, 63dB is thus alloted to each of thermal, distortion, and interference noise.

This amount of noise allotment is still divided to be alloted to the respective causes of noise.

(2) Noise allotment table

A typical noise allotment table for a 6 GHz band system is given below.

Noise Allotment (6 GHz System)



[II]-8 Calculation of Occupied Frequency Band Width

1. Equations for Calculation of Occupied Frequency Band

(1) 29.7 MHz ∿ 470 MHz

a) Single channel (in the case of FM)

$$f_B = 2fm \qquad (m < 1)$$

$$f_B = 2f_d + 2fm \quad (1 \leqq m < 10)$$

$$f_B = 2f_d + 4fm \quad (10 \leqq m)$$

b) Multichannel (in the case of SS-FM)

$$f_B = 2fm \qquad (m < 1)$$

$$f_B = 2(L_c, f_{do}) + 2fm \quad (1 \leqq m < 0)$$

$$f_B = 2(L_c, f_{do}) + 4fm \quad (10 \leqq m)$$

c) Multichannel (in the case of SS-PM)

$$f_B = 2fm \qquad (m < 1)$$

$$f_B = 2L_c(f_x \cdot mc) + 2fm \quad (1 \leqq m < 10)$$

$$f_B = 2L_c(f_x \cdot mc) + 4fm \quad (10 \leqq m)$$

f_d : Maximum frequency deviation

f_{do}: Effective frequency deviation (kHz) caused at test tone level

L_c : Maximum load coefficient corresponding to the number of channels

fx : $fx = \sqrt{\dfrac{fa^2+fa \quad fb+fb^2}{3}}$

fa : Lowest frequency of baseband (0 when there is orderwire circuit which modulates as it is the voice band)

fb : Highest frequency of baseband

m_c : Effective phase deviation (radians) caused at test tone level

(2) More than 890 MHz

1) In the case of SS-FM

a) When the channel capacity is less than 120 channels:

$$f_B = 2(L_c \cdot f_{do}) + 2f_m \qquad \text{When } f_b > 552 \text{ kHz}, \; f_B = 552 \text{ kHz}.$$

$$= 2f_d + 2f_m$$

b) Exceeds 120 channels:

When $L_n \leqq \frac{1}{2.6}\left(\frac{552}{f_{do}} + 10\right)$: $f_B = 20f_{do} + 1104$

When $L_n > \frac{1}{2.6}\left(\frac{552}{f_{do}} + 10\right)$: $f_B = 2 \times 2.6\, L_n \cdot f_{do}$

c) When preemphasis is employed:

To be obtained by equations mentioned in items a) and b) above but by using f''_{do} instead of f_{do} and

$$f''_{do} = f_{do}\sqrt{\frac{1}{N}\sum_{i=1}^{N} f_{di}^{2}}$$

(When 8dB preemphasis as per CCIR Recommendation 275 is employed and the system is fully equipped, this compensation of $f_{do} \rightarrow f''_{do}$ is not required.)

2) In the case of SS-PM

As in the case of SS-FM but $f_x \cdot m_c$ is instead of f_{do} and

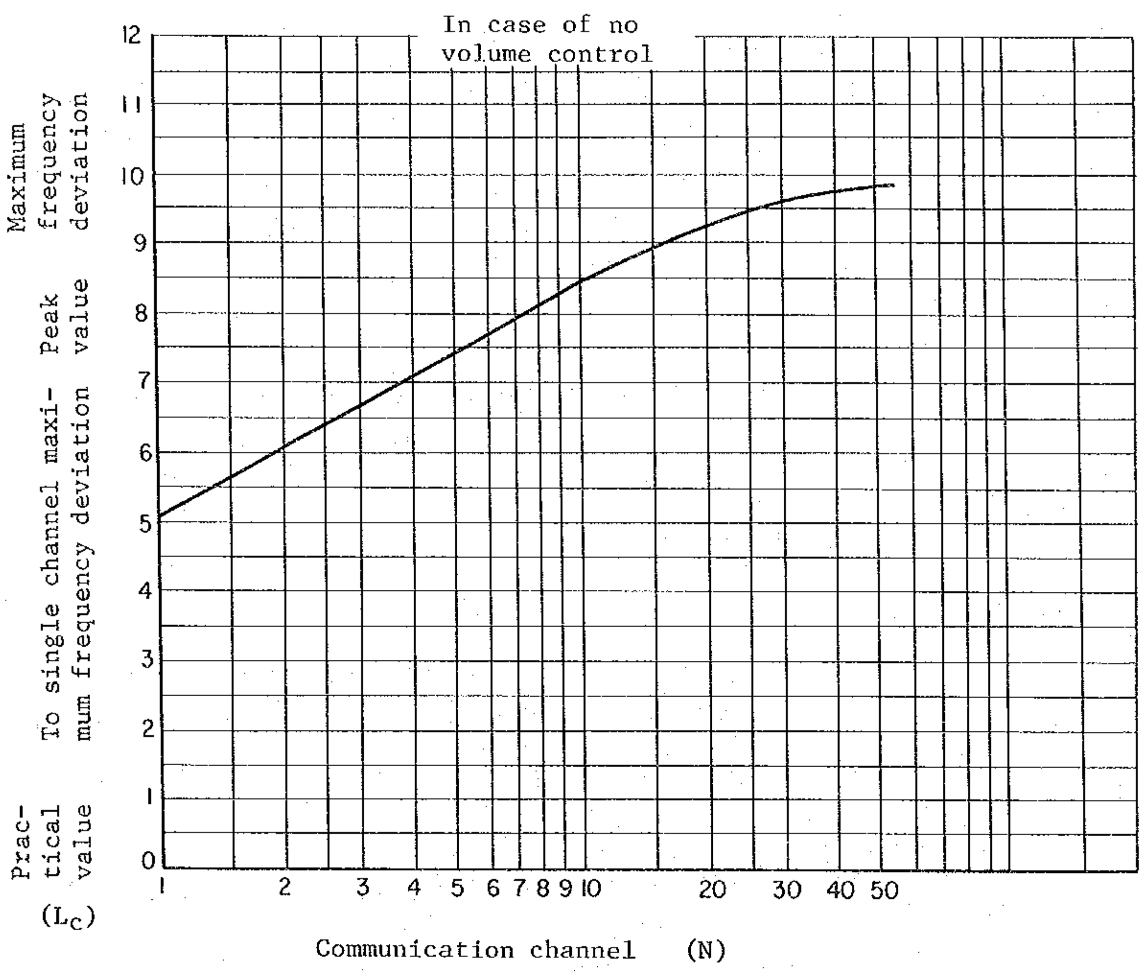
N : Number of channels

f_{di} : Relative effective value of frequency deviation on each channel relative to the effective value of the frequency deviation caused at the test tone level (0dB)

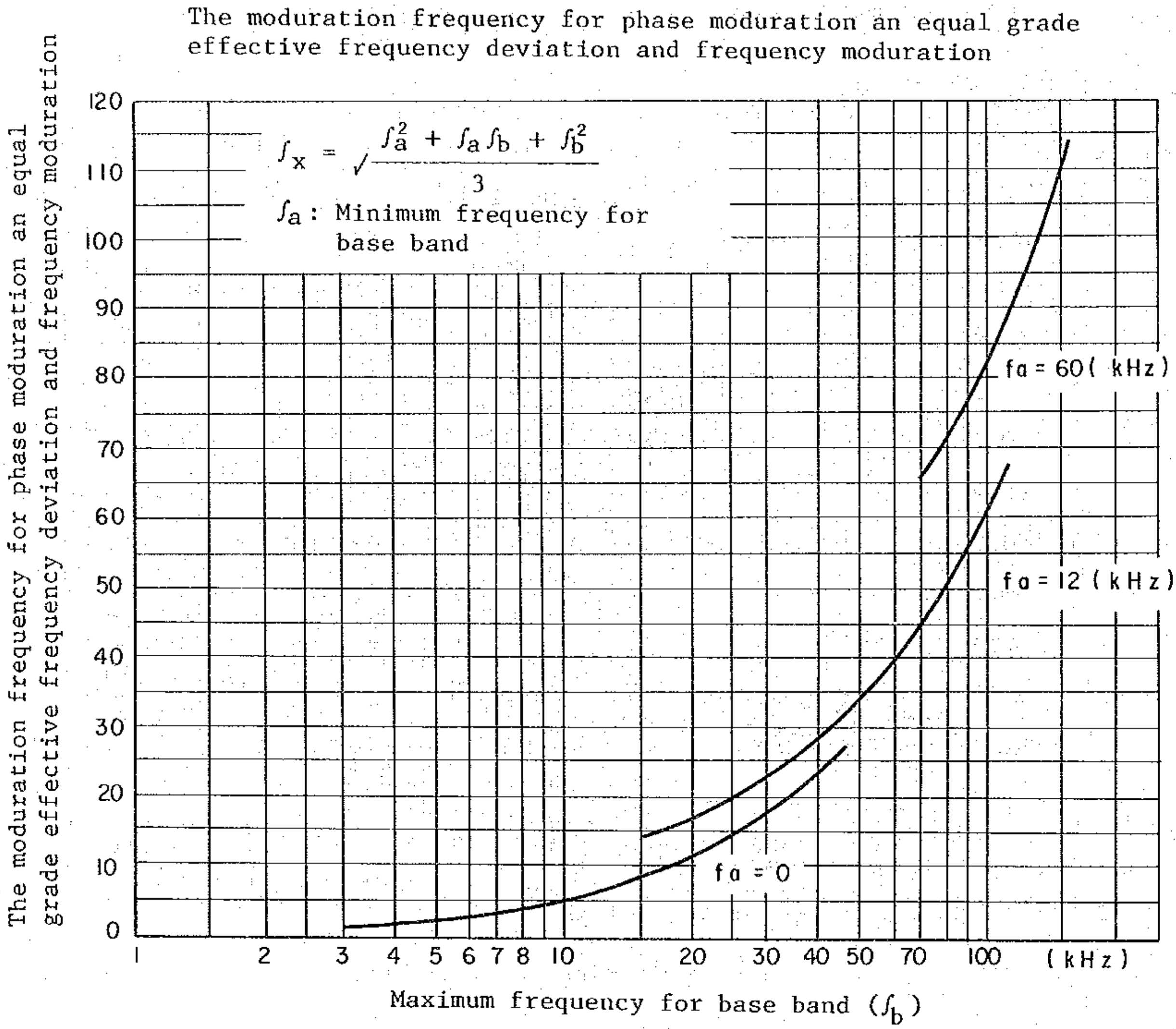
L_n : Equivalent power corresponding to the number of channels

