

IMPERIAL COLLEGE OF SCIENCE AND TECHNOLOGY

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CONSIDERATIONS RELATING TO THE PLANNING AND PERFORMANCE
OF MICROWAVE LINKS

by

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ABSTRACT

Microwave radio relay links is the underlying purpose of this work. Although such links, except for the so-called over the horizon systems, are restricted to line-of-sight, there are cases where the intervention of elevated ground is not prohibitive. This is so where the diffracted field arising from such an obstacle is high enough to maintain the necessary signal-to-noise ratio and keep the system working according to the specified rules and recommendations.

Factors to be taken into account when planning microwave radio relay links are examined in detail with particular attention to the propagation of microwaves in line-of-sight systems. Some comments on C.C.I.R. recommendations and military systems are made and some passive repeater systems are briefly described. The first part of the work ends with the planning of a microwave link at 900 MHz between the roof of Imperial College to a point on the S.R.D.E. site at Christchurch. In this planning each of the relay stations is individually examined taking into account the field survey results.

The second part is devoted to an experimental investigation of obstacle diffraction, using an optical model with a gas laser as the source. Results of measurements for knife-edge and cylinders (both smooth and rough) are given and the effect of radius of curvature and surface roughness is shown. These experimental results are compared with existing theoretical solutions.

The final part of the work is concerned with devising an experimental technique for measuring the phase of the field, since this is important in determining the performance of practical antennas, particularly if the antenna is very near to a rough surface as could be the case in military applications. Two methods based on the same principle, namely interferometry and holography, are investigated. The uniformity of the

phase distribution over the entire aperture is found to be not appreciably disturbed or at least is satisfactorily maintained within the acceptable limits specified for microwave antennas. This implies that the results obtained with the optical model, using a different method of detection, give a reliable indication of the performance of practical antennas.

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CHAPTER ONE

INTRODUCTION

1.1. Background

Microwave radio relay links are communication systems operating at carrier frequencies above about 1000 MHz. The distance between adjacent stations is normally restricted to line-of-sight and intermediate stations as relays are provided in order to cover a large distance. High gain antennas are used to concentrate the transmitter power towards the receiving station, since point to point communications are involved.

Many problems are encountered when planning any sort of radio or microwave link and the most fundamental is, perhaps, that of determining the propagation loss between the transmitting and receiving stations. A compromise between low cost and high operating standards of a link can be achieved by accurately predicting this propagation loss which is dependent on many factors. For example, a ground elevation between two adjacent stations of a line-of-sight microwave link may be a prohibitive factor. On the other hand, careful selection of the relay station sites may result in a considerable reduction of the necessary repeaters for a particular link.

A straightforward relation exists between the propagation loss and the distance between stations if free space conditions are assumed. This is usually the starting point when estimating the actual propagation loss involved. There are, however, two primary processes which, besides distance, heavily influence the strength of the field at the receiving antenna by either contributing to the loss or, as in some cases, reducing it. These processes are the diffraction of the beam by obstacles such as buildings, hills or mountains, which obstruct the line-of-sight between the stations and the scattering of the beam by

inhomogeneities in the atmosphere through which the line-of-sight is supposed to pass. It is not unusual to determine the final route of a microwave link or the exact position of a transmitting station by mainly considering these two processes.

The adoption of ultra high frequency bands for broadcasting purposes has also introduced some problems which have their origin in the diffracted or scattered field beyond the transmitter horizon.

Although both of the above processes are very important in any communication planning it is the diffraction problem which is of interest in this thesis. When estimating the diffracted field it is usual to imagine diffracting obstacles to be replaced by ones of idealised shape for which this field is known; such as knife-edges, wedges, slabs and cylinders.

1.2. Basic Configuration of a Microwave Link

A microwave link, like all radio relay systems, consists of terminal and repeater stations. Two stations and the transmission path in-between is called a radio section or more commonly a "hop". Signals in a suitable form are fed into the transmitter of the first terminal and are derived at the other terminal in approximately the original shape. Repeaters are arranged at appropriate distances between the terminals, when the length of the link or the nature of the intervening terrain makes it necessary, in order to compensate for the path loss by amplifying the signal.

An electromagnetic carrier wave cannot simultaneously transmit two signals in opposite directions. Consequently, for two-way links, two independent one way links on separate working frequencies are required. The second link is normally identical and parallel to the

first one and is usually installed at the same time.

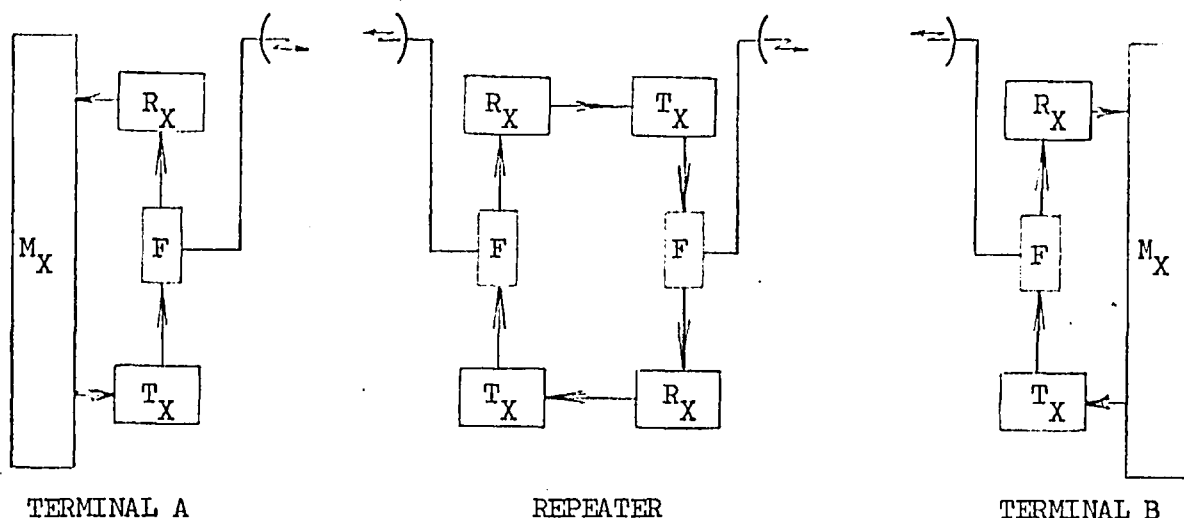


FIGURE 1.1

A BLOCK DIAGRAM OF A MICROWAVE LINK

A block diagram of a two-way link is given in figure 1.1 where two terminal stations, with their multiplex (M_x) equipment, and a repeater station are shown.

At a repeater station the basic signal transmitted does not have to be processed into its original form if only an amplification to compensate for the path loss is required. In this case the through-connection from receiver to transmitter may be arranged at microwave or intermediate frequencies and avoid the necessity for extra equipment. However, if the traffic requirements and the network configuration require the original signal form, a through-connection at baseband is necessary and demodulation and modulation equipment is indispensable. For a two-way repeater, two transmitters (T_x), two receivers (R_x) and four antennas are needed. The requirement for four can be reduced to two if each antenna can be used for simultaneous transmission and reception by means of appropriate filters.

A two-way terminal requires a transmitter, a receiver, two antennas

(which can be frequently combined in one), a modulator and a demodulator. In some cases, where a part of the information carried is derived for local purposes or branching off in a third direction, a mixed arrangement of terminal and repeater is used. Frequently several links are used in parallel through the same route and system, when the capacity required is very large. In such cases branching filters are employed wherever feasible for multiple use of the antennas.

1.3. The Available Frequencies and Their Characteristics

The international allocation of frequency bands for specific services and areas is carried out by the International Telecommunication Union and contained in special publications. Operating frequencies for microwave links are at present restricted to the following bands:

790 - 960	MHz	with some restrictions.
1.7 - 2.3	GHz	
3.8 - 4.2	GHz	also allocated to satellite systems.
4.4 - 5.0	GHz	in Europe mainly for military purposes.
5.85 - 8.5	GHz	some frequencies exclusively for satellite systems.
10.7 - 13.25	GHz	
14.4 - 15.4	GHz	

Although the frequency bands included in the publications go above 15 GHz up to 40 GHz, they are not at present employed for commercial communication systems. The general international regulations provide national authorities with a basis for allocating radio frequency channels to specific services.

The main advantages of the lower frequencies is that reasonable

transmitting powers can be achieved more easily and cheaply than at higher frequencies while the expected fading due to atmospheric conditions is less. This permits longer repeater sections to be planned if a high-gain, stable antenna can be installed. The installation of such an antenna is, however, very expensive. On the other hand, the rapidly improving technology for reasonable transmitter powers in the higher frequency bands together with the high-gain small-size antennas, which present low windage area, makes these bands more and more competitive. The main disadvantage of the highest frequency bands (above 11 GHz) is the associated loss due to scattering by rain and other precipitation.

For radio relay systems it is necessary to adhere to the radio frequency channel arrangements recommended by the C.C.I.R., in order to avoid interference across national frontiers and to facilitate the interconnection of systems between neighbouring countries.

Due to the shortage of the allocated bands repetition of carrier frequencies in each system is inevitable. In planning a microwave link the frequencies are so allocated as to minimise their number, while a "dog-leg" arrangement, as in figure 1.2, is used to avoid "over-reach"

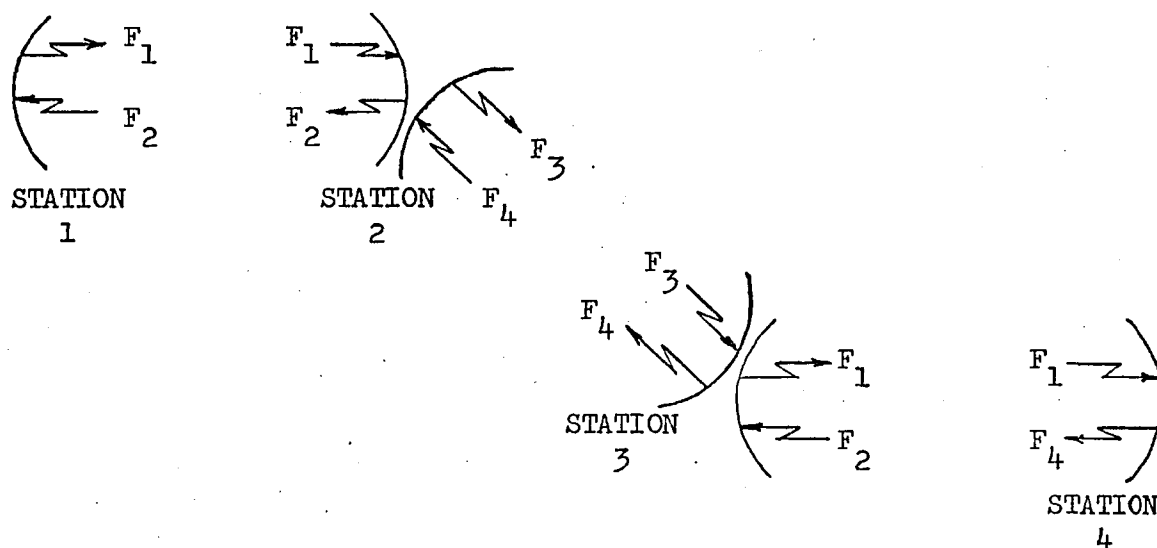


FIGURE 1.2

A "DOG-LEG" ARRANGEMENT

interference at remote stations. In this arrangement the same frequencies can be used at every other station, since the directivity of the antennas minimises the possibility of interference.

1.4. Objectives

The work described in this thesis can be divided into two separate but correlated major parts. The purpose of the first part is to study the propagation of microwaves and apply the theory to a practical problem by planning a microwave link between two specific points in the United Kingdom. The second part deals with the diffraction, particularly, from the experimental point of view, and investigates the diffracted field both in amplitude and phase.

Microwave propagation is described in chapter two, together with the C.C.I.R. recommendations applicable to microwave links, as well as the techniques employed in using passive repeater stations. The various factors determining the signal level in the system is also investigated in the same chapter.

The planning of a microwave link on 900 MHz * is the subject of chapter three. The roof of Imperial College (elevation 183') and a point on the S.R.D.E. site at Christchurch (height 18' above ground level) constitute the terminals of the link. The repeater station locations, the transmitter power, the dish sizes and the overall signal-to-noise ratio over the path are eventually decided. Although the use of frequencies below 1 GHz is exceptional nowadays and generally used for local purposes, where many allowances can be made, the planning in this work has been dealt with as if the link were to be incorporated in some existing international network; so, every standard for this

* This was related to a requirement at Imperial College in connection with another research project.

purpose had to be met. All sites but one (on Isle of Wight) have been visited for visual inspection and field surveys made to ascertain their suitability. Nevertheless, information concerning type of soil, temperature and prevailing weather conditions has not been obtained and have been assumed as normal.

In chapter four the diffraction by knife-edges and cylinders (both smooth and rough) of various radii and roughness is examined. Diffraction measurements, undertaken on an optical path, are given in graphical form and compared with existing theoretical solutions. The model path, using a helium-neon laser as source, facilitates a high degree of control over the diffraction experiments due to its small size. The operation of this model and its limitations are described in the same chapter.

Finally in chapter five, the phase of the field is investigated using interference and holographic methods. These methods are also described together with their difficulties and limitations. The optical model, used for diffraction measurements, is appropriately modified to carry out phase measurements. The results of these measurements are given in photographs and are appropriately discussed.

CHAPTER TWO

MICROWAVE LINKS

2.1. General

Equal performance of radio relay systems and cable systems has been established long ago. To this purpose the CCIR recommendations have to be met. Once these recommendations are met, there are certain advantages in using radio relay instead of cable systems. Most of these advantages rely on flexibility of capacity and position and construction-maintenance costs.

The planning of a radio relay link is carried out following a specific procedure which is more or less always the same for every case. As a start the following factors have to be considered:

- a. Traffic density.
- b. Possible future extensions.
- c. Quality and reliability requirements.
- d. Cost analysis.

Also it is important to identify the centres to be interlinked and the overall capacity of the system.

These considerations will help to decide what kind of a radio link is economically preferable. After that, it is necessary to decide on the final location of the terminals and the number and location of the repeater stations in-between. Proper locations for both terminals and repeaters may be selected using topographical maps with contour lines. Dish sizes can then be fixed and hop attenuation computed. With the typical propagation conditions of the area in mind, the actual quality of the link with relation to the quality required can be checked.

A field survey will finally ascertain the real possibilities of the

solution proposed. This survey should include the reflection points and the ridges should be taken into consideration, if some areas have been left obscure in the map survey.

Propagation tests, for a sufficiently long period of time, will give better evaluation of the system reliability in the most difficult cases. If a proposed path lies along the coast or in a terrain unfavourable for propagation, this kind of test is absolutely necessary.

2.2. Microwave Site Selection

Line-of-sight transmission is needed to install communication on a microwave frequency. Designing a link between any two points, all possible routes are studied with emphasis on the contour and geological structure of the intervening terrain. This is primarily carried out by the study of available topographical maps of largest possible scale and latest date so as to contain the more recent details. From this data a topographical profile chart is prepared for each path showing the elevation of the terminals and all obstacles including hills, trees, water masses, buildings and other structures that might possibly interfere with the line-of-sight transmission.

Additional information can be obtained from field surveys, as extensive as possible, or aerial surveys wherever the former cannot be employed. Information gathered during the field survey is very important and critical in deciding whether or not a location or even the path at all is suitable for the link. Exact geographical site location is determined taking into account the type of soil for rigid construction of towers and building, the existence of any access road or the costs of providing one, the clearance needed to be done before the site can be developed and the area needed for site installations.

Other information obtained during field surveys is the availability,

type, distance and reliability of power supplies, existence of local obstructions to determine minimum antenna heights, distance of nearby airports, temperature differences to provide for heating and ventilation and prevailing weather conditions in the area including the average rainfall, pressure and water vapour as well as the amount of ice and snow, if any. Maximum expected wind velocity is of great importance for the rigidity of narrow beam antennas.

Most of the above information cannot be obtained from map study or even from local authorities. Thus personal field survey by the designer is important for the exact site location.

2.3. Microwave Propagation

2.3.1. Characteristics of a Microwave Link

Microwave radio beams, like all electromagnetic beams, exhibit many of the properties of a beam of light:

- a. They travel in relatively straight lines.
- b. They are reflected by reflecting surfaces.
- c. They are refracted due to varying densities of the atmosphere.
- d. They are diffracted as they pass over transverse terrain, unless the clearance between the line-of-sight and the terrain is large enough to assume free-space conditions.

Depending upon radio section length, path employed and propagation characteristics, radio relay systems can be distinguished as line-of-sight systems and over-the-horizon systems. Utilising the former, hops of ⁽¹⁾ about 50 kilometres long are usually adopted although sections of even twice this length may exceptionally be accepted where the conditions are favourable. In line-of-sight systems intervening terrain or other obstacles do not usually affect the wave propagation between the antennas.

Hops of a distance up to about 500 kilometres can be obtained by

over-the-horizon systems, using high power transmitters, where line-of-sight propagation is seriously obstructed or other factors dictate this kind of communication. The use of over-the-horizon systems is restricted to special applications while line-of-sight systems are commonly used to ensure high system performance.

2.3.2. Free Space Propagation

For propagation calculations the isotropic radiator is taken as a reference antenna,⁽²⁾ because it forms a well defined standard and the calculations are simplified by its introduction. Used as a receiving antenna its function is to absorb power from the radiation field in which it is situated. The amount of power that it absorbs, in relation to the power density in the field, is determined by its effective aperture or effective area (A_{eff}) given by:

$$A_{\text{eff}} = \frac{\lambda^2}{4\pi} \quad (2.1)$$

where λ is the wavelength in metres of the incident radiation.

The power flux density at a distance d metres from an isotropic transmitting antenna of power P_T watts is given by:

$$P_{\text{av}} = \frac{P_T}{4\pi d^2} \quad \text{watts/m}^2 \quad (2.2)$$

It follows that an isotropic receiving antenna situated in a radiation field with power density P_{av} , will deliver into its load a power P_R given by:

$$P_R = P_{\text{av}} \frac{\lambda^2}{4\pi} = P_T \left(\frac{\lambda}{4\pi d} \right)^2 \quad (2.3)$$

When two isotropic antennas, which radiate uniformly in all directions, face each other separated by a distance d , in free space, the ratio of the transmitted power to the received power is called "free space loss" or "transmission loss" and is given by:

$$\frac{P_T}{P_R} = \left(\frac{4\pi d}{\lambda}\right)^2 \quad (2.4)$$

It is customary to refer to the above quantity in decibels as:

$$L = 10 \log \frac{P_T}{P_R} = 20 \log \frac{4\pi d}{\lambda} \quad (2.5)$$

which yields:

$$L = 21.98 + 20 \log \frac{d}{\lambda} \text{ [dB]} \quad (2.6)$$

It is convenient to note that in calculating transmission loss for a more general case the free space loss, for two isotropic antennas one wavelength apart, is almost precisely 22 dB. Equation 2.6 can be recast into a form more suitable for practical purposes by converting the distance d between the antennas from metres to kilometres. Then:

$$L = 82 + 20 \log \frac{d}{\lambda} \text{ [dB]} \quad (2.7)$$

where d is now in kilometres and λ in metres.

Sometimes the transmission loss is referred to as "propagation attenuation" and is defined as the inverse quantity the "transmission equivalent" T , which is the logarithmic expression of "free space transmission ratio" $\frac{P_R}{P_T}$. So we have:

$$L_{\text{(dB)}} = -T_{\text{(dB)}} = -10 \log \frac{P_R}{P_T} \quad (2.8)$$

Figure 2.1 gives in nomogram form the graphical presentation of equation 2.7 and allows a direct reading of the free space transmission loss as a function of distance.

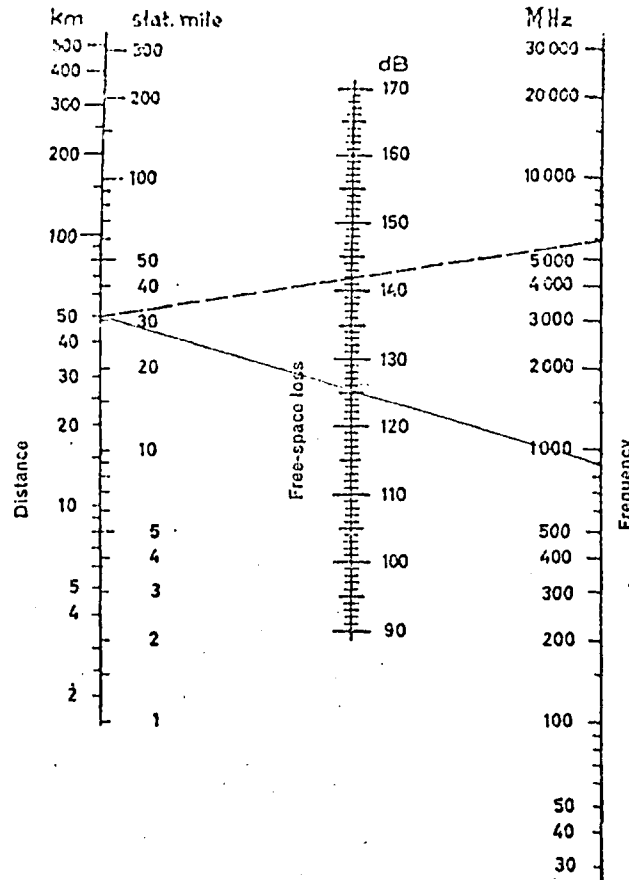


FIGURE 2.1

FREE SPACE LOSS BETWEEN ISOTROPIC RADIATORS AS A
FUNCTION OF DISTANCE AND FREQUENCY

From this it will be seen that the loss increases by 6 dB if the frequency or the distance is doubled.

The gain of a microwave antenna is generally defined as being compared to that of an isotropic antenna. To a first approximation, the gain of an antenna is given by:

$$G = \frac{4\pi A_{\text{eff}}}{\lambda^2} \quad (2.9)$$

or expressed in decibels:

$$G = 10 \log n \frac{4\pi A}{\lambda^2} \quad (2.9)$$

where A is the real area of the antenna and n is its gain factor given by the ratio A_{eff}/A (≈ 0.5 for parabolic antennas)⁽³⁾.

When the gain of the transmitting and receiving antennas are respectively g_T and g_R , in the direction of each other, the power in the receiving antenna will be:

$$P_R = P_T g_T g_R \left(\frac{\lambda}{4\pi d} \right)^2 \quad (2.10)$$

and the transmission loss is given by:

$$\frac{P_T}{P_R} = \left(\frac{4\pi d}{\lambda} \right)^2 \frac{1}{g_T g_R}$$

or expressed logarithmically: (2.11)

$$L' = 20 \log \frac{4\pi d}{\lambda} - G_T - G_R$$

This can be written in terms of the antenna area, as in equation 2.9 to yield:

$$L' = 10 \log \frac{d^2 \lambda^2}{A_T A_R}$$

and if $A_T = A_R = A$: (2.12)

$$L' = 20 \log \frac{d \lambda}{A}$$

2.3.3. Propagation in the Atmosphere

Propagation in free space or in an ideal atmosphere with a constant refractive index, is rather a mathematical concept and never takes place in the lower atmosphere where radio waves propagate in the case of radio relay systems. The relative permittivity of the air (ϵ) is greater than unity varying considerably in time and location. Wave propagation, therefore, may not always follow a straight line, even if there are no obstacles in the line-of-sight.

In most cases a microwave beam, which tends to travel in a straight line, is bent slightly downwards due to refraction depending on the variation of the atmospheric properties. In fact, a microwave path which lies within a few degrees of the horizontal differs negligibly from an arc of a circle. This phenomenon is the basis of the effective-earth-radius method, employed in investigating the propagation in the atmosphere: The straight line propagation may be maintained if a correction of the actual earth curvature is provided.⁽⁴⁾ To do that we have to assume that the refractive index of the air is a linear function of the height above the ground introducing therefore the concept of an effective radius of the earth, which is the true radius R multiplied by the factor:

$$K = \frac{1}{1 + \frac{R}{2} \frac{\Delta_n}{\Delta_h}} \quad (2.13)$$

where $\frac{\Delta_n}{\Delta_h}$ is the gradient at the refractive index. This index is

dependent upon the gradient of the temperature and the pressure of the air with height, but to a much greater extent upon the humidity.

With normal atmospheric conditions for at least 50% of the time, a value of $K = 4/3$, is normally assumed while the effective radius

is taken as $R = 8,500$ km. In practice, K varies from a minimum of $5/12$ for less than 0.1% of the time, up to infinity. These various values are generally used to make templates of plastic or cardboard and can be used as an over or underlay to profiles that have been plotted on special graph papers.

Profiling the terrain underlying the path on a K -earth profile paper, straight line propagation is assumed for this particular value of K , because earth curvature and atmospheric refraction effects have been taken into account in preparing the K -earth profile paper. Terrain effects by obstruction, reflection or diffraction can then be treated in a more straightforward manner.

The expression for K as a function of atmospheric refraction conditions in a particular area can be derived with the aid of figure 2.2 where:

- AB = Section of earth surface of distance d .
- AC = Circular path traversed by a microwave ray starting horizontally from point A.
- R = Radius of the earth centred at O.
- R' = Radius of curvature of the path centred at O_1 .
- h = Distance between the point at the end of the ray path and the ground point directly below.
- A'C' = Transformed ray path into a straight line centred now at a point O'_1 located an infinite distance below the diagram.
- A'B' = Modified earth surface section of distance d centred now at a point O' and having a radius KR .

Point C' is located directly above B' on the ground.

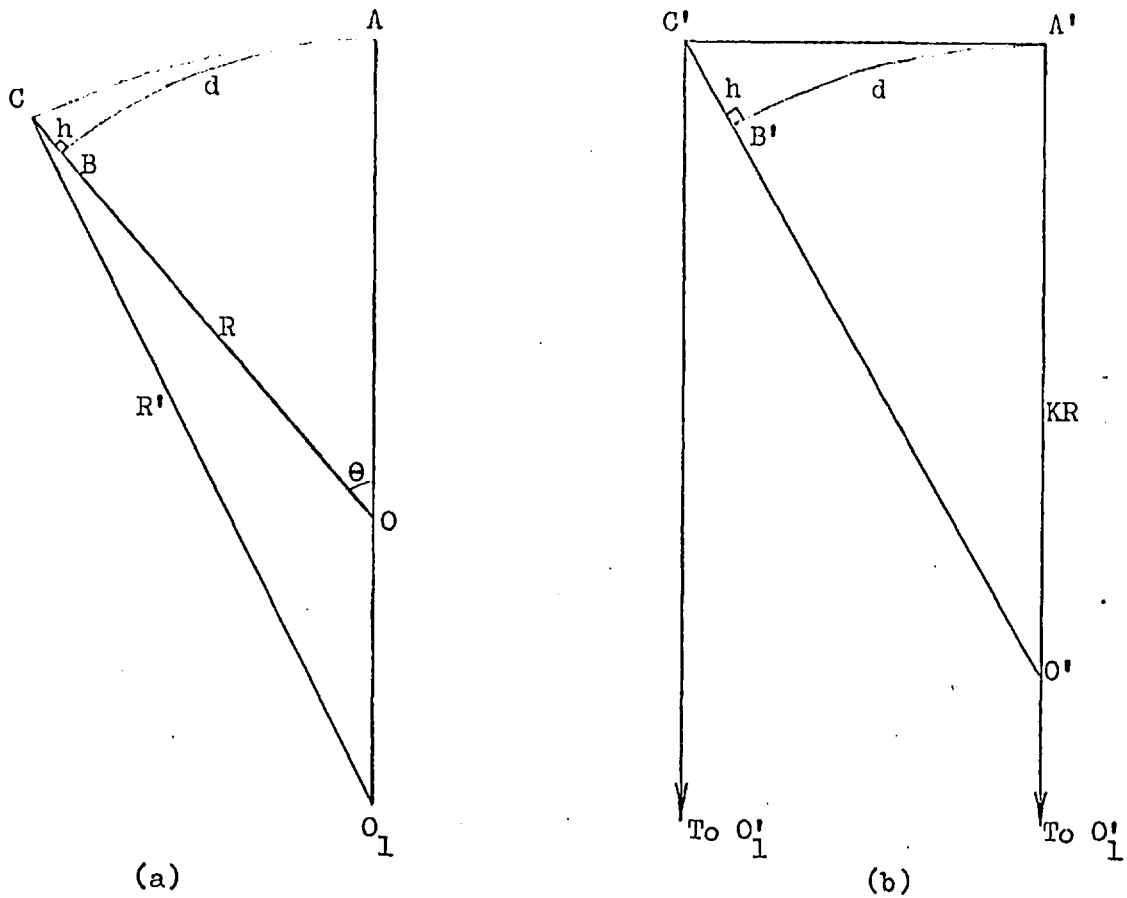


FIGURE 2.2

ATMOSPHERIC REFRACTION TREATMENT

The ratio factor K between the actual earth radius and the modified earth radius must be given a value so that the distance h has the same relationship to the distance d in both diagrams. From triangle OO_1C using the cosines rule:

$$(R + h)^2 = R'^2 - (R' - R)^2 - 2(R + h)(R' - R)\cos \theta \quad (2.14)$$

Replacing θ by its equivalent d/R and $\cos \theta$ by a series expansion, we have:

$$\cos \theta = 1 - \frac{d^2}{2R^2} \quad (2.15)$$

obtained by terminating the series after the first two terms, since it is assumed that $d \ll R$.

Upon substituting equation 2.15 into equation 2.14 we obtain:

$$R^2 h^2 + 2R^2 R' h - R' h d^2 - R' R d^2 + R h d^2 + R^2 d^2 \quad (2.16)$$

Using the conditions $h \ll R'$ and $D \ll R'$ we can simplify equation 2.16 with negligible error to:

$$2R^2 R' h - R' R d^2 + R^2 d^2 = 0 \quad (2.17)$$

Solving for h yields:

$$h = \frac{(R' - R)d^2}{2R'R} \quad (2.18)$$

Replacing R' by KR and allowing R' to become infinite we obtain:

$$h = \frac{d^2}{2KR} \quad (2.19)$$

Equivalence between figure 2.2a and figure 2.2b requires a value of K such that both equation 2.18 and equation 2.19 will yield the same value for h corresponding to any given value of d . This value of K is given by:

$$K = \frac{R'}{R' - R} \quad (2.20)$$

where R' is dependent upon the atmospheric refraction conditions in a particular area, as well as the refractive index of equation 2.13.

Another useful relation is:

$$K = \frac{10^6}{R(dM/dh)} = \frac{R'}{R' - R} \quad (2.21)$$

where M is the refractive modulus,⁽⁵⁾ relating the curvature of a microwave beam to that of the earth and given by:

$$M = (n - 1 + \frac{h}{R})10^6 \quad (2.22)$$

Its value is usually about 300.

The curve representing the ratio:

$$\frac{dM}{dh} = \frac{A}{T} \frac{dp}{dh} - \frac{A}{T^2} (p + \frac{2Be}{T}) \frac{dT}{dh} + \frac{AB}{T^2} \frac{de}{dh} + \frac{10^6}{R} \quad (2.23)$$

is called the M-profile and is widely used in radio meteorological observations. In equation 2.23 A and B are constants, T is the absolute temperature, p is the atmospheric pressure in millibars and e is the water-vapour pressure in millibars.

As mentioned before, it has become customary to adopt $K = 4/3$ as a standard working value because it has been found, in the region of middle latitudes, a strong tendency for K to lie in the neighborhood of 1.3 or 1.4. This led to what is known as the $4/3$ -earth atmosphere, in which the condition $K = 4/3$ is satisfied exactly over a range of elevations extending from the ground to about one kilometre.

A particular case is that of a well-mixed atmosphere indicating an air mass which has been thoroughly mixed by convection, eddy turbulence and molecular diffusion. Such an atmosphere has a value of $K = 6/5$. The condition of $K = 1$ corresponds to a homogeneous atmosphere.

2.3.3.1. Atmospheric Absorption

Although atmospheric absorption effects need only be considered at frequencies above a few GHz⁽⁶⁾, they are briefly studied in the following paragraphs.

The absorption of microwaves in an atmosphere without water drops is mainly caused by oxygen molecules through magnetic interaction and water vapour interacting with the radiation field electrically. Absorption by

both of these gases has a great significance at millimetric wavelengths because oxygen molecules have a resonance line at $\lambda = 0.25$ cm and a resonance band near $\lambda = 0.5$ cm, while water vapour affects wavelengths around 0.16 cm and to a smaller extent wavelengths at 1.25 cm. Together their absorption becomes very small for frequencies below 8 GHz, where an average attenuation of less than 0.01 dB/km may be considered normal, giving a loss of less than 0.5 dB for a 50 km path.

Raindrops have a more severe influence upon practical radio circuits using frequencies above 5 GHz. This influence depends on rain density and is basically due to scattering and absorption by the water drops. Moderate rainfall causes absorption - at even 15 GHz - of no more than 0.1 dB/km. Heavy rainfall of 15 mm/h is noticeable at 7 GHz producing about 10 dB loss on a 50 km -link if the area rain extends that far. Torrential rain affects even 4 GHz and could interrupt an 8 GHz link. In regions where heavier rains prevail, distances above 50 km should be avoided for frequencies above 5 GHz.

2.3.4. Reflection and the Fresnel-zone Concept

Free-space loss, dealt with in section 2.3.2., gives only an approximate value for the actual loss encountered for microwave propagation above the earth. Besides atmospheric refraction and absorption, mentioned in previous section, the loss can be markedly affected by reflections at the terrain as well as by ray diffraction. Reflections cause the major disturbance if the line-of-sight between the antennas is not obstructed by obstacles.

The transmitted energy reaches the receiver R directly along a straight line and indirectly by reflection on the ground.

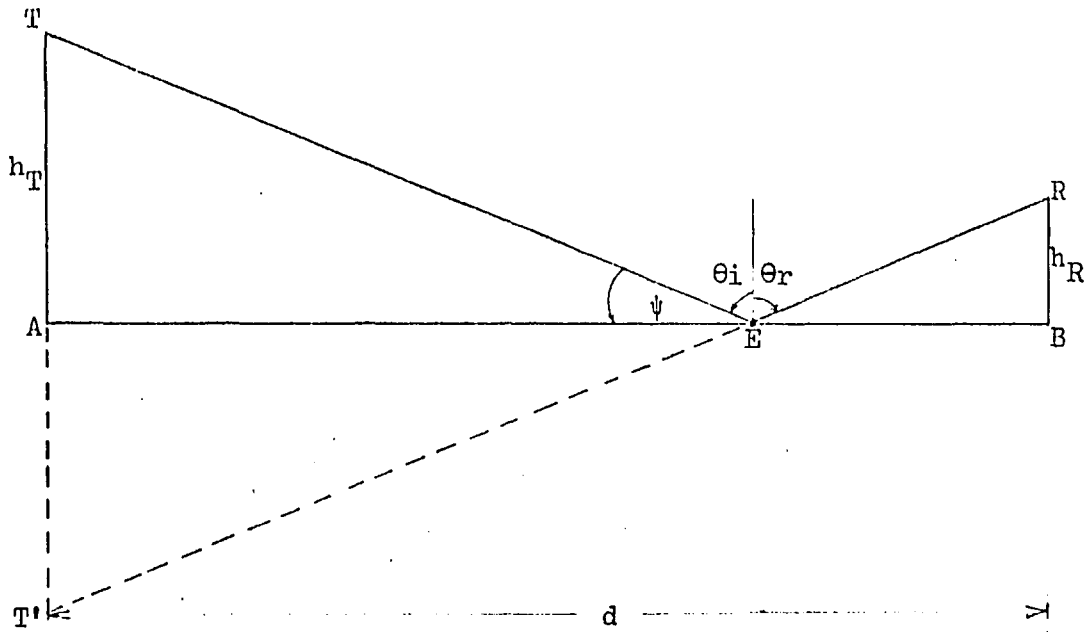


FIGURE 2.3

GROUND REFLECTION GEOMETRY

The path difference between direct and reflected waves is given by:

$$\delta = \sqrt{d^2 + (h_T + h_R)^2} - \sqrt{d^2 + (h_T - h_R)^2} \quad (2.24)$$

Considering $h_T, h_R \ll d$, since T and R must be close to the reflecting plane if appreciable power is to flow along the reflection path assuming a comparatively narrow beam, equation 2.24 becomes:

$$\delta = \frac{2h_T h_R}{d} \quad (2.25)$$

and the angle ψ becomes very small so that the reflection coefficient of the ground is - 1 for both horizontally and vertically polarized waves.

An approximate solution - adequate for engineering purposes - for propagation over a spherical earth, is to regard reflection as occurring at a plane tangential to the reflecting point and the reflected wave

to be divergent. The effective antenna heights ($h_{T_{eff}}$, $h_{R_{eff}}$) are now smaller compared to the actual heights (h_T, h_R), in accordance with their distances (d_T, d_R) from the reflecting point, and are given by:

$$h_{T_{eff}} \approx h_T - \frac{d_T^2}{2KR}$$

$$h_{R_{eff}} \approx h_R - \frac{d_R^2}{2KR}$$
(2.26)

where K and R are as described in section 2.3.3.

For the relatively short hops employed in ground-to-ground links the effect of the curvature of the earth is small compared with that of ground irregularities.

With a path difference δ equal to half a wavelength, the reflected ray arrives at the receiver with the same phase as the direct ray since its phase delay is composed of the 180° phase shift at the reflecting point plus 180° phase shift caused by the difference in path length. In such a case, direct and indirect rays are added. With a full wavelength path difference and a resulting phase shift of $360^\circ + 180^\circ$ the two rays tend to cancel each other. In general with a path difference of $(2n + 1)\frac{\lambda}{2}$ ($n = \text{integer}$) direct and indirect waves are added, while they cancel each other if the difference is $(2n + 2)\frac{\lambda}{2}$.

Constant-path length rays are obtained by reflection at points described by an ellipse having as its major axis the line between the antennas and its minor axis given approximately by:

$$b = \sqrt{\frac{(2m-1)\lambda d}{4}}$$
(2.27)

where m is an odd integer, λ is the wavelength and d is the distance between the antennas. When these ellipses are revolved around their

major axis d , they describe ellipsoids and their intersection by a plane perpendicular to the axis d , reveals a family of concentric circles with their centre on the axis d . Their radii in metres are approximately⁽⁷⁾:

$$F_{\eta} \approx 548 \sqrt{\frac{\eta d_T d_R}{(d_T + d_R) f}} \quad (2.28)$$

where d_T , d_R are in kilometres, f in MHz and η has the value 1 for the innermost circle, 2 for the second circle from the centre and so on.

The approximation in equation 2.28 introduces negligible error as long as F_{η} remains small compared to both d_T and d_R , a situation generally encountered in practice.

Fresnel-zones are defined through equation 2.28. The area enclosed by the F_1 circle is called the first Fresnel-zone, that between the F_1 and F_2 circles the second Fresnel-zone and so on. From equation 2.28 is also deduced that all Fresnel-zones in a given plane have the same area.

For a given terrain, the order of the zone reaching the ground, and consequently the path difference between the direct and the reflected ray, is a function of the antenna height or wavelength or distance between antennas. With perfectly reflecting ground and equal power of both rays, full cancellation of the received signal can occur if one of the above factors or their simultaneous result is unfavourable. A link subject to strong power fluctuations due to ground reflections, would be unacceptable for a radio relay system, because reflection points on the ground are not stable and their location as well as their reflection coefficients can change with the season, weather conditions, vegetation etc.. Strong ground reflections are produced from large bodies of water, deserts, wide plains or larger agricultural regions with uniform cultivation.

Diffuse reflection is exhibited by rough surfaces and in this case laws for specular reflection are no longer applicable. Huygens' principle must be used to examine the reflection mechanism in this case.

In order to estimate the roughness of the terrain the Rayleigh criterion can be used, according to which a given plane is smooth for a specific wavelength (λ) if the height of the obstacle is:

$$h \leq \frac{\lambda}{16\psi} \quad (2.29)$$

where ψ is the angle, as in figure 2.3, in radians.

Rough or not significantly reflecting can be considered a surface for which

$$h \geq \frac{\lambda}{8\psi} \quad (2.30)$$

For $\psi = 0.5^\circ = 8.72 \times 10^{-3}$ rad - a figure occurring in practice - h has to be at least 15λ , which is about 5 metres for a frequency of 900 MHz.

2.3.5. Obstacles, Diffraction and Fresnel-zone Clearance

For links of only local importance, mostly operating at frequencies below 1 GHz, a path with some obstructions might be acceptable for economic reasons.

Light wooden structures or woods obstructing a small part of the path are obstacles of smaller importance for frequencies below 1 GHz, but large buildings, hills and mountains must be considered as high-loss obstacles throughout the frequency range.

In accordance with Huygens' principle, the field at a receiving antenna can be regarded as arising from the joint action of a system of Huygens' sources⁽⁸⁾ distributed over a surface interposed between

this antenna and the transmitting one. If the passage of radiation through any portion of this surface is obstructed, diffraction takes place and the resultant field at the receiving antenna may be seriously affected even when the obstruction does not actually cut the line-of-sight path. This effect becomes more pronounced the closer the obstruction approaches the line-of-sight and eventually becomes large attenuation - defined as shadow loss - when the obstacle obstructs the line-of-sight.

Diffraction loss described in detail in Chapter four, can only be predicted in a few cases. Such a case happens when the obstruction forms an edge very long compared to the wavelength and placed perpendicularly to the propagation direction. This is the case of single knife-edge diffraction the loss of which can be derived from the Fresnel-Kirchhoff theory and is additional to the free-space propagation loss. Figure 2.4 illustrates this in nomogram form from which one can immediately acquire the diffraction loss, for the interesting range of distance and frequencies, if the path profile is given. If there is more than one obstacle, as frequently happens, the loss can be approximately estimated by graphic interpolation, as in figure 2.4a. The loss is estimated as resulting from one obstacle of height H rather than two obstacles O_1, O_2 .

Other cases where diffraction loss can be evaluated easily are when it is possible to represent the terrain schematically by a smooth sphere or a cylinder. In these cases loss is further increased compared with that due to knife-edge diffraction. In practice the surface of the obstacle is rough for microwave frequencies resulting in lower values of loss compared with those for smooth obstacles.

In planning a line-of-sight microwave link, an optimum Fresnel-zone clearance is provided as a compromise between performance and cost. Although first Fresnel-zone (F_1) clearance is not regarded as optimum by all authors or organizations, it is widely accepted as such, while a clearance slightly exceeding $0.5 F_1$ has often to be accepted, if tower

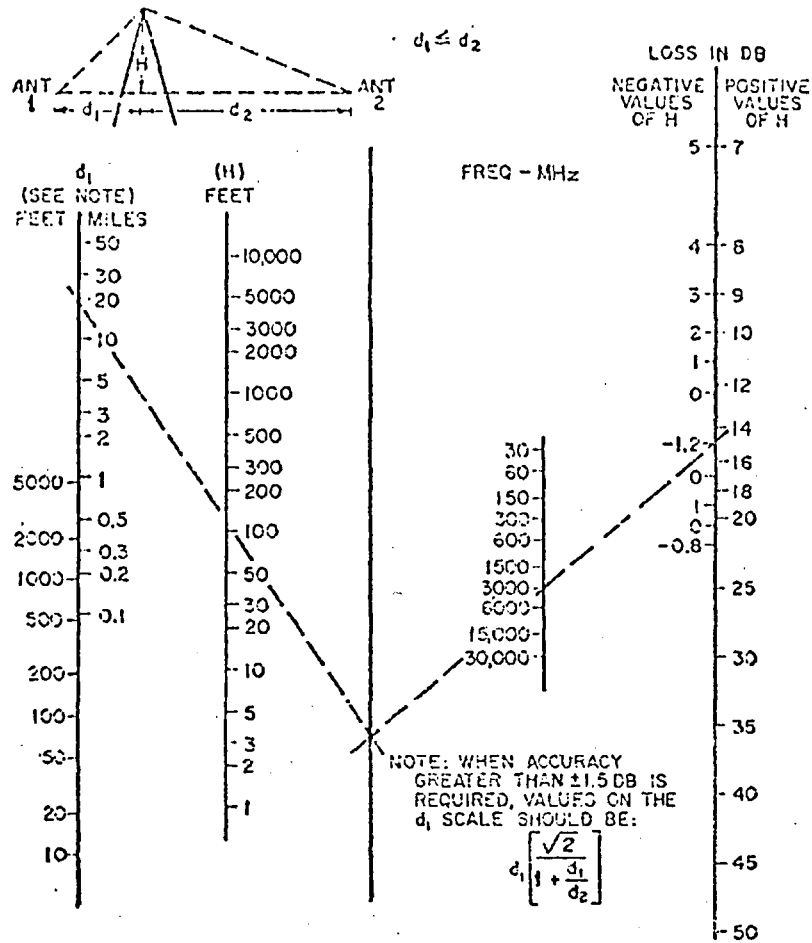


FIGURE 2.4

NOMOGRAM FOR ESTIMATING DIFFRACTION LOSS DUE TO
OBSTACLES IN THE PROPAGATION PATH

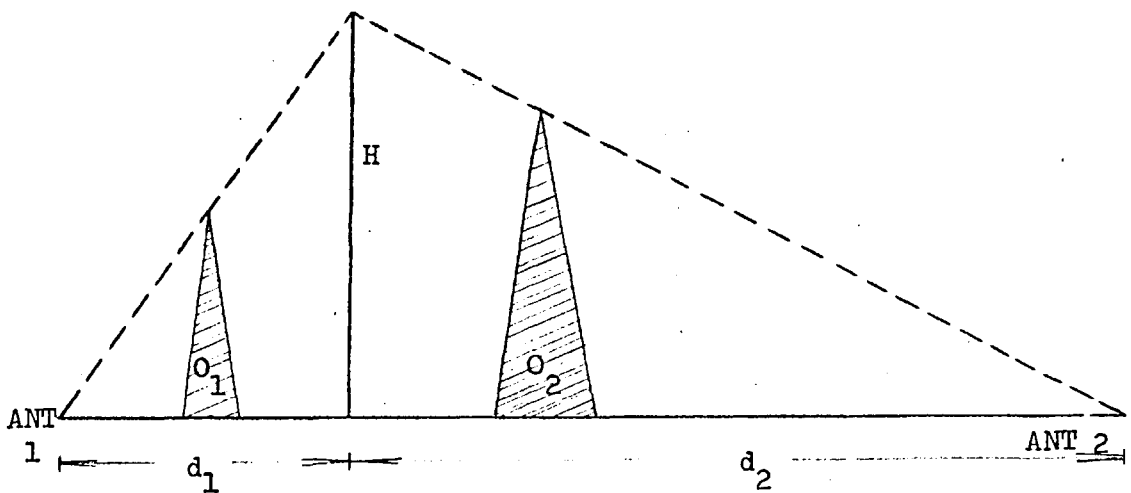


FIGURE 2.4a

DIFFRACTION BY MORE THAN ONE OBSTACLE

installation cost is to be kept low. In such a case atmospheric deflection of the propagation path, even on rare occasions, must be taken into account so that the clearance never becomes smaller than $0.5 F_1$; if not, diffraction loss may become large and the signal level at the receiving antenna may fall below the recommended standards.

2.3.6. Fading and Diversity Systems

If the atmosphere were well-mixed⁽⁷⁾ along the transmission path and both antennas were properly located so as to ensure adequate Fresnel-zone clearance over the intervening terrain and avoidance of any reflections, then only the wave following the line-of-sight path would reach the receiving antenna. The induced signal would be a faithful reproduction of that driving the transmitting antenna. The only loss encountered in this ideal case is the free-space transmission loss given by equation 2.7.

Although the proper antenna location is an objective which can be achieved, the atmosphere is not usually well-mixed throughout the transmission path. As a result, various types of disturbance to propagation may be encountered causing a loss in received signal, which is said to fade and the loss is said to occur due to fading.

Fading may be caused by:

- a. absorption and scattering by fog, rain and other precipitation.
- b. a variable refractive index profile, resulting in reduction of Fresnel-zone clearance.
- c. interference between two or more signals arriving at the receiving antenna by different paths, of which at least one is time dependent.
- d. deflection of radiation by a duct resulting from irregular variation of the refractive index of the atmosphere.

There is no commonly accepted nomenclature in use for different types of fading phenomena. We may distinguish between attenuation fading and interference fading or slow and fast fading correspondingly. Attenuation or slow fading is generally caused by (a) and (b) above, while interference or fast fading happens due to (c) and (d) above.

Deflection of transmitted radiation may result in a received signal too small to be usable. This may happen if the effective-earth-radius factor K becomes temporarily smaller than its standard value (due to an unusually small vertical refractivity gradient along the propagation path) giving rise to a microwave path of smaller curvature than normal. As a result the Fresnel-zone clearance over the intervening terrain may become less than enough to permit line-of-sight communication. This type of fading develops and subsides slowly, because the refractive gradient of the atmosphere cannot change rapidly. This is why it is sometimes referred to as slow fading. Microwave paths subject to this kind of fading are those passing relatively close to the ground in a region where the ground is intensely heated by the sun. More Fresnel-zone clearance than is needed for free-space transmission under normal atmospheric conditions is usually an effective cure for slow fading. Reduction of the clearance to a level where fading becomes appreciable is then unlikely to occur. Heights for the transmitting and receiving antennas are no longer determined on the basis of a $4/3$ -earth radius, but on the lowest effective earth radius known to occur with appreciable frequency along the given path.

Fast or interference fading may happen when the received signal is not a single ray, but a number of distinct and separate rays, each arriving via a slightly different path. This type of fading is also referred to as multi-path fading, for obvious reasons. Turbulence and unwell-mixed atmosphere as well as ground reflections are responsible

for these different ray-paths, the lengths and shapes of which are continually changing by small amounts corresponding to many wavelengths at microwave frequencies.

The most effective technique in minimising fast fading effects is to keep the transmitted power high enough and the length of the hop short enough to maintain a usable level at received signal in the presence of the most severe fade likely to occur. In addition one or more diversity methods, described below, may be employed in the most severe cases.

For the simple case of a link in which a large part of the energy reaches the receiver after one ground reflection, the following techniques may be employed to exclude or mitigate the effect of reflection, when the ground profiles and other factors are favourable.

Figure 2.5 shows a link over a highly reflecting ground.

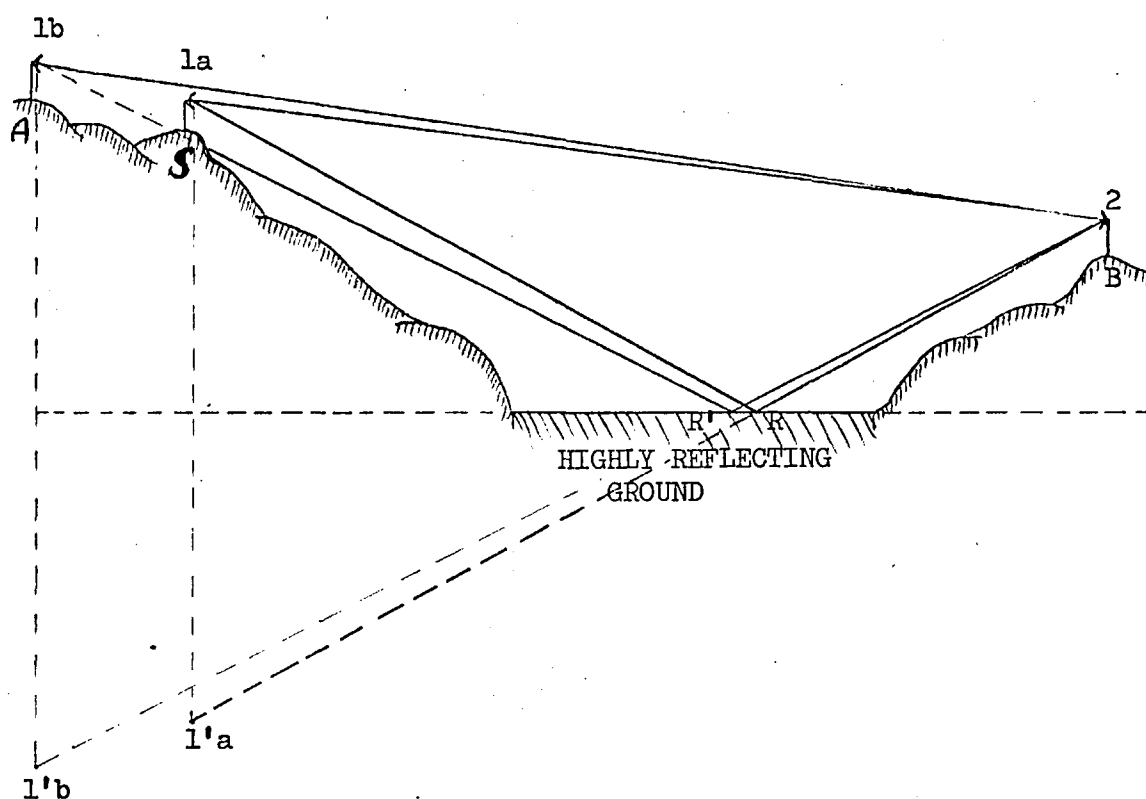


FIGURE 2.5

SCREENING THE REFLECTED PATH

Locating the antennas on sites S and B, specular reflection takes place causing a severe fading in the received signal. Removing one antenna from site S to site A, the reflected path is screened by hill S, which does not interfere with the line-of-sight. Fresnel-zone clearance above hill S must be provided in such cases; but even if it is not fully accomplished, diffraction loss that might occur will probably not exceed a few dB and is negligible compared with the fading reduction achieved.

Another technique is shown in figure 2.6(a) and is often called

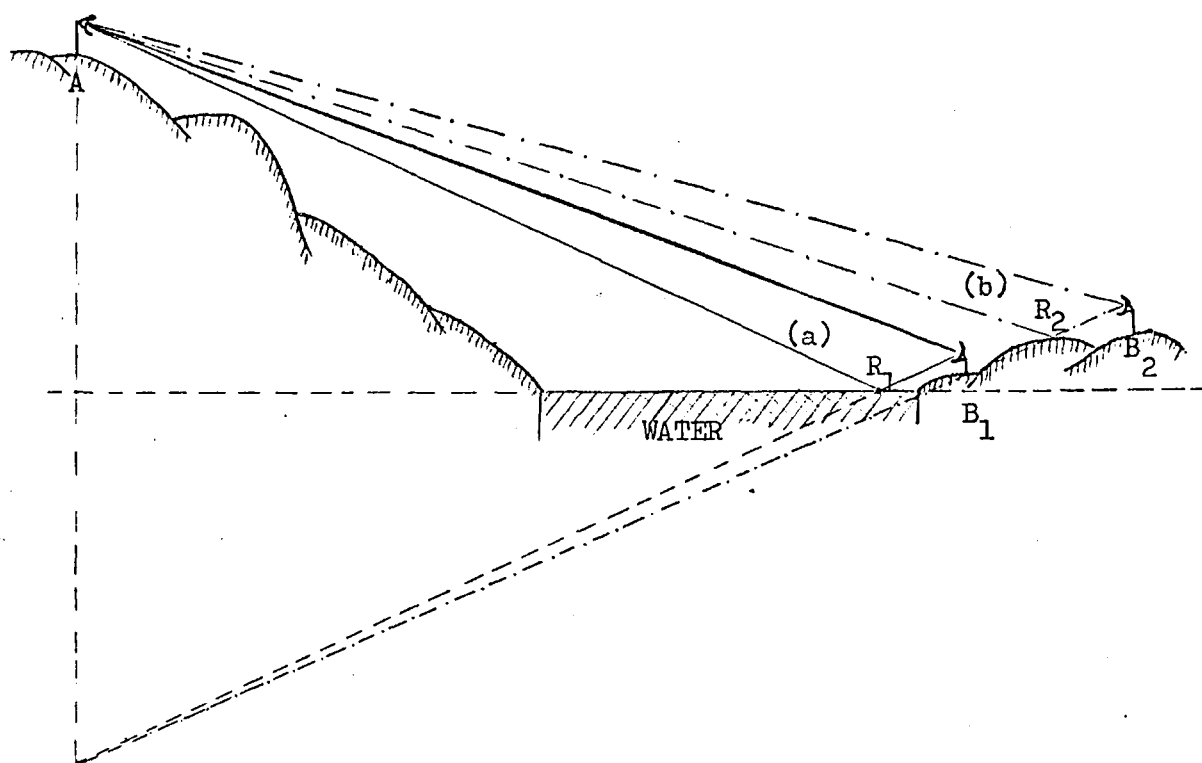


FIGURE 2.6

- a. HIGH-LOW LINK METHOD
- b. REFLECTION POINT OUTSIDE THE WATER

a high-low link technique. If the reflection is unavoidable it is important to make the path difference between direct and indirect rays as small as possible. This can be achieved by arranging the reflection

point to be as near as possible to one end of the path (R_1 in figure 2.6) by making the height difference of the antennas rather large. This might, however, cause some troubles because the ray has now to pass through various layers of the atmosphere. By locating one antenna, as at B_2 in figure 2.6, we can move the reflection point to a medium with bad reflection properties, such as the ground at point R_2 .

Slow and fast fading may occur simultaneously or separately, although there is a connection between slow fading due to reduction of Fresnel-zone clearance and fast fading due to ground reflection.

Fading intensity generally increases with the frequency and with the length of the path. The fading in the individual sections of a link comprising many repeater stations is only partially correlated and hence at a given time each section shows a different behaviour. Simultaneous occurrence of shallow fading is relatively frequent, whereas peak values do not usually occur simultaneously in practice. A margin between 5 and 10 dB in addition to the free space loss of all radio sections in the system is generally accepted for engineering purposes.

2.3.6.1. Diversity Systems

Fast or multi-path fading is frequency dependent, because with equal path difference the phase shift is frequency dependent. If the same frequency is used on two slightly different paths, phase shift results as with two frequencies on the same path. To reduce the effects of fast fading, two different frequencies for the same link can be employed or two properly spaced antennas for the same frequency. Combination of the signals after amplification or selection of the better one by automatic switching, results in reasonable signal variations depending on the devices used.

These methods, customarily employed for reducing the fading effects⁽⁹⁾,

are known as diversity techniques and are referred to as two or more independent methods provided for the signal to reach a given receiver from a given transmitting station.

If two different carrier frequencies are employed the technique is known as frequency diversity, whereas with two different antennas, properly spaced, we have space diversity.

Frequency diversity is usually regarded as the most attractive, whereas space diversity is the best known alternative and is usually preferable whenever fading is mainly due to interference between line-of-sight and ground reflected waves. Two different transmitters driving the same antenna and two different receivers using a single antenna are used for the same signal - which modulates two somewhat different frequencies - when frequency diversity is employed. Employing space diversity, two separate antennas are typically operated at the receiving site spaced either vertically or horizontally, although vertical separation is preferable for both economical and efficiency reasons, particularly in the presence of a strong ground reflected wave.

Polarization diversity is another method employed to avoid fading; this is not suitable for use in line-of-sight microwave links because, in practice, fading is not sensitive to the plane of polarization of the waves even in the case of ground reflections, since the angle of incidence is nearly 90° .

2.4. Passive Repeaters

2.4.1. General

When the line-of-sight between stations is obstructed and a direct link is impossible, a passive repeater⁽¹⁰⁾ can be employed for the purpose of redirecting the incident energy and so establishing a reasonable path. The use of a passive repeater introduces a considerable loss and incurs high costs if its performance is to be reliable. It

means that a passive repeater is only used when an alternative is neither possible nor satisfactory.

Three types of passive repeater are commonly used:

- a) Single-mirror system.
- b) Double-mirror system.
- c) Passive antennas system.

All of these types can be used when the passive repeater is in the distant (Fraunhofer) field of the transmitting and receiving antennas, whereas the method for the Fresnel or near-zone field is more appropriately thought as an antenna system or periscope antenna. Generally passive repeaters of the mirror type have dimensions of many wavelengths.

2.4.2. Single-Mirror System

This method, shown in figure 2.7, is useful when the transfer angle ϕ_t is greater than about 60° , because for smaller angles the length of the mirror becomes excessively large for a given projected area. For 100% efficiency the reflector surface must be flat to within $\pm \lambda/16$, so that the reflection is specular ($\theta_i = \theta_r$) and all the incident power,

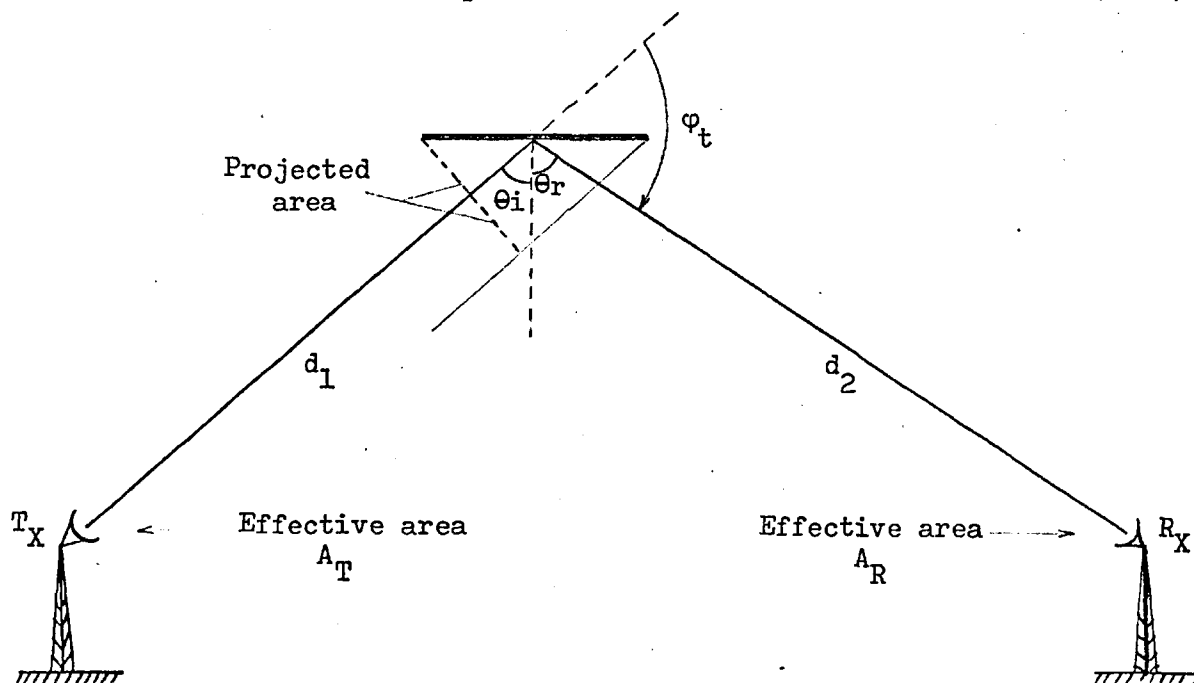


FIGURE 2.7

SINGLE-MIRROR SYSTEM

except the negligible heat loss, is reflected. In this case the radiation patterns are those of a uniformly illuminated aperture.

2.4.3. Double-Mirror System

When the transfer angle is so small (less than about 60°) that the single-mirror system becomes uneconomical the double-mirror system, as shown in figure 2.8, should be used.

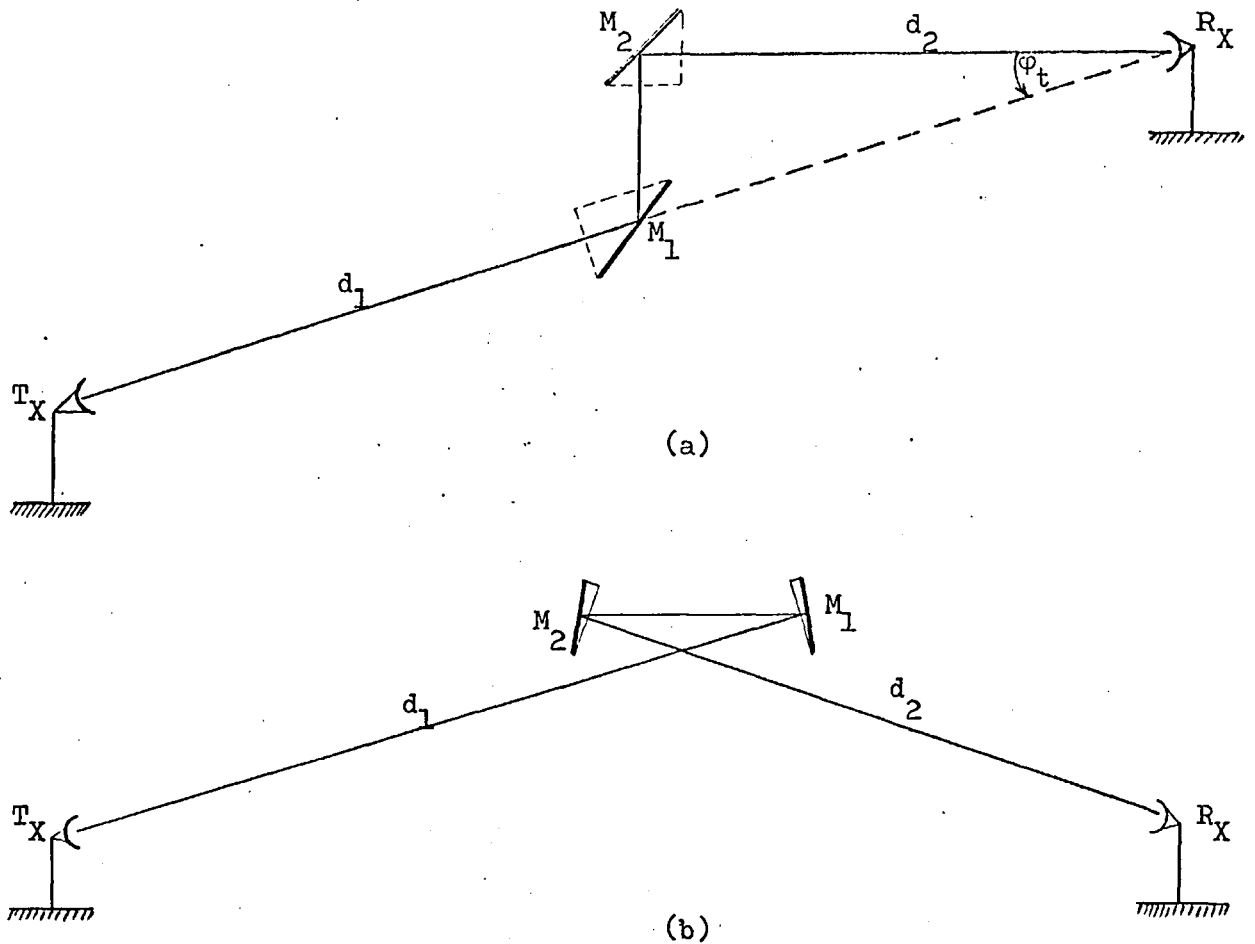


FIGURE 2.8

DOUBLE-MIRROR SYSTEMS

The difference between these two types of twin reflectors are very small.

In this case the loss between the reflectors will be small provided that their separation does not exceed the Rayleigh range and the energy between them travels in perfectly parallel rays. On the other hand,

their sizes are so related that the projected areas for both the incident and the final exit rays are the same.

2.4.4. Loss Introduced by a Passive Repeater

The loss or attenuation affecting the propagation between parabolic antennas with one interposed passive repeater of an effective area A_{RP} is given by:

$$L_1 = 10 \log \frac{d_1^2 d_2^2 \lambda^4}{A_{RP}^2 A_T A_R} \quad [\text{dB}] \quad (2.31)$$

where d_1, d_2 are the lengths of the two parts of the path, A_T and A_R are the effective areas of the transmitting and receiving antennas and the effective area A_{RP} of the passive repeater is related to the geometrical area by:

$$A_{RP} = A_{\text{geom}} \sin \frac{\phi_t}{2} \quad (2.32)$$

The loss that would result if free propagation occurred directly between the two antennas at a distance $d_1 + d_2$ from each other is given by:

$$L' = 10 \log \frac{(d_1 + d_2)^2 \lambda^2}{A_T A_R} \quad [\text{dB}] \quad (2.33)$$

Therefore, the loss introduced by the repeater can be calculated as:

$$L_{RP} = L_1 - L' = 10 \log \frac{d_1^2 d_2^2 \lambda^4}{A_{RP}^2 A_T A_R} \frac{A_T A_R}{(d_1 + d_2)^2 \lambda^2} \quad [\text{dB}] \quad (2.34)$$

or

$$L_{RP} = 20 \log \frac{d_1 d_2 \lambda}{A_{RP}(d_1 + d_2)} \quad [\text{dB}] \quad (2.34)$$

where λ , d_1 , d_2 are in metres.

Converting the distances d_1 , d_2 into kilometres we have:

$$L_{RP} = 20 \log \frac{\lambda}{A_{RP}} \frac{d_1 d_2}{d_1 + d_2} 10^{-3} \quad [\text{dB}] \quad (2.35)$$

In practice it is useful to employ a reflector only if $d_1 \ll d_2$ or vice versa. For $d_1 = 50$ km, d_2 should normally not exceed 1 km, if the loss or the size of the reflectors are to be kept within reasonable limits.

2.4.5. Passive Antennas System

Instead of a mirror system two series-connected antennas may be used as shown in figure 2.9. The path loss is then the sum of the loss

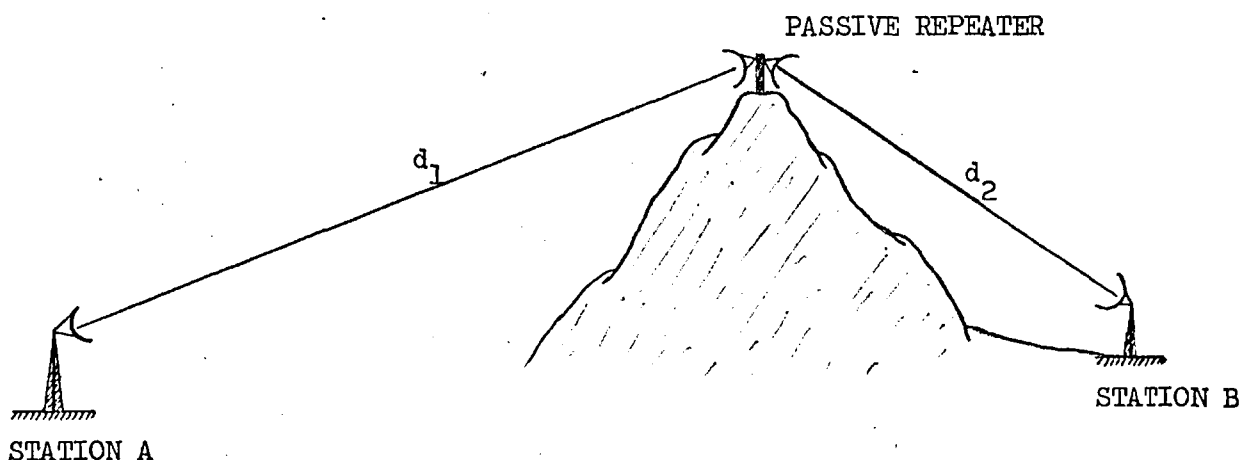


FIGURE 2.9

PASSIVE ANTENNAS SYSTEM

in dB over the two paths d_1 , d_2 plus the loss due to the connection

between the passive antennas.

2.5. C.C.I.R. Recommendations

2.5.1. Hypothetical Reference Circuits

During the X Plenary Assembly⁽¹¹⁾ (Geneva 1963 - Recommendations N.391/392) the hypothetical circuits given in figures 2.10(a) and 2.10(b) have been recommended by the C.C.I.R., for small and large capacity radio relay-systems respectively, which use frequency division multiplexing.

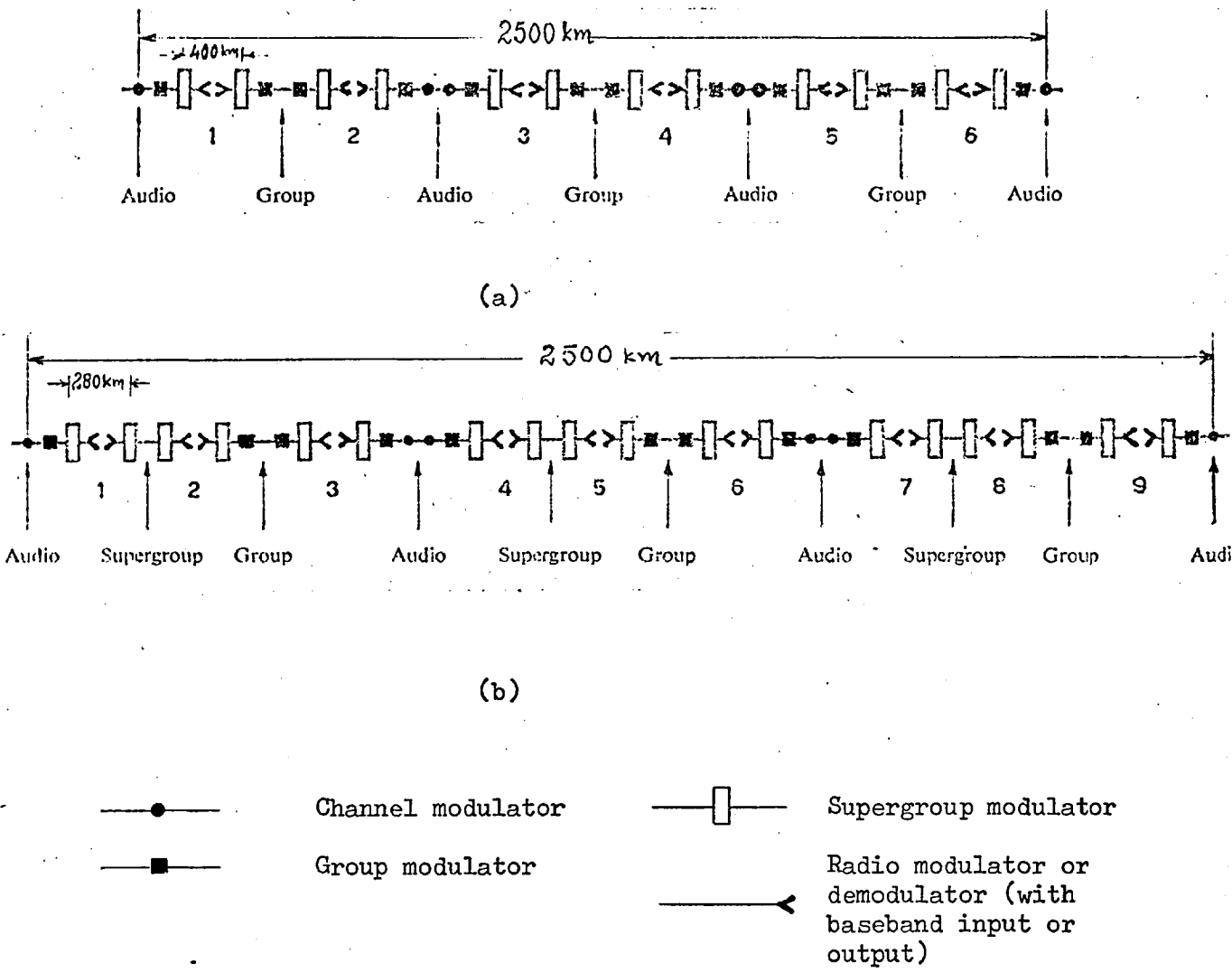


FIGURE 2.10

HYPOTHETICAL REFERENCE CIRCUIT FOR RADIO-RELAY
SYSTEMS USING FREQUENCY-DIVISION MULTIPLEX

According to the recommendations, these circuits should include six or nine sets of radio modulators and demodulators for each direction of transmission, which should divide the circuits into six or nine homogeneous sections of equal length.

2.5.2. Noise Specifications

At the above Plenary Assembly (Geneva 1963 - Recommendations N.393/395) the C.C.I.R. has recommended that, at the end of the 2500 km hypothetical reference circuit for frequency division multiplex radio-relay systems, the noise power at a point of zero relative level in any telephone channel should not exceed the values given below:

1. 7500 pW psophometrically weighted mean power in any hour.
2. 7500 pW psophometrically weighted one minute mean power for more than 20% of any month.
3. 47500 pW psophometrically weighted one minute mean power for more than 0.1% of any month.
4. 1000000 pW unweighted (with an integrating time of 5 ms) for more than 0.01% of any month.

In order to evaluate the noise contribution due to only a portion of the hypothetical reference circuit, it has been established that the noise power, as per points (1) and (2) above and the percentage of time as per points (3) and (4) will be regarded as proportional to the number of homogeneous sections involved.

In the case of real circuits of length $D(\text{km})$ between 280 and 2500 km, whose composition does not differ appreciably from the hypothetical one, the specifications are as follows:

1. 3D pW mean power in any hour.
2. 3D pW one-minute-mean power for 20% of a month.
3. 47,500 pW one-minute-mean power for $\frac{D}{2500} \times 0.1\%$ of a month.

Finally, as a planning objective for real circuits, whose length D

is greater than 50 km but less than 280 km, the weighted noise power should not exceed:

1. $50\sqrt{D}$ pW mean power in any hour.
2. $50\sqrt{D}$ pW one-minute-mean power for more than 20% of any month.
3. 47,500 pW one-minute-mean power for more than $\frac{280}{2500} \times 0.1\%$ of any month.

Moreover, it is noted in the recommendations that:

a. the noise due to frequency-division multiplex equipment is not included. The C.C.I.T.T. allows 2500 pW mean value for this noise in any hour on a 2500 km hypothetical circuit.

b. at the demodulation points of the real circuit (or at the junctions between the homogeneous sections of a hypothetical circuit) the telephone channels, groups, supergroups and master groups are interconnected at random; whereas the noise coming from the different sections is power-additive.

c. during the busy hour, the multiplex signal can be represented by a uniform-spectrum signal, the mean power absolute level of which, at a zero relative point, is $(-15 + 10 \log_{10} N)$ dBm for 240 channels or more, and $(-1 + 4 \log_{10} N)^{10}$ dBm for numbers of channels between 12 and 240; N is the total number of channels of the system.

2.5.3. Comments on the C.C.I.R. Noise Specifications

The allowance of 7500 pW mean noise power in any hour, together with the 2500 pW allowed for the terminal equipment, corresponds to a signal-to-noise ratio of 50 dB for a 2500 km circuit. For circuits shorter than 2500 km, but longer than 280 km, an analogous signal-to-noise ratio has to be maintained, since the permitted noise power for both the transmission medium and the terminal equipment has to be reduced in proportion to the circuit length.

Considering radio-relay allowance only (7500 pW) the signal-to-noise

ratio is increased to 51.25 dB and this is the figure (or in proportion for shorter circuits) to be achieved when the terminal equipment is not taken into account. In real circuits this figure is obtained from the permissible noise of 3 pW/km , when the circuit length lies between 280 and 2500 km.

However, for radio-relay circuits shorter than 280 km a different recommendation was made. The admissible noise power was increased to $50\sqrt{D}$, where D is the circuit length in km, in order to make allowance for the fact that even with virtually zero circuit length some basic noise remains unchanged from the terminal equipment as a fixed contribution. The extreme variability of a short-length circuit is another reason why the C.C.I.R. recommends different noise allowances, providing some relaxation for systems of only a few sections. Statistically it would have been better to have specified the average of the noise exceeded for several hours.

As it is technically and economically impossible to have a circuit usable for a hundred percent of the scheduled operating time without interruption or deterioration of performance, some relaxation is permitted by the C.C.I.R. recommendations. Provided the hourly mean noise power is maintained, the average noise power of 3 pW/km may be exceeded minutewise for 20% of any month. Furthermore, the one-minute mean noise power may exceed 47,500 pW for a period of less than $D/2500 \times 0.1\%$ of any month, where D is the circuit length in km. This, for a length of 2500 km, gives a signal-to-noise ratio equal to 43.25 dB, so permitting an 8 dB increase in the noise level or an 8 dB reduction in the signal level - compared with the hourly mean noise - for not more than 0.1% of the time.

Complete loss of signals due to fade outs, circuit interruptions and equipment outages are not considered in the recommendations. Proper

standby switching arrangements, with a transfer-switching time depending on the requirements, can be used to minimize interruptions and equipment outages.

The terms "worst hour" and "worst month" may usually be used when estimating noise power, since clearly "any hour" and "any month" must include the worst ones. The hours when the noise is greatest are usually those in which the fading is most severe. These hours will sometimes be different from the busy hours.

The restriction "any telephone channel" indicates the possibility of noise variation in different channels of the same system. The top channel is usually more noisy due to the frequency dependence of most noise sources.

For local circuits having a restricted length, some deviation from the recommendations may be permissible if, for economic and geographical reasons, the international standards of performance cannot be met without undue expenditure. In such cases, less than the recommended signal-to-noise ratio between terminals may be the objective, but with the growing interconnection of networks, performance exceptions become more and more unacceptable.

2.5.4. Some Comments on Military Systems

Radio-relay systems serving only military purposes have sometimes to comply with more severe requirements than the commercial ones although the opposite is more frequent in practice, since military communications rarely depend on only one means. The deciding factors are the nature of the operations relying on the circuit, the secrecy demands and the availability as well as the reliability of alternative methods.

Due to the uncertainty of 100% reliability of any method of communication, particularly in wartime, almost no military operation

relies on a single means. Wire communications as well as messengers and optical means, whenever possible, are used in parallel with radio communications to ensure the delivery of a message which is usually very important. This in connection with reasons of secrecy, which demand low transmitter power even at the cost of signal quality, allows a reduction of design objectives when planning military radio-relay systems. The minimum objective of 50 dB signal-to-noise ratio between terminals, as mentioned in section 2.5.3., might be reduced to 40 dB for some military applications. Other compromises such as circuit interruptions, equipment failures, standby switching arrangements, transfer switching time etc., may also be more tolerable than in commercial systems.

Contrary to this, if the importance of the operations dependent upon the system is great and the communications heavily rely on this, then the requirements may become more severe than for commercial systems. Shorter repeater sections are used in these cases to obtain the minimum objective of 50 dB signal-to-noise ratio, while standby switching arrangements, with very short transfer switching time are always available.

It is usual to use existing commercial systems for military purposes during wartime, while the opposite is rather exceptional in peacetime. In the latter case the military system has to be planned in such a way as to comply with the commercial system requirements.

In order to keep down the transmitter power, particularly in those sections adjacent to the enemy, the transmission loss is kept down by shortening the section while the antenna gain at the receiving side (feeder loss included) is kept as high as possible. Also, during site selection, every effort is made to reduce the likelihood of the enemy picking up transmissions.

2.6. Signal Level Factors

2.6.1. General

Besides stability and reliability, the transmission performance is the main characteristic of a radio-relay system. This is essentially dependent on thermal noise and distortion, both being closely related to the characteristics of the transmission system and of the equipment employed.

To a first approximation thermal noise and distortion are independent of one another and their contributions can be treated separately. Their magnitude and influence can be calculated within quite close limits of accuracy from the system and equipment characteristics.

2.6.2. Signal-to-Thermal Noise Ratio

If a simple link is assumed, consisting of one radio relay section only, the output signal-to-thermal noise ratio (R_o) is given by the formula:

$$R_o = P - L_{FT} + G_T - L + G_R - L_{FR} - C - B - Z + V + W$$

(2.36)

where the terms, all expressed in dB or dBm, indicate:

- a. P the transmitter power, ordinarily given in watts (P_T), converted⁽¹²⁾ into dBm by means of the conversion scale below:

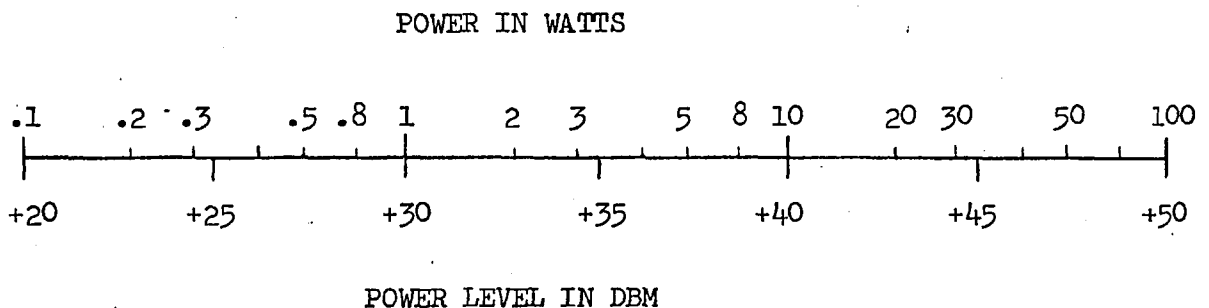


FIGURE 2.11

POWER CONVERSION SCALE

b. G_T and G_R are the Transmitting and Receiving antenna gains correspondingly, given by their geometrical area and the wavelength through the formula in equation 2.9. It is common practice to make G_T and G_R the same ($G_T = G_R = G$).

c. L_{FT} and L_{FR} are the Transmitting and Receiving antenna Feeder losses respectively, calculated from the length and the characteristic attenuation of the selected feeder type at the frequency concerned. Information is given in the manufacturer's literature. Coaxial cables are suitable as antenna feeders for frequencies up to 3 GHz and can be used up to about a length of 100 metres, as long as the loss does not become excessive for the particular case. The loss of a modern coaxial cable, with the branching filter loss included, amounts to about 4.5 dB/100 m at frequencies below 1 GHz, whereas it reaches approximately 7 dB/100 m at 2 GHz.

d. L is the propagation loss for the particular distance and frequency, determined according to equation 2.5. Shadow loss, in the case of grazing or a slightly obstructed path, determined according to section 2.3.5., is included in this term. The fading margin, as required by the system specifications, may be included in this factor; but in a multi-section link it is usually considered separately, as it is in this work.

e. The last five terms, C , B , Z , V and W determine the so-called thermal receiver noise described in detail in the following section.

2.6.3. Thermal Noise on the Receiving Side

The starting point for the evaluation of the thermal noise on the receiving side, is the constant C corresponding to the input noise power⁽¹³⁾ for room temperature ($T_0 = 293^{\circ}\text{K}$) and bandwidth $B = 1$ Hz. This is given by:

$$C = kT_0 = -174 \text{ dBm} \quad (2.37)$$

where k is Boltzman's constant ($= 1.38 \times 10^{-23} \text{ J/}^\circ\text{K}$). For a bandwidth B different from 1 Hz, the input power is:

$$N_R = kT_o B$$

or expressed in dBm

$$N_R[\text{dBm}] = -198.6 + 10 \log B + 10 \log T_o \quad (2.38)$$

The bandwidth $B = 3.1 \text{ KHz}$ of the signal channel (or 5 MHz for television) is usually taken as the basis for noise power calculations. This corresponds to an increase of the noise level of 35 dBm, thus giving an N_R equal to -139 dBm.

The noise characteristic of a receiver is expressed by multiples of KT_o . The value of the total noise power developed at a matched receiver input, assuming a bandwidth B , is given by:

$$N_{R_t} = zkT_o B$$

or expressed in dBm

$$N_{R_t}[\text{dBm}] = -198.6 + 10 \log B + 10 \log T_o + 10 \log z \quad (2.39)$$

for $B = 3.1 \text{ KHz}$ and $T_o = 293^\circ\text{K}$, $N_{R_t} = -139 + 10 \log z = -139 + Z$.

The quantity $Z = 10 \log z$ is called the receiver noise figure⁽¹⁴⁾, given by the manufacturer, and indicates the amount of deterioration of signal-to-noise ratio which occurs when signal and noise pass through the equipment. The product zT_o gives the effective noise temperature (T_{eff}) of the receiver.

The signal-to-noise ratio is modified by the factors V and W . The former one, called "the improvement factor", expresses the influence of the modulation system used and is treated in section 2.6.3.1.. The latter,

called "the weighting factor" or "the psophometric constant", corrects for the frequency response characteristic of the human ear or eye, according to the performance considered. To that purpose a weighting or psophometric filter, with characteristics defined by the C.C.I.T.T. or C.C.I.R., is inserted in the channel to be measured. In the case of speech weighting the defined filter gives a 2.5 dB improvement of the signal-to-noise ratio at its output.

2.6.3.1. Effect of Modulation System

When a single-sideband, suppressed carrier, 100% amplitude modulated signal is used, only the amplitude variations of the carrier, caused by the noise, are considered because the "phase modulation" does not affect the amplitude modulation (AM) receivers. In this case the unweighted signal-to-noise ratio at the receiver output is given by:

$$(S/N)_o = \frac{P_R}{KzTB} \quad (2.40)$$

where P_R is the receiver input power, K is Boltzman's constant, z is the receiver noise factor, T is the absolute temperature (usually equal to 293°K) and B is the channel bandwidth. The above system will be the reference system in the following paragraphs.

If double-sideband AM were used, the symmetrical band should also be considered. The signal voltages would be synchronous and add voltage-wise when demodulated, whereas the noise voltages, statistically distributed, would add power-wise. In this case the signal-to-noise ratio would be two times that given in equation 2.40, corresponding to an improvement $V = 3\text{dB}$.

Due to the particular type of detection (discriminator) the noise appearing at the receiver input of a frequency modulation system acts

differently⁽¹⁵⁾. The amplitude variations are now eliminated by the limiter, while the phase variations persist being capable of acting on the discriminator designed for either frequency or phase modulation. If frequency modulation is used the maximum frequency deviation ($\Delta\omega$) caused by a noise component of frequency ω is given by:

$$\Delta\omega = \omega m \quad (2.41)$$

where m is known as the modulation index.

The discriminator transforms this $\Delta\omega$ into a voltage output:

$$E_o = A_1 \Delta\omega = A_1 \omega m \quad (2.42)$$

where A_1 is a constant.

This voltage is proportional to the noise modulation frequency (see figure 2.12) if the modulation index remains constant.

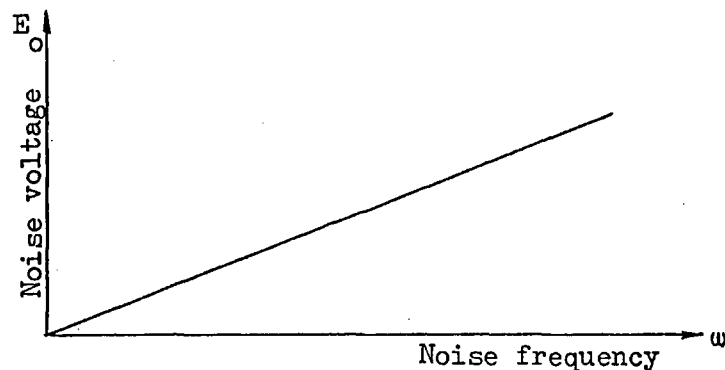


FIGURE 2.12

TRIANGULAR NOISE

This is usually known as "the triangular FM noise voltage law". In calculating the noise and signal powers, a deviation $\pm \Delta\omega$ is considered in order to take into account the action of the symmetrical band $(\omega_o - \omega)$ where ω_o is the central frequency.

Assuming the modulation spectrum covers the band $B = \omega_2 - \omega_1$ the

output noise and signal powers are respectively given by:

$$N_o = A_2 \frac{kzT}{P_R} \int_{\omega_1}^{\omega_2} \omega^2 d\omega = A_2 \frac{kzT}{P_R} \frac{\omega_2^3 - \omega_1^3}{3} \quad (2.43)$$

and

$$S_o = A_2 \frac{\Delta^2 \omega}{2} \quad (2.44)$$

where A_2 is a constant. The output signal-to-noise ratio is then given by the formula:

$$(S/N)_o = \frac{3}{2} \frac{P_R}{kzT} \frac{\Delta^2 \omega}{\omega_2^3 - \omega_1^3} \quad (2.45)$$

If the spectrum starts at $\omega_1 = 0$, then $B = \omega_2$ and:

$$(S/N)_o = \frac{P_R}{kzTB} \frac{3}{2} \left(\frac{\Delta \omega}{\omega_2} \right)^2 \quad (2.46)$$

The factor $\frac{P_R}{kzTB}$ is the output signal-to-noise ratio for a SSB - AM system covering the spectrum from 0 to ω_2 . Thus, the improvement factor is:

$$V = 10 \log \frac{3}{2} \left(\frac{\Delta \omega}{\omega_2} \right)^2 \quad (2.47)$$

This factor represents the typical triangular noise and applies to all single-channel modulation. In radio relay systems it mainly refers to television modulation where $B = \omega_2$ is equal to 5 MHz or 10 MHz according to line standards. In the case of multi-channel modulation, each

telephone channel covers a spectrum $B = \omega_2 - \omega_1 = \delta\omega = 3.1$ KHz while its mean frequency $\omega_m = (\omega_1 + \omega_2)/2$ may be several hundred KHz. In this case ω can be considered constant and equal to ω_m under the integration sign of equation 2.43 and the noise output becomes:

$$N_o = A_2 \frac{kzT}{P_R} \omega_m^2 (\omega_2 - \omega_1) \quad (2.48)$$

The signal-to-noise ratio in a channel centred on ω_m is:

$$(S/N)_o = \frac{P_R}{kzT\delta\omega} \frac{1}{2} \left(\frac{\Delta\omega}{\omega_m}\right)^2 = \frac{P_R}{kzTB} \frac{1}{2} \left(\frac{\Delta\omega}{\omega_m}\right)^2 \quad (2.49)$$

As in equation 2.46, the factor $\frac{P_R}{kzTB}$ corresponds to the SSB-AM system covering the spectrum $\delta\omega = \omega_2 - \omega_1 = B$. So, the improvement factor for this case is:

$$V = 10 \log \frac{1}{2} \left(\frac{\Delta\omega}{\omega_m}\right)^2 = 10 \log \frac{1}{2} \left(\frac{\sqrt{2} \Delta\omega_{rms}}{\omega_m}\right)^2 \quad (2.50)$$

where $\Delta\omega_{rms}$ is the root mean square deviation generally used by the CCIR and equal to $\frac{\sqrt{2}}{2}$ of the peak channel deviation $\Delta\omega$.

From equations 2.4, 2.49 and 2.50 we obtain:

$$(S/N)_o = P_T \left(\frac{\lambda}{4\pi d}\right)^2 \frac{1}{kzTB} \frac{1}{2} \left(\frac{\sqrt{2} \Delta\omega_{rms}}{\omega_m}\right)^2 \quad (2.51)$$

Taking into account the antenna gains (G) and feeder losses (L_F) and assuming $G_T - L_{TF} = G_R - L_{RF} = G' = 10 \log g'$, equation 2.51 gives the signal-to-thermal noise ratio for a voice channel ($B = \delta\omega = 3.1$ KHz)

centred on ω_m

$$(S/N)_o = P_T \left(\frac{\lambda}{4\pi d} \right)^2 \frac{g^2}{k_z T \delta \omega} \frac{1}{2} \left(\frac{\sqrt{2} \Delta \omega_{rms}}{\omega_m} \right)^2 \quad (2.52)$$

Equation 2.52 expressed in dB or dBm gives equation 2.36 if the weighting factor W is added.

2.6.3.2. Pre-emphasis and De-emphasis

In order to compensate for the noise increase with the frequency (figure 2.12), which results in large differences in the signal-to-noise ratios in the different channels, it would be necessary to increase the deviation ($\Delta\omega$) in proportion to ω_m . This is achieved by applying amplitude pre-emphasis at the transmitter on the baseband signal resulting in the transmission of the top modulation frequencies with a higher deviation than the bottom ones. By a complementary de-emphasis after demodulation the baseband response is recovered.

For radio-relay systems the following C.C.I.R. recommendation applies, for multi-channel telephony:

TABLE 2-I

ω_m / ω_{max}	0	0.2	0.4	0.6	0.8	1
dB with respect to standardized values	- 4	- 3.4	- 2.1	0.1	+ 2	+ 4

The amount of improvement obtained when wide-band modulation is applied (frequency modulation, pulse modulation) is sometimes called "wide-band gain" and is realized only if the signal remains at a level above the noise level by a certain amount. This level is called the "threshold" level of the receiver and for F.M. systems is about 10 to 13 dB. The

reason is that a considerable number of noise peaks reach this order of level above the r.m.s. noise and the limiter, preceding the demodulator, suppresses only the weaker of the two signals. If a noise peak exceeds the signal amplitude it reaches the demodulator while the signal is suppressed.

This effect is caused by any interfering frequency that reaches the receiver; therefore, the IF-bandwidth must be taken into account when determining the threshold power of the receiver input.

2.6.3.3. System Figure of Merit

A main feature for any radio-relay transmit-receive equipment is the so-called "system figure of merit" which is the logarithmic ratio of the transmitter power to the characteristic noise of the system, referred to the signal channel bandwidth and measured at the receiver output.

By means of terms in equation 2.36 this can be expressed, in dB or dBm, as:

$$S = P - C - B - Z + V + W \quad (2.53)$$

and in terms of equation 2.52 this is given as the ratio:

$$S = P_T \cdot \frac{w}{k_z T \delta \omega} \cdot \frac{1}{2} \left(\frac{\sqrt{2} \Delta \omega_{rms}}{\omega_m} \right)^2 \quad (2.54)$$

where w stands for the weighting factor (1.78 for speech).

The difference between system figure of merit, given by the manufacturer, and path loss (antenna gains and feeder losses inclusive) gives directly the signal-to-noise ratio referred to the signal channel.

2.6.4. Transmission Distortion and Intermodulation Noise

In practice there are other sources of noise to be taken into account due to the signal distortion caused by the transmission system. This distortion, which is partially independent and partially a function of the modulation frequency, produces either a direct distortion of the signal to be transmitted or an intermodulation noise between the channels, which is added to the thermal noise of the particular circuit. The direct distortion is of importance for television signals.

Frequency independent distortion, depending on the signal amplitude, is due to non-linearities of the baseband amplifiers and of the modulators and demodulators and is not very difficult to calculate. On the contrary, frequency dependent distortion is difficult to calculate and is mainly due to non-linearities in the phase characteristics of the R.F. and I.F. circuits; it is also due to amplitude variations within the pass-band of the filters and amplifiers.

Typical noise figures for design purposes are:

- a. 100 pW per modulator-demodulator pair.
- b. 70 pW per amplifier.

CHAPTER THREE

A MICROWAVE LINK BETWEEN LONDON AND CHRISTCHURCH

3.1. General - Profiles

Between London (I.C.) and Christchurch (S.R.D.E.) more than one path or route can be selected comprising three or more microwave sections; in other words using two or more repeater stations. Each of them has its own advantages and disadvantages whereas some seem to be quite unsuitable.

Instead of studying each route as a whole (terminal to terminal) from the very beginning, it was considered advantageous to study reasonable sections separately and then to combine the most favourable and consider them when integrated in various routes before proposing the best. Although this required many profiles to be drawn, every possibility was examined and the optimum combination obtained. To avoid confusion in identifying every section, the following nomenclature has been used:

- L, represents the terminal at Imperial College (TQ267792) in London.
- S, represents the terminal at the S.R.D.E. site (SZ182928) at Christchurch.
- W, represents a repeater station at Rowridge (SZ447866) on the Isle of Wight.
- A₁, represents a repeater station at Leith hill (TQ139431) on the North Downs S.E. of Guildford.
- A₂, represents a repeater station at Holmbury hill (TQ104429) on the North Downs S.E. of Guildford.
- A₃, represents a repeater station at Gibbet hill (SU900359) North of Haslemere.
- A₄, represents a repeater station at a site (SU725437) North of Alton.

B_1 , represents a repeater station at Linch hill (SU849173) on the South Downs.

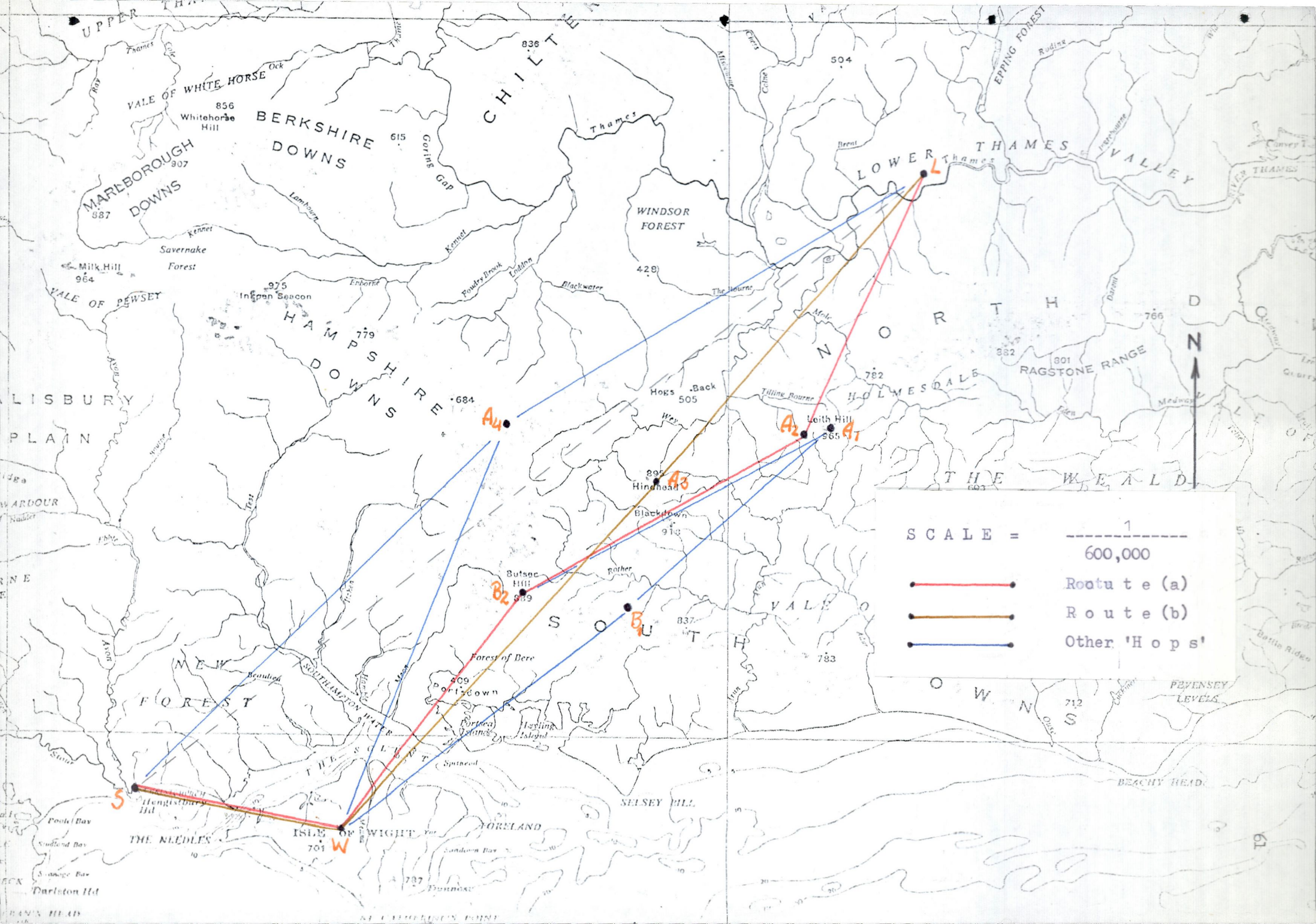
B_2 , represents a repeater station at Butser hill (SU720207) on the South Downs.

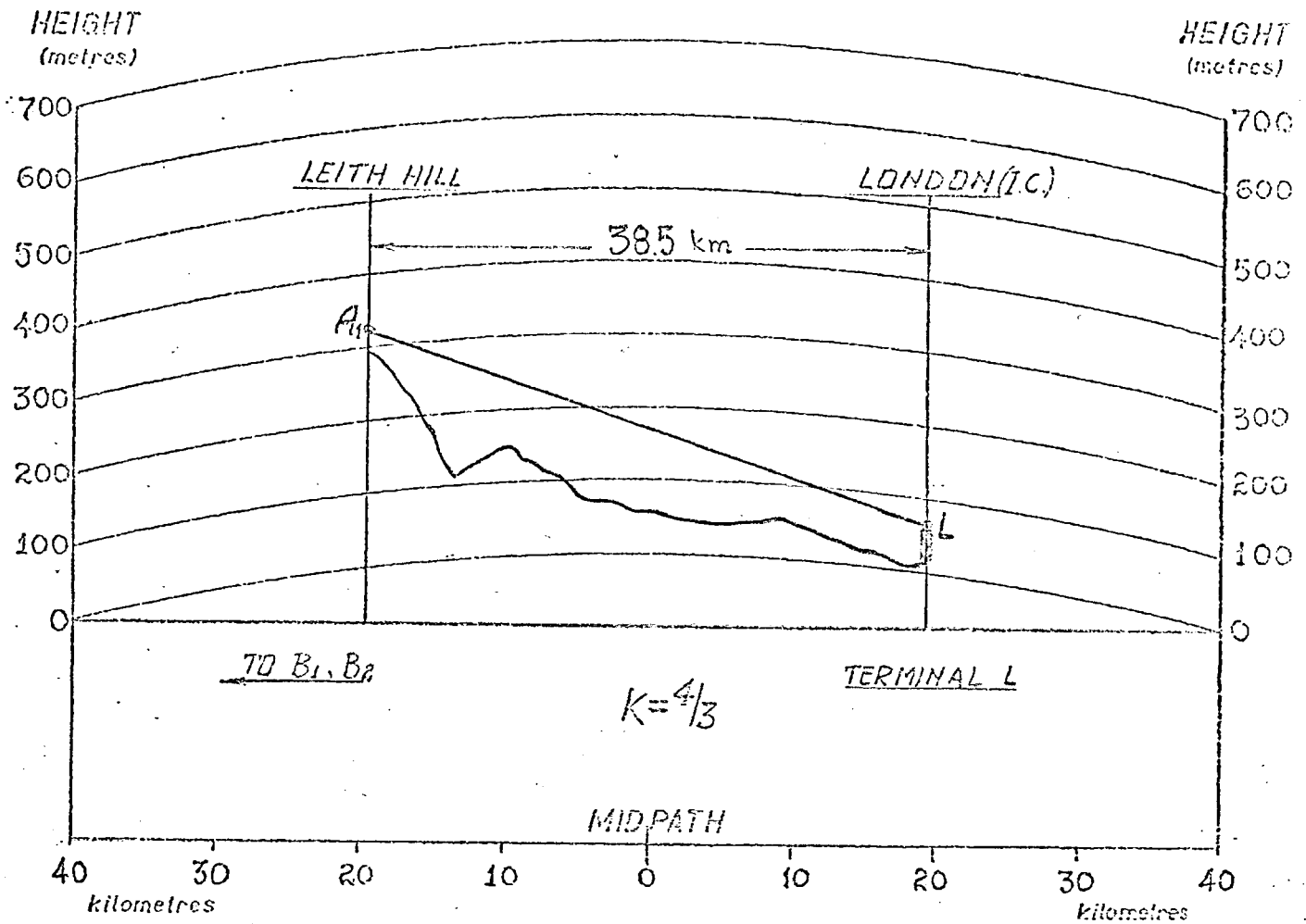
Using this nomenclature every hop is identified by the stations at its ends; so that hop A_1B_2 is the microwave section between the station at Leith hill and the station at Butser hill. The following hops seem to be the most reasonable and have been studied in this work:

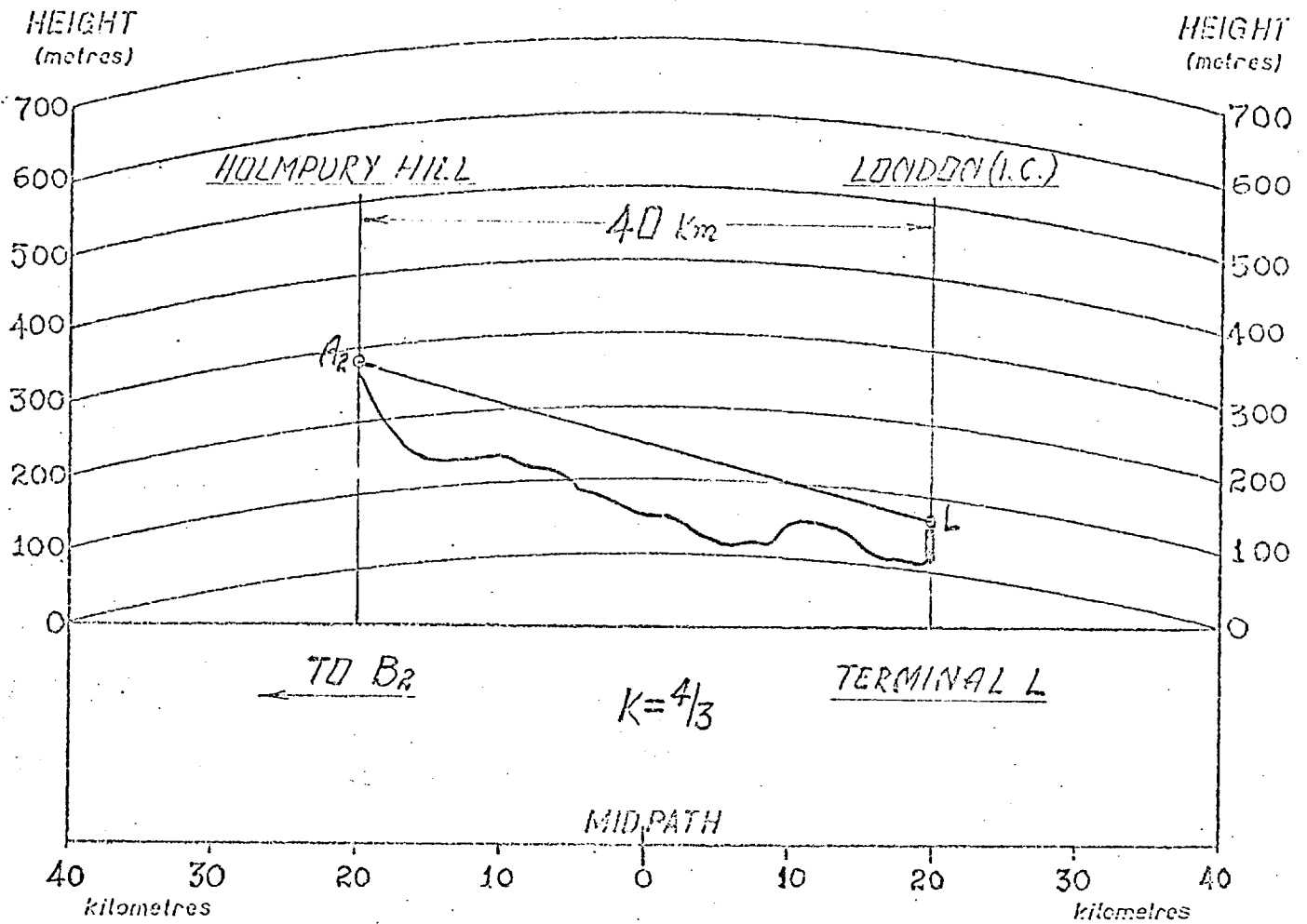
1. Hop LA_1 38.5 kilometres
2. Hop LA_2 40.0 kilometres
3. Hop LA_3 57.6 kilometres
4. Hop LA_4 65.3 kilometres
5. Hop A_1B_1 39.0 kilometres
6. Hop A_1B_2 48.0 kilometres
7. Hop A_2B_2 44.5 kilometres
8. Hop A_3W 67.2 kilometres
9. Hop A_4W 64.2 kilometres
10. Hop B_1W 50.5 kilometres
11. Hop B_2W 44.0 kilometres
12. Hop B_2S 60.0 kilometres
13. Hop WS 27.0 kilometres

Profiles on a 4/3-earth profile paper have been drawn for all these hops, using the available topographical Ordnance Survey maps of scale 1:25,000 with contour lines every 25 feet, published by H.M.S.O.. These profiles are given in the following pages.

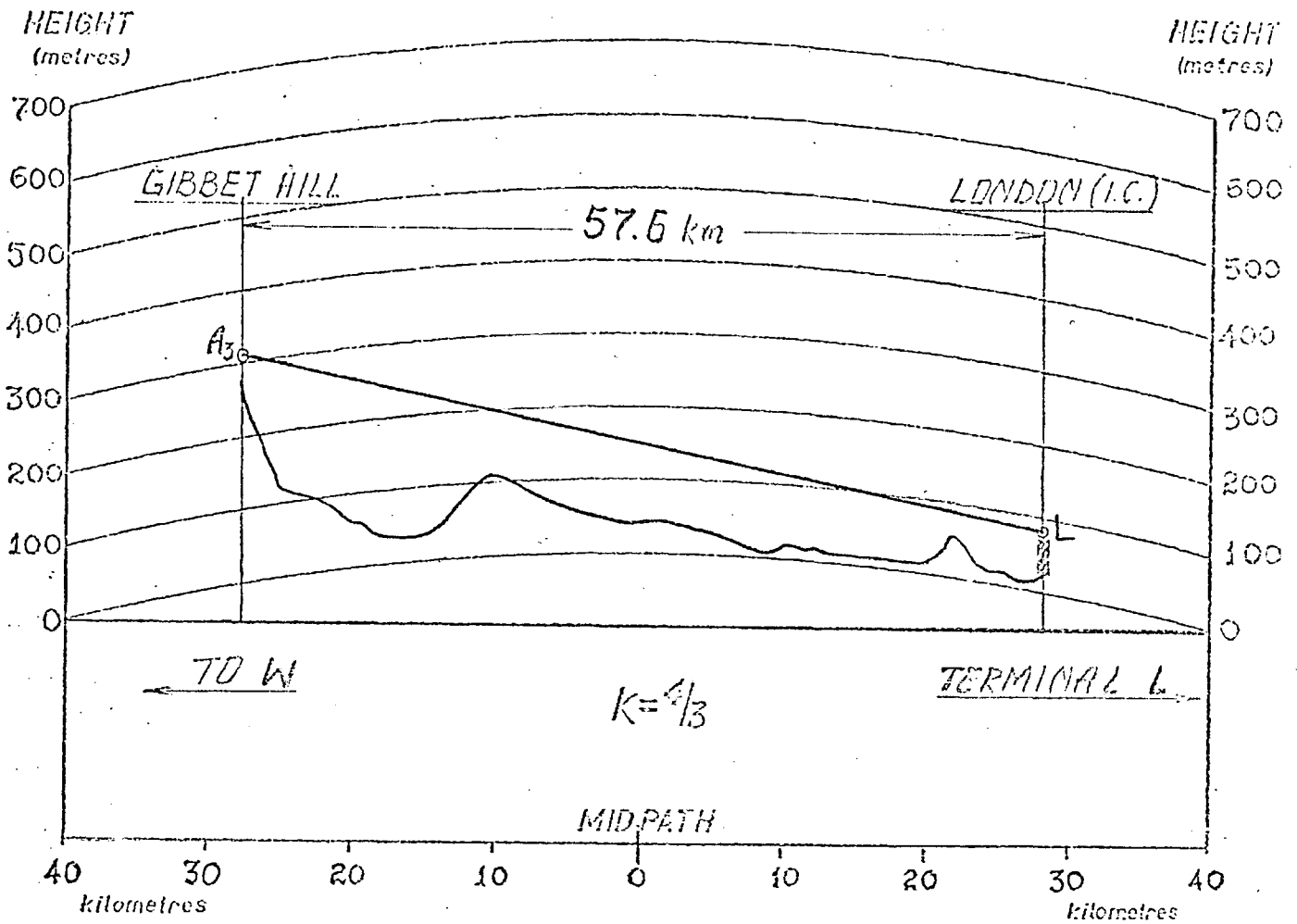
From equations 2.7 and 2.28 the free space loss (L) and the first Fresnel-zone clearance for midpaths (F_{lm}) and other important positions (F_1), have been computed. The results are listed below together with the necessary mast heights (M) to meet these clearances.

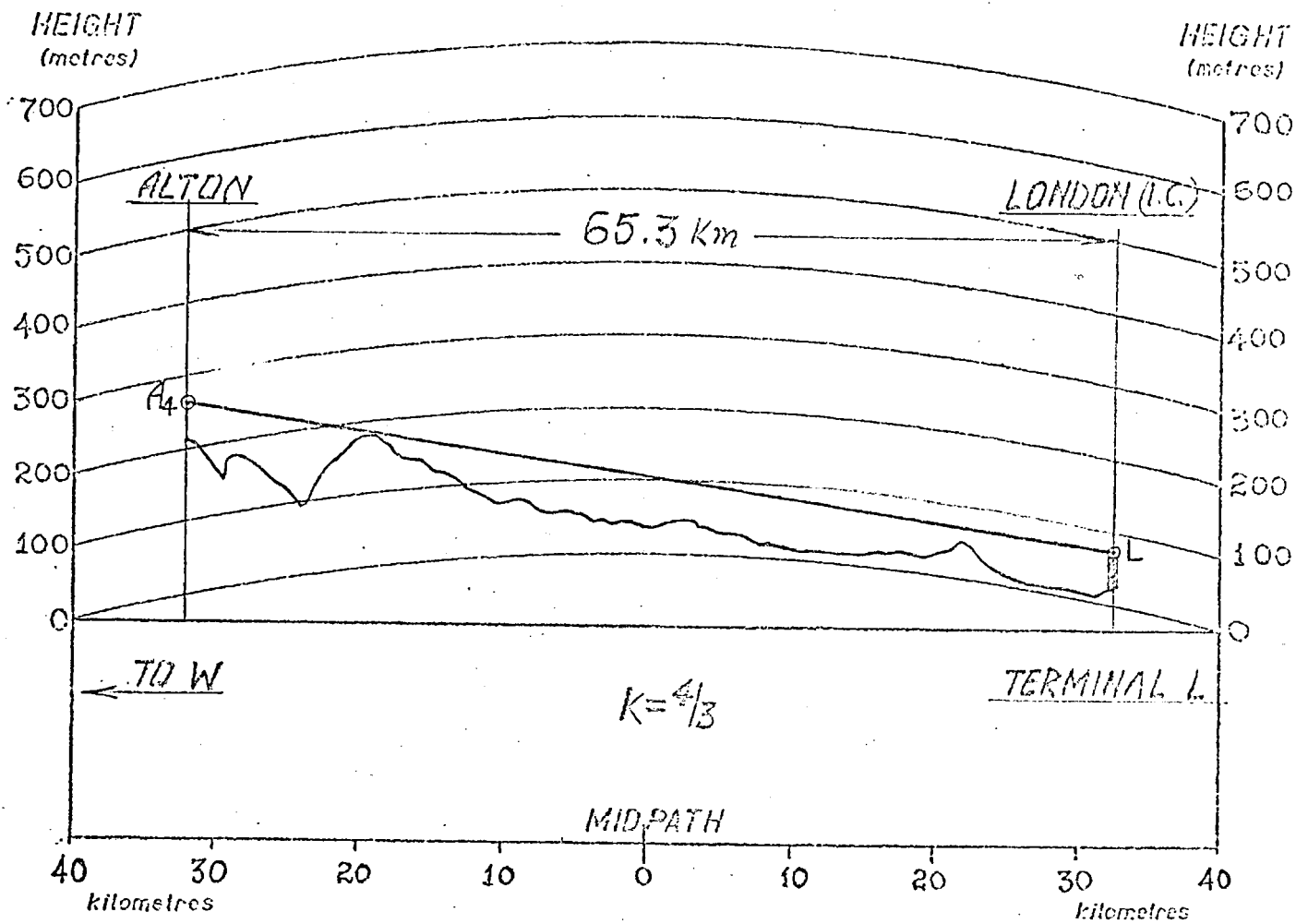


HOP LA₁

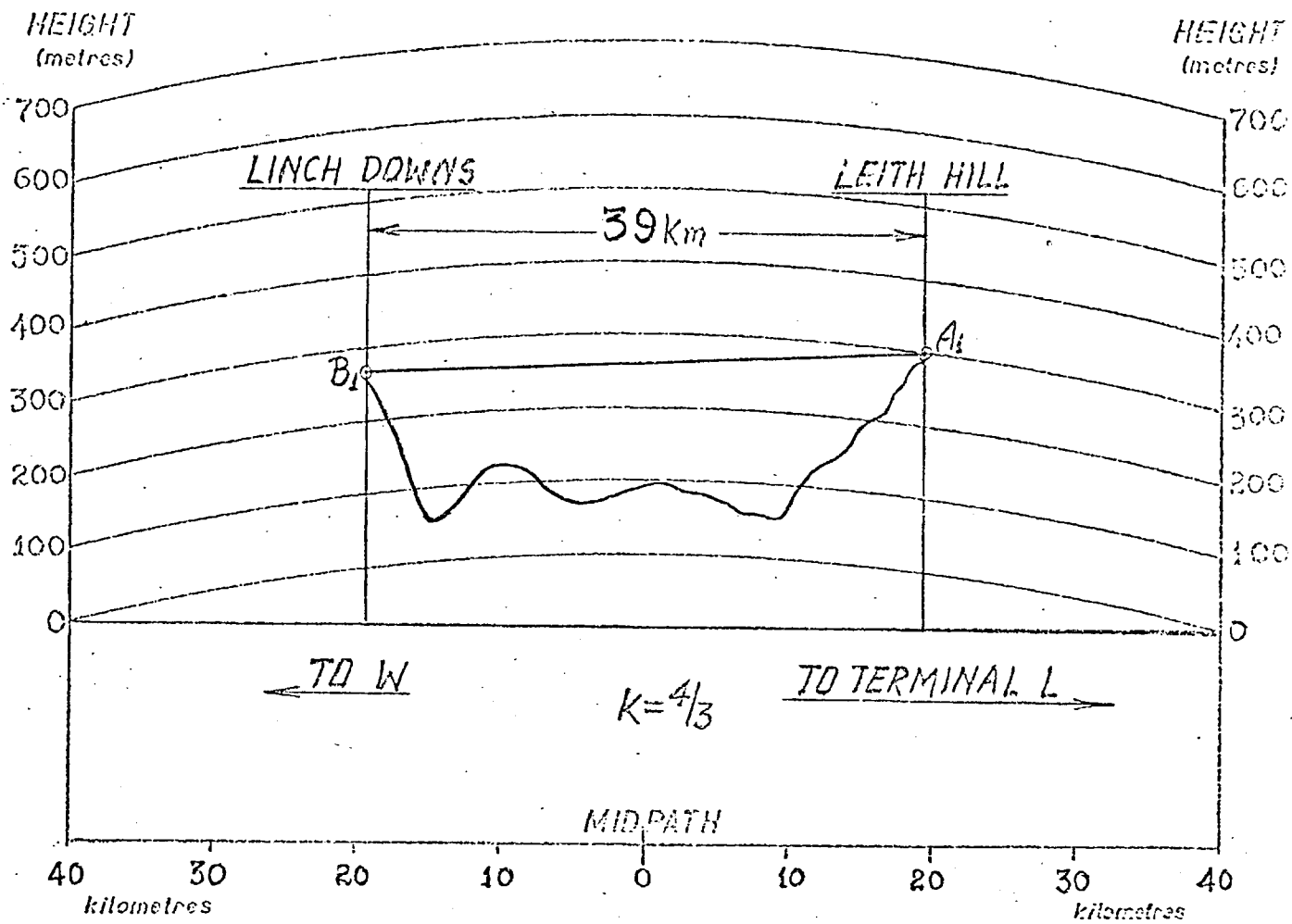
HOP LA₂

HDP LA₃

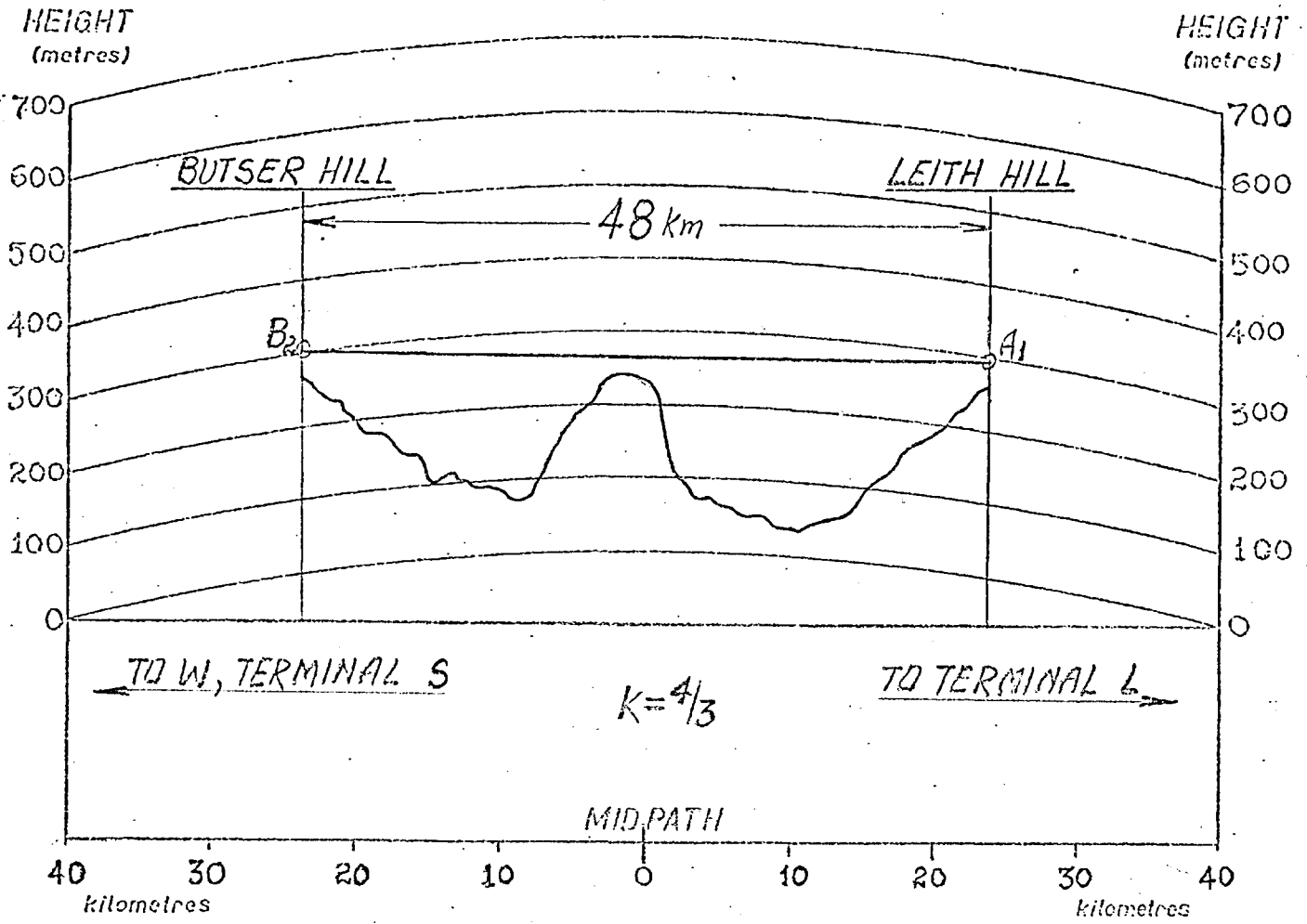


HOP LA

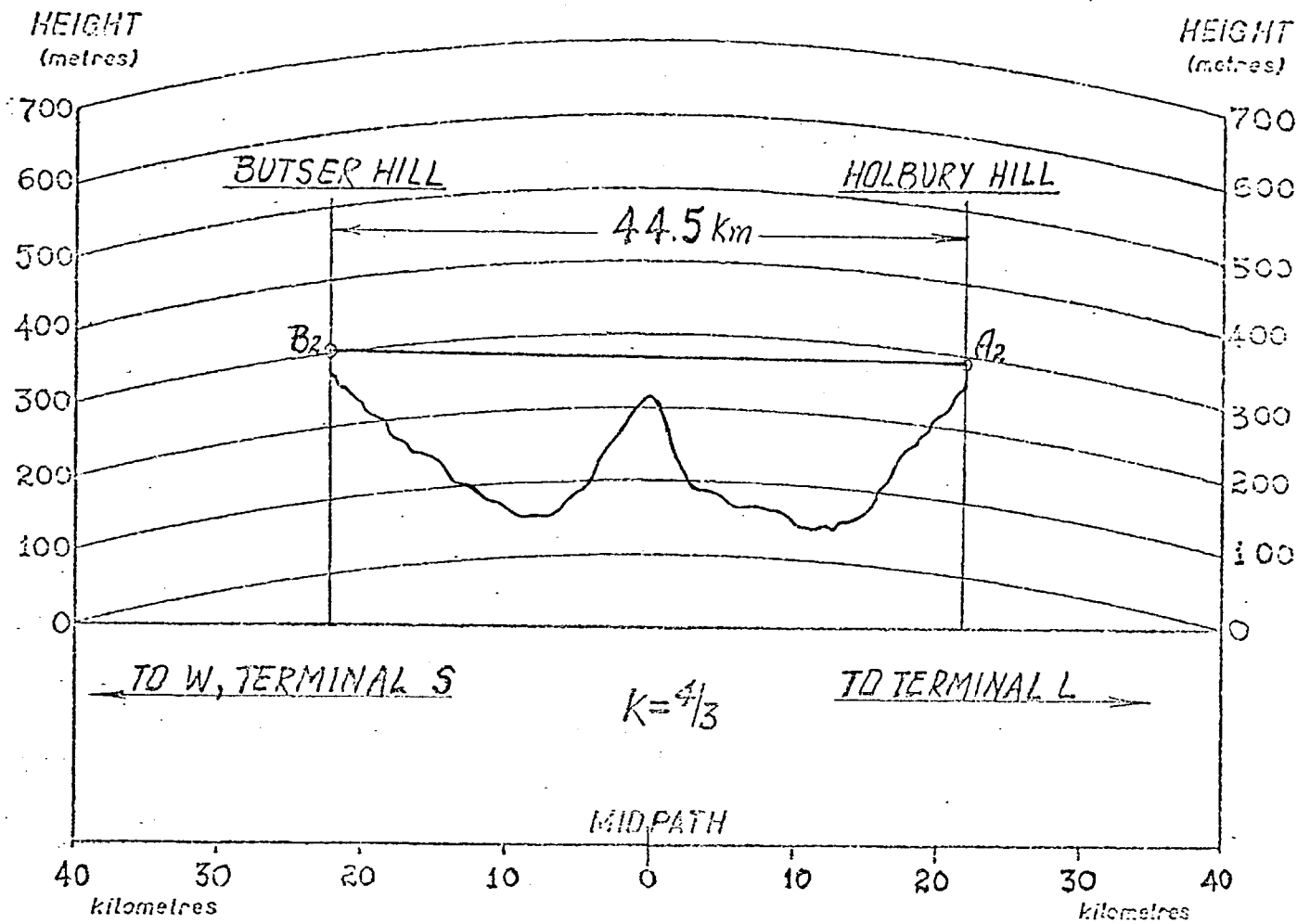
HOP $A_1 B_1$



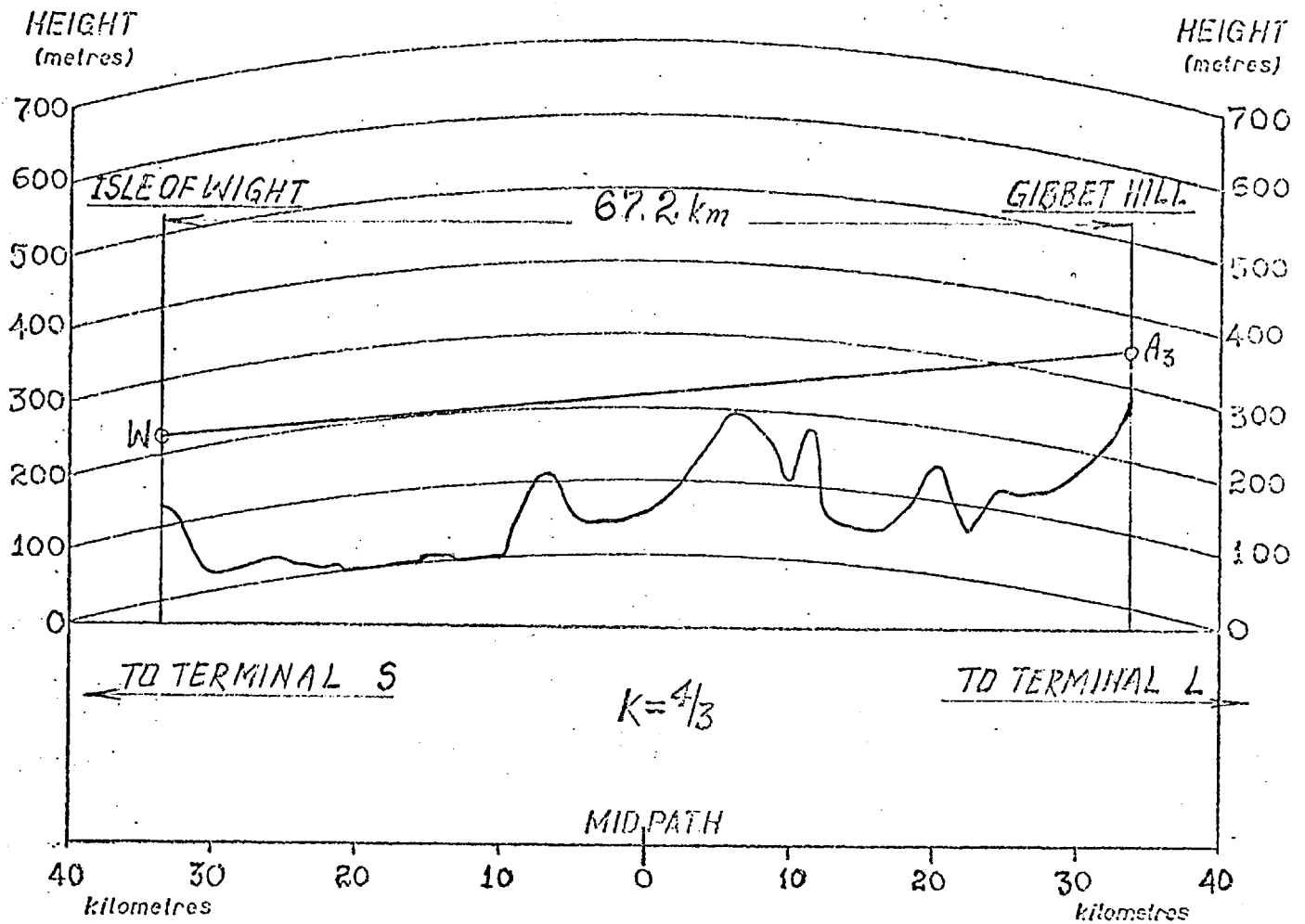
HOP $A_1 B_2$



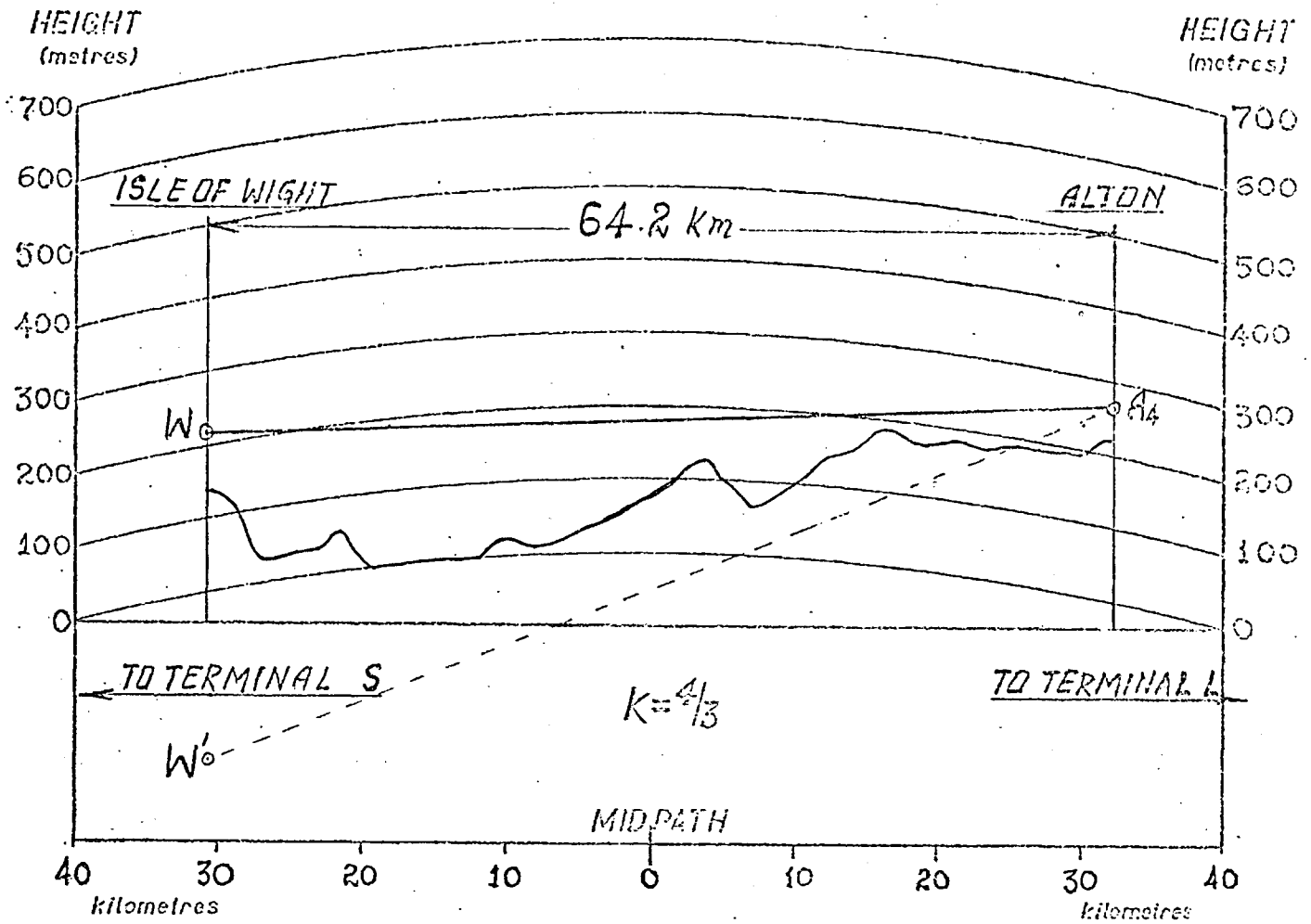
HOP A_2B_2



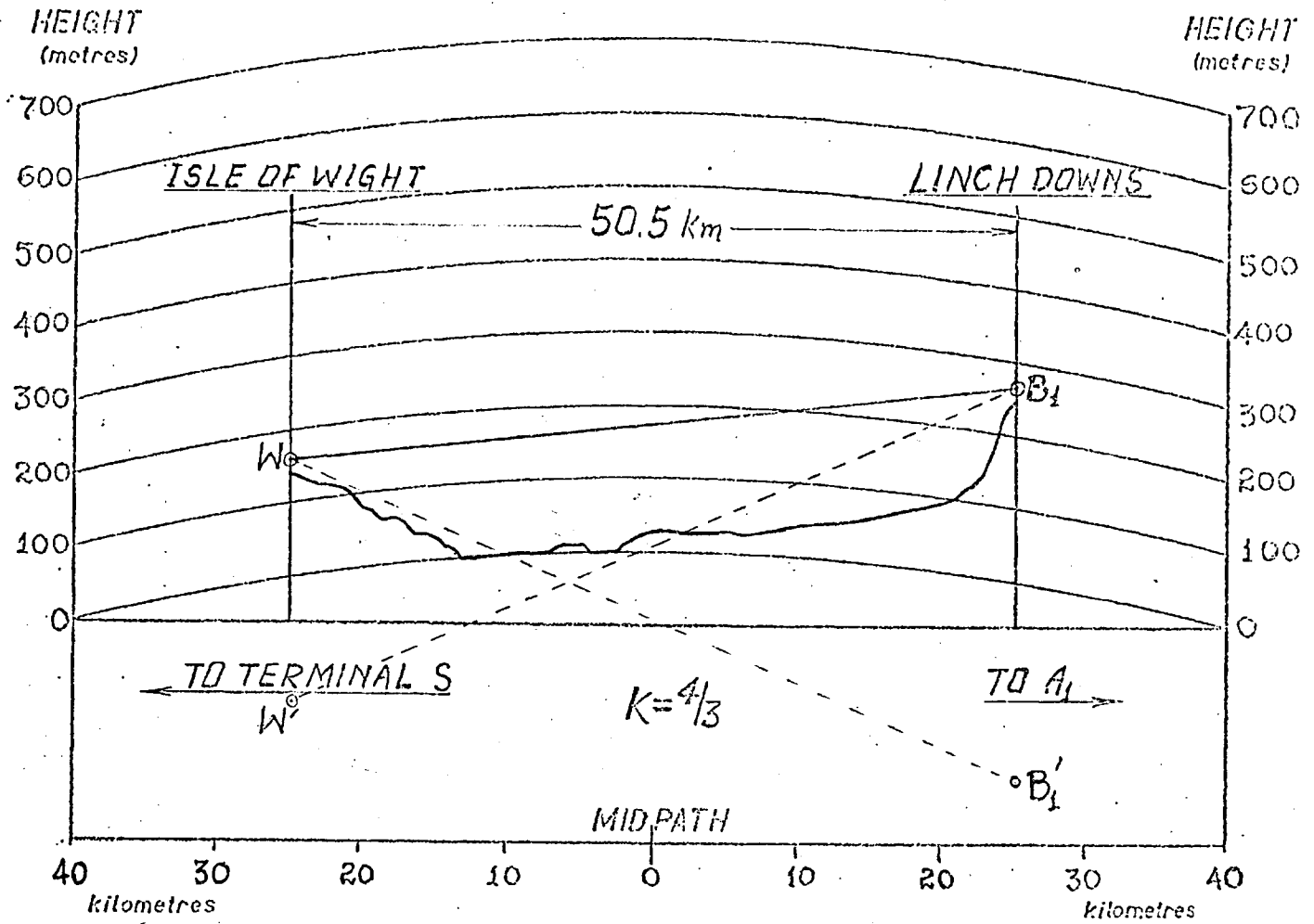
HDP A₃W

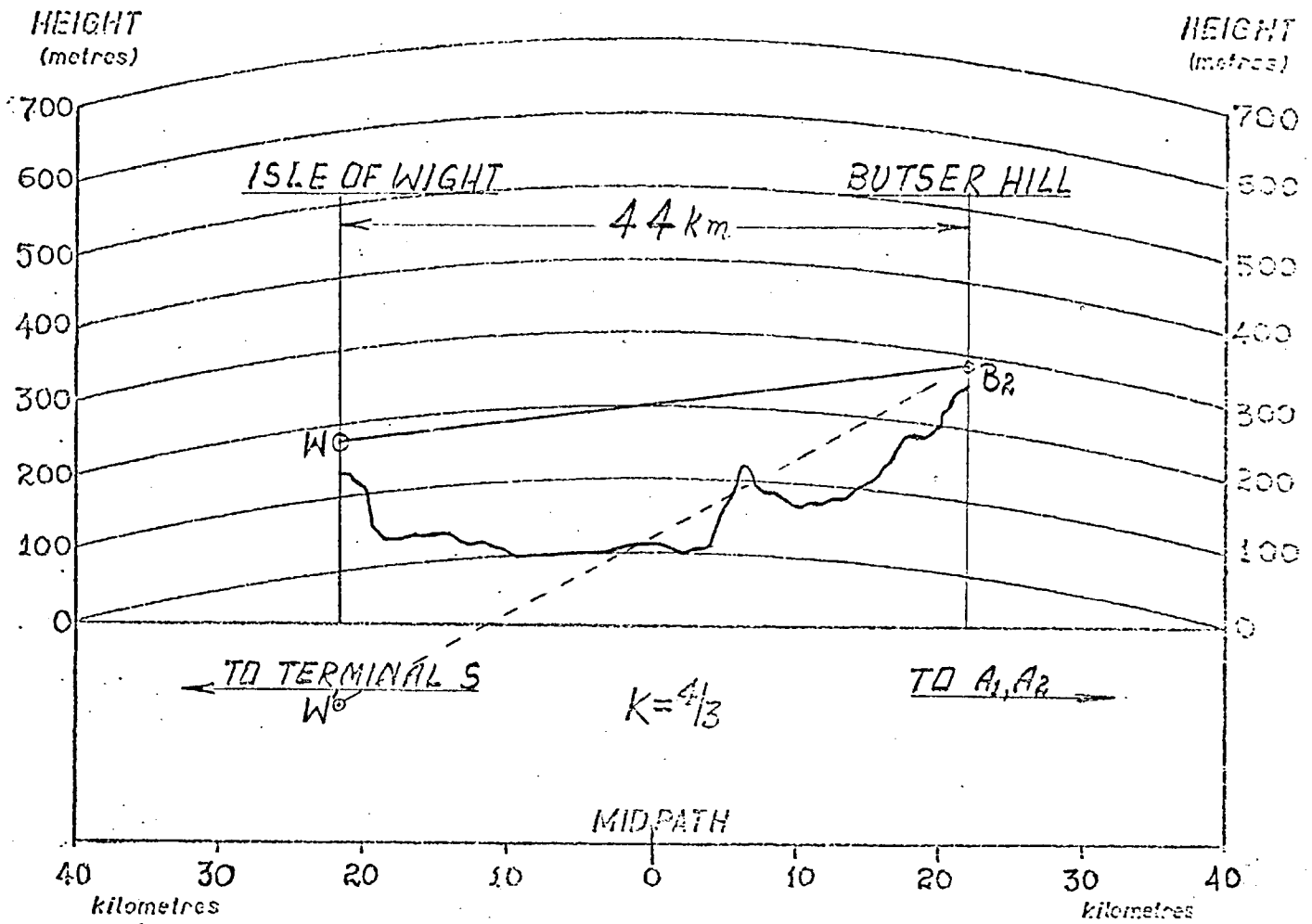


HOP A₄W

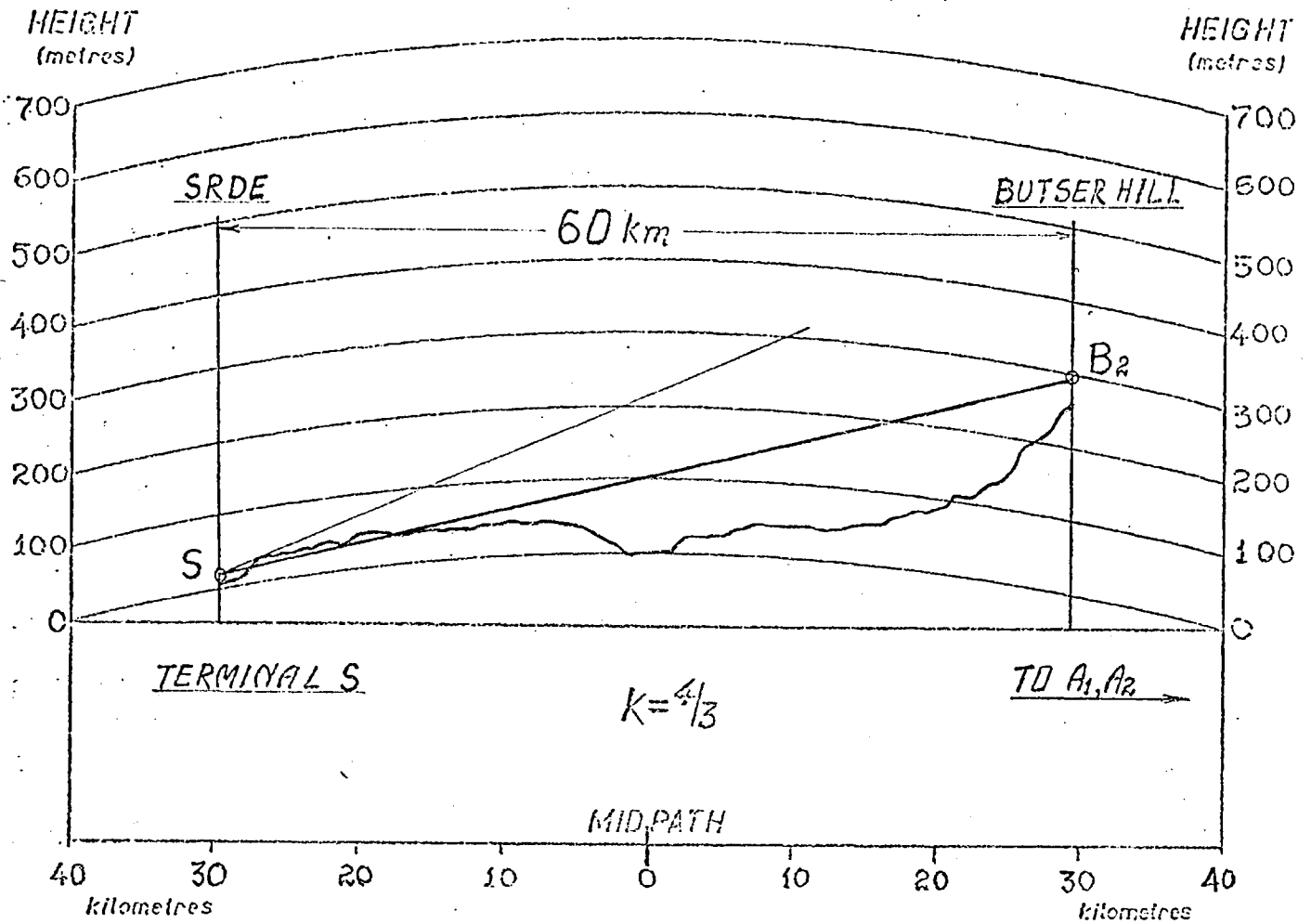


HOP B_1W

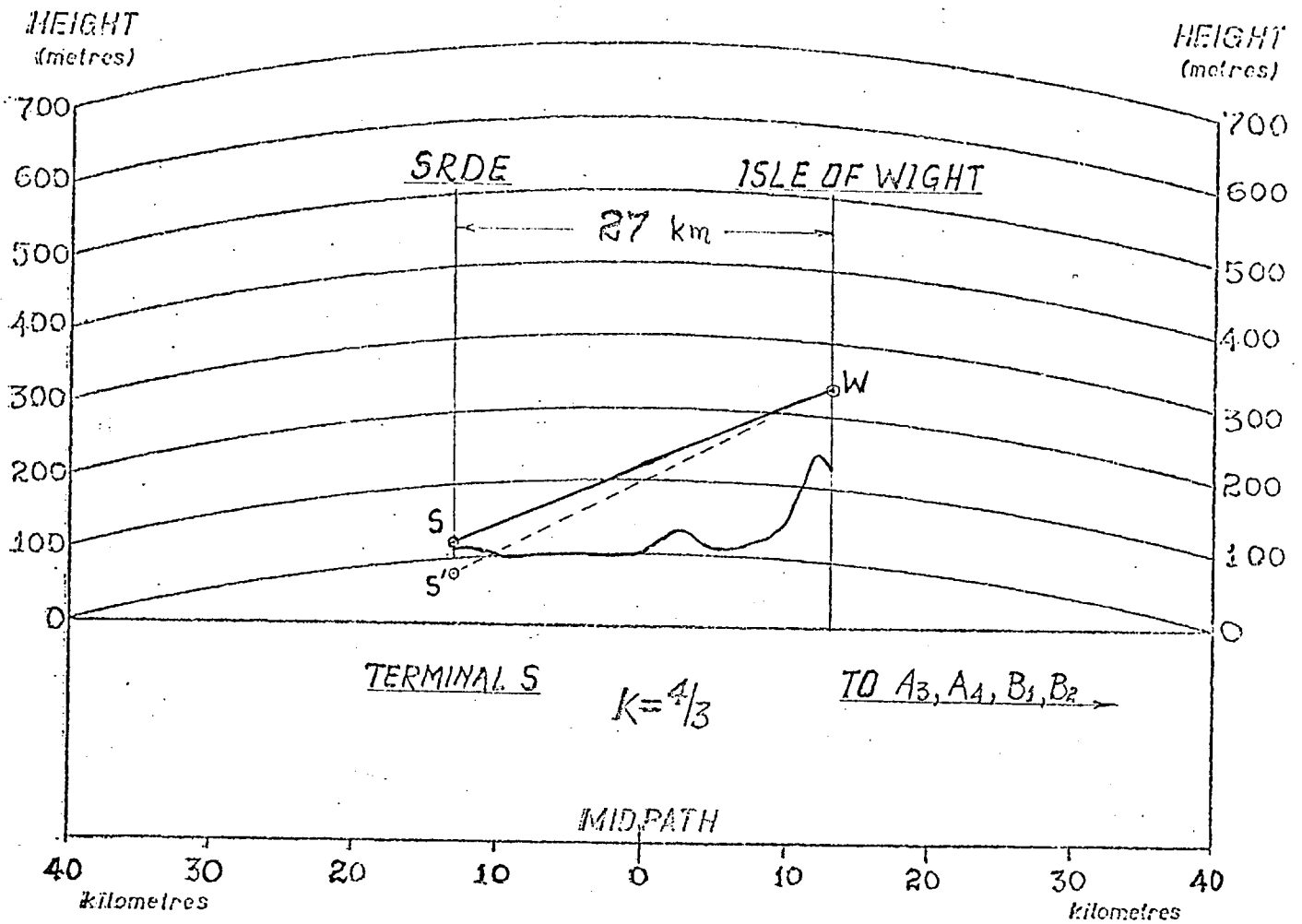


HOP B_2W 

HOP. B₂S



HDP WS



HOP LA₁10 km from station A₁

10.5 km from station L

$$L = 123.3 \text{ dB}$$

$$F_{lm} = 57 \text{ m}$$

$$F_1 = 50 \text{ m}$$

$$F_1 = 51 \text{ m}$$

$$M_L = 15 \text{ m}, \quad M_{A_1} = 15 \text{ m}$$

HOP LA₂

6.8 km from terminal L

10 km from terminal L

9.8 km from station A₂

$$L = 123.6 \text{ dB}$$

$$F_{lm} = 57.8 \text{ m}$$

$$F_1 = 43.5 \text{ m}$$

$$F_1 = 50.0 \text{ m}$$

$$F_1 = 49.5 \text{ m}$$

$$M_L = 10 \text{ m}, \quad M_{A_2} = 18 \text{ m}$$

HOP LA₃18 km from station A₃

7.6 km from terminal L

$$L = 126.8 \text{ dB}$$

$$F_{lm} = 69.4 \text{ m}$$

$$F_1 = 64 \text{ m}$$

$$F_1 = 47 \text{ m}$$

$$M_L = 30 \text{ m}, \quad M_{A_3} = 55 \text{ m}$$

HOP LA_4 11.9 km from station A_4

31 km from terminal L

10 km from terminal L

$$L = 128 \text{ dB}$$

$$F_{1m} = 74 \text{ m}$$

$$F_1 = 57 \text{ m}$$

$$F_1 = 73.5 \text{ m}$$

$$F_1 = 53 \text{ m}$$

$$M_4 = 50 \text{ m}, \quad M_{A_4} = 110 \text{ m}$$

HOP A_1B_1 10 km from station B_2

$$L = 123.4 \text{ dB}$$

$$F_{1m} = 57 \text{ m}$$

$$F_1 = 50 \text{ m}$$

$$M_{A_1} = 3 \text{ m}, \quad M_{B_1} = 3 \text{ m}$$

HOP A_1B_2 22 km from station B_2

$$L = 125.2 \text{ dB}$$

$$F_{1m} = 63.3 \text{ m}$$

$$F_1 = 63 \text{ m}$$

$$M_{A_1} = 75 \text{ m}, \quad M_{B_2} = 93 \text{ m}$$

HOP A_2B_2

$$L = 124.5 \text{ dB}$$

$$F_{1m} = 61 \text{ m}$$

$$M_{A_2} = 25 \text{ m}, \quad M_{B_2} = 65 \text{ m}$$

HOP A₃W

26.6 km from station W

27 km from station A₃22 km from station A₃

$$L = 128.1 \text{ dB}$$

$$F_{1m} = 75 \text{ m}$$

$$F_1 = 73 \text{ m}$$

$$F_1 = 73.3 \text{ m}$$

$$F_1 = 70.2 \text{ m}$$

$$M_{A_3} = 120 \text{ m}, \quad M_W = 140 \text{ m}$$

HOP A₄W13 km from station A₄16 km from station A₄28.7 km from station A₄

$$L = 127.7 \text{ dB}$$

$$F_{1m} = 73.2 \text{ m}$$

$$F_1 = 58.5 \text{ m}$$

$$F_1 = 63.3 \text{ m}$$

$$F_1 = 72.7 \text{ m}$$

$$M_{A_4} = 93 \text{ m}, \quad M_W = 110 \text{ m}$$

HOP B₁W

4.5 km from station W

$$L = 125.6 \text{ dB}$$

$$F_{1m} = 65 \text{ m}$$

$$F_1 = 79 \text{ m}$$

$$M_{B_1} = 3, \quad M_W = 3$$

HOP B₂W16 km from station B₂

$$L = 124.4 \text{ dB}$$

$$F_{1m} = 60.6 \text{ m}$$

$$F_1 = 53.3 \text{ m}$$

$$M_{B_2} = 3, \quad M_W = 3$$

HOP B₂S

$$L = 127.1 \text{ dB}$$

$$F_{1m} = 70.7 \text{ m}$$

$$M_{B_2} = 30 \text{ m}, \quad M_S = 92 \text{ m}$$

HOP WS

$$L = 120.2 \text{ dB}$$

$$F_{1m} = 44.5 \text{ m}$$

16 km from station S

$$F_1 = 46.5 \text{ m}$$

1 km from station W

$$F_1 = 18 \text{ m}$$

$$M_W = 3, \quad M_S = 3$$

Reasonable mast heights for each station are shown on the profiles and have been taken as the basis for further discussion.

3.2. Characteristics of the Radio Sections

Studying each section individually we may notice the following advantages and disadvantages:

HOP LA₁

From London (I.C.) to Leith hill a distance of 38.5 kilometres, introduces a comparatively small free space loss of 123.3 dB. Adequate terrain clearance is obtained over all the hop except at a distance 10.5 kilometres from terminal L, where a clearance less than the first Fresnel-zone is obtained if the buildings existing at that point are to be taken into account. Nevertheless, the line-of-sight is not obstructed and a clearance of more than 0.6 of the first Fresnel-zone is maintained for a 4/3-earth atmosphere.

Terrain reflections will not introduce significant interference fading due to the terrain roughness and antenna locations. Some site development clearance at the Leith hill location will be necessary and the cost of the antenna tower and of a short access road will be considerable; but the cost of supplying power to the site will not be high since there is a nearby power station.

In conclusion this hop seems to be reasonable but its eventual incorporation in the route must be considered in accordance with the possible use of hops A_1B_1 or A_1B_2 and after comparison with hop IA_2 .

HOP IA_2

A free-space loss of 123.6 dB for its 40 kilometres makes this hop similar to the previous one from this point of view. The same applies considering Fresnel-zone clearance and terrain reflection effects; but the cost of an access road and antenna tower construction will be considerably lower due to the characteristics of the site. Site development clearance is unnecessary and the cost of supplying power to the site (A_2) will be low.

HOP IA_3

A relatively high loss of 126.8 dB is introduced by a distance of 57.6 km. To avoid diffraction loss due to intervening terrain a very tall tower has to be constructed at Gibbet hill (A_3) site with a rigidity suitable for use with the narrow beam antennas. The cost of supplying power to the site (A_3) must be taken into account.

On the other hand, the location of station A_3 almost midway between terminal L and station W makes it possible to link the terminals by using only two repeater stations. Terrain reflections will be insignificant and site development clearance is unnecessary. A very short access road is needed but its cost will be low.

Even with a high located antenna at station A_3 some diffraction loss

might be expected on extreme atmospheric deflection of the path towards the terrain.

HOP LA_4

High loss is also introduced by this hop (128 dB). On the other hand adequate Fresnel-zone clearance is not obtained at two points along the path: at Richmond hill (10 km from terminal L) and at Beacon hill (11.9 km from station A_4). The existence of large buildings at the former and trees at the latter makes the situation more unfavourable. The existence of a tower on site A_4 and the availability of an access road and power supply are some advantages to the section but not enough to compensate for its disadvantages. A highly elevated antenna at I.C. might make this hop very suitable but the cost for a stable antenna would be high.

For the situation as it is, this hop seems quite unsuitable.

HOP A_1B_1

Low transmission loss (123.4 dB) and good propagation conditions may be expected if this hop is adopted. Highly elevated terrain at both stations can give line-of-sight even with antennas mounted a few metres over the ground, but the cost of site development clearance, access road construction and the supplying of power will be very high.

With its two stations on the North Downs (A_1) and South Downs (B_1) this hop interlinks terminal L and repeater W avoiding diffraction losses due to these two transverse Downs. On the other hand, these two sites must be developed from scratch because no tower or other installation exists on them.

HOP A_1B_2

The existence of a tower on Butser hill (B_2) in connection with a loss of 125.2 dB due to its 48 km length give an advantage to this hop; but a midpath clearance smaller than $0.4 F_1$ even with normal atmospheric

conditions make this hop unsuitable.

HOP A_2B_2

A midpath clearance of more than 0.7 of the first Fresnel-zone, combined with the existence of a tower on Butser hill (B_2) and a free-space loss of only 124.5 dB make this a very good section. No terrain reflections are expected, and the low cost of developing the A_2 site, as mentioned in describing hop LA_2 , give one more advantage to this hop over hop A_1B_2 .

HOP A_3W

A transmission loss of 128.1 dB is introduced by this hop of distance 67.2 kilometres between its stations. A clearance of only 0.45 of the First Fresnel-zone is obtained at a point distant 27 kilometres from station A_3 . Nevertheless, the loss introduced by diffraction by the terrain at that point will be small for normal atmospheric conditions, as it can be estimated from the nomogram in figure 2.4.

On the other hand, station A_3 has the advantage of being located almost midway between terminal L and repeater W as mentioned before (HOP LA_3).

Cost of tower, access road and power supply at station A_3 will be very high, and must be seriously considered. Adopting this hop, only three hops need be employed and one tower on Gibbet hill (station A_3) will be necessary.

HOP A_4W

A relatively high free space loss (127.7 dB) and an inadequate terrain clearance ($0.25 F_1$) at a point 13 kilometres from station A_4 , are the main disadvantages of this hop. On the other hand it can only be used in combination with hop LA_3 which has proved to be undesirable. As a result its great advantage of two existing towers at both ends can

not be made use of.

HOP B₁W

Good terrain clearance and a relatively small free-space loss are the main advantages. Some ground reflections (probably specular) and cost of tower, access road etc. at B₁ station are its disadvantages.

HOP B₂W

Many advantages are obtained with this hop namely; relatively small free-space loss (124.5 dB), terrain clearance adequate over all the path, screening for the ground reflected rays and existence of antenna towers at both ends.

Its main disadvantage is the necessity of three more hops for a link between terminals L and S.

HOP B₂S

From the path profile it can be seen that the line-of-sight propagation between station B₂ and terminal S is seriously obstructed near terminal S by the terrain. A diffraction loss of more than 20 dB has been estimated as introduced by this obstacle; this combined with a transmission loss of 127 dB makes this hop quite unsuitable, unless a very high mast with a very stable antenna could be constructed at terminal S. This in turn, would be very expensive.

HOP WS

A small distance (27 km) between its ends gives a correspondingly low transmission loss of only 120.2 dB. Adequate terrain clearance over all the path and a high-low technique for the ground reflection avoidance, make this hop quite desirable.

The situation of the terminal S at a low position near the coast in connection with the comparatively elevated terrain at its north and north-east sides, makes the use of hop WS unavoidable unless a high tower is constructed at site S.

3.3. Equipment Data

The narrow-band system FM 12/24-800 (manufactured by Siemens and Telefunken - Germany) has the following technical characteristics and can be taken as the basis for the system employed in the link:

Frequency band	: 610 to 960 MHz	
Transmit-receive spacing		
1 antenna	: 40 MHz	
2 antennas	: About 8 MHz	
Transmitter power	: About 10 Watts	
Modulation frequencies	: 4 to 108 KHz	
	150 ohms balanced	
Frequency tolerance	: $\pm 2.10^{-4}$	
R.M.S. frequency deviation		
4 channels	: 45 KHz	
12 channels	: 14 KHz	referred to test tone level 0 dBmo
24 channels	: 10 KHz	
Pre-emphasis	: 6 dB	
Receiver noise figure	: ≤ 12 dB	
Intermediate frequencies	: 50 and 35 MHz	
System figure of merit		
4 channels	: ≥ 172 dB	
12 channels	: ≥ 159 dB	
24 channels	: ≥ 152 dB	

3.4. Possible Routes Between the Terminals

According to the characteristics of the described hops, two routes seem to be quite reasonable for a microwave radio relay link between the terminals L and S.

- (a) The first consists of four hops using the repeater stations A_2

on Holmbury hill on the North Downs, B_2 on Butser hill on the South Downs and W at Rowridge on the Isle of Wight.

(b) The second consists of only three hops making use of the repeater stations on Gibbet hill (A_3) north or Haslemere and at Rowridge (W) on the Isle of Wight.

To estimate their transmission performance and to relate it to the planning rules given in CCIR Recommendation No. 395 (noise in real circuits) a list for each of them has been made as shown in tables 3-I and 3-II. From the path loss and the system figure of merit the overall thermal noise of the system is obtained.

Due to the small length (D) of both routes, the formula $50\sqrt{D}$ has to be applied as recommended by the CCIR and described in the previous chapter.

A system figure of merit equal to 159 dB, corresponding to a 12-channel system, has been used in calculations, while the transmitter power has been assumed as given by the manufacturer (10 watts). The system figure of merit is also given by the manufacturer in the equipment specification and can be calculated using equations 2.53 or 2.54.

Route (a) [Table 3-I]

Applying formula $50\sqrt{D}$ pW to route (a) a recommended mean-noise power of 623.5 pW is obtained, for a length of 155.5 kilometres, which must not be exceeded in any hour, or for more than 20% of any month.

On the other hand, the calculated noise power for the data given, with the severe assumption of 9 dB simultaneous fading in all hops, amounts to 225 pW in any channel. If 100 pW, due to modulator-demodulator and $70 \times 4 = 280$ pW due to amplifiers are added, an overall noise of 605 pW is obtained. This is inside the CCIR recommendation of 623.5 pW and corresponds to a signal-to-noise ratio

TABLE 3-I

ROUTE (a)

SIGNAL-TO-NOISE RATIO ON A 900 MHz RADIO-RELAY LINK COMPRISING THREE REPEATER STATIONS.

SYSTEM FIGURE OF MERIT 159. FEEDER LOSS 4.5dB/100m. 12 FEET PARABOLAS. GAIN 24.75 dB.

No.	Hop	Distance (km)	Free Space Losses (dB)	Antenna Heights (m)	Antenna Feeder's Length (m)	Feeder Losses (dB)	Antenna Gains (dB)	Path Losses (dB)	Signal-to-Noise Ratio (dB)	Noise Power (pW)
1	LA ₂	40	123.6	15 - 10	30 + 30	2.7	49.5	76.8	82.2	6
2	A ₂ B ₂	44.5	124.5	10 - 30	30 + 45	3.4	49.5	78.4	80.6	8.7
3	B ₂ W	44	124.4	25 - 40	40 + 45	4.3	49.5	79.2	79.8	10.5
4	WS	27	120.2	30 - 7	45 + 22	3.0	49.5	73.7	85.3	3
		<u>155.5</u>								28.2 8 x 225.6 ≈ 226 + 380 606 pW

$$50\sqrt{155.5} \approx 623.5$$

TABLE 3-II

ROUTE (b)

SIGNAL-TO-NOISE RATIO ON A 900 MHZ RADIO-RELAY LINK COMPRISING TWO REPEATER STATIONS.

SYSTEM FIGURE OF MERIT 159. FEEDER LOSS 4.5dB/100m. 12 FEET PARABOLAS. GAIN 24.75 dB.

No.	Hop	Distance (km)	Free Space Losses (dB)	Antenna Heights (m)	Antenna Feeder's Length (m)	Feeder Losses (dB)	Antenna Gains (dB)	Path Losses (dB)	Signal-to-Noise Ratio (dB)	Noise Power (pW)
1	LA ₃	57.6	126.8	15 - 78	30 + 90	5.4	49.5	82.7	76.3	23.5
2	A ₃ W	67.2	128.1	78 - 90	90 + 105	8.8	49.5	87.4	71.6	69.2
3	WS	27	120.2	30 - 7	45 + 22	3.0	49.5	73.7	85.3	3
		<u>151.8</u>								<u>95.7</u> x
										<u>8</u>
										765.6 +
										<u>380</u>
										1145.6 pW

$50\sqrt{151.8} \approx 616$

of 62.5 dB for the top channel.

Route (b) [Table 3-II]

Using the same procedure as in route (a) we obtain a recommended noise power of 616 pW, while that calculated for route (b) is $765.6 + 100 + 280 = 1145.6$ pW. An assumption of 9 dB simultaneous fading has also been made in these calculations.

The calculated noise is well outside that recommended by CCIR. This is in addition to a very high cost for developing station A_3 , as mentioned in describing section A_3W .

Therefore, the system using route (a) is recommended with 12 feet parabolas at all stations and 10 watts transmitter power.

CHAPTER FOUR

OBSTACLE DIFFRACTION INVESTIGATION

4.1. General

Over the last few decades many scientists have undertaken the study, both theoretically and experimentally, of the obstacle diffraction problem. For most of the recent experimental work a model using a laser source has been utilized.

Although there exist a few differences between the results of the various workers they are generally not very important. In fact, they only appear when the problem becomes increasingly complicated and major approximations have to be made.

Approximations are also made even for the simplest case*, but here the errors are minimal and certainly very small compared with experimental inaccuracies where such results are compared with theory. Considering the diffracting obstacle to be uniform in the direction transverse to the direction of propagation, in order to simplify the analysis, is one approximation. Assuming a perfectly conducting ground and ignoring ground reflections between the obstacle and the transmitting or the receiving points, introduces other kinds of approximations, which generally do not alter the situation significantly.

It has been stated⁽¹⁶⁾ that it is pointless to increase the body of diffraction knowledge by developing solutions for obstacles of completely arbitrary shapes, which are inapplicable to other even closely related configurations. It is important, however, to use other than the knife-edge diffraction solution in order to estimate the field diffracted by obstacles more commonly encountered in practice. This leads to the conclusion that a set of solutions which apply to basic idealized configurations is required. These solutions can then be extended to deal

* namely the knife-edge case.

with more complicated shapes. King⁽¹⁷⁾ suggested that such a basic set of diffracting obstacles should include the knife-edge, the cylinder, the wedge and the flat and cylindrical slabs. He has developed approximate solutions for diffraction by all these shapes and has compared them with existing rigorous solutions and experimental results.

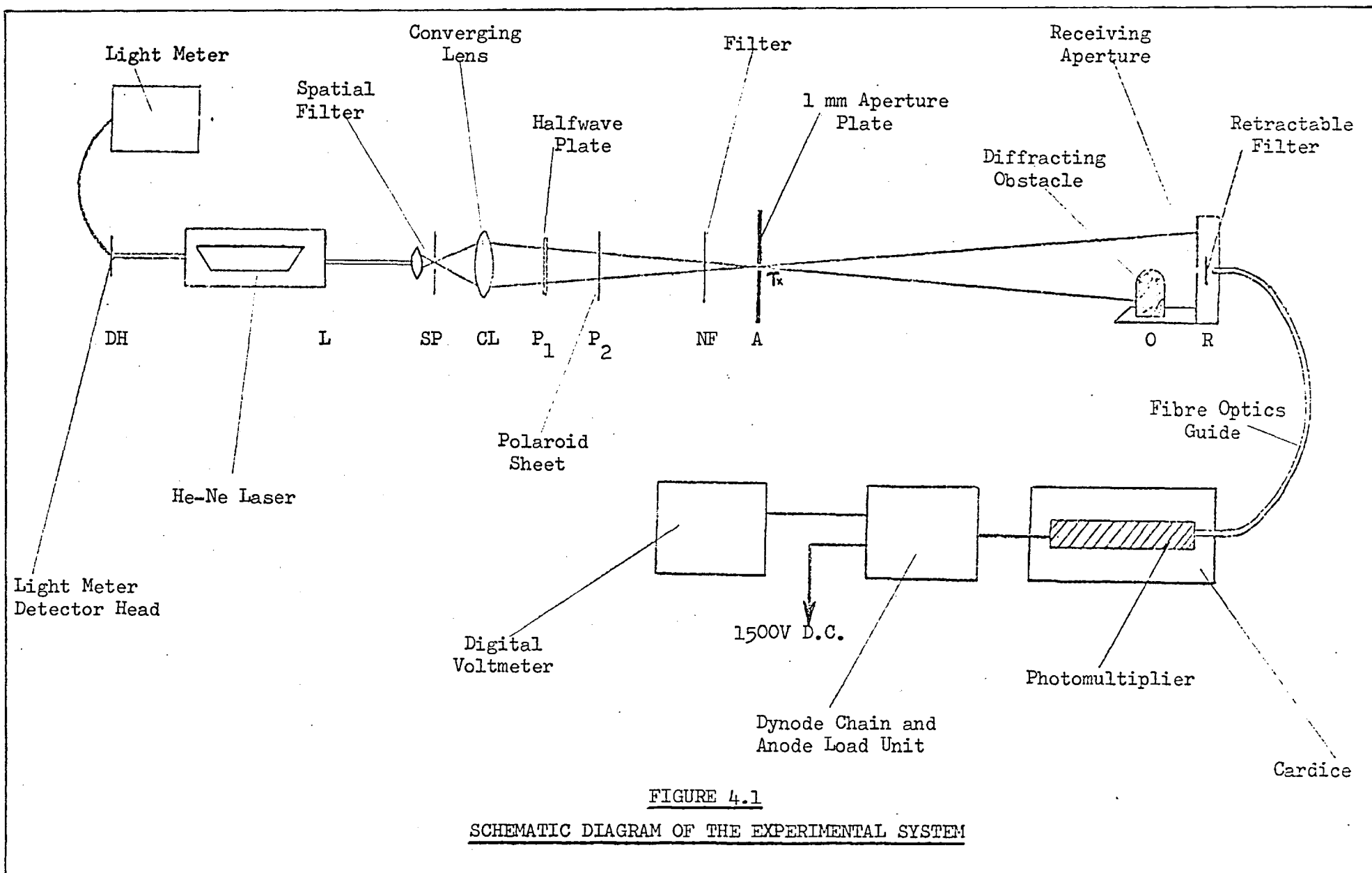
This chapter deals with knife-edge diffraction and diffraction by circular cylinders and compares theoretical solutions with experimental results obtained in the laboratory by means of a model using a laser source. Particular attention has been paid to the measurement of the field diffracted by cylinders with various roughnesses since these are cases most similar to the problems encountered when planning line-of-sight microwave links, including the effect of obstacle diffraction.

4.2. The Experimental System

It has been usual to utilize an optical path with a laser source when investigating diffraction phenomena and other aspects of propagation, particularly in the microwave region. This is due to the high degree of control required over these experiments and is the best known alternative to an actual situation.

The source used by some of the investigators was a helium-neon laser operating at a wavelength of 0.6328μ and was also used in the model to be described. The scale factor between the model and microwave frequencies is about 10^5 , which being very large, causes some difficulties when modelling obstacles. At optical frequencies machining tolerances and difficulties in measurement make it impossible to accurately model the surfaces of obstacles, rough or smooth.

The model is shown schematically in figure 4.1 and a photograph given in figure 4.2. A continuous single mode laser beam of about 8 mw is transmitted to the receiving aperture (R) after passing through



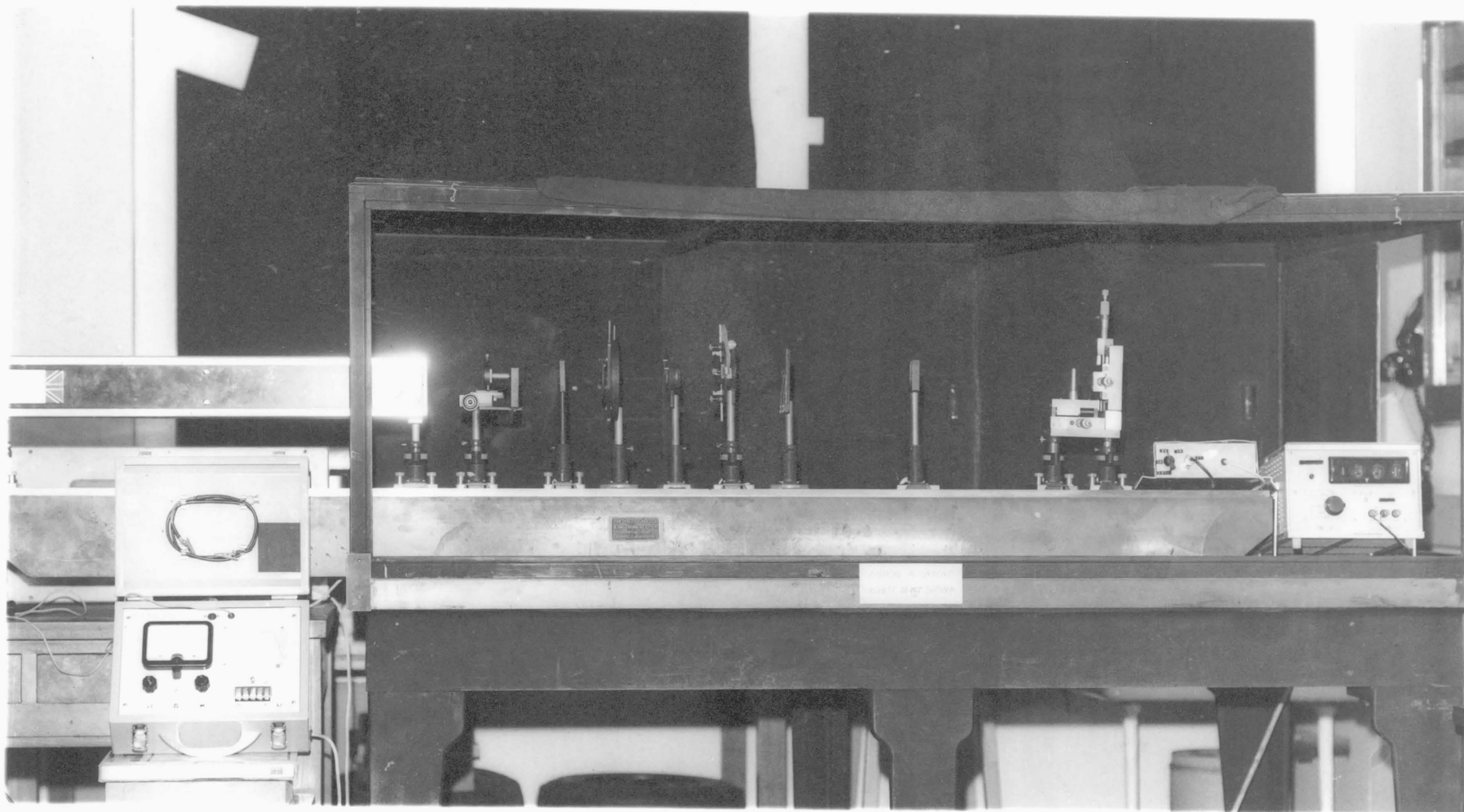


FIGURE 4.2

THE OPTICAL MODEL USED FOR DIFFRACTED FIELD MEASUREMENTS

various beam controlling components.

A spatial filter (SP) is used to focus the beam through a very small aperture (about 40λ) in order to reduce its bandwidth and eliminate any secondary beams arising from optical surfaces before (SP). The diverging beam is then converged and focussed to a point T_x by means of a converging lens (CL), which allows the point T_x to move along the bench by adjusting the relative positions of (SP) and (CL).

At point T_x a 1 mm aperture plate eliminates secondary beams arising from optical surfaces set between L and A and locates the virtual source of radiation i.e. the virtual transmitter. The laser beam itself is horizontally polarized but depending on the orientation of the obstacles we have either vertical or horizontal polarization relative to the obstacle. To avoid turning the obstacle, when measurements for both polarizations are needed, a sheet of birefringent mica (halfwave plate) between two glass plate (P_1), together with a polaroid sheet (P_2) is used to rotate the plane of polarization to any desired angle.

For high light levels, attenuation of the beam power may be achieved by inserting neutral-density filters with various transmission coefficients between the polaroid sheet and the 1 mm aperture plate as at (NF) in figure 4.1.

The diffracting obstacle (O) and the scanning receiving aperture (R) are mounted on a combined obstacle and receiver unit, which is fixed relative to the laser source. The obstacle and the receiving aperture can be moved independently in all required directions and every movement is controlled and monitored by standard micrometric heads with a resolution of 10μ .

The receiving aperture diameter is about 100λ , whereas the diameter of an actual microwave antenna is about 20λ . However, there is no consequent

significant error in the amplitude of the measured field and this would be true even for aperture diameters up to about 300λ . In the deep shadow even larger apertures are permissible without significant error (17).

Behind the receiving aperture a neutral density filter of 22 dB attenuation may be introduced or retracted by a lever when such an attenuation is needed.

The obstacle and receiving aperture unit together with the micrometer heads, a mounted cylinder as obstacle and other details are shown in figure 4.3.

The light received by the small aperture is carried to a photomultiplier tube by means of a fibre-optics guide mounted behind the receiving aperture or the retractable filter when in use.

The photomultiplier tube itself is contained in a thermally isolated box and supplied with 1500 v d.c. through a dynode chain and anode load unit. Five load resistors are available, (1 K Ω , 5 K Ω , 10 K Ω , 100 K Ω , 1 M Ω with a 1% tolerance) for use with various light levels at the receiving aperture. For normal work the most suitable value is 5 K Ω while the higher ones are used when low light levels require increased meter sensitivity. The voltage across the load resistor is directly proportional to the received light intensity, when working in the photomultiplier's linear range, and is measured by a digital voltmeter. When other than the 5 K Ω load resistor is used the measured voltage is translated to the value that would appear if a 5 K Ω load were used.

To reduce the noise current of the photomultiplier, particularly when only low signal currents are obtainable, the housing box is filled with cardice (solid CO₂) which reduces the temperature and subsequently the dark current of the photomultiplier.

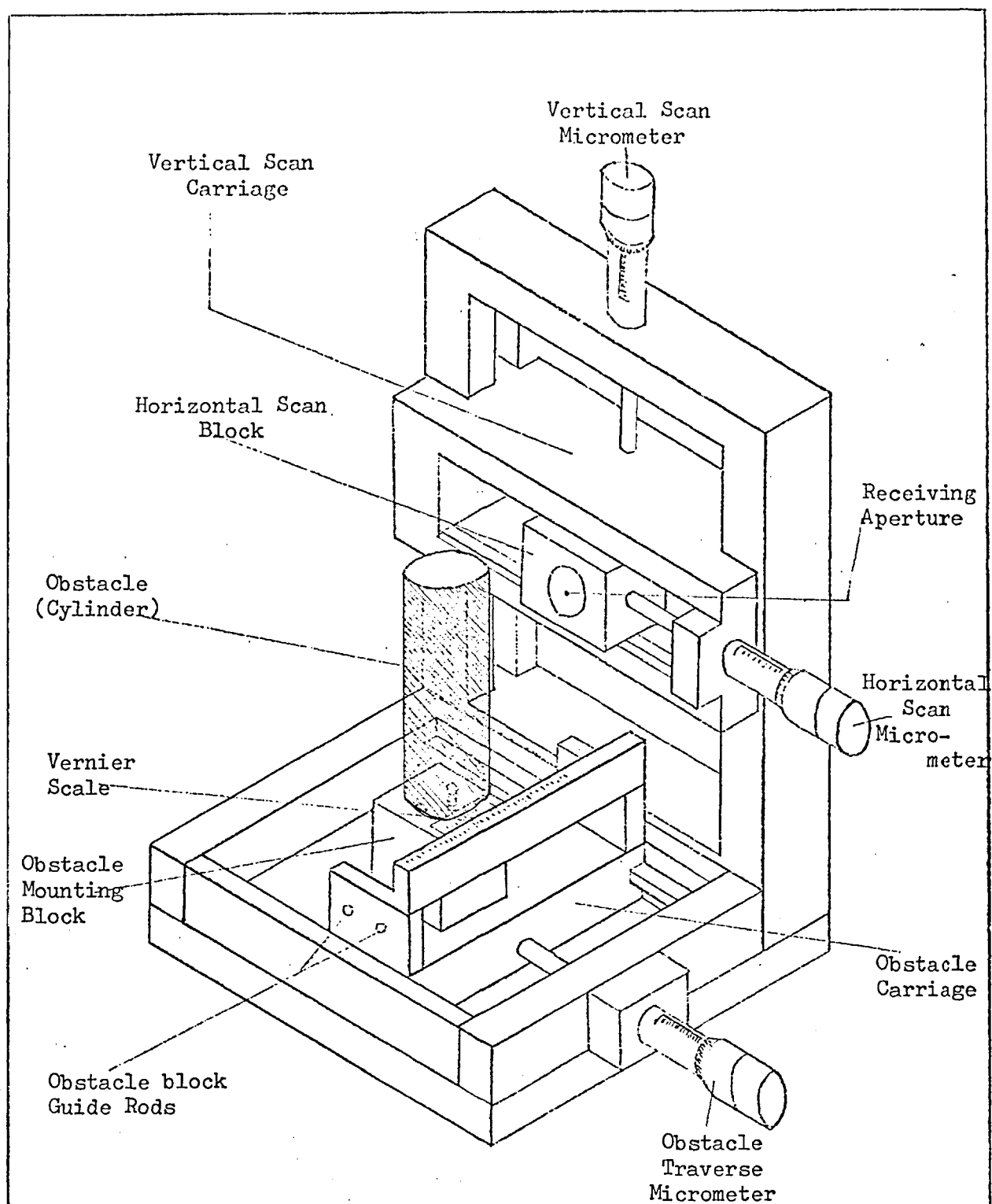


FIGURE 4.3

OBSTACLE AND RECEIVING UNIT

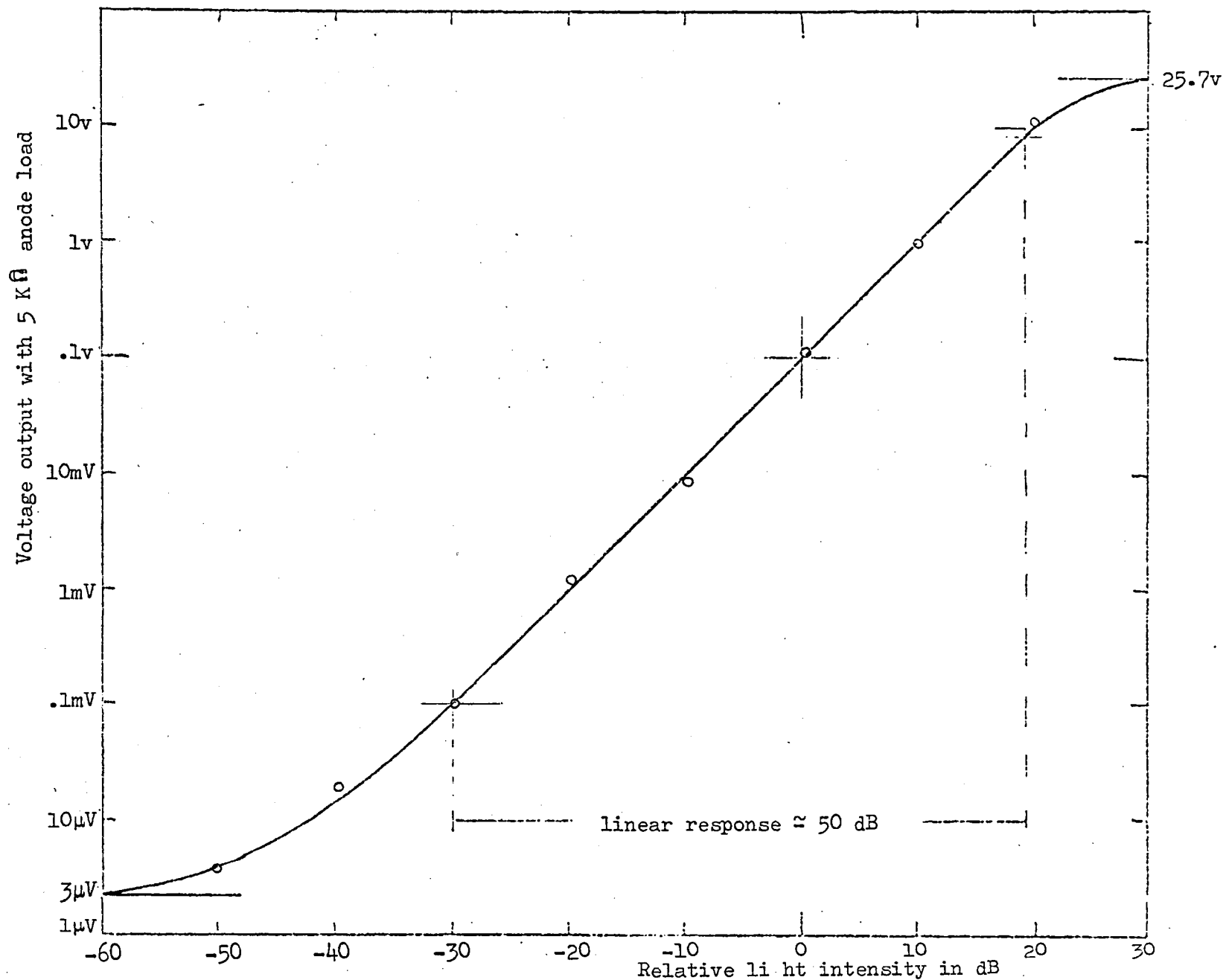


FIGURE 4.4
PHOTOMULTIPLIER CALIBRATION CURVE

A calibration curve of the photomultiplier tube, given in figure 4.4, was obtained by inserting neutral density filters of known transmission in position (NF) of figure 4.1 and comparing the measured output voltage to the expected value due to the filter attenuation. Depending on the filters available and their possible combinations the extent of the linear part can be found to a sufficient degree of accuracy. Besides the manufacturer's data given for the filters, their transmission was tested by means of a light meter prior to the calibration measurements. An output of about 100 mV with a $5\text{ K}\Omega$ load was taken as the starting point, obtained by a combination of filters in front of the transmitting aperture A. Then the light intensity relative to this was calculated by the transmission of a range of filters added to or subtracted from the combination.* It is estimated that a linear operating range of about 50 dB of the photomultiplier, extended to about 70 dB by means of the retractable filter behind the receiving aperture, can be used for measuring the diffracted field which in any case is negligible when attenuated by more than 60 dB from its unobstructed value.

Besides receiving aperture limitations, mentioned above, the beam taper and width as well as its directivity have to be considered when modelling a path, if measurements are to be compared with theoretical predictions based on isotropically radiating antennas. Budin⁽¹⁸⁾ and King⁽¹⁹⁾ have investigated diffraction by a directional Gaussian beam. It has been shown that for a large range of possible combinations of beam amplitude taper and beam width in the plane of the obstacle the error involved in the measurements, as compared with the theoretical predictions, is very small.

Finally, to avoid errors due to the instability of the laser source, its output is monitored by means of a light meter whose detector head (DH in figure 4.1) is aligned with the rear output of the laser.

* Finally, the input (light intensity) - output (voltage) characteristic, as in Figure 4.4 was plotted.

4.3. Knife-Edge Diffraction

4.3.1. Theoretical Aspects

Figure 4.5 illustrates a simple example of knife-edge diffraction. To keep the problem two-dimensional the transmitting source T is assumed to be an infinite line source parallel to an infinite straight line which defines the crest of a ridge whose sides fall steeply downward on both sides of the crest.

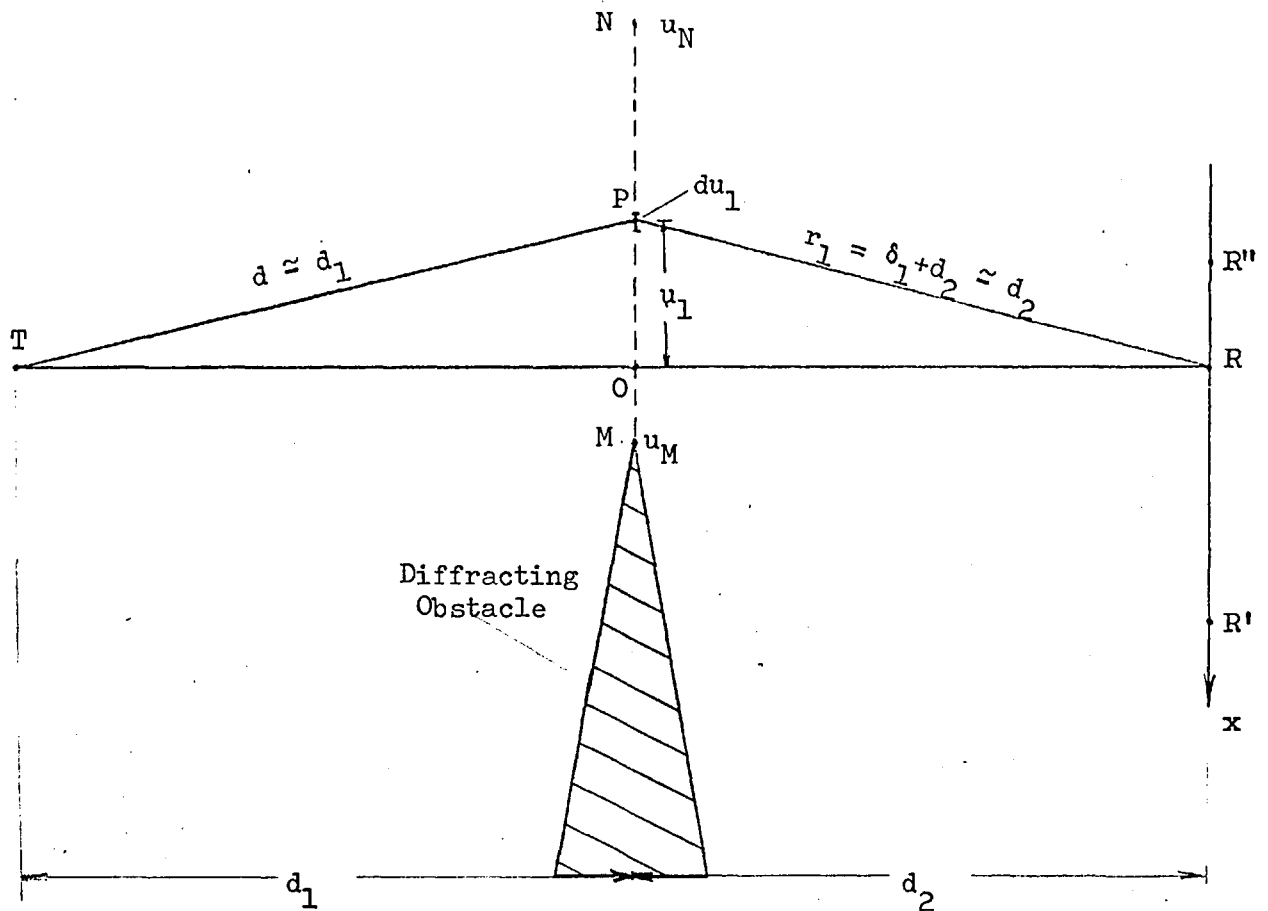


FIGURE 4.5.

KNIFE-EDGE DIFFRACTION (CLASSICAL APPROACH)

The application of Huygens' principle, mentioned in chapter two, to the aperture MN above the diffracting obstacle is the classical approach to diffraction problems in general. It is assumed here that the field in this aperture is unperturbed by the presence of the diffracting

obstacle. In making this approximation the error introduced when estimating the diffracted field is very small. This error is negligible when considering the illuminated region field or the shadow field for small diffraction angles, because the perturbed Huygens' sources, which extend only a few wavelengths above the crest, contribute to the distant field only by a very small amount.

Assuming the ideal case of a perfectly-absorbing knife-edge as the diffracting obstacle in figure 4.5, the field at R can be regarded as equivalent to the radiation from a set of Huygens' sources spread over the aperture above M. In this case an elemental source du of the wave front contributes to the field at R by a quantity:

$$dE = \frac{A(u)}{f(r)} e^{-jkr} du \quad (4.1)$$

where

k is the phase factor equal to $2\pi/\lambda$

r is the distance to the receiving point R and

$A(u)$ is the amplitude of the incident wave at P

which, for $d_1, d_2 \gg \lambda$, may be approximated by:

$$A(u) = A_0 e^{-j \frac{ku^2}{2d_1}} \quad (4.2)$$

if A_0 is the amplitude of the incident wave at O.

The amplitude of the signal reaching point R may be regarded as independent of the variations of the distance r over the aperture plane since, according to classical optical theory, only a small part of this plane contributes to the received signal. In this case the distance r from all elemental sources to the receiving point R may be approximated by d_2 . However, the same does not apply to the individual relative phases

which must be taken into account.

Comparing the line-of-sight path with that through source du_1 we have a difference δ_1 of:

$$\delta_1 = r_1 - d_2 \approx u_1^2 \frac{d_1 + d_2}{d_1 d_2} \quad (4.3)$$

Replacