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Microwave Principles

1.1 Introduction

Microwaves are radiowaves of very short wavelength. Basically radiowaves are electromagnetic waves the term radio being introduced from the concept of radiation of electromagnetic waves. Accordingly there should not be any lower or higher limits of the wavelengths. However, it is usual to confine the wavelength of radiowaves within certain limits to have electronics systems to generate, radiate and detect the radio waves, while infrared and optical waves are covered by systems called optoelectronic systems.

Table I
Radio wave bands

Frequency band	*Frequency range*	*Wavelength range*
Extra Low Frequency (ELF)	< 3 KHz	> 100,000 m
Very Low Frequency (VLF)	3 – 30 KHz	100,000 – 10,000 m
Low Frequency (LF)	30 – 300 KHz	10,000 – 1,000 m
Medium Frequency (MF)	300 – 3000 MHz	1,000 – 100 m
High Frequency (HF)	3 – 30 MHz	100 – 10 m
Very High Frequency (VHF)	30 – 300 MHz	10 – 1 m
Ultra High Frequency (UHF)	300 – 3000 MHz	100 – 10 cm
Super High Frequency (SHF) or Microwave	3 – 30 GHz	10 – 1 cm
Extra High Frequency (EHF) or Millimeterwave	30 – 300 GHz	10 – 1 mm

Microwaves have wavelengths which are limited to include wavelengths in the centimeter and millimeter wavelength regions. The entire frequency spectrum of radiowaves upto microwaves and millimeterwaves are, in fact, divided into a number of frequency bands as indicated in Table I, in which microwave and millimeter-wave bands cover the shortest wavelengths. The relation between the frequency f and wavelength λ being $f\lambda = c$, where c is velocity of propagation of the radiowave, which is equal to that of lightwaves in free space, 3×10^{10} cm/sec = 3×10^{8} m/sec.

1.2 Microwave band

A glance at the various frequency ranges indicates that the UHF band and the SHF range constitute the microwave frequency range. However, instead of microwave the term 'microwaves' include both the microwave and millimeterwave band. Like wise, the term "millimeterwaves" include both millimeterwave and submillimeter wave [> 300 GHz].

It may be mentioned here that the unit of frequency which was originally called cycles per second is now called. Hertz, with the code Hz, after the name of Heinrich Hertz, who first demonstrated the existence of electromagnetic waves at microwave, (λ = 66 cm) 30 years after the theoretical prediction of electromagnetic waves by James Clerk Maxwell in the year 1858.

During world war II, the development of Microwave Radars in the frequency range 1–12 GHz prompted the defence engineers to subdivide the microwave range into four bands with the codes L, S, C, and X using the letters, in a way so that the bands may be kept secret through this code letters. These four bands cover the frequency ranges as shown in Table II.

Table II
Radar microwave bands

Frequency band	*Frequency range*
L	1 – 2 GHz
S	2 – 4 GHz
C	4 – 8 GHz
X	8 – 12 GHz

Subsequently, in recent years microwaves at frequencies higher than 12 GHz were generated and used for Radar and communication systems and those are also subdied into K_u, K and K_a bands as shown in Table III, which also includes the radar microwave bands, *L*, *S*, *C*, and *X*.

Table III
Microwave bands

Frequency Band	*Frequency Range*
L	1 – 2 GHz
S	2 – 4 GHz
C	4 – 8 GHz
X	8 – 12 GHz
Ku	12 – 18 GHz
K	18 – 26.5 GHz
Ka	26.5 – 40 GHz

1.3 Millimeterwave band

It may be noted that the *Ka* band covers a part of the millimeter-wave band which starts from 30 GHz. However, the manufacturers of millimeter wave products extended the millimeter wave band down to 18 GHz as the same infrastructure for the production of millimeterwave products may be used to manufacture products down to 18 GHz. There is thus a overlap of microwave and millimeterwave bands over the frequency range 18 – 40 GHz covering *K* and *Ka* band of which 18 – 30 GHz is microwave and 30 – 40 GHz is millimeterwaves, as indicated in Table I. Manufacturers of Microwave products are now, in fact, developing products in the range 1 – 40 GHz, while Millimeterwave manufacturers are developing products in the range 18 – 300 GHz and even beyond 300 GHz. Thus there is a scope of choice between the Microwave and Millimeterwaves manufactures in the overlap bands *K* and *Ku*. Recently, some manufactures like Hewlett Packard are developing both Microwave and Millimeterwave products.

In view of the commercial interest millimeterwave manufacturers are now covering the band 18 – 300 GHz and even beyond 300 GHz upto about 400 GHz in the submillimeterwave region, and the millimeterwave band has been subdivided like that of microwave into a number of subbands as shown in Table IV, which also includes the *K* and *Ka* bands, included in Table III for the Microwave bands.

Table IV
Millimeterwave bands

Band	Frequency range
K	18 – 26.5 GHz
Ka	26.5 – 40 GHz
Q	33 – 50 GHz
U	40 – 60 GHz
V	50 – 75 GHz
E	60 – 90 GHz
W	75 – 110 GHz
F	90 – 140 GHz
D	110 – 170 GHz
G	140 – 220 GHz
H	170 – 260 GHz

It may be noted that there are overlaps of frequency bands between the *Ka* and *Q*, *Q* and *U*, *U* and *V*, *V* and *E*, *F* and *W*, *W* and *F*, *F* and *D*, and *D* and *G* bands.

In recent years, the entire spectrum of radio starting from ELF/VLF/LF/HF bands waves, may be integrated as shown in Fig. 1, covering Microwave and Millimeterwave bands, in which the frequency ranges in the UHF, SHF and EHF bands are indicated as 'Microwaves'.

1.4 Historical background

Historically microwave signal was first generated by Henrich Hertz in 1888 at 66 cm wavelength (454.5 MHz) while millimeter-wave signal was first generated by Sir J. C. Bose in 1895 at 5 mm wavelength (60 GHz). Subsequently, in 1890s, Bose also generated microwave signal at wavelengths upto 2.5 cm (12 GHz) wavelength. Besides these, Sir J. C. Bose also developed the world's first solid state point contact detector working at millimeterwave, infrared and optical wavelengths, using Galena (PbS) crystal as the detector material. Bose also developed world's first horn antenna and waveguide radiator for microwave and millimeter wave bands. In 1899 Sir J. C. Bose developed a highly sensitive iron-mercury detector in which a U-tube, made of glass, filled up with mercury, was used for a fine control of mercury contact pressure to optimise the sensitivity of the detector. Subsequently, in 1901 Marconi employed Sir J.C. Bose's technique of the highly sensitive iron-mercury detector.

Often the millimeterwaves, submillimeter waves ($f > 300$ GHz) and infrared waves are covered in an integrated manner under "infrared and millimeterwave" although the technologies for millimeterewaves as well as that of submillimeterwaves may be quite different from that required for infrared, where Optoelectronics Technology is prevalent. The reason for such an integration is that, some of the components are similar in design for millimeterwaves, submillimeterwaves and infrared, which are, in fact, borrowed from the systems used for optical waves, and called quasi-optical component. It may be mentioned here that many of the quasioptical components used for microwave and millimeterwave bands were first developed by Sir J. C. Bose in 1890s, to prove the quasi-optical properties of millimeterwave and microwave[2]. Sir J. C. Bose, in fact, developed a number of quasioptical components for microwave and millimeterwave bands, such as dielectric lens for focusing the wave to the detector, wire grid polariser for polarising the wave and also for analysing the polarised waves and curved cylindrical grating for accurate measurement of wavelength of microwave and millimeterwave. In addition to these, Sir J. C. Bose also worked for the first time on the response of living and nonliving meterials to microwave and millimeterwave bands, a subject which has now grown to a high level for the studies of microwave hazards to living beings, animals and vegetation and also for microwaves diathermy therapy of tumour and cancer. From the above history of Science about Sir J. C. Bose, who worked pioneeringly on a wide range of fields in Radio Science covering URSI commissions A. B. C. D. E. F. and K, it appears that Sir J.C. Bose may be called the father of Radio Science.

The difficulty of generating which, however, Bose generated by using his 'open resonator' coherent millimeterwaves and the limited distance of propagation of millimeterwaves upto line of sight distance apparently shifted the interest from millimeter waves to longer waves to wave length longer than a meter.

However, investigations of the millimeterwaves were continued and stimulated by scientific and millitary developments. For the purpose, feasibility of generating millimeterwaves of wavelength 1 cm (30 GHz) was demonstrated by Cleeton and Williams in studies of ammonia molecular resonances in the early 1930s. Subsequently in 1930 the observation by L. A. Hyland of the Naval Research Laboratory (NRL), about the reflection of radio signals from overflying aircrafts, stimulated the investigation of radar techniques for detecting flying aircrafts in NRL, Fort Monmouth and elsewhere in United States. During this period excellent work was also done independently in England with high priority as there was the threat of war in Europe with impending raids by bomber aircraft. This early connection between aircraft and radar is continued even today for detecting aircrafts from remote ground locations or from ships and also from airborne radars for detecting other aircrafts, ships and ground features and for developing countermeasures.

During early stages of World War II, however, the most requirements about aircraft related radar was found to be met at microwave frequencies upto 10 GHz. At the same time, an attempt to build a radar at millimeterwave around 10 mm wavelength range, at the M. I. T. Radiation Laboratory, failed due to a wrong choice of the exact wavelength which happened to be near the water vapour absorption line.

During World War II J. H. Van Vleck predicted theoretically the oxygen absorption band at 60 GHz ($\lambda = 5$ mm), at which Sir J. C. Bose first generated millimeterwaves. NRL used this band to develop secure communication system for millitary applications, which is still being used round the world. It is interesting to note that the 60 GHz band is now being used for a number of other application areas like Wireless LAN developed in Japan, Doppler radar for speed monitoring of cars and also for intersatellite communication. After World War II millimeterwave work continued at NRL for millitary applications and at Bell Telephone Laboratories (TL) for telecommunication applications. In 1950s Hughes aircraft company successfully extended the frequency range of coupled–cavity travelling wave tubes to millimeterwaves. In 1960s, intensive work in millimeterwave technology was done at BTL on the development of solid state components required in the development of underground repeaters using low loss overmoded waveguides, which was also first developed by Sir J. C. Bose.

In 1970s, Hughes began manufacturing a solid state sweep generator which apparently triggered a chain of development in Hughes in the Millimeterwave Technology area of a full range of devices, components and instruments covering 18 to 160 GHz with the work continuing to extend the frequency range to shorter millimeter wavelength range. Recently, all this activities of Hughes has been taken over by Millitech, U.S.A., which developed a unique focal plane array acting as a 'millimeterwave eye' for high resolution passive imaging of objects at millimeter waves around 94 GHz.

1.5 Velocity of waves propagation in media

It may be mentioned here that although the velocity of propagation of radio waves and that of light wave are the same and equal to $c = 3 \times 10^{10}$ cm/sec when the medium is vacuum, in reality the propagation medium is usually not vacuum but may be the dielectric media like Teflon, Polysterene, Polyetheline, in which case the wave is slowed down, reducing the free space wavelength, λ_o, to dielectric wavelength λ_d, the two being related by the equation

$$\lambda d = \frac{\lambda_o}{\sqrt{\varepsilon_r}} \qquad \ldots(1)$$

where ε_r = relative dielectric constant of the medium.

Even for propagation through air, we have to take into account the relative dielectric constant ε_r for the air medium, which is dependent on the humidity or watervapour density in the propagation path and temperature of the atmosphere. During rain or fog events, the effective relative dielectric constant may increase significantly to reduce the velocity of propagation of radio waves compared to that in clear air environment. The reduction of the velocity of propagation relative to that in free space causes some excess delay due to the atmospheric medium which may affect the estimation of target range by radar observations.

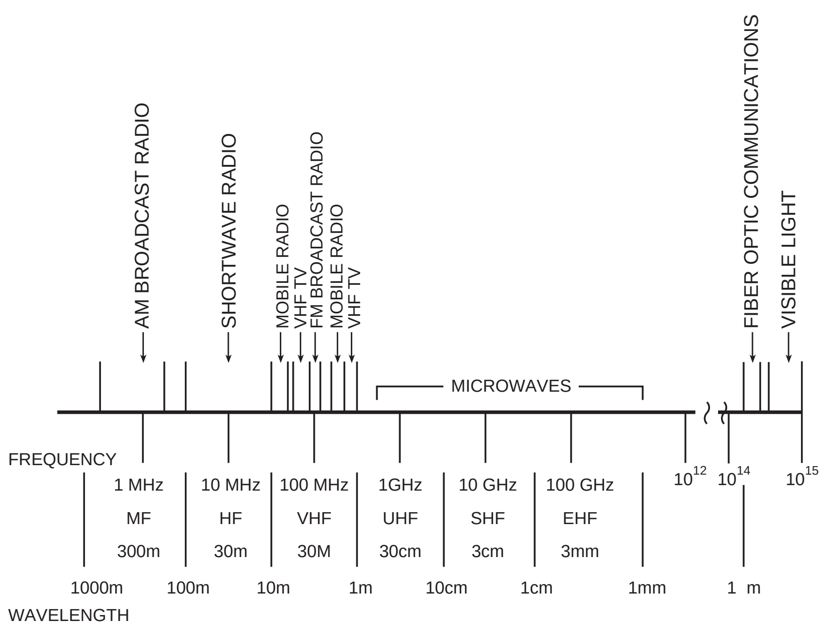

Fig. 1.1.

1.6 Advantages of microwave line of sight communication

The microwave systems for radar and communication are becoming more and more important in Navigation, ground based line of sight communication, satellite communication, and other fields like remote control and remote sensing.

The various reasons pertaining to the wide use of microwaves are underlined as below :

(*a*) Broader band width

(*b*) Improved directivity and higher gain of antennas.

(*c*) Higher Reliability and

(*d*) Lower Transmitter Power requirements

(*a*) **Band Width :** Telegraph, speech, music and video signals may be transmitted directly through wireline or cable as a baseband signal covering baseband width for each type of signals. We know that a telegraph channel requires a bandwidth at 120 Hz, the speech requires a bandwidth of 2–3 kHz, a music channel requires a bandwidth of the order 10–20 kHz, each one as the base bandwidth, whereas the bandwidth required by TV channel is very high, being 7 MHz, to include the video bandwidth of 5 MHz together with the bandwidth of the frequency modulated audio signal. Now the entire frequency range upto 1000 MHz or 1 GHz is crowded with various radio channels, telegraph channels, TV channels etc. Also with the advent of frequency modulation we have to go in for higher bandwidths in order to achieve a high S/N ratio. Thus the band congestion was becoming a major problem in radio communication. Hence keeping in view all these things, we have to go in for microwave frequencies. In microwave range, the number of channels that can be accommodated simultaneously in any single microwave band is very large.

(*b*) **Improved Directivity and Gain of Antennas :** At microwave frequencies, size of a dipole is very small. Hence we may have an aerial array consisting of a greater number of dipoles giving very high directivity and gain compared to an omnidirectional antenna. However, to have a wide bandwidth of the antenna, instead of an array, a parabolic reflector antenna with a horn feed is universally employed at microwaves to have high directivity and gain, the two antenna parameters being proportional to each other.

(*c*) **Reliability :** In Radio communication in the MF/HF band, the propagation of the radiowaves takes place by reflections from *F* and *E* layers of the ionosphere, whose concentrations at any location varies widely with time and weather condition. Hence the signal received varies widely with time and weather. Hence the energy received by the receiver is not of uniform strength but there are great time variations in the strength of received signal, though the transmitted power may be constant, thus giving rise to Fading Effect. But at microwave frequencies, there is less fading since the propagation of microwaves from transmitter to receiver takes place by line of sight propagation through the neutral atmosphere. As the frequency increases, the reception becomes clearer in microwave ranges with still less fading, as the fading at microwaves occurs only in the presence of atmospheric turbulence created by winds in the atmosphere while the ionosphere is transparent to microwaves allowing stable propagation between satellite and earth based microwave communication links.

(*d*) **Power Requirements :** At microwaves, the power requirements of the transmitter become very small as compared to that at MF/HF, due to the high gain of the antennas at microwaves. The basic configuration of the transmitter as well as the receiver at microwaves is similar to that in MF/HF bands but the active and passive components used are different. The only disadvantage of microwaves being the need for a number of repeater stations, as a microwave line of sight link over the ground cannot have a range greater than about 40–60 km due to curvature of the earth's surface. For earth-space communication with satellites, the range is about 3 orders of magnitude higher, being about 36,500 km for geostationary satellites with circular equatorial orbit. Microwaves have a widerange of other application areas like consumer, Industrial, Biomedical and Chemical applications, electron paramagnetic resonance and Navigational Aid.

1.7 Other Applications of Microwaves

Microwave energy has, in fact, a heating effect just as any other form of energy. Because of very short wavelength of microwaves and consequent compact circuitry, this heating effect has got a wide variety of non-military or civilian applications including Microwave ovens, Microwave diathermy and Microwave drying. Microwave ovens are, in fact, used for very quick and uniform home cooking of food materials, since the food is cooked by the waves penetrating inside the entire food material to all depths without any overheating on the outside surface of the material. Likewise, Microwave diathermy machines also produce heat inside the muscle without overheating of the skin. Microwave drying machines are used in the printing industry, Microwave curing of Rubber, Tire and Asphalt from socked water, in Microwave material processing like thawing, drying and heating and also in Agricultural industries for vacuum drying of sliced parsely root, cotton seed, microwave treatment of seed, drying of field corns and heating of corn fields, warming of plants, for use as insectisides in grain, Tobacos and Pecan Weevil, and for pasteurisation of milks. Even for Forest products microwaves are applied for drying of Veneer and Bending of Wood.

1.8 Wave Propagation and Noise

Line of sight propagation at microwave and millimeterwave bands, in atmospheric medium, usually produces an excess attenuation over and above that expected for propagation through free space in vacuum. The atmospheric water vapour, liquid water content and temperature are, in fact, the main determining factors controlling the excess attenuation in the atmosphere. Typical values of the atmospheric excess attenuation in the microwave and millimeter-wave bands is shown in Fig. 1.2 (*a*) which also shows the contributions of rain and fog on the excess attenuation. The figure shows that in microwave line of sight links, the atmospheric excess attenuation in clear atmosphere is as low as 0.01 dB/km for frequencies upto 10 GHz. However, for the excess rain attenuation, it may be as high as 10 dB/km at 10 GHz for a rain rate of 150 mm/hr. The

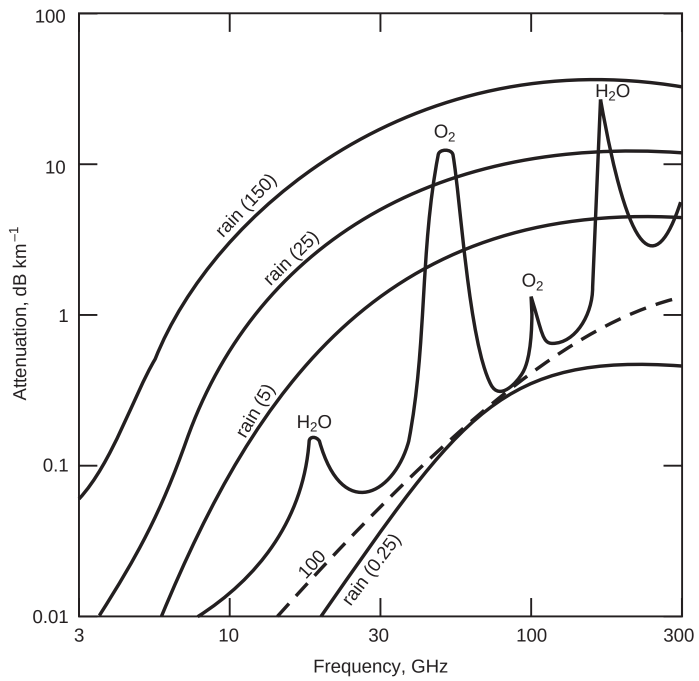

Fig. 1.2(a).

figure also shows that in clear atmosphere without rain there are a few absorption lines of oxygen and water vapour molecules of the atmosphere, being around 22 and 183 GHz for water vapour, and at 60 and 120 GHz for oxygen. The figure also shows that the excess attenuation due to fog is very low and starts only beyond about 20 GHz, the amount being comparable to rain attenuation for a rain rate of only 0.25 mm/hr. In clear air the attenuation increases with an increase of water vapour content as shown in Fig. 1.2 (*b*), which

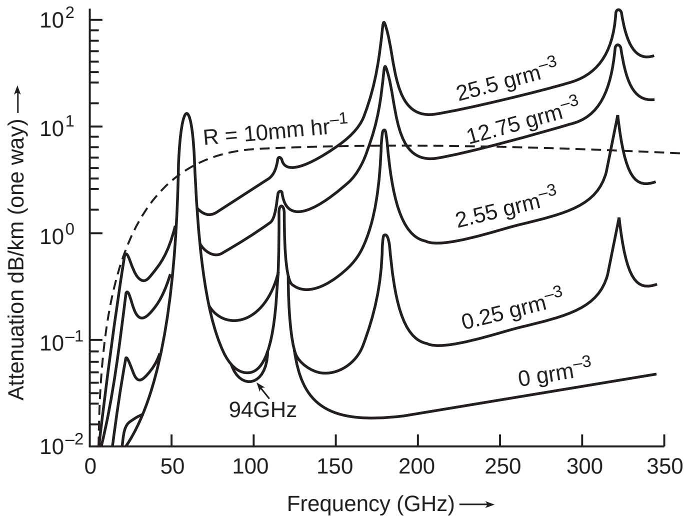

Fig. 1.2(*b*).

also shows that the excess attenuation above 100 GHz due to rain rate of 10 mm/hr is about 10 dB/km which is around the clear air attenuation with water vapour density of 25.5 gm m^{-3}.

For satellite communication, where the ray path is slanting, the attenuation is less, particularly, at frequencies between the absorption lines. But the amount increases rapidly with water vapour content or the humidity of the atmosphere. For frequencies above 300 GHz in the submillimeterwave, infrared and optical regions, the atmospheric excess attenuations in clear weather and in fog are shown in Fig. 1.2 (*c*).

For clear weather the excess attenuation at submillimeterwave is more than 100 dB/km, while at infrared and optical bands, it is very low, in the range 0.1–1.0 dB/km as shown in Fig. 1.2 (*c*), which also shows that for fogy weather with a water vapour density of 0.1 g/m^3 the excess attenuation becomes as high as 100 dB/km in the infrared and optical regions.

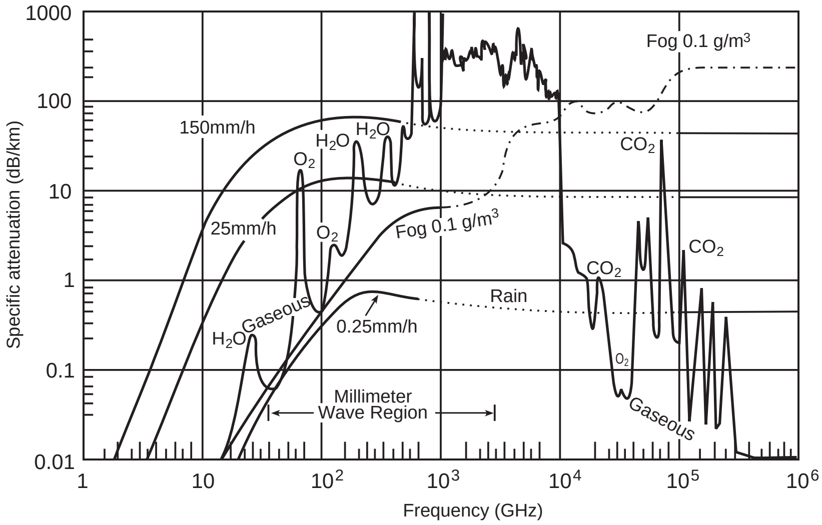

Fig. 1.2(*c*).

Fig. 1.3 (a) shows how the excess attenuation in atmosphere for a vertical propagation between satellite and the earth varies with

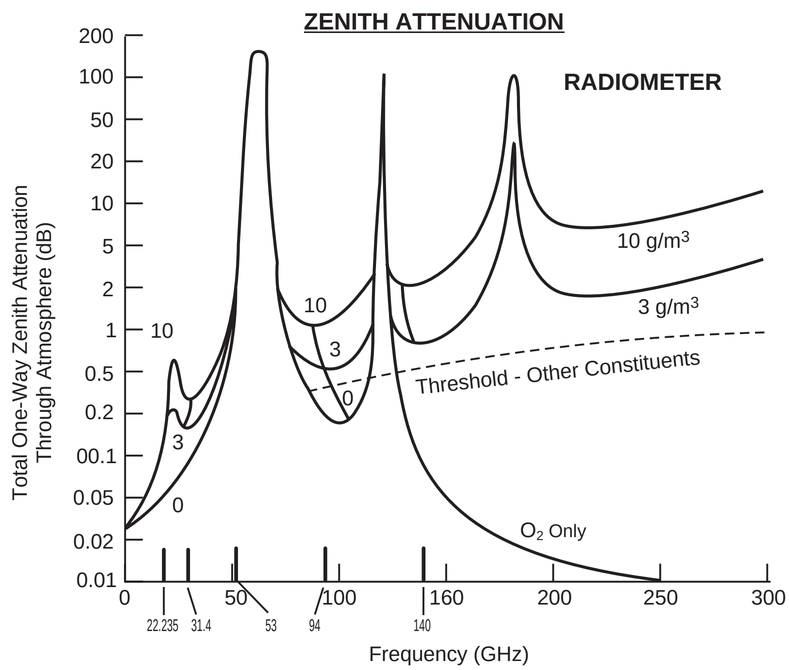

Fig. 1.3(*a*).

frequency in the microwave and millimeterwave bands, showing the window frequency's around 30, 94 and 140 GHz at which the excess attenuation attains a minimum value. The figure also shows the dependence of attenuation on the water vapour content upto 10 gm/m^3. The effect of water vapour content on the transmittance of satellite signals along vertical path between earth and satellite is shown in Fig. 1.3 (*b*) from which it is clear how the transmittance or transmissivity of satellite signals is nearly zero at the oxygen absorption lines 60 and 120 GHz and also at the water vapour absorption line at 183 GHz and how the transmittance at the atmospheric windows decreases with increasing amounts of water vapour at different humidity regions from polar to tropical regions. The figure also indicates how the window frequency shifts to lower frequencies with increasing amounts of water vapour. The dependence of the window frequency on the atmospheric humidity and temperature has, in fact, beam studied using theoretical atmospheric models of microwave and millimeter wave absorption from which it appeared that shifts may be significant in humidity regions, like the Eastern Regions of India including the coastal regions of Bay of Bengal and the Assam region.

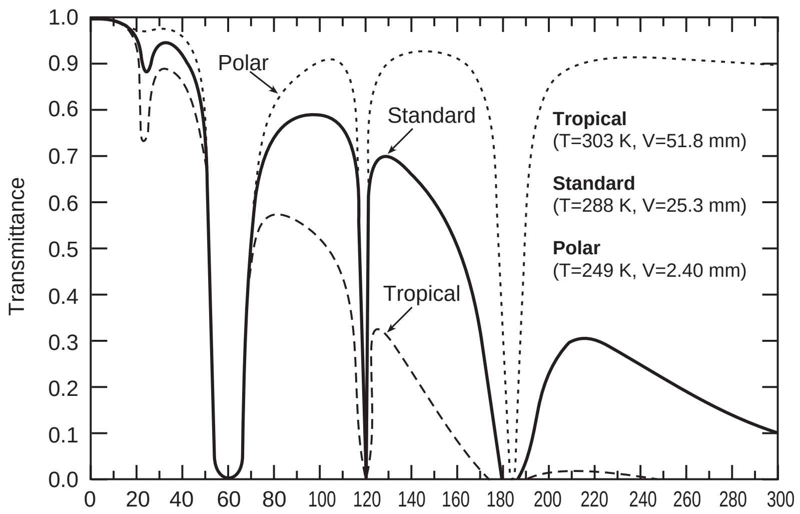

Fig. 1.3(*c*).

The spectrum of transmittance over the entire range from microwave to visible region for a clear atmosphere is shown in Fig. 1.3 (*c*), from which it appears that the windows in microwaves including millimeter and submillimeter waves cover wider octaves

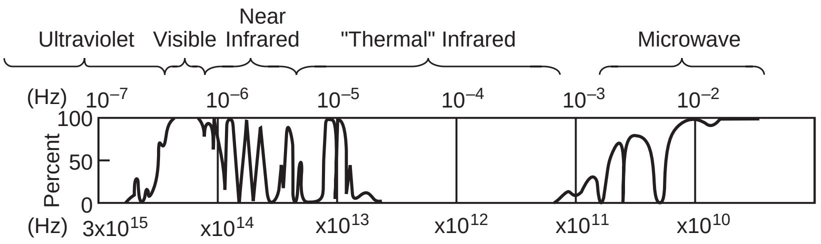

Fig. 1.3(*c*).

than that at optical or infrared regions. However for a given octave (doubling of frequency), the bandwidth at optical region would be orders of magnitude greater than that at microwaves, because of very short wavelength or very high frequency of the optical region. For propagation paths at a slant angle the attenuation at any frequency will increase with the slant angle from the zenith This will be evident from the thermal emission noise of the atmospheric constituents, water vapour, and oxygen, as shown in Fig. 1.4, as both the attenuation and thermal emission noise increases with the density of the atmospheric constituents. The larger the water vapour density or oxygen density the larger would, in fact, be the

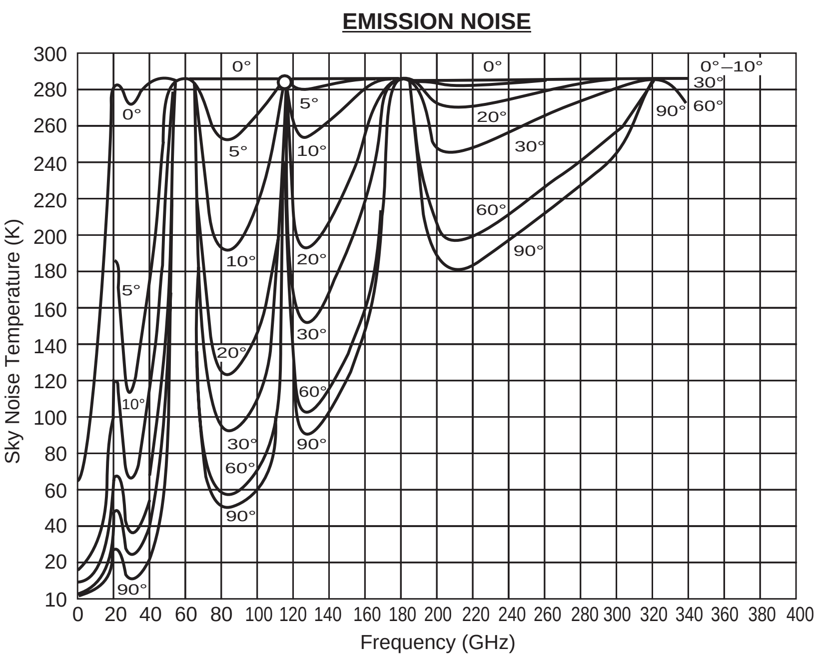

Fig. 1.4.

thermal noise power emitted. The antenna temperature due to thermal emission noise, N_a, is expressed as antenna noise temperature T_a, the two being related by the equation $N_a = k\ T_a\ B$, where k = Boltzman's constant = 1.38×10^{-23}, T_a = antenna temperature and B = bandwidth of the receiver. From the figure we find that for elevation angle 0°, which is used for horizontal propagation path in line of sight link, the antenna noise temperature is 290 K = 27°C, which is equal to the temperature of the earth's surface. As the elevation angle increases upto 90° upto the vertical path, the antenna temperature near the atmospheric window frequencies decreases reaching minima values for the vertical path, while the antenna temperature remains unaltered near the resonance lines of oxygen at 60 and 120 GHz and that of water vapour at 183 Ghz.

1.9 Bending of Ray Path Due to Atmospheric Refractivity Gradient

For a line of sight (LOS) link at microwave and millimeterwave bands the refraction in the atmosphere due atmospheric refractivity height gradient may bend the line-of-sight (LOS) ray path. Fig. 1.5 shows the configuration of as LOS link installed over the curved earth's surface along with the refractivity structure of the atmosphere. The density of the atmosphere decreases exponentially with height. As a result, the line of sight ray path, for free space, shown dotted in ray path no. 1, will become a bent one as indicated in the ray path no. 2. The curvature of the ray path depends on the refractivity gradient of the atmosphere with height. Higher the gradient more will be the curvature of the ray path. If the gradient is such that the curvature of the ray path becomes equal to that of the earth's surface, then the ray path will be as shown in path no. 3, with the transmitted beam being tangential to the ray path. For a higher gradient, as the curvature, of the tangential beam would produce ray path no. 2, a beam with higher elevation angle α may be used to have the link operative, as shown in ray path No. 4.

In practice, the refractivity gradient is subject to variation from time to time to produce a variation of curvature of the ray path, resulting in loss of beam alignment. That is why, the antenna beam cannot be made very sharp so that the variability of the curvature of the ray path due to be variation of refractivity gradient, may be accomodated, using some wide beam pattern as shown in Fig. 1.5.

The refractivity of the atmosphere is related to the density and temperature of the atmosphere through the empirical relation given by

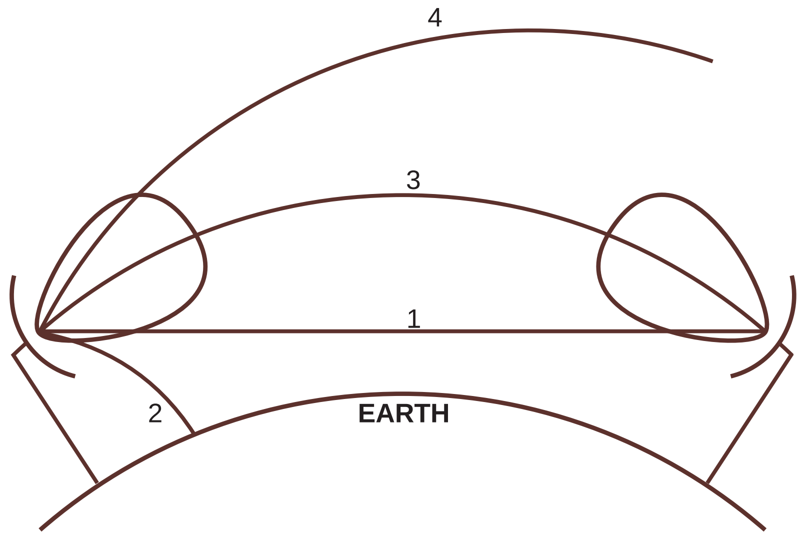

Fig. 1.5.

$$N = (n - 1)\, 10^6 = \frac{77.6\, p}{T} + \frac{3.73 \times 10^5\, e}{T^2} \qquad ...(1.0)$$

where n = index of refraction or scaled up refractive index

p = atmospheric pressure in mbar (millibar)

T = absolute temperature of atmosphere

A typical value of the scaled up index of refraction, n, near the surface of the earth is 1.0003, the corresponding refractivity being $N = (n - 1)\, 10^6 = 300$. The index of refraction generally decreases with height except that around a refractivity inversion layer. The decrease of the index of refraction is really exponential, but for heights near the ground it may be assumed to be linear. For a standard atmosphere, the water vapour pressure at sea level is 10 millibar and it decreases at a rate of 1 millibar per-thousand feet.

This corresponds to a decrease of index of refraction with height at a rate 4×10^{-8}/m. Due to curvature of the earth the layers of constant refractive index for a particular height will also have the same curvature as that of the earth, as shown in Fig. 1.6(a). The curvature of the atmospheric layers bends the ray path of microwave link downwards making it curved as shown in Fig. 1.6(b), the curvature of the ray path depending on the height gradient of refractive index, $\frac{dn}{dh}$. Had the atmospheric refractive index been uniform and independent of height, the ray path will be straight as shown in link path 1 with the link distance, d. In reality, due to

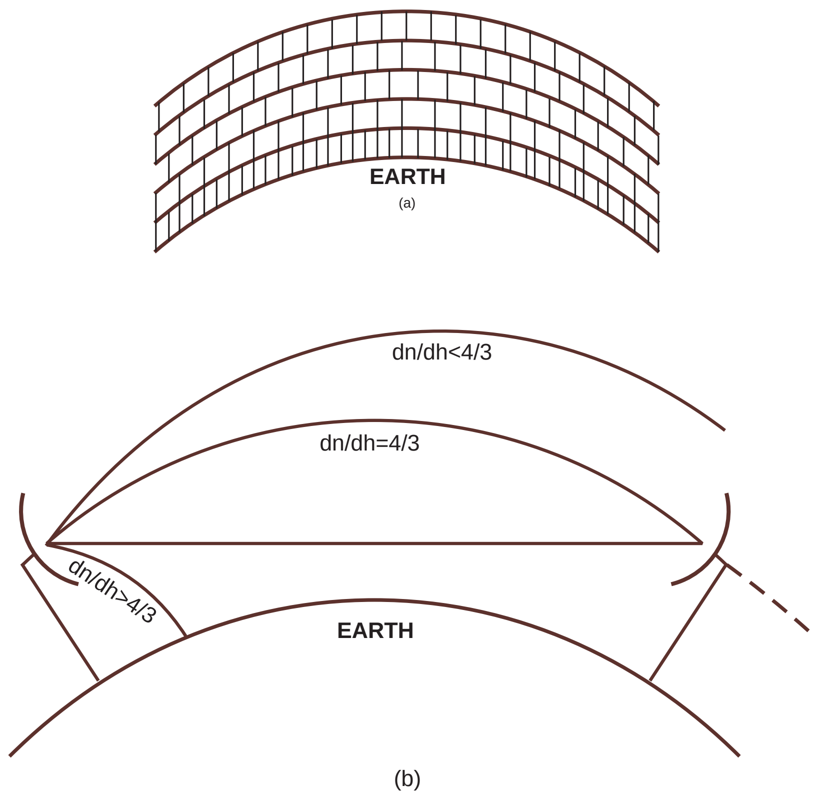

Fig. 1.6.

dependence of refractive index on height, the ray path becomes curved which may be as shown in link paths 2, 3 and 4, varying with the height gradient of refraction index. For the best ray path 2, the line-of-sight distance of the link is increased to $d_2 > d_1$. For a height gradient $-\frac{4}{3}$ the link distance, in principle, may be very large as the curvature of the ray path becomes equal to that of earth. In practice, the height gradient of 4/3 may be maintained upto a certain distance over the surface of the curved earth, which in fact, determines the effective link distance d_3. For a height gradient of $n < \frac{4}{3}$, the curvature of the ray path would be less than that of earth, for which the ray will be diverted away from the receiving antenna, as shown in the link path d_4. Due to the curvature of the ray path it is often difficult to estimate the maximum link distance of a microwave link. To make the estimation easier, we can assume a modified equivalent earth's radius with lower curvature so that the equivalent ray path may become straight as shown in Fig. 1.7 (*a*). If the earth's radius

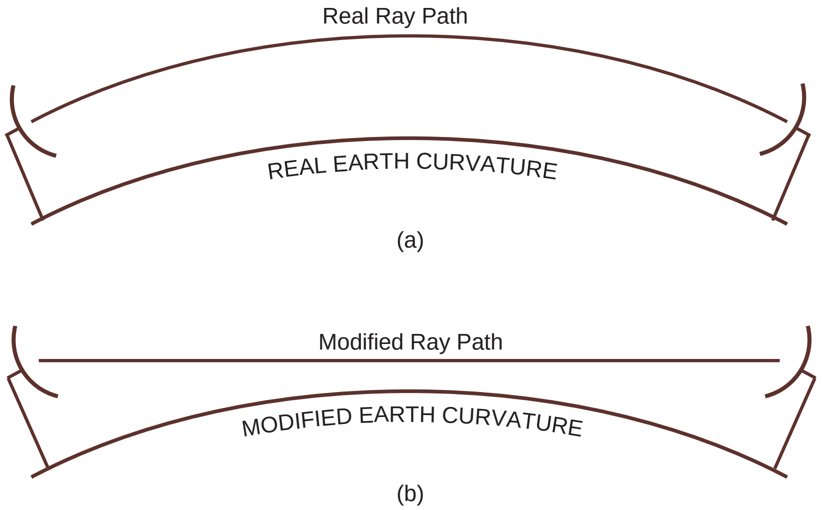

Fig. 1.7.(a) (b)

be a, then it can be shown by using Snell's law of refraction that the modified equivalent earth's radius, R_m, would be given by $R_m = K_a$ where K_a is given by

$$K = \frac{1}{1 + a\ (dn/dh)} \qquad \text{...(1.2)}$$

To estimate the link path distance d, the geometry of the problem is drawn as shown in Fig. 1.7(*b*), from which we have

$$\left[\frac{d}{2}\right]^2 = h[2\,R_m - h] \cong 2\,R_m\,h,$$

as << 2 R_m

$$\therefore \qquad \frac{d}{2} = \sqrt{2\,R_m h} = \sqrt{2\,Kah}$$

or
$$d = 2\,\sqrt{2\,Kah} \qquad \text{...(1.3)}$$

$$a = 3440 \text{ nautical miles} = 6450 \text{ kilometers}$$

from which we have, for $\dfrac{dn}{dh} = \dfrac{4}{3}$, the link path distance d given by

$$d = 2.46\,\sqrt{h}\ (ft) \text{ nautical miles (nmi)} \qquad \text{...(1.4)}$$

or
$$d = 260\,\sqrt{h} \text{ kilometers (km)} \qquad \text{...(1.5)}$$

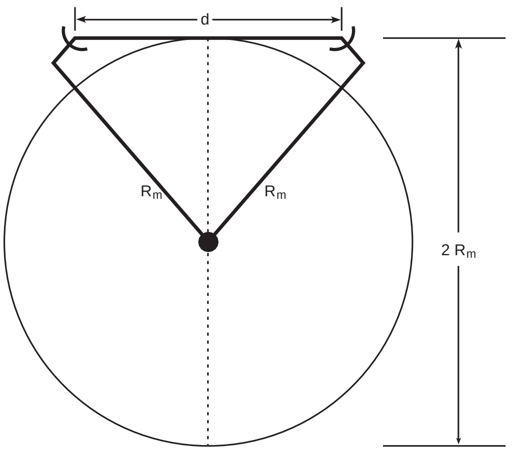

Fig. 1.7.(c)

1.10. Estimation of Microwave Link Parameters

For a microwave line of sight link, after estimating the maximum line of sight distance from equation (1.3) the requirements of transmitter power and receiver sensitivity can be estimated assuming the loss of radiated power with distance, following the inverse square law, assuming if the atmospheric attenuation to be is negligibly small at the microwave frequency chosen for the link. Also, the transmitter power has to be estimated by taking into account the gains of the radiating and receiving antennas. The gain of an antenna, in fact, arises from the focussing effect of the parabolic dish antenna with horn fed used in microwave links. The situation, is, in fact, similar to the increase of the intensity of the light beam produced by torch lights due to the focussing effect of the parabolic optical reflector of the torch, with a torch bulb located at the focal point, of the reflector, to convert the omni-direction radiation pattern of the torch bulb, into a narrow light beam of much higher intensity. The intensity of the light beam due to the focussing effect of the reflector depends on the diameter of the parabolic optical reflector of the torch. At microwaves, the torch bulb is replaced by a horn feed radiator located at the local point of the parabolic metallic dish antenna acting as a reflector for microwaves.

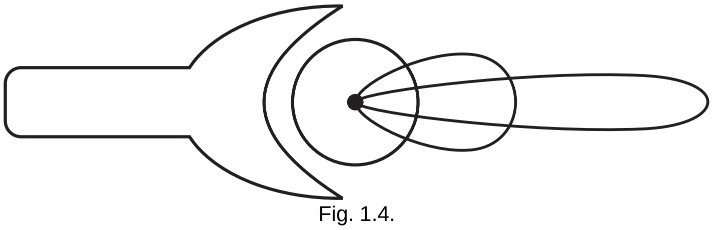

Fig. 1.4.

In the torch light, the larger the diameter of the optical parabolic reflector the narrower will be the beam produced by focussing, to get a brighter beam at a given distance, as shown in Fig. 1.8, in which the light distribution pattern 1 is omnidirectional, due the torch bulb alone, beam pattern 2 is due to parabolic optical reflector, D_1 while the sharper beam pattern 3 is due to the larger diameter of the parabolic reflector 2.

1.10.1 Pattern and gain of microwave parabolic dish antennas

In microwave parabolic dish antenna, the torch bulb is replaced by a horn feed radiator of microwaves, while the dish is made metallic of good conductivity with diameter of the dish made much larger than that of torch light as required due to the much longer wavelength of microwaves compared to that for torch light. The structure of a microwave dish antenna is shown in Fig. 1.9. In this the microwave feed horn located at the focal point of the dish, is not however, omnidirectional but the beam of the horn feed is broad enough to cover the full area of the parabolic dish as shown in Fig. 1.9 with no power being lost by radiating millimeterwave microwaves in the direction away from the dish area. The width of the microwave beam can be estimated by using the same equation as used for estimating the beamwidth for the torch light beam, the only difference between the light and microwave dish being the wavelength.

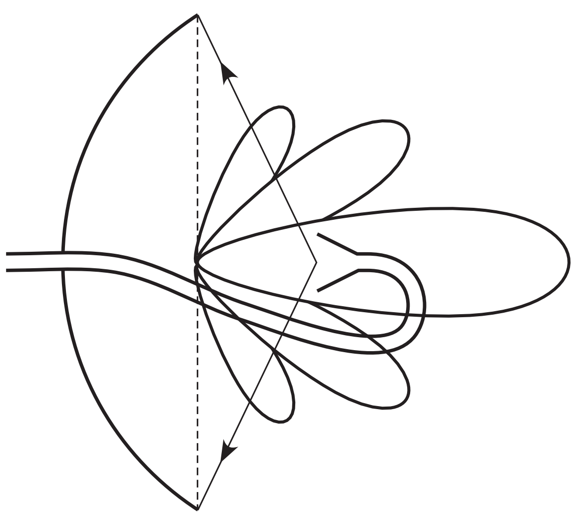

Fig. 1.9.

Thus, the beamwidth of the parabolic dish antenna is given by

$$\theta = 1.22\ \lambda/D \text{ radians} \qquad \text{...(1.6)}$$

$$= 70\ \lambda/D \text{ degrees} \qquad \text{...(1.7)}$$

and it can be shown that the gain G of the antenna, which determines the power flux of the beam at the beam centre at a given distance compared to that of the omnidirectional pattern without the dish, is given by

$$G = \frac{4\,\pi\,A}{\lambda^2}$$

where, $A = \frac{\pi\,D^2}{4}$ is the front mouth or aperture area of the dish.

1.10.2. Effective radiated power and power received

The use of the dish effectively increases the transmitter power by the factor G along the axis of the beam to produce an effective transmitter power, P, given by

$P = P_T\,G_T$, where P_T is the real power of the microwave transmitter. P is called the Effective Isotropic Radiated Power (EIRP). We can then calculate the power flux along the beam centre reaching the receiving end by using the inverse square law of the fall of power with distance, as in optics. Thus the power flux, P_f, at the receiving antenna aperture at a distance from the transmitting dish antenna is given by $P_f = \frac{P_T\,G_T}{4\,\pi\,d^2}$. If the diameter of the receiving dish antenna is also the same as that of the transmitting dish antenna, each being of aperture area being given by a $A = \frac{\pi\,D^2}{4}$, then the power, P_r, picked up by the receiving dish antenna is given by

$$P_r = P_f \times A = \frac{P_T\,G_T}{4\pi d^2} \times A = \frac{P_T\,G_T}{4\pi\,d^2}\,\frac{\pi\,D^2}{4}$$

Fig. 1.10 shows the configuration of typical microwave link with parabolic dish antennas.

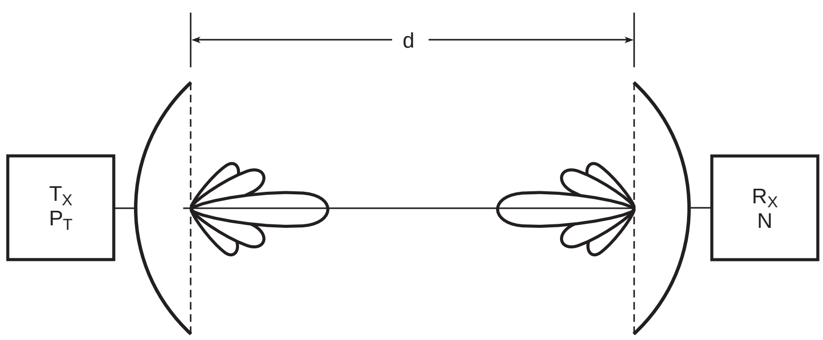

Fig. 1.10.

It can be shown from the reciprocity theorem that the gain of a parabolic dish antenna is related to its area A by the equation

$$G = \frac{4\,\pi\,A}{\lambda^2} = \frac{4\pi}{\lambda^2}\;\frac{\pi\,D^2}{4} = \frac{\pi^2\,D^2}{\lambda^2} \qquad ...(1.9)$$

from which we have

$$P_r = \frac{P_T\,G_T}{4\pi d^2} \times A \quad P_r = \frac{P_T}{4\pi d^2}\;\frac{\pi^2\,D^2}{\lambda^2}\;\frac{\pi\,D^2}{4}$$

$$= \frac{P_T}{4d^2}\;\frac{\pi^2\,D^4}{4\lambda^2}$$

$$= \frac{P_T\,\pi^2\,D^4}{16\,d^2\lambda^2} \qquad ...(1.10)$$

$$\cong \frac{10\,P_T\,D^4}{16\,d^2\,\lambda^2},\ \text{as}\ \pi^2 \cong 10$$

or

$$P_r = \frac{P_T\,D^4}{1.6\,d^2\lambda^2} \qquad ...(1.11)$$

If the diameters of the Transmitting and receiving dishes are different, D_1 and D_2 respectively, than we have

$$P_r = \frac{P_T\,D_1^2\,D_2^2}{1.6\,d^2\,\lambda^2} \qquad ...(1.12)$$

Now the wavelength λ is related to frequency f through the relation

$$f\,\lambda = c \qquad ...(1.13)$$

or

$$\lambda = c/f \qquad ...(1.13)$$

Where c is the velocity of propagation of the microwaves, that may be assumed to be equal to that of light propagating in free space, as the effect of the atmospheric environment, through which the microwave propagates, is negligible and may be neglected for initial estimation of the link parameters. For a more accurate estimation of the link parameter one has, however, to take into account the effect of the most adverse atmospheric weather environment, including the highest rate rain events, on the microwave propagation to ensure reliability of the link in the worst atmospheric environmental conditions.

The frequency of the microwave link is decided on the allocation of the frequency band by the ministry of communication for a particular service for which the link will be used. If the allocated frequency be f then the equation for power received P_r is given by equation (1.12) as

$$P_r = \frac{P_T D_1^2 D_2^2}{1,6\, d^2 \lambda^2} = \frac{P_T D_1^2 D_2^2}{1.6\, d^2 (c/f)^2}$$

$$= \frac{P_T D_1^2 D_2^2 f^2}{1.6\, d^2 c^2} \qquad ...(1.14)$$

This received power P_r available at the output terminal of the receiving antenna should be sufficiently larger than the inherent noise level of the receiver, so that the signal to noise ratio, (S/N) before demodulation produce the signal to noise ratio, $(S/N)_D$, after demodulation, which may be high enough to produce an acceptable signal to noise ratio of the demodulated signal from the microwave receiver, to be sent to the customer-user of the link.

1.10.3. Receiver noise level and system noise

Limitation of signal to noise ratio due to the noise level of the receiver is to be estimated in designing a link. The noise figure of the receiver F is defined by the equation

$$F = \frac{(S/N)_i}{(S/N)_o} \qquad ...(1.15)$$

where $(S/N)_i$ is the signal to noise ratio at the receiver input and $(S/N)_o$ is the signal to noise ratio of the receiver output before demodulation, Equation (1.15) can be rewritten as

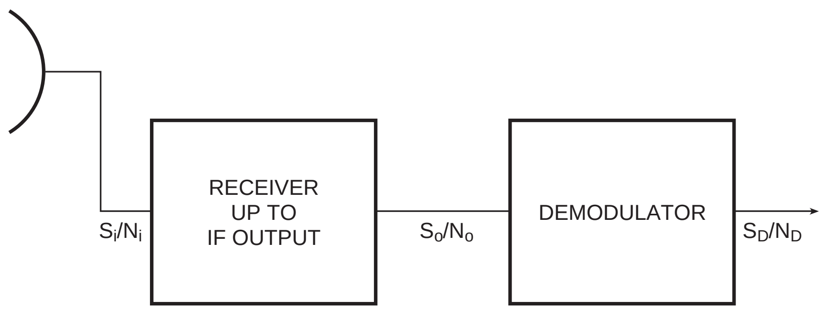

Fig. 1.11.

$$F = \frac{S_i}{S_o} \times \frac{N_o}{N_i}$$

$$= \frac{N_o}{gN_i} \qquad ...(1.16)$$

where $g = \frac{S_o}{S} =$ gain of the receiver before demodulation. As there may be a mixer with a local oscillator to convert the incoming signal frequency to intermediate frequency, the conversion gain of the mixer is multiplied with the gain of the 1F amplifier to get the receiver gain before demodulation. In case, where as a low noise RF amplifier is used before the mixer, the gain of the RF amplifier is also included in the receiver gain.

Now $$N = g\,N_i + N_r \qquad ...(1.17)$$

as, N is the sum of noise power produced in the receiver N_r plus the noise power at the input of the receiver amplified g times by the receiver. Using the relation (1.16) in the equation (1.17) we have

$$F = \frac{gN_i + N_r}{gN_i} = 1 + \frac{N_r}{gN_i}$$

$$= 1 + \frac{N_r}{g} \times - \frac{1}{N}$$

$$= 1 + \frac{N_{ri}}{N_i} \qquad ...(1.18)$$

where $N_{ri} = \dfrac{N_r}{g}$ is the receiver noise power referred to its input. The noise power N_i at the receiver input is normally assumed to be the thermal noise power produced by the resistance of the receiver matched to the antenna radiation resistance. This is valid for ground based line of sight links at microwaves. If the receiver input resistance R_i is matched to a signal generator having the same output resistance R_g, as the receiver input resistance R_i, then the S_i/N_i may be estimated by referring to Fig. 1.12.

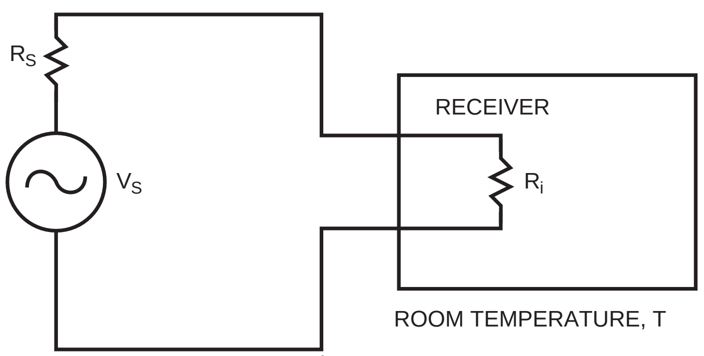

Fig. 1.12.

$$S_i = i^2 R_i$$

where i is the microwaves current flowing between the generator and the receiver

or $$S_i = \left[\frac{V_s}{R_s + R_i}\right]^2 R_i$$

where V_s is the microwave voltage generated at the signal source of the generator.

or $$S_i = \frac{V_s^2 R_i}{[R_s + R_i]^2}$$

For optimum power transfer of noise power from the signal source to the receiver input we make $R_i = R_s$, when we have

$$S_i = \frac{V_s^2}{(2R_i)^2} R_i = \frac{V_s^2}{4 R_i}$$

The thermal noise voltage produced by R_s at the environment of room temperature T, can be shown to be given by $V_n = \sqrt{4kTBR_s}$.

where $k = 1.38 \times 10^{-23}$ is the Boltzman's constant and B is the receiver bandwidth.

$\therefore$ Noise power at the receiver input N_i is given by

$$N_i = \frac{V_n^2}{(R_s + R_i)^2} \times R_i$$

$$= \frac{4kTBR_s}{4 R_s^2} \times R_s \text{ under matched condition with } R_i = R_s$$

or $$N_i = kTB, \qquad \text{...(1.19)}$$

which is independent of R_s or R_i.

Using this relation in equation (1.18) we have

$$F = 1 + \frac{N_{ri}}{N_i} = 1 + \frac{N_{ri}}{kTB}$$

$$\therefore \qquad \frac{N_{ri}}{kTB} = F - 1$$

or $$N_{ri} = (F - 1)\, kTB \qquad \text{...(1.20)}$$

$$= k.\,(F - 1)\, T.\, B = k\, T_R B$$

where $(F - 1)\, T = T_R =$ noise temperature of the receiver.

For microwave line of sight link the total noise power referred to the input of the receiver is given by

$$N = N_i + N_{ri} = kTB + (F - 1)\, kTB = FkTB \qquad \text{...(1.21)}$$

by combining equations (1.19) and (1.20).

For microwave link, as the dish antenna is located on the ground and antenna beam is directed horizontally over the ground towards the receiving dish antenna also installed at ground, the antenna radiation resistance will have an effective temperature equal to the temperature, T, near the ground environment for which $N_i = kTB$

and equation (1.21) is thus valid. However, in case the antenna beam is directed upwards, as insatellite communication links connecting the satellite with the ground station by a microwave link, N_i may be much less than kTB, as the antenna is looking towards the sky where the sky temperature is much lower than that of the environment near the ground as the contribution of the sky noise is the thermal noise of atmospheric water vapour in the sky at a very low temperature together with some amount of galactic noise. Both of these noise sources are much worker than the noise from the ground level atmospheric environment due to water vapour and that from the earth's surface, which are around the room temperature T, in case of line of sight ground based microwave links.

1.10.4. Receiving system noise temperature for satellite link

For estimation of the lower noise level picked up by antenna with its beam pointed upwards, as in Satellite Communication we make use of the antenna temperature T_a which is the equivalent temperature, to which the antenna radiation resistance or the terminal resistance, must be raised to account for the noise picked up by antenna from the sky. In that case $N_i = k\ T_a B$ instead of kTB and thus the total noise power referred to the input of the receiver, is given by

$$\begin{aligned} N = N_i + N_{ri} &= k\ T_a\ B + k\ (F-1)\ TB \\ &= k\ T_a\ B + k\ T_R B = k\ (T_a + T_R)\ B \\ &= k\ T_{sys}\ B \end{aligned} \quad \text{...(1.22)}$$

where $T_a + T_R = T_{sys}$ is the equivalent system noise temperature of the satellite receiving system.

1.10.5 Complete link design equations

Fig. 13 shows a microwave link set up for which the design equations are summarised as follows :

Antenna beam widths $\theta_1 = 70\dfrac{\lambda}{D_1}$, and $\theta_2 = 70\ \dfrac{\lambda}{D_2}$

Antenna gains $G_1 = \dfrac{\Pi^2\ D^2}{4}$ and $G_2 = \dfrac{\Pi^2 D^2}{4}$.

Power received at the receiving end for the down link operating at f_1.

$$P_{r_1} = \frac{P_{T1}\ D_1^2\ D_2^2}{1.6\ d^2\ \lambda^2} \qquad \frac{P_{T1}\ D_1^2\ D_2^2\ f_1^2}{1.6\ d^2 c} \quad \text{...(1.23)}$$

For the uplink at f_2 the satellite receiving system will have the received power, P_{r_2}, given by

$$P_{r_2} = \frac{P_{T2}\, D_1^2\, D_2^2}{1.6\, d^2\, \lambda^2} = \frac{P_{T_2}\, D_1^2\, D_2^2\, f_2}{1.6\, d^2 c} \qquad \text{...(1.24)}$$

For two way simultaneous links, the two frequencies are required to be used for forward or uplink and backward or down link, so that the high transmitter power at any one end may not leak into the highly sensitive receiving system at the same end due to insufficient isolation between the transmitter and receiver. The forward and reverse transmission frequencies are, in fact, separated by a minimum amount required to avoid the problem of high power transmission leaking into the receiver of the same end. The difference of the frequencies f_1 and f_2 can not, in fact, be made large primarily for the reason that the same dish antenna with its feed used for the up/down link cannot be developed to cover a widely separated up/down link frequencies f_2 and f_1. For this reason, the frequency difference between f_1 and f_2 for the downlink and uplink, respectively, is kept within about 3 GHz.

The signal power S, before demodulation at the receiver input is often called the carrier power C and it must be, at least, 20 dB above the total receiver noise power both referred to the receiver input, so that after demodulation C/N can reach a value of the order of $(S/N)_D = 40 - 50$ dB, acceptable to the consumers. The increase in the $(S/N)_D$ ratio after demodulation of C/N is due to the nonlinear processing mechanism of the demodulator and the ratio $p_r = \frac{(S/N)_D}{C/N}$ is called the processing gain of the demodulator. The value of p_r depends on the type of modulation and the modulation index.

1.10.6. Worked out problem on link design

Example 1.1. *A microwave line of sight link at 10 GHz is to be designed for communication over a distance of 10 km. If the diameter of the dishes at the two ends of the link are the same and equal to 3m, calculate the transmitter power required when the receiver noise figure at the receiving end is 10 dB, receiver bandwidth is 1 GHz and the C / N ratio is required to be 40 dB.*

Ans. From equation for power received P_{ri} we have

$$P_{ri} = \frac{P_T\, D_1^2\, D_2^2\, f^2}{1.6\, d^2\, c^2} = \frac{P_T\, D^4\, f^2}{1.6\, d^2\, c^2}$$

where $D_1 = D_2 = D = 3\text{m}$

$$f = 10 \text{ GHz} = 10 \times 10^9 \text{ Hz}$$

$$d = 30 \text{ km} = 30 \times 10^3 \text{ m}$$

$$c = 3 \times 10^8 \text{ m/sec}$$

$$\therefore \qquad P_{ri} = \frac{P_T \times 3^4 \times [10^2 \times 10^9]^2}{1.6 \times (30 \times 10^3)^2 \times [3 \times 10^8]^2}$$

$$\approx \frac{P_T \times 3^4 \times 10^{20}}{1.6 \times 3^2 \times 10^8 \times 3^2 \times 10^{16}} = \frac{P_T\, 10^{-4}}{1.6}\, W$$

Total noise power, N, at the receiver input is given by

$$N = F\,kTB$$

$$10 = 4 \times 1.38 \times 10^{-23} \times 300 \times 10^9$$

where $T = 273 + 27 = 300$ for the environmental temperature of 27°C.

$$= 4 \times 1.38 \times 300 \times 10^{-23} \times 10^9\, W$$

$$\cong 10 \times 4 \times 10^{-21} \times 10^9\, W$$

$$C = 40 \times 10^{-12 \text{ watt}}$$

where C is the carrier power.

As, the carrier power, C, is to be 40 dB above N we should have

$$C = 40 \times 10^{-12} \times 10^4 = 40 \times 10^{-8} \text{ Watt}$$

Assuming, $P_{ri} = C$, we have

$$\frac{P_T}{1.6} \times 10^{-4} = 40 \times 10^{-8}$$

$$\therefore \qquad P_T = 40 \times 1.6 \times 10^{-8} \times 10^4 \text{ W}$$

$$= 64 \times 10^{-4} \text{ W} = 6.4 \times 10^{-3} \text{ W} = 6.4 \text{ mW}.$$

Example 1.2. *A satellite communication link for earth station has got a parabolic dish antenna of diameter 10 m fed from a microwave transmitter at 6 GHz having a power of 3 kW. If the overhead geostationary satellite has a microwave dish antenna of diameter 2 m, calculate the C/N ratio at the satellite receiving terminal for the up link having a noise temperature 300 K and a bandwidth of 500 MHz.*

Solution. We have for the uplink at 6 GHz

$$P_{ri} = \frac{P_T\, D_1^2\, D_2^2\, f^2}{1.6 \times d^2\, c^2} = \frac{3 \times 10^3 \times 10^2 \times 2^2 \times (6 \times 10^9)^2}{1.6 \times (36500 \times 10^3)^2\, (3 \times 10^8)^2}$$

$$= \frac{3 \times 2^2 \times 6^2 \times 10^{23}}{1.6 \times (3.65)^2 \times 10^{14} \times 3^2 \times 10^{16}}$$

$$= \frac{12 \times 36 \times 10^{23}}{1.6 \times 13.32 \times 9 \times 10^{30}} = \frac{432 \times 10^{23}}{191.81 \times 10^{30}} \text{ W}$$

$$= 2.252 \times 10^{-7} \text{ W} = 0.2252\ \mu \text{ W}$$

$$N = k\ T_{sys}\ B = 1.38 \times 10^{-23} \times 300 \times 500 \times 10^{6}$$

$$\cong 4 \times 10^{-21} \times 5 \times 10^{8} = 20 \times 10^{13} \text{ W}$$

$$= 2 \times 10 - 12 \text{ W}$$

$$\therefore \qquad \frac{C}{N} = \frac{P_{ri}}{N} = \frac{2.252 \times 10^{-7}}{2 \times 10 - 12} \cong 1 \times 10^{5} = 50 \text{ dB}$$

Example 1.3. *If the transmitter power of the satellite terminal is 200 W, calculate the C/N ratio of the down link at 4 GHz, at the earth station, assuming the noise temperature of the receiving system of the earth station to be 50 K with the receiver band width is 500 MHz.*

Solution. We have for the down link at 4 GHz

$$P_{ri} = \frac{200 \times 2^2 \times 10^2 \times (4 \times 10^9)^2}{1.6 \times (36500 \times 10^3)^2 \times (3 \times 10^8)^2} \text{ W}$$

$$= \frac{2 \times 2^2 \times 10^4 \times 4^2 \times 10^{18}}{1.6\ (3.65 \times 10^7)^2 \times 3^2 \times 10^{16}} \text{ W}$$

$$= \frac{128 \times 10^{22}}{1.6 \times 13.32 \times 10^{14} \times 9 \times 10^{16}} \text{ W}$$

$$= \frac{128 \times 10^{22}}{191.81 \times 10^{30}} \text{ W} = 0.6673 \times 10^{-8} \text{ W}$$

$$N = k\ T_{sys}\ B$$

$$= 1.38 \times 10^{-23} \times 50 \times 500 \times 10^{6}$$

$$= 1.38 \times 25 \times 10^{-23} \times 10^{9} \text{ W}$$

$$= 1.38 \times 25 \times 10^{-14} \text{ W}$$

$$\frac{C}{N} = \frac{P_{ri}}{N} = \frac{0.6673 \times 10^{-8}}{1.38 \times 25 \times 10^{-14}}$$

$$= \frac{6.673 \times 10^{-9}}{34.5 \times 10^{-14}}$$

$$= \frac{6.673 \times 10^{-9}}{3.45 \times 10^{13}}$$

$$= 1.9342 \times 10^{4}$$

$$= 40 + 2.865 \text{ dB} \cong 43 \text{ dB}$$

1.11. Application of Ray and Wave Theories in Microwave Links

Ray theory is often applied in microwave links for convenience. However, as microwaves are basically electromagnetic waves, application of wave theory would be more appropriate.

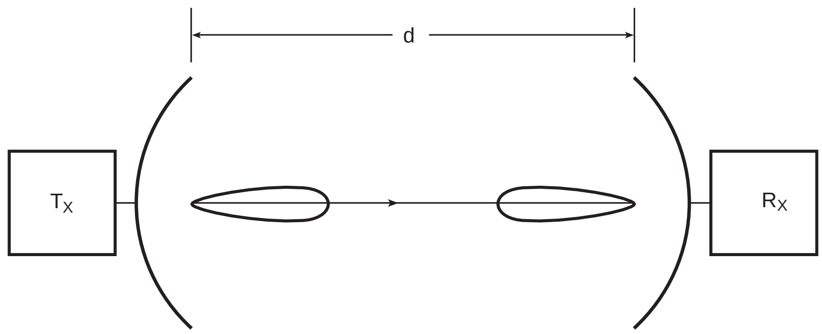

Fig. 1.13.

1.11.1. Microwave antenna pattern and wave theory

In microwave parabolic dish antennas application of the ray theory indicates how the dish antenna produces parallel ray paths by reflection of the radial ray paths produced by the feed of the dish antenna located at the focal point of the dish as shown in Fig. 1.14. The parallel beam is received by the receiving dish antenna to reconvert the parallel ray paths to the radial ray path fed to the feed of the receiving dish antenna, shown in Fig. 1.14. According to ray

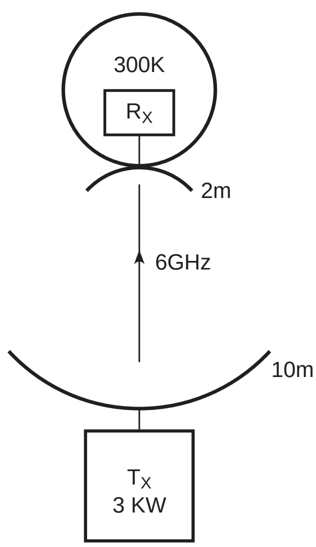

Fig. 1.14.

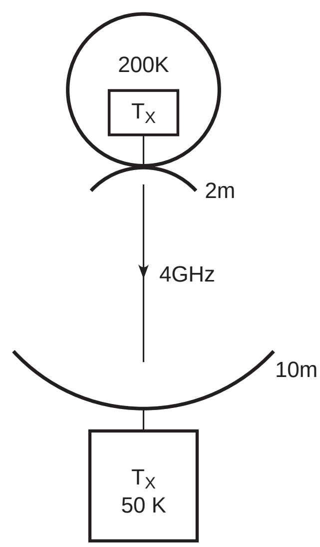

Fig. 1.15.

theory, this figure is correct. However, in reality the ray theory is not valid. For, if a parallel ray path is produced by a parabolic reflector antenna at the transmitting end, the microwave beam would then be really be parallel and, therefore, the intensity of the signal will be independent of the distance between the transmitter and receiver. This is, not possible, as it is against the inverse square law of fall of power flux or power per unit area with distance. This is resolved, if we bring in the wave theory in place of the ray theory, in which case, the radial ray paths indicate the path of wave propagation from the horn, the wave being reflected from the parabolic reflector to reach in parallel paths to the aperture area of the parabolic reflector as shown in Fig. 1.14 to produce a plane wave front covering the aperture area. The ray paths from the horn to the aperture as shown in the Fig. 1.15, in fact, indicates the paths of wave propagation from the horn to the aperture, with the total path length from the horn feed to the aperture, of the parabolic reflector being equal for each ray path, due to reflection in the parabolic reflector, thus producing a plane wavefront at the aperture of the parabolic reflector, bounded by the circular aperture area of the dish. We know from optics, how a plane wavefront bounded by a circular aperture, produces a diffraction pattern ahead of the aperture. The aperture of the parabolic antenna acts as such a circular aperture of area being illuminated by the plane wave front generated by the horn feed through reflection in the parabolic reflector. The main beam width, θ, of the antenna is obtained from optical wave theory, as given by

$$\theta = 1.22 \frac{\lambda}{D} \text{ radians} \qquad \text{...(1.25)}$$

$$= 70\ \lambda/D \text{ degrees} \qquad \text{...(1.26)}$$

where λ is the wavelength of the microwave and D is the diameter of the aperture. As the beamwidth cannot be 0°, with finite diameter D of the aperture, the antenna beam for the finite diameter D will not be parallel but will be conical with the width θ as given by

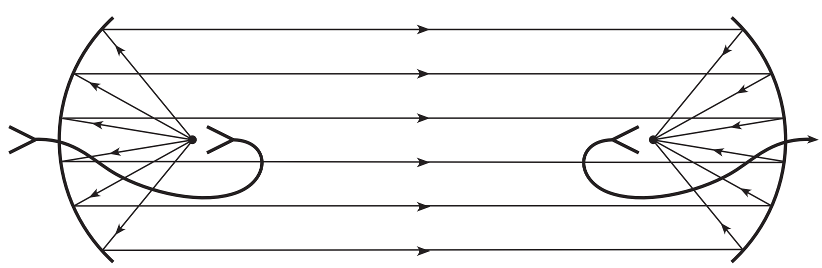

Fig. 1.16.

equation (1.26) and shown in Fig. 1.16 and, therefore, the inverse square law of fall of power flux with distance will become effective. The parabolic reflector only concentrates the beam into a narrow conical beam of width θ, from the wide beamwidth of the horn feed, to produce an antenna gain, which is defined as the ratio of power fluxes at fixed distance from the parabolic dish antenna to that of an omnidirectional antenna, both being fed from the same microwave power source.

Besides these, the side lobes of the antenna radiation pattern are like the secondary maxima of the diffraction patterns in optics. Thus the optical wave concept is directly applicable to estimate the radiation pattern of microwave parabolic dish antennas.

1.11.2. Obstruction of Microwave line of sight propagation and wave theory

In a microwave link the propagation of the microwave occurs along the line of sight path between the transmitting and receiving antennas. If the line of sight propagation is obstructed by a building or hill top, the link will be stopped. For such obstructions of the link, the ray theory suggest that the building top B_2, should just cross the line of sight path TR, as shown in Fig. 1.17. However, in reality the link path is found not to being stopped by the building top B_2, as the propagation of microwave in the line of sight path is governed by wave theory, according to which we have to obstruct at least the Freshnel first half period zone, the boundary of which is ellipsoidal as shown dotted in Fig. 16. For the purpose, it is important to estimate the boundary of the Freshnel 1st half period zone from wave theory. The obstruction of the microwave starts when the height of the building touches the lower boundary of the Freshnel 1st half period zone as shown in the building top B_1 and the obstruction is completed when the building top reaches the height B_3, the upper boundary of the Freshnel 1st half period zone.

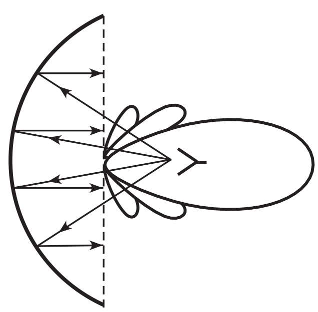

Fig. 1.17.

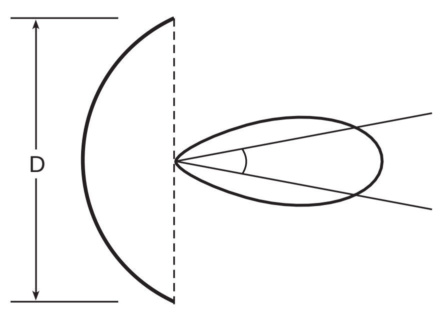

Fig. 1.18.

In the building top is somehow arranged to move gradually from B_1 to B_3 the received microwave power P_r will be gradually blocked as shown in Fig. 1.18.

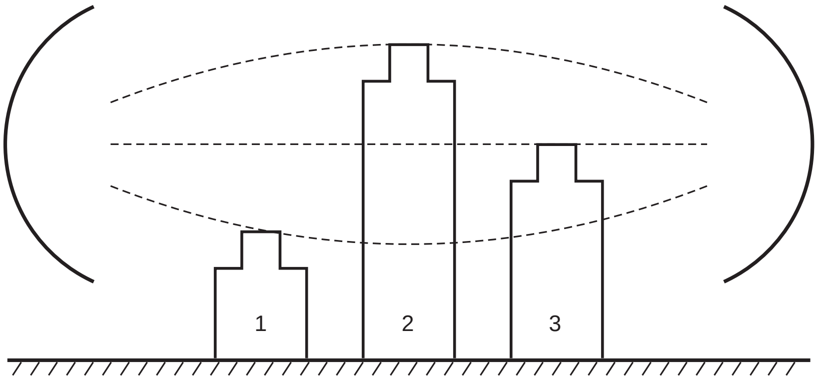

Fig. 1.19.

From wave theory it can be shown that the diameter of the Freshnel 1st half period zone along the line of sight propagation path is maximum at the centre of the link path, where its diameter d is, in fact, given by $d = \sqrt{r\lambda}$, where r is the range of the link or the distance between the transmitting and receiving antennas. Therefore, precautions must be taken so that the heights of the antennas of the link should be high enough to avoid any obstruction of the line of sight first half period zone by any building top.

It may be mentioned here that the concept of Freshnel first period zone, which determines the propagation of the microwave

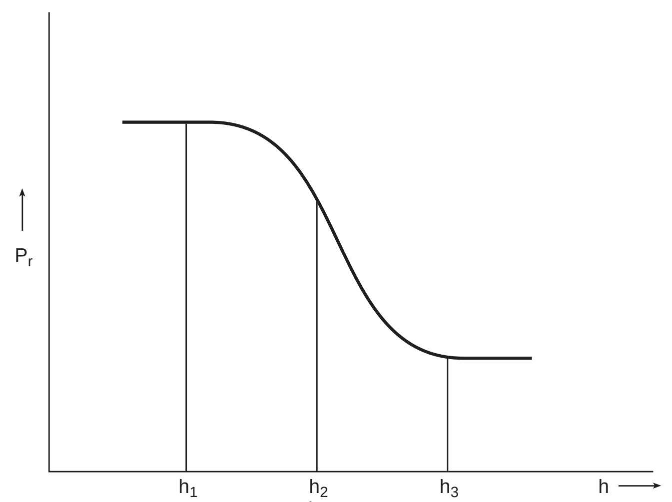

Fig. 1.20.

beam is similar to the line of sight propagation of an optical ray. At any point of the ray path the wave plays its role by producing concentric circular half period zones in a plane orthogonal to the direction of propagation of the ray as shown in Fig. 1.18.

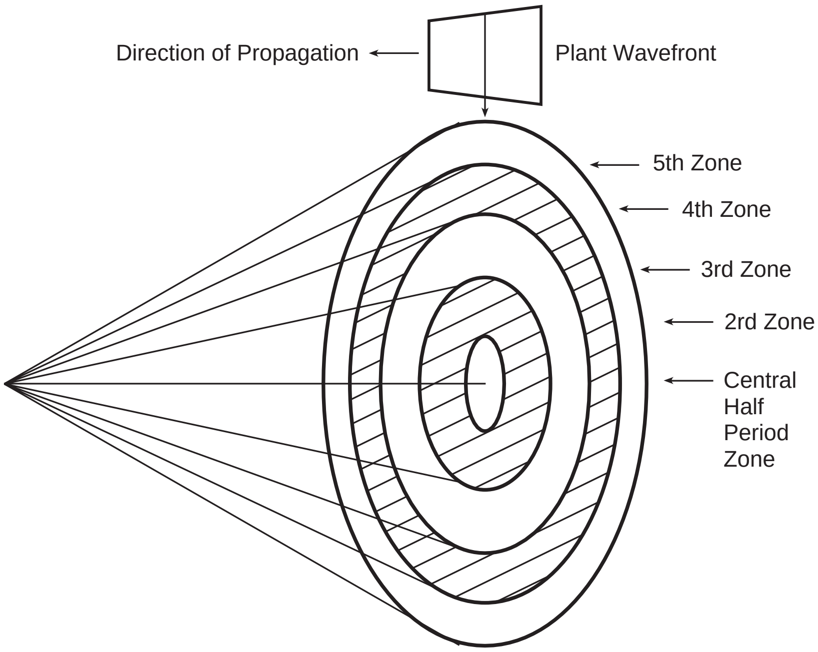

Fig. 1.21. Half period zones for light waves. Light from 3rd zone is conelled by the 2nd zone, while light from the 5th zone is concelled by the 4th zone, retaining only the light from the 1st half period zone, which propagates, in a straight line path along the direction of propagation.

The path difference between ray paths 1 & 2, between 2 & 3 and between 3 & 4 are each equal to $\lambda/2$ so that the secondary waves radiated from the consecutive half period zones, cancel each other in the forward direction, retaining only the 1st half period zone for the propagation of the wave in the forward direction. Thus to stop the light beam one has to obstruct only the 1st half period zone, which is otherwise maintained for the propagation of light beam along a straight line path to have rectilinear propagation of light beam. However, the diameter of the 1st half period zone for an optical wave is very small, being almost like a geometrical point, as the wavelength is very short of the order of 4000–8000 Angstroms or 0.4—0.8 micrometer making $d = \sqrt{r}\lambda$ very small, like a point. Thus the optical beam, due to the 1st half period zone being like a point, it may be called a ray. However, at microwaves the wavelength λ is much larger being about 1 m to 1 mm corresponding to 300 MHz to 300 GHz respectively. The distance or range r of the link is also much larger, being about 10–5 km, at microwaves compared to that of the optical case. In view of this, while the optical beam due to 1st half period zone may be called a ray, at microwaves the wavelength λ being much larger, about 1 mm to 1 m corresponding to 300 GHz to 300 MHz frequency respectively, ray theory cannot be used. The diameter of the Fresnel 1st period zone at the link centre of a microwave link at 1 GHz (λ = 30 cm), for instance, may be as large as 120 m as may be calculated using the relation $d = \sqrt{r\lambda}$, indicated below :

$$
\begin{aligned}
d &= \sqrt{r\lambda} = \sqrt{50 \times 10^3 \text{ m} \times 30 \text{ cm}} \\
&= \sqrt{50 \times 10^3 \times 0.30} = \sqrt{5 \times 0.3 \times 10^4 \text{ m}} \\
&= \sqrt{1.5 \times 10^4 \text{ m}} \\
&\cong 1.2 \times 10^2 \text{ m} = 120 \text{ m}
\end{aligned}
$$

The boundary of the Freshnel half period zone is really an ellipsoside of revolution with its axis along the ray path connecting the centres of the transmitting and receiving parabolic reflectors as shown in Fig. 1.22 (*a*). Clearance of the 1st half period zone may, in fact, be a great problem for microwave link installations in Hilly areas, where hill tops are often located in between the Transmitter and Receiver, obstructing a part or the whole of the Freshnel 1st half period zone.

The height of the antennas are raised sufficiently to ensure clearance of the 1st half period zone from the Hilltops located in an intermediate position between the Transmitting and Receiving terminals, as shown in Fig. 1.22 (*b*).

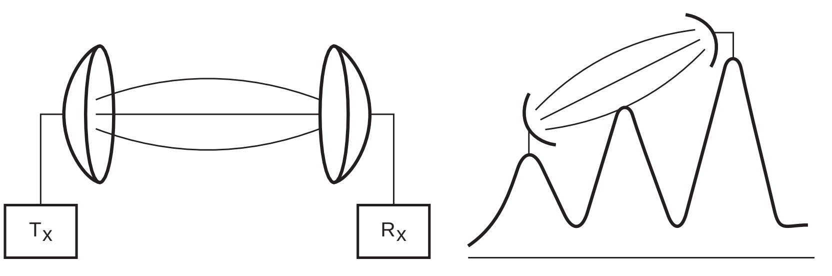

Fig. 1.22.

1.11.3. Ground reflected component of line of sight link and wave theory

In a practical Microwave link set up, the direct ray path is usually interfered with the ground reflected component as shown in

Fig. 1.23. From wave theory both the direct and reflected ray paths are both ellipsoides of revolution, indicating the boundaries of the Freshnel 1st half period zones as shown dotted in Fig. 1.23. As a result the ground reflection point *G* will become elliptical in shape on the ground instead of being a geometrical point *G*, expected for the reflected ray path. Thus the reflectivity or reflection property of the ground involved in the ground reflection path is dependent on the large elliptial area on the ground instead of a geometrical point *G* on the ground.

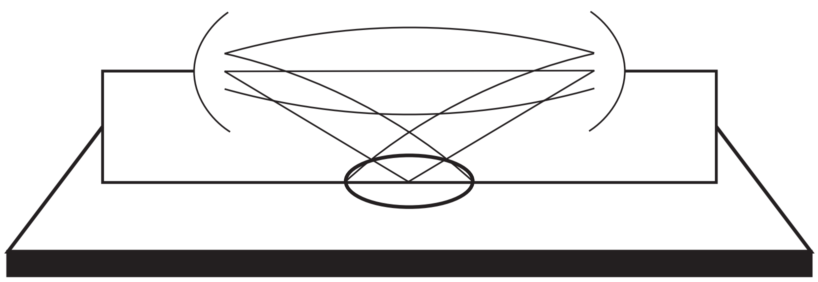

Fig. 1.23.

1.11.4. Bending of Microwave ray path due to diffraction of ray paths expected from wave theory

For communication beyond a high altitude hilltop, often a repeater station is installed at the hill top, as shown in Fig. 1.24 (*a*).

However, if the hill top is not accessable it may still be possible to set up a microwave link across the hill top by using the diffraction effect, as shown in Fig. 1.24 (*b*), by the dotted ray path at the receiving end beyond the hill top. As the diffraction effect is based on the wave theory, the amount of downward bending and the diameter of the Freshnel 1st half period zone may be estimated from the wave theory.

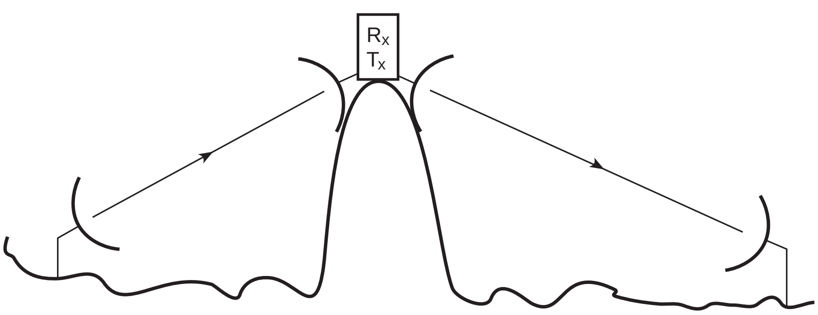

Fig. 1.24(*a*)

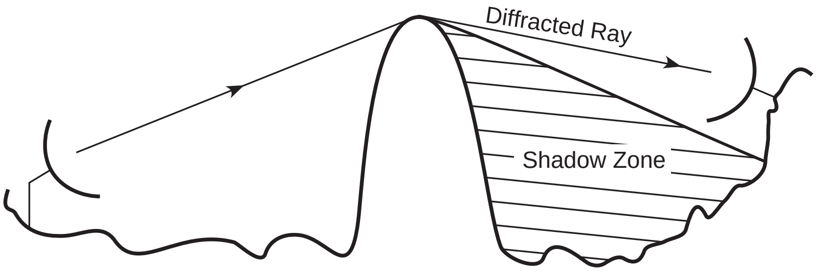

Fig. 1.24(*b*)

1.11.5. Effect of ground reflection on the signal strength at the receiving end of a line of sight microwave link

The phase of the ground reflected component is reversed due to reflection and, therefore, the reflected component tends to cancel the direct ray path of the line of sight link, thereby reducing the power flux at the receiving terminal. However, as the total path length of the reflected component is somewhat larger than the direct ray path, the cancellation of the direct ray path by the reflected components becomes partial and even become additive depending on the path difference. The difference of the path lengths between the direct and reflected ray paths depends on the heights of the transmitting and receiving antennas and an estimate of the path difference and the corresponding phase difference may be made by refering to Fig. (1.25), where

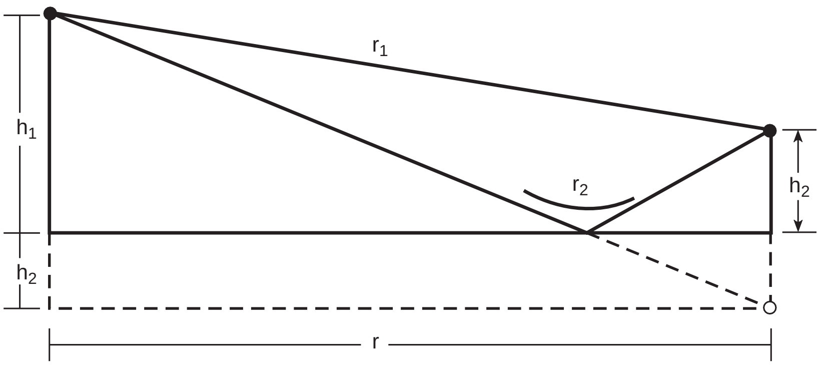

Fig. 1.25.

$$r_1^2 = (h_1 - h_2)^2 + r^2$$
$$r_2^2 = (h_1 + h_2)^2 + r^2$$

Usually $r >> h_1 + h_2$ and, therefore, also $r >> h_1 - h_2$

Using these, we have

$$r_1^2 = r^2 \left[1 + \frac{(h_1 - h_2)^2}{r^2} \right]$$

or
$$r_1 = r \left[1 + \frac{(h_1 - h_2)^2}{r^2} \right]^{1/2}$$

$$r \left[1 + \frac{1}{2} \frac{(h_1 - h_2)^2}{r^2} \right]$$

or
$$r_1 \; r + \frac{1}{2} \; \frac{(h_1 - h_2)^2}{r} \qquad ...(1.27)$$

Likewise,
$$r_2 = r + \frac{1}{2} \; \frac{(h_1 + h_2)^2}{r} \qquad ...(1.28)$$

$\therefore$ The path difference, r_d, is given by

$$r_d = r_2 - r_1 = \left[r + \frac{1}{2} \; \frac{(h_1 + h_2)^2}{r} \right] - \left[r + \frac{1}{2} \; \frac{(h_1 - h_2)^2}{r} \right]$$

$$= \frac{1}{2} \frac{(h_1 + h_2)^2 - (h_1^2 - h_2)^2}{r} = \frac{1}{2} \frac{4 \, h_1 \, h_2}{r}$$

$$= \frac{2 \, h_1 \, h_2}{r} \qquad ...(1.29)$$

The corresponding phase difference between waves in the paths r_2 and r_1 is given by

$$\phi \; d = \frac{2\pi}{\lambda} \times r_d = \frac{2\pi}{\lambda} \times \frac{2 \, h_1 \, h_2}{r} = \frac{4\pi \, h_1 \, h_2}{\lambda r} \text{ radians} \qquad ...(1.30)$$

If E_1 and E_2 are field strengths due to direct and reflected waves, then the resultant electric field at the receiving terminal is obtained from the equation $E_r = E_1 \cos \omega t - E_2 \cos (\omega t - \phi d)$. The negative sign of the 2nd term is due to the phase difference of π produced by reflection of the wave from ground.

Now $E_1 \cong E_2 = E_o/r$ where E_o is the field strength at unit distance, E_o/r being the field strength at the distance r of the receiving terminal following the inverse distance law of fall of field strength with distance, corresponding to inverse square law fall of power with distance. The assumption $E_1 \cong E_2$ is due to the fact that the distances r_1 and r_2 are not sufficiently different to make any sizeable difference between the amplitudes E_1 and E_2, although the difference in r_1 and r_2 is sufficient to produce a difference in phase or phase path of the signals E_1 and E_2, being received at the receiving terminals. As a result, the resultant signal E_r at the receiving terminal is given by

$$\begin{aligned} E_r &= \frac{E_o}{r}\cos\omega\,t - \frac{E_o}{r}\cos(\omega\,t - \phi\,d) \\ &= \frac{E_o}{r}\;[\cos\omega\,t - \cos(\omega\,t - \phi\,d) \\ &= \frac{E_o}{r}\,2\sin\frac{\omega t + \omega t - \phi d}{2}\sin\frac{\omega t - \omega t + \phi d}{2} \\ &= \frac{2E_o}{r}\;\sin\;\frac{\phi d}{2}\sin\left[\omega t - \frac{\phi d}{2}\right] \end{aligned}$$

From which the peak field strength is given by

$$E_p = \frac{2E_o}{r}\;\sin\;\frac{\phi d}{2}$$

or

$$E_p = \frac{2E_o}{r}\sin 2\pi\,\frac{h_1 h_2}{\lambda r} \qquad ...(1.31)$$

At the receiver if

$$\frac{2\pi\,h_1\,h_2}{\lambda\,r} << 1$$

then

$$\sin\frac{2\pi h_1\,h_2}{\lambda r} \cong \frac{2\pi h_1\,h_2}{\lambda r} \qquad ...(1.32)$$

and we then have $E_p = \dfrac{2E_o}{r}\;\;\dfrac{2\pi h_1 h_2}{\lambda r^2}$

$$= \frac{4\pi h_1\,h_2}{\lambda r^2}\,E_o \qquad ...(1.33)$$

The field strength, in fact varies with distance r in a sinusoidal manner due to the repeated phase reversals of the reflected component with increasing distance, as shown in Fig. 1.26(*a*), for a typical case, following the equation

$$E_p = \frac{2E_o}{r}\sin\frac{2\pi\,h_1 h_2}{\lambda r} \qquad ...(1.34)$$

from which Relative field strength is given by

$$E_r = \frac{E_p}{E_o} = \frac{2}{r}\sin\frac{2\pi h_1\,h_2}{\lambda r}$$

In Figure 1.26(*b*), the $\dfrac{2\pi h_1 h_2}{\lambda}$ is double of that in Figure 1.25 (*a*).

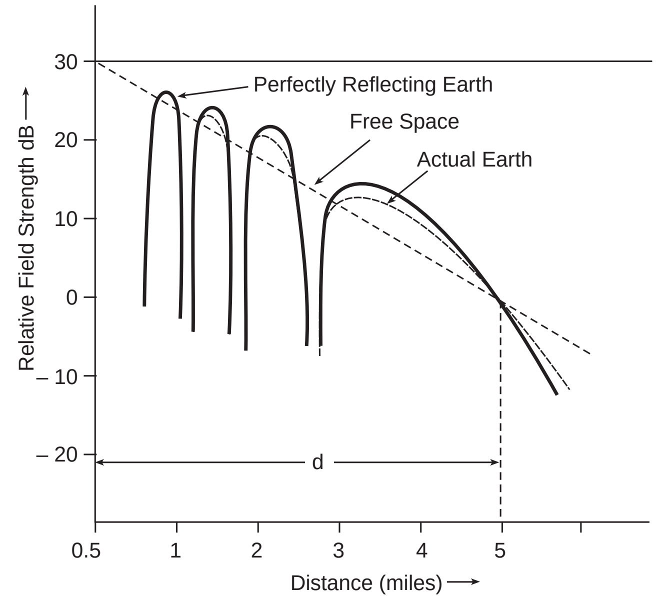

Fig. 1.26(*a*) Field strength as a function of distance, flat earth.

Equation (1.34) indicates that the sinusoidal variation may occur due to variations of the angle $\frac{\phi d}{2} = \frac{2\pi h_1 h_2}{\lambda r}$. This the variation of h_1, h_2, λ or r may produce a similar sinusoidal variation of field

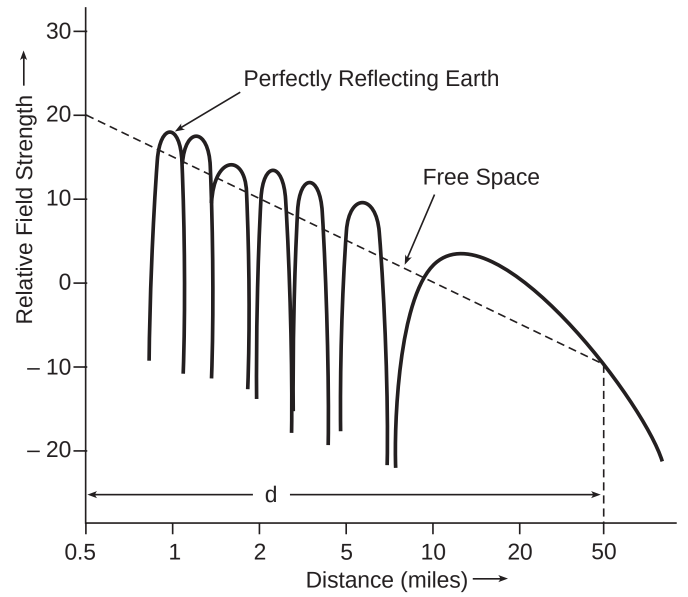

Fig. 1.26(*b*)

strength. If the free space field strength, as shown by the dotted lines in Fig. 1.26 (*a*) and 1.26 (*b*), are compared with the combined field strength of the direct and reflected waves, then the maxima of the sinusoidal variation curve is found to be 3 dB above or double the value for the free space link. The doubling is, in fact, due to path difference between the direct and reflected waves becoming multiples of π with the additional phase change of π on reflection of the wave at the ground. The doubling is, in fact, due to path phase difference between the direct and reflected waves becoming multiples of π with the additional phase change of π on reflection of the wave at the ground, thus making a total phase path difference equal to multiples of 2π, so that the direct and reflected waves may add up in phase to double the field strength and qua draple the power at the receiving end.

Regarding the minima, of the sinusoidal variation of relative field shown in Fig. 1.26(*a*) and 1.26(*b*), the field strength may reach zero under ideal conditions of perfectly reflecting ground. In practice, the ground reflectivity is less than unity and, therefore, the minimum level produced by cancellation of the direct and reflected fields will not be total, making the minima above the ideal value of zero.

For the design of microwave links all the parameters h_1, h_2, λ and r are considered to obtain one of the maxima of the field strength. For the purpose the range r cannot be changed, changes in h_1 or h_2 may be made to get the maximum field strength at the receiving end, which will be about double the free space value. The values of h_1 and h_2 are also decided to avoid any obstruction by a building or hill top in between the transmitter and receiver. It may be mentioned here that excepting the links for communication between hill top, all ground based microwave links have got to take account of the ground reflected components. For, the distance is large and the antenna beam width is not sharp enough to avoid the ground reflected component. However, at higher microwaves or millimeterwaves, the shorter wavelength allows the antenna beamwidth to be sharp enough to avoid the ground reflected component, particularly if the distance is less than 10 km.

For a microwave link

if $$h_1 = h_2 = 100 \text{ m}, \ \lambda = 3 \text{ cm} = 0.03 \text{ m}$$

we have

$$E_p = \frac{2E_o}{r} \sin \frac{2\pi h_1 h_2}{\lambda r}$$

$$E_p = \frac{2E_o}{r} \times \sin \frac{2\pi \times 100 \times 100}{0.03 \times 10 \times 10^3}$$

$$= \frac{2E_o}{r} \times \sin \frac{2\pi \times 10^4}{0.03 \times 10^4}$$

$$= \frac{2E_o}{r} \times \sin 2\pi \times 33.3$$

$$= \frac{2E_o}{r} \times \sin 2\pi \times (32 + 1.33)$$

$$= \frac{2E_o}{r} \times \sin (2\pi \times 1.33)$$

$$= \frac{2E_o}{r} \times \sin (2 \times 180 \times 1.33)$$

$$= \frac{2E_o}{r} \times \sin 234° = \frac{2E_o}{r} \times 0.809$$

$$= 1.618 \frac{E_o}{r}$$

For free space propagation without ground reflection

$$E_p = \frac{E_o}{r}$$

Thus the ground reflection is utilised to increase the field strength. However, the field strength is critically dependent on the heights of the antennas. For instance, if the heights be reduced to 90 m, we have

$$\frac{2Hh_1h_2}{\lambda r} = \frac{2\pi \times 90 \times 90}{0.03 \times 10^4} = \frac{2\pi \times 8000}{0.03 \times 300} = 2\pi \times 27$$

$$\sin \frac{2\pi h_1 h_2}{\lambda r} = \sin 2\pi \times 27 = \sin 2\pi = 0$$

The field strength thus drops to zero with only 10 m decrease in the heights of the antennas.

For a 10 m increase of the heights of the antennas, we have

$$\sin \frac{2\pi h_1 h_2}{\lambda r} = \sin \frac{2\pi \times 110 \times 110}{0.03 \times 10^4}$$

$$= \frac{\sin 2\pi \times 12100}{300}$$

$$= \sin 2\pi \times 0.333 = \sin 2 \times 180 \times 0.333$$

$$= \sin 119.88 = 0.867$$

$$\therefore \qquad E_d = \frac{2E_o}{r} \times 0.867 = 1.734 \frac{E_o}{r}$$

The field strengths thus increased compared to that for $h_1 = h_2 = 100$ m for which $E_d = 1.618 \times \frac{E_o}{r}$. h_1 and h_2 can be increased further to reach the maximum value of $E_r = \frac{2E_o}{r} \times 1$.

Likewise, the field strength can also be increased by decreasing the heights h_1 and h_2 below 90 m to reach the next maximum of field strength at a lower height, provided, the lowering of the height of the antennas does not create any new problem of obstructing of the line of sight wave Freshnel half period zone by any building top, located in between the transmitting and receiving stations.

The field strength can be converted to power flux, p_f, of the signal through the relation

$$p_f = \frac{E_r^2}{120\pi}$$

where $$p_f = \frac{P_T\, G_T}{4\pi r^2}$$

P_T being the transmitter power and G_T is the gain of the transmitting antenna. This conversion may be required to estimate the C/N ratio at the receiving end.

It may be mentioned here that link distance in all the equations connected with wave theory has been referred to as r, while in equations based on ray theory such as that in equation (1.14) the link distance is referred to as d instead of r.

1.11.6. Two way microwave link

Microwave links are usually required to be developed as a simultaneous two way link line that of telephone. For the purpose, the same parabolic antenna is used at each terminal of the link to economise the cost of the link set up. However, to utilise a single parabolic antenna for the two way link, some problems have to be solved. For instance, as both the transmitter and receiver should be aperative at each of the link terminals, the frequency of transmission should be made different from the frequency of reception at each terminal, so that transmitted power at any one terminal may not leak into the receiver at the same terminal. However, as the transmitter power is orders of magnitude higher than the power of the received signal, separation transmitter power is orders of magnitude higher than the power of the received signal, separation transmitter and receiver frequencies has to be made sufficiently high so that the receiver input bandpars filter may not allow the transmitter leakage power to pollute the receiver. Such widely separated transmitting and receiving frequencies may again create a problem of accomodating. Such widely different frequencies by a single parabolics antenna. To avoid this problem, a special microwave ferrite component, called the circulator is used, as shown in Fig. 1.27 (*a*). The circulator effectively circulates microwaves entering it through one of four entries 1, 2, 3 or 4, in the clockwise direction. The circulator can also be designed to circulate microwaves in anticlockwise direction. In Fig. 1.27 (*a*), the microwave transmitter power enters the circulator through the entry 2 and then starts circulating clockwise, when the power is released to the antenna through entry 3, acting as a releasing terminal. If the antenna is properly matched to the terminal 3 then no part of the transmitter power is returned through the transmitter power is returned through the terminal 3 to get out through terminal 4 to pollute the receiver. However, matching of the antenna cannot be made perfect, and, therefore, some part of the power released through terminal 3 will be returned back, the amount depending on the extent of matching as indicated by VSWR of antenna. The returned power then leaks into the receiver through terminal 4. In the same way, due to imperfect matching of the receiver to the terminal 4, part of the leakaged signal to the receiver is returned and released through terminal 1, where, however it may be perfectly matched to avoid any further return of the leakage signal towards the terminal 2, 3 or 4. Regarding the power received by the antenna through terminal 3, it is circulated to be released through terminal 4 to the receiver. If the receiver is properly matched to terminal 4, the full received power gets into the receiver. In practice, however, perfect matching is not possible and, therefore, some part of the received power is returned back to terminal. Thus, the circulator effectively separates out or isolates the transmitter power from the receiver, the amount of isolation being typically about 40 dB. Some additional isolation of about 50 dB may be provided by the bandpass filter at the receiver input if the transmitting and receiving frequencies are separated by about 3 GHz. Thus, the total isolation between the transmitter and receiver may become as large as (40 + 50) dB = 90 dB, which is good enough for satellite communication links at microwaves. For terrestrial line of sight link where the range to be covered is much less so much of isolation may not be required. The configuration of a two way microwave terrestrial link is shown in Fig. 1.27 (*b*). The transmitting and receiving frequencies are f_1 and f_2 respectively, separated by about 3 GHz. It may be mentioned here that the requirement of the amount of

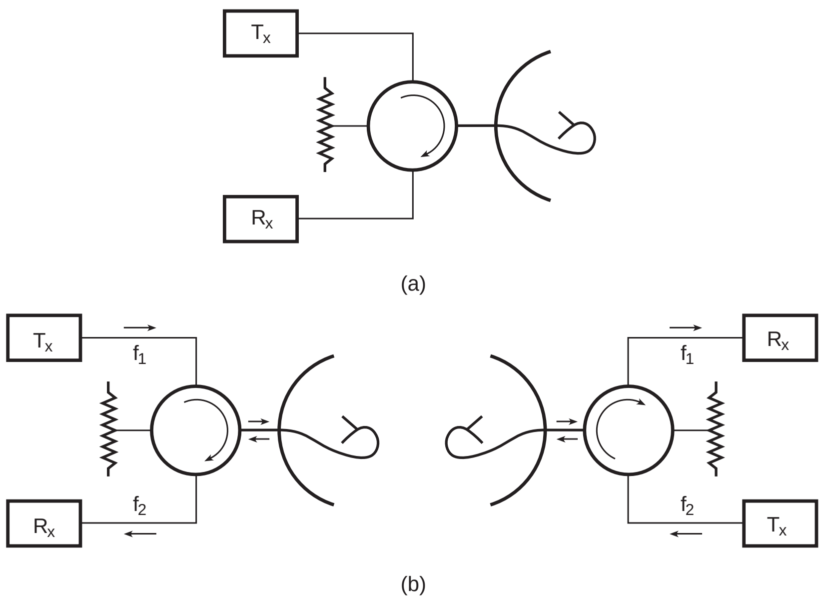

Fig. 1.27.

isolation depends on the antenna gain. For a high gain antenna, the Transmitter power can be kept small, for the same Effective Isotropic Radiated Power (EIRP). The high antenna gain also pickes up the signal with high gain to produce a larger received signal power for the same power flux of the received signal at the antenna. This further reduces the requirement of the amount of isolation between the transmitter and receiver. Reduction of the requirement of isolation amount allows a circulator with lower isolation to be used for which the cost of the circulator will also be reduced.