L_c : Maximum load coefficient

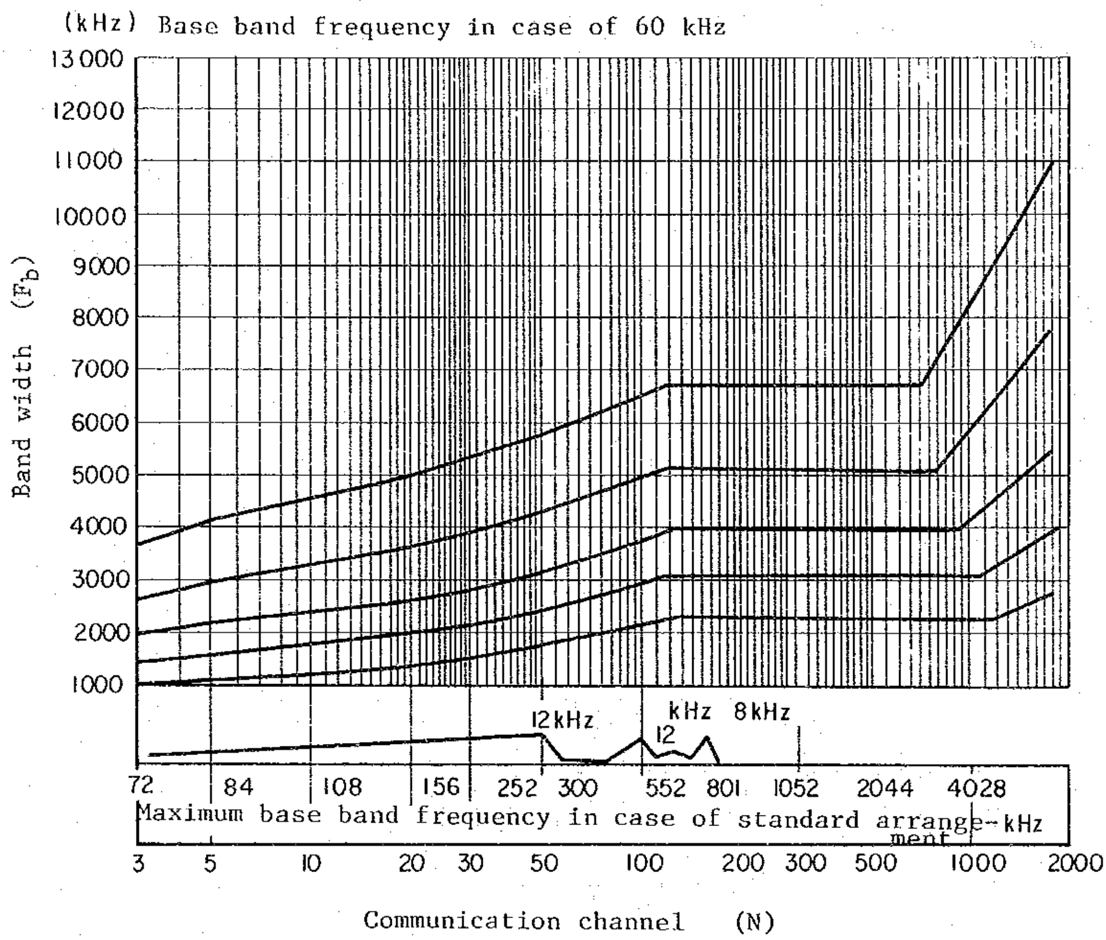
The ratio for maximum frequency deviation between single channel maximum frequency deviation (Maximum load coefficient)



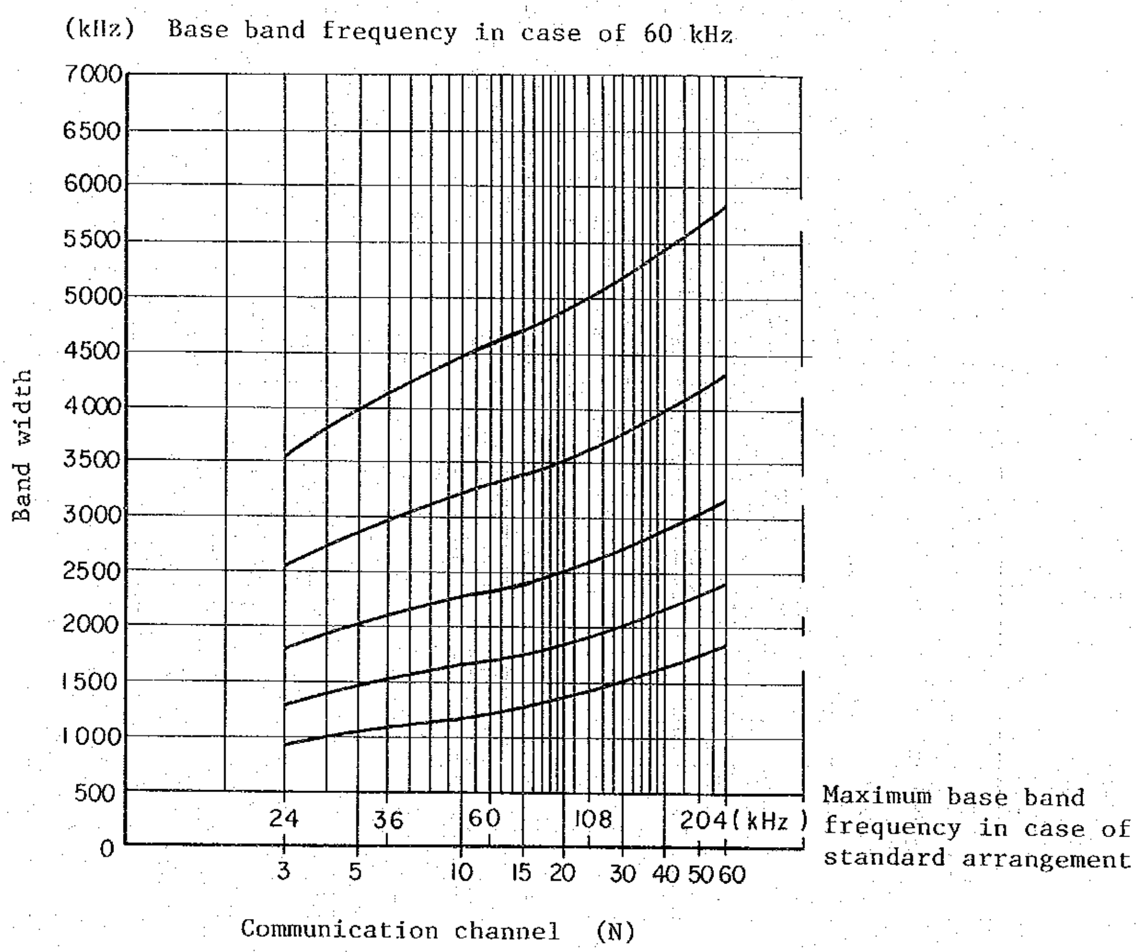
The moduration frequency for phase moduration an equal grade effective frequency deviation and frequency moduration



The calicuration for band width (SS-FM and SS-PM)



The calicuration for band width (SS-FM)



The curve of equal volume and maximum load coefficient

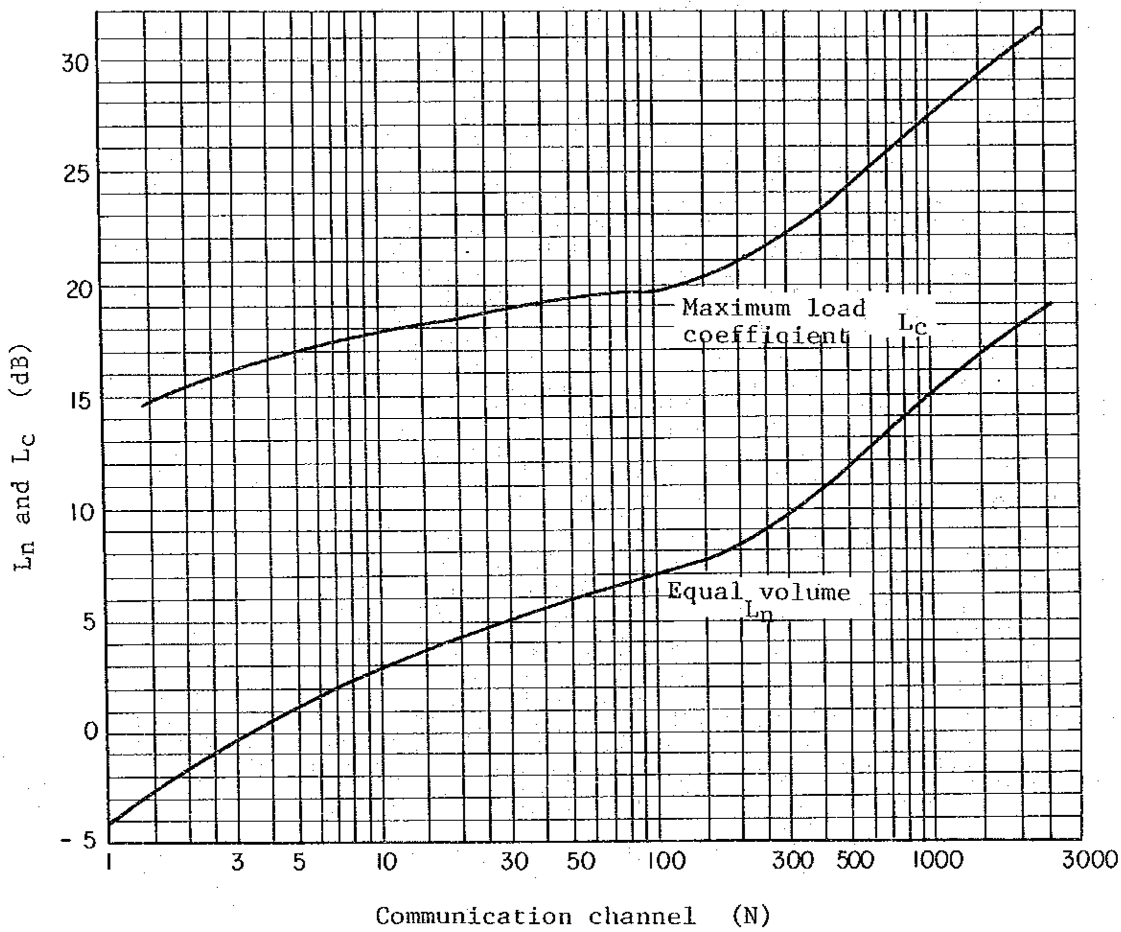
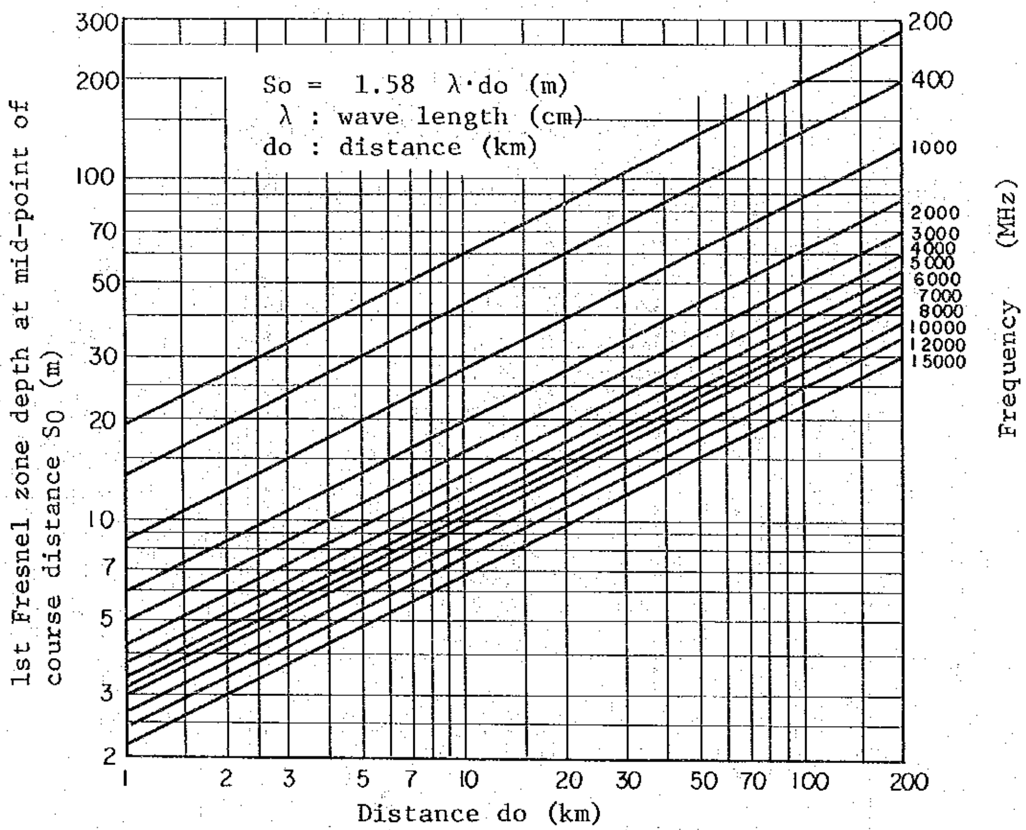
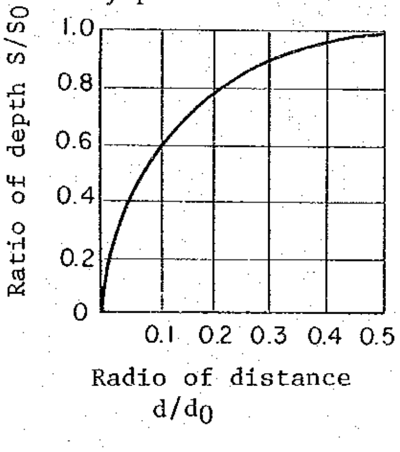
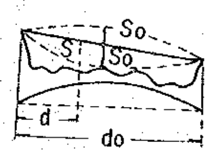


Chart for depth of 1st Fresnel zone



1st Fresnel zone depth at any point





S : 1st Fresnel zone depth at any point (m)

d_0: course distance

d : distance to nearer side (km)

Free space loss by iso-tropic gain ant.

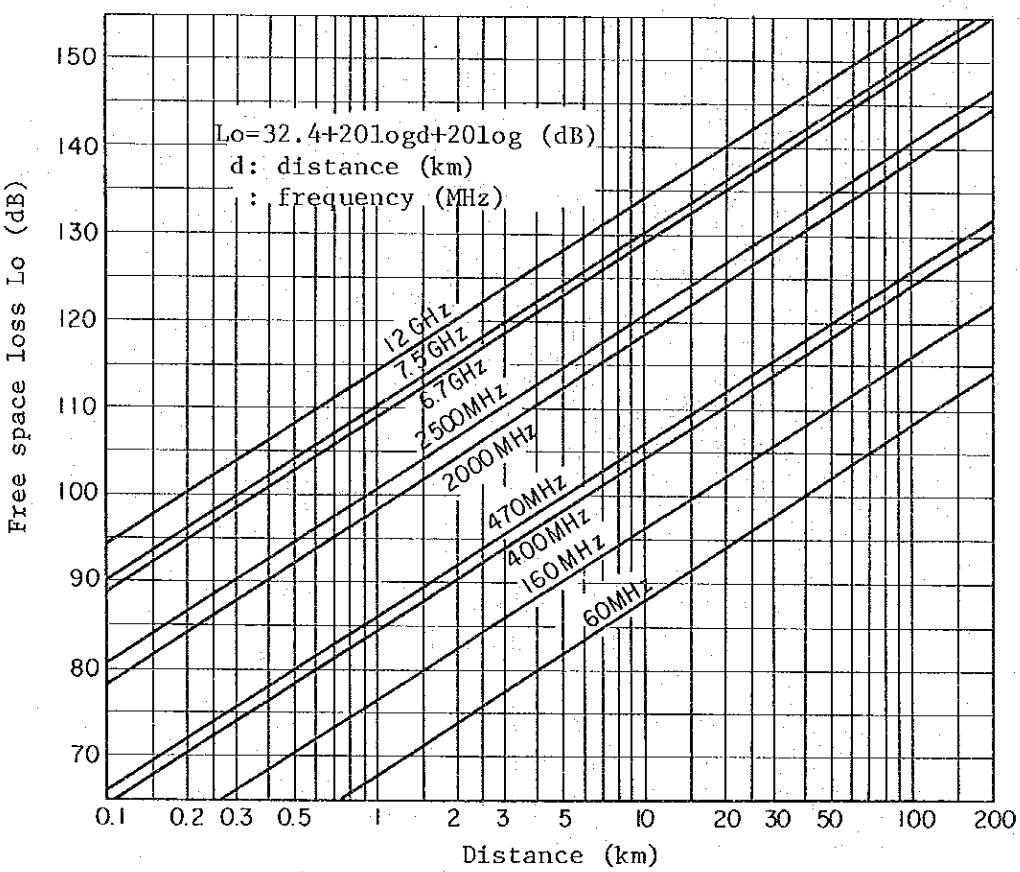
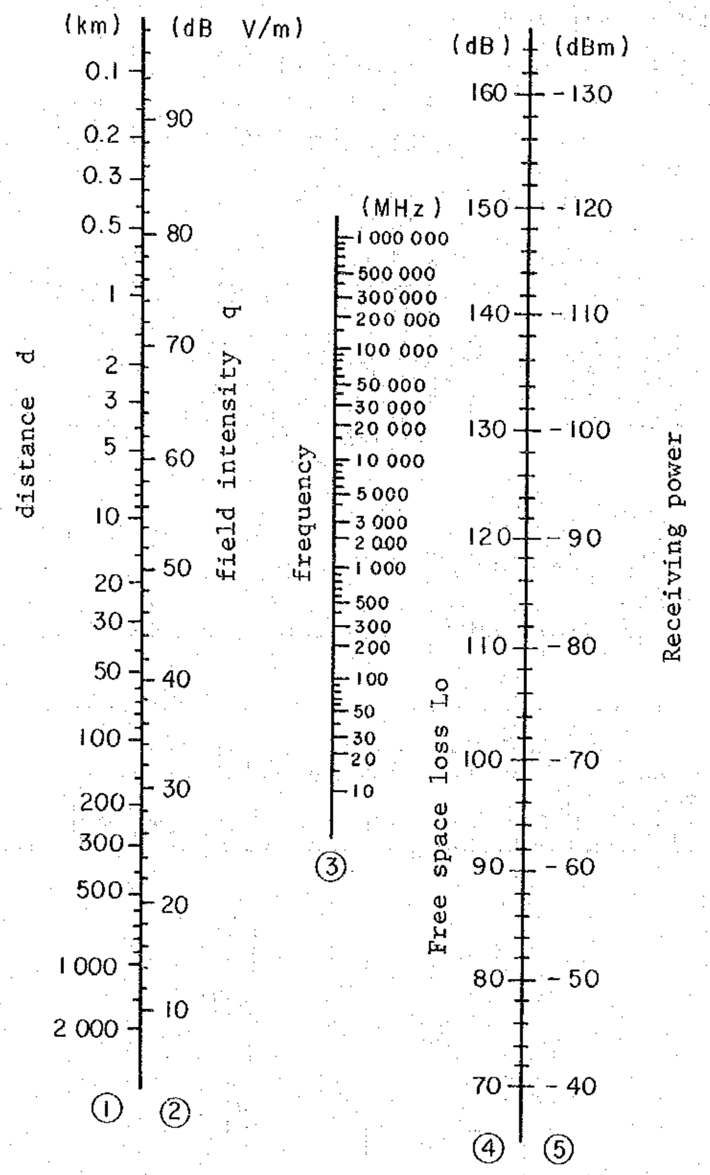


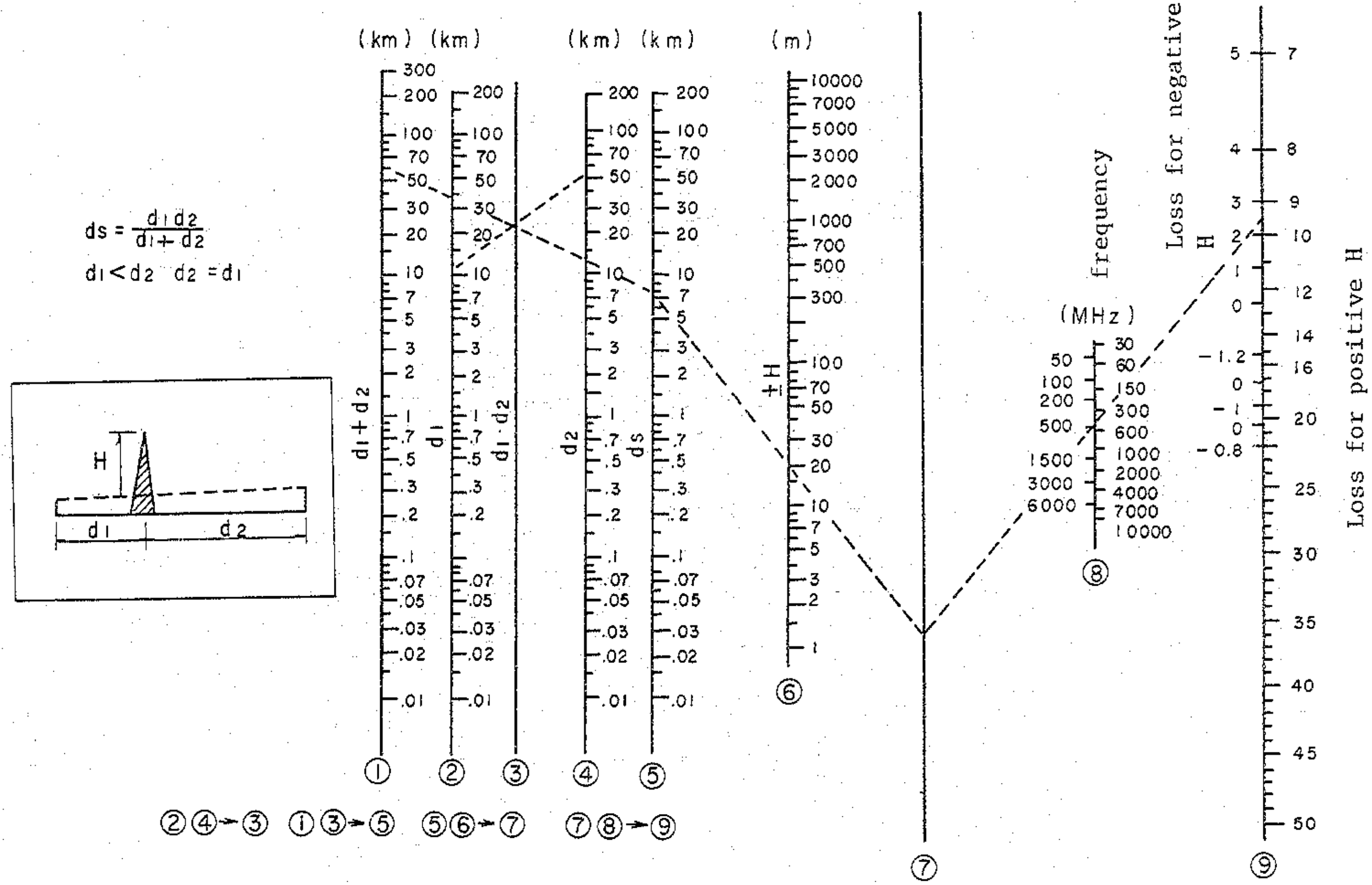
Chart for free space loss, field intensity and receiving power

Lo = 32.4 + 20logd + 20log
Lo : free space loss
d : distance
: frequency (MHz)

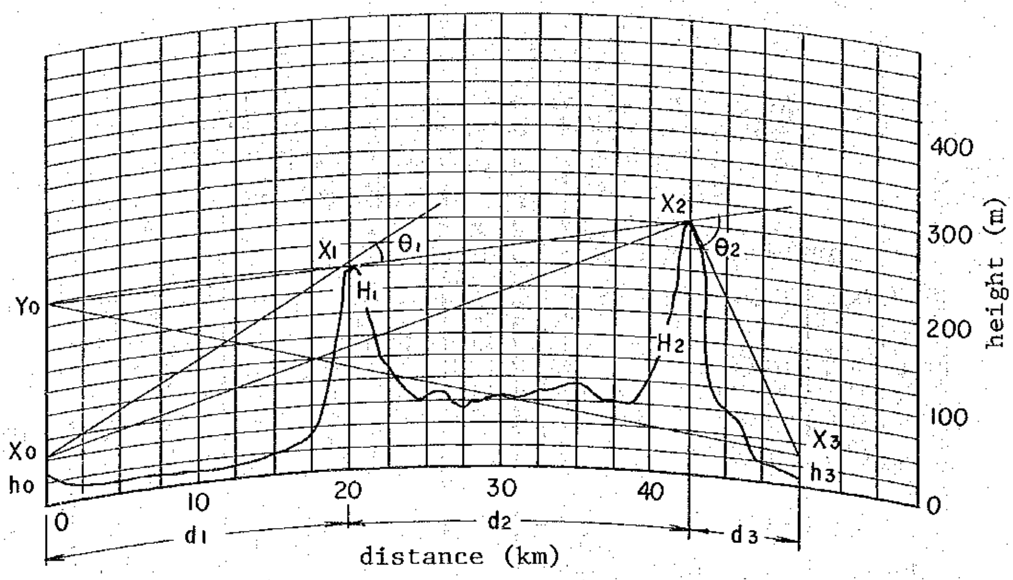


The number of field intensity corresponds to 1 watt radiated power.

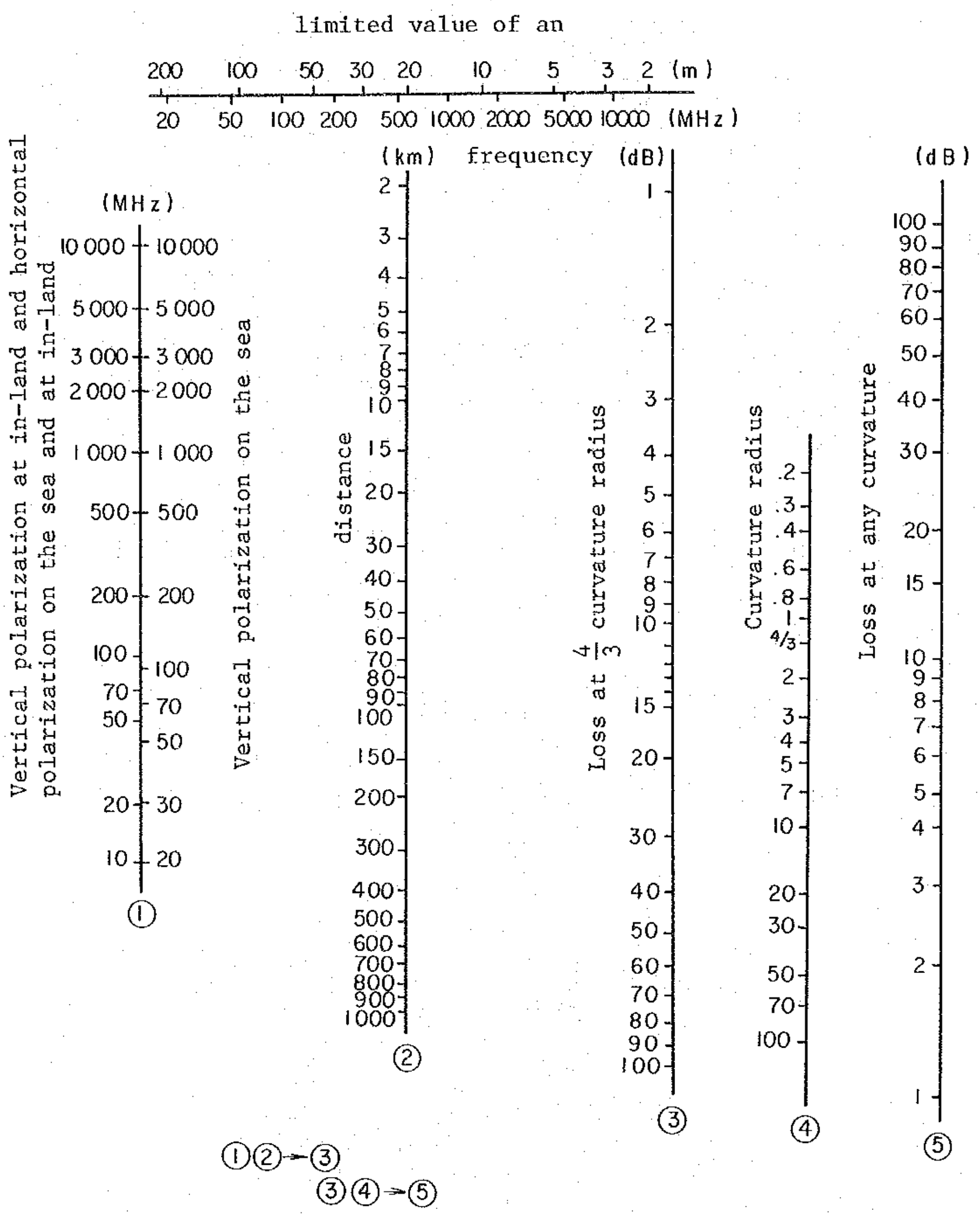
Loss due to diffraction of knife edge



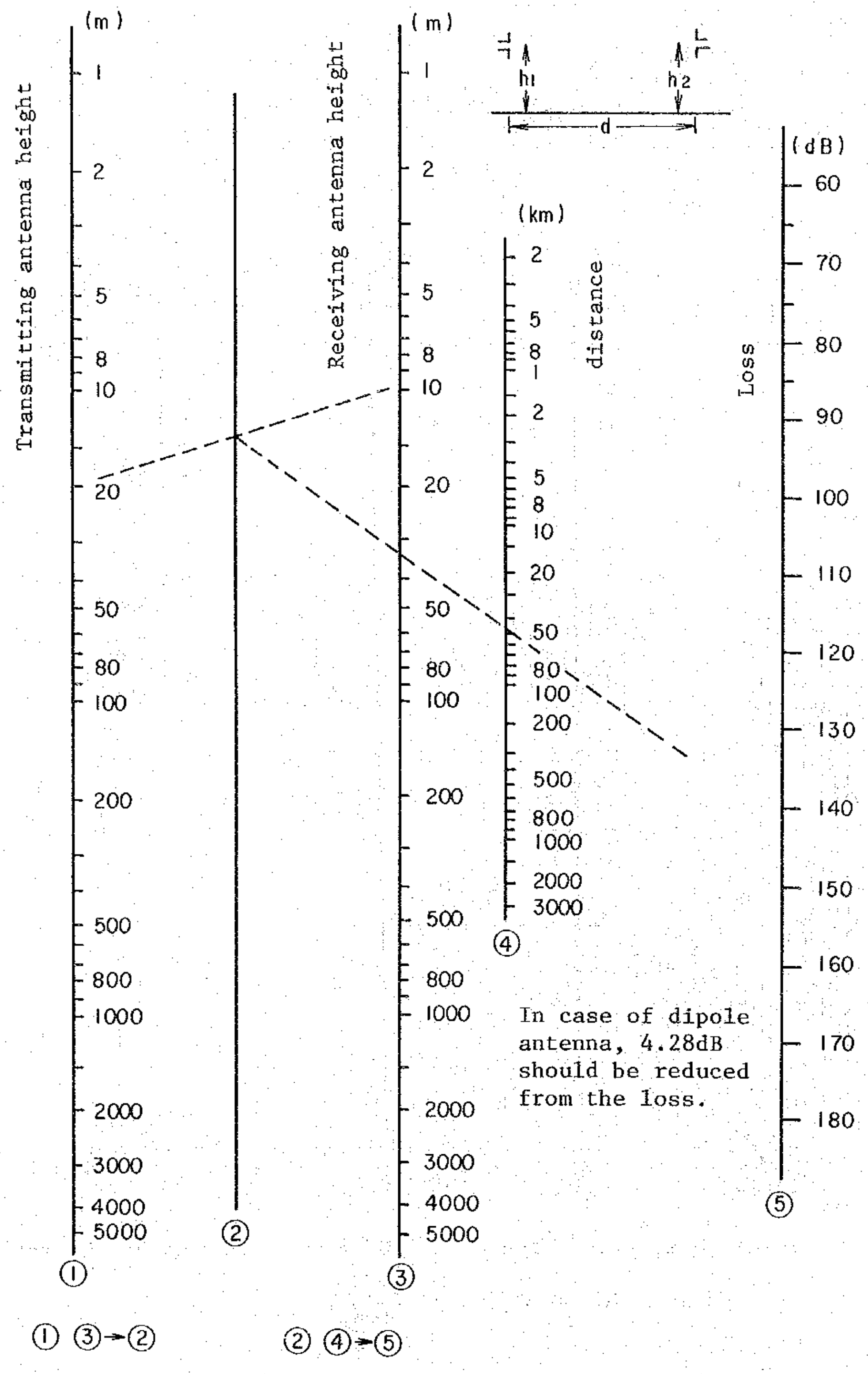
Loss due to diffraction of knife edge



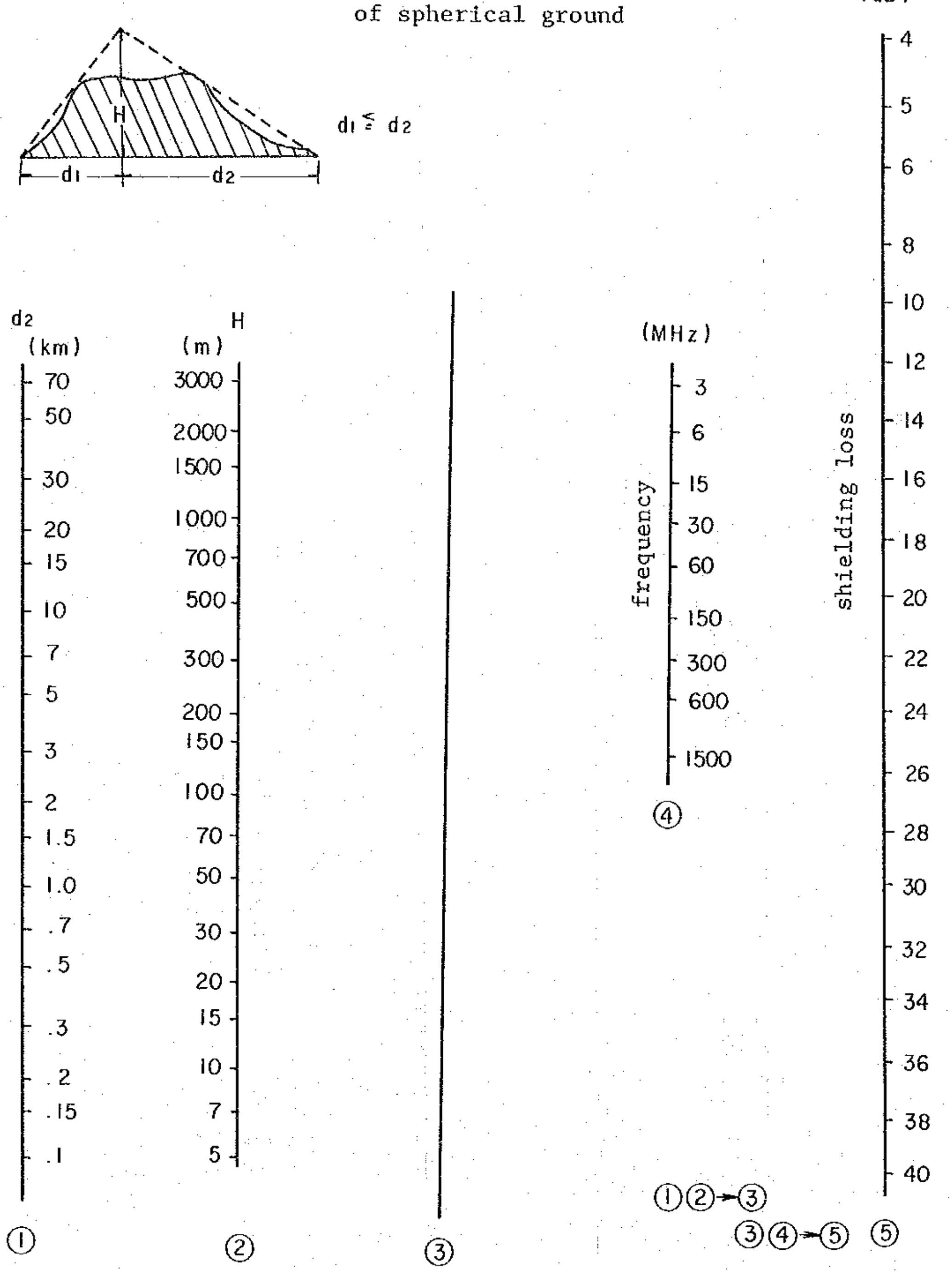
Additional loss due to spherical ground when antenna height is lower than the limited value



Flat grand propagation loss

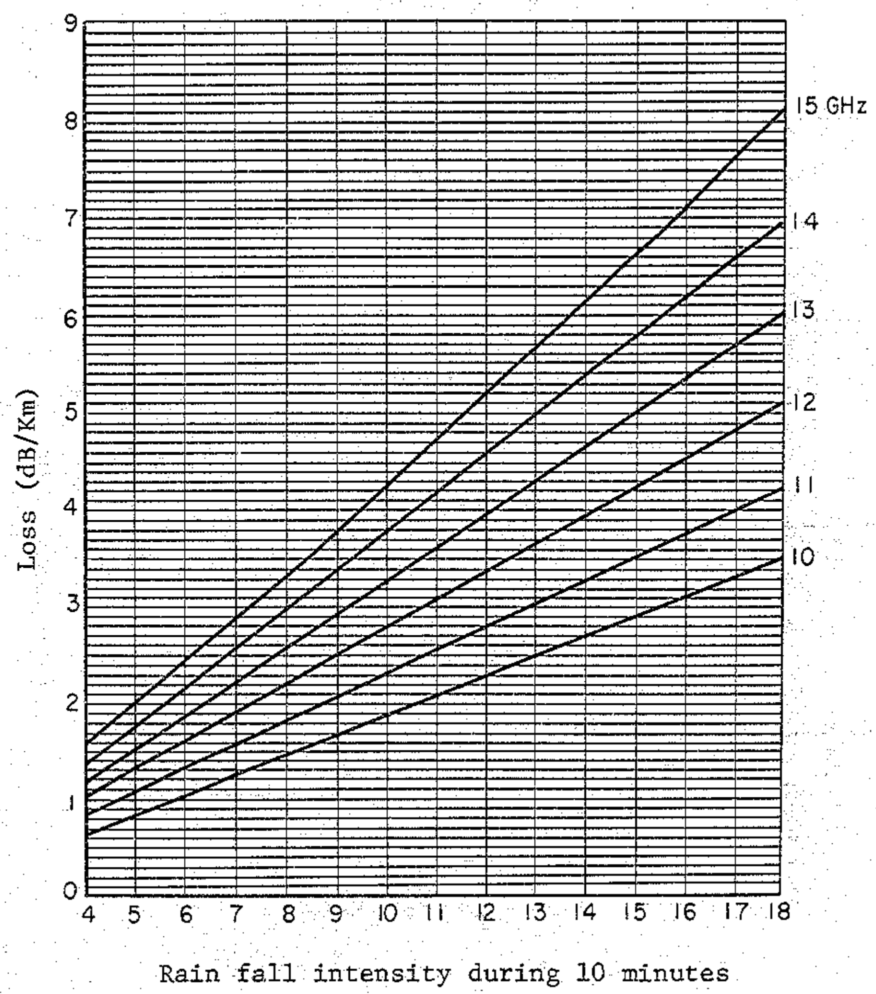


Loss due to shielding effect of spherical ground

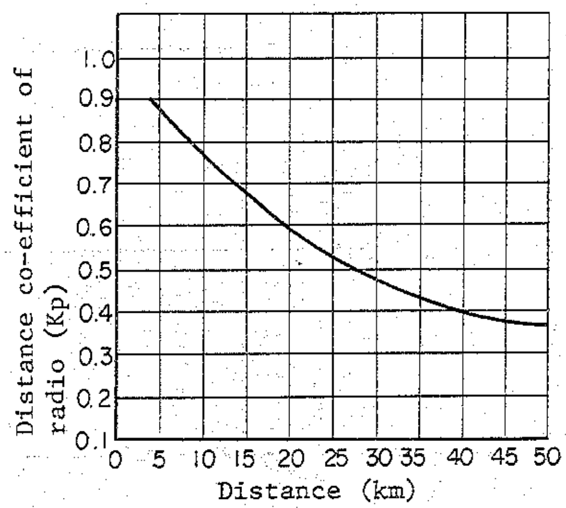


Note: An accuracy within ±1.5dB can be expected by the scale conversion of d_1 to $d_1 \left(\frac{\sqrt{2}}{1+\frac{d_1}{d_2}}\right)$

Chart for radio wave loss due to rain

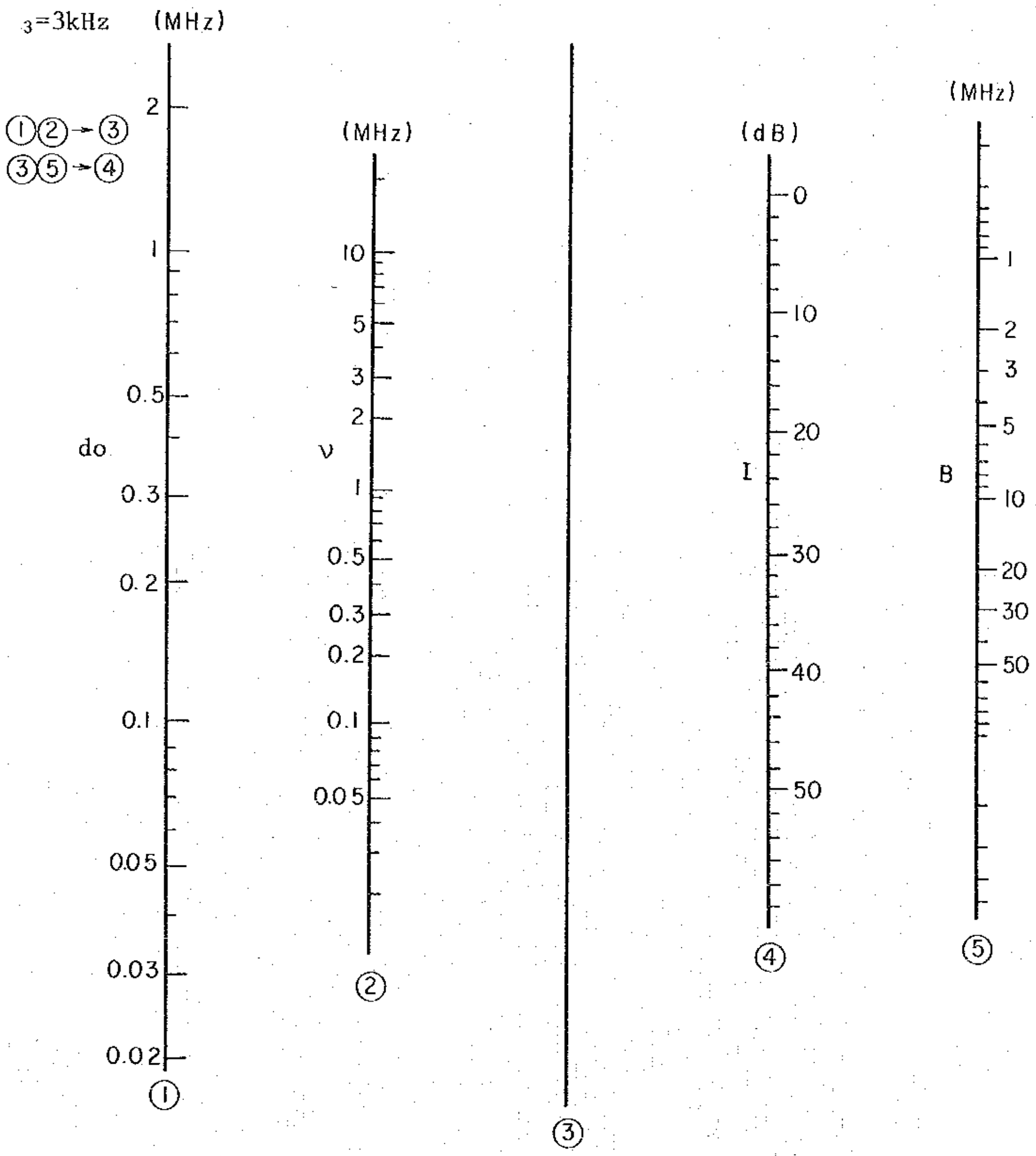


Distance co-efficienc of radio wave loss due to rain



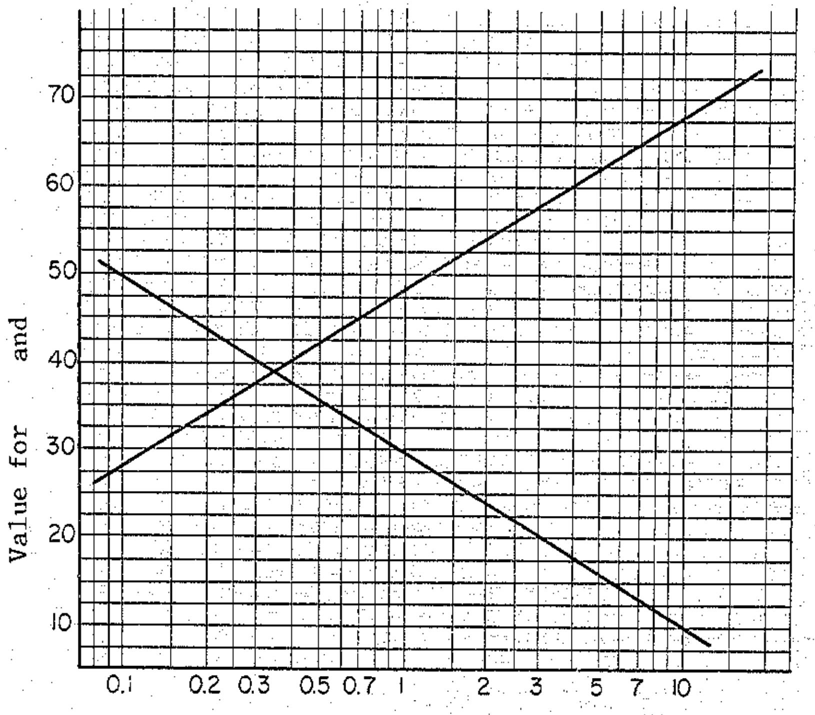
SS-FM improvement factor

$10\log \frac{do^2 \cdot B}{\nu^2 \cdot}$



Gain of parabolic antenna

G=20logD+20log +18
G: Gat or Car(dB)
D: diameter of reflector (m)
: frequency (GHz)

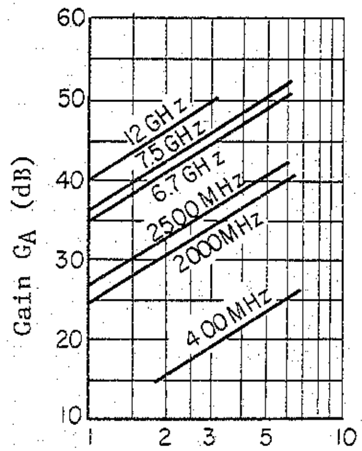


Diameter of reflector

Frequency in GHz

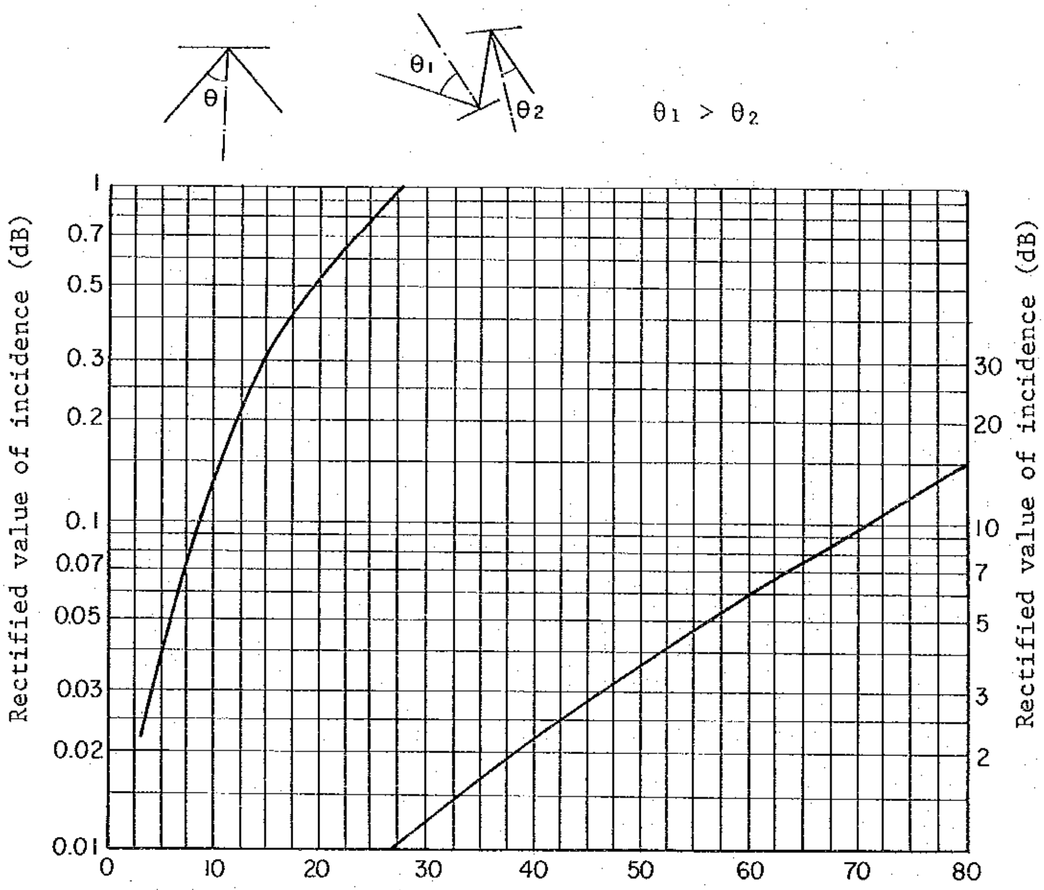
Gain can be obtained by G = -

Gain of parabolic antenna



Diameter of reflector
(m)

Rectification of reflector gain



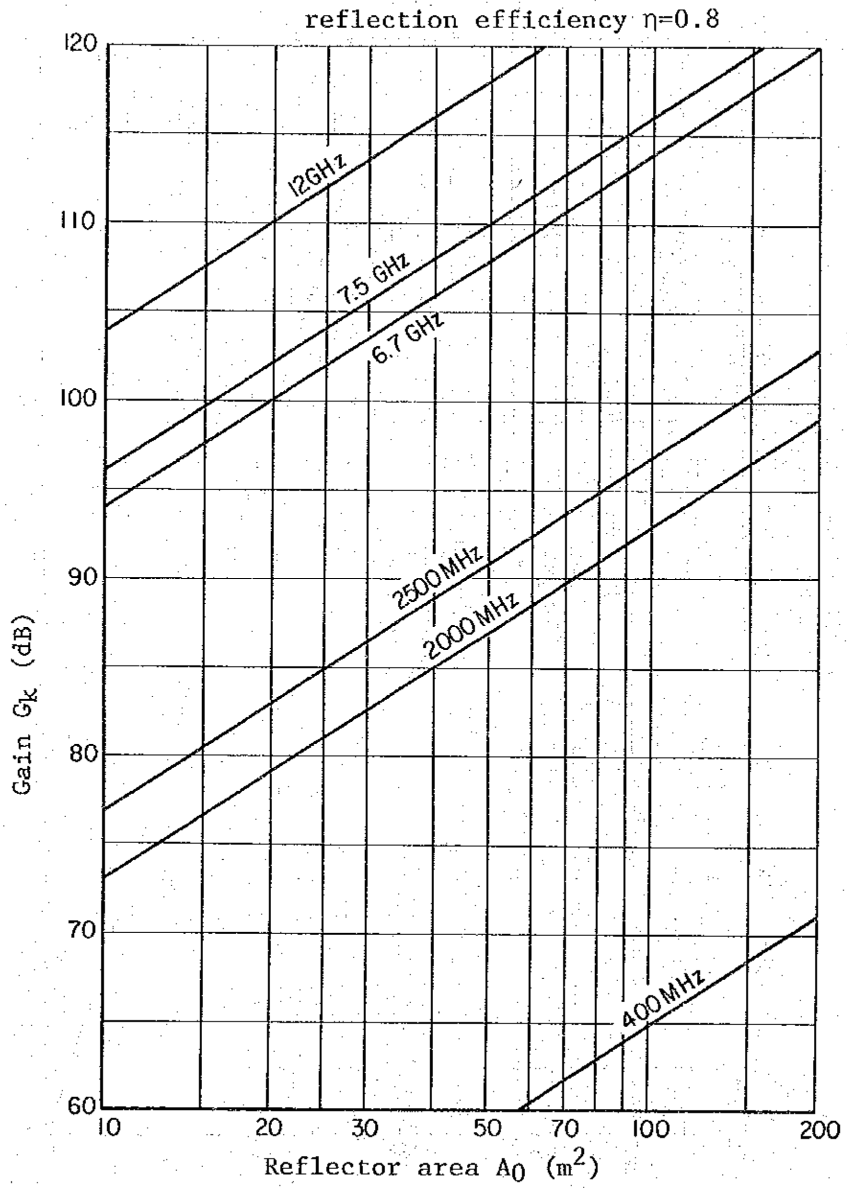
θ_1 should be taken in case of double reflectors

$G_k = 20\log A_0 + 40\log \ +41$

G_k: reflector gain (dB)

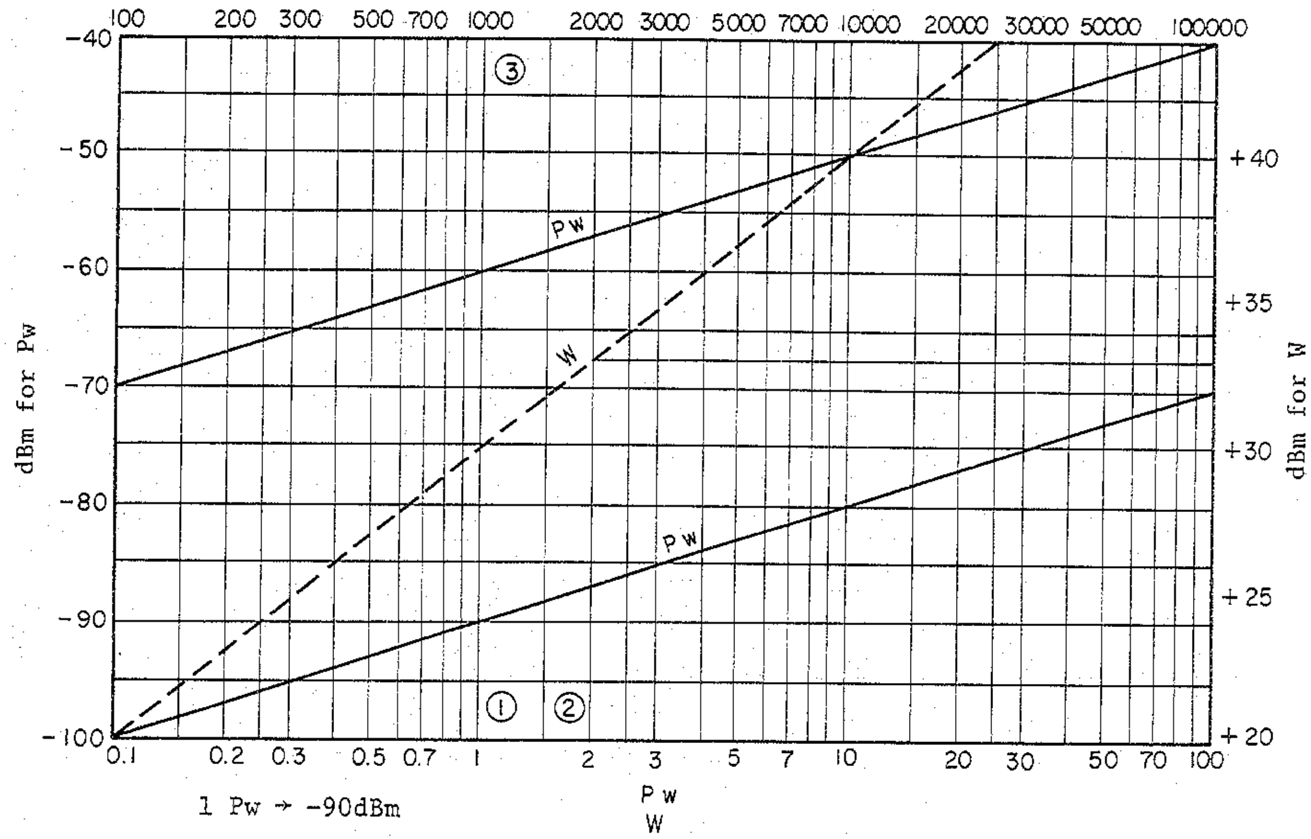
A_0: reflector area (m^2)

: frequency (GHz)



Double reflector requires the rectification of -1 ∿ -3dB

PW
W dBm chart



1 Pw → -90dBm

2 W → 33dBm

1000Pw → 60dBm

Typical radiation pattern of parabolic reflector

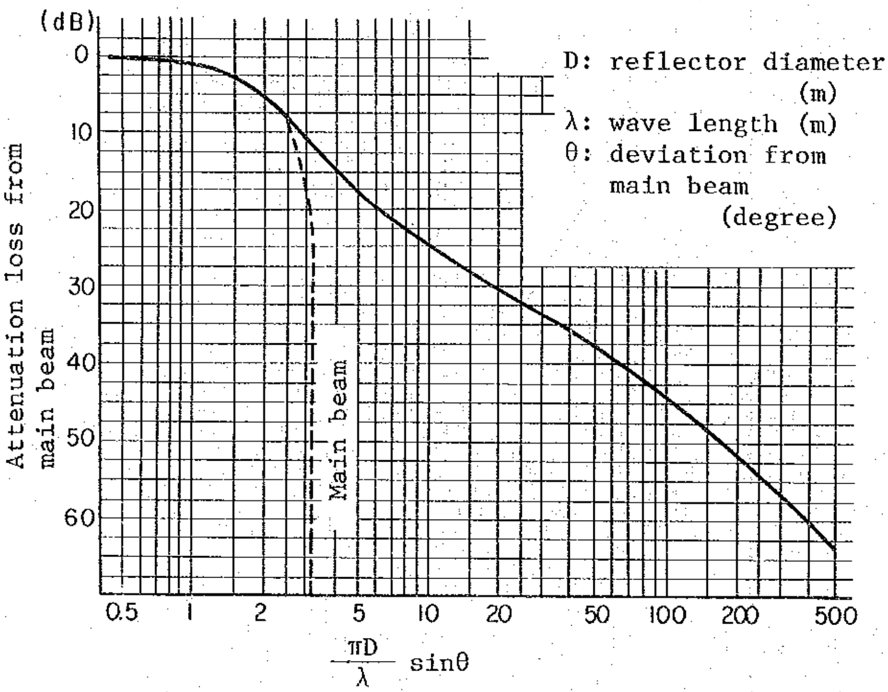
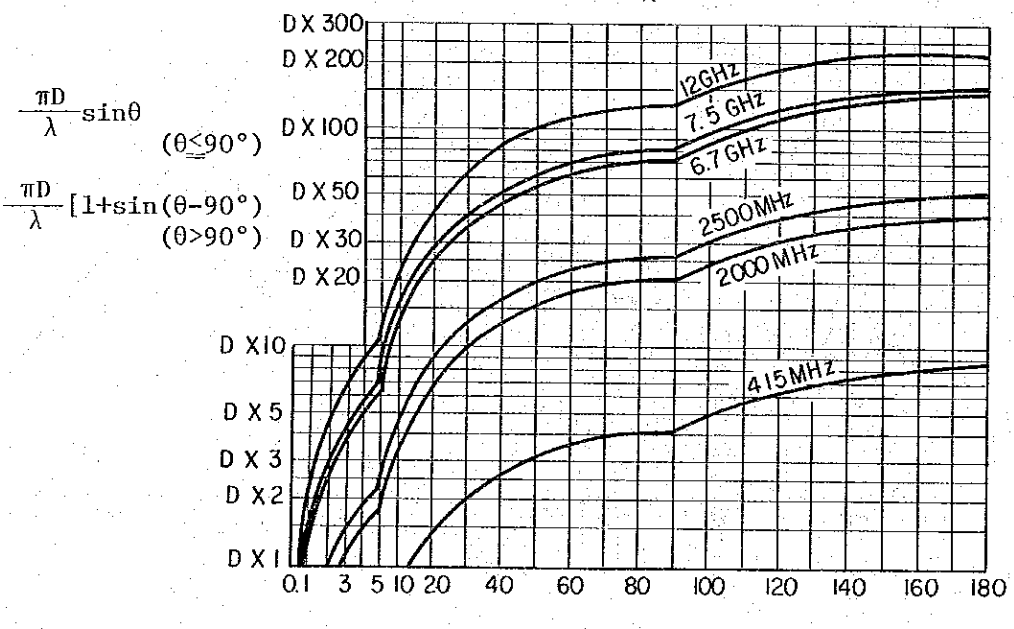


Chart for $\frac{\pi D}{\lambda}\sin\theta$



Angle from main beam (degree)

Minimum effective antenna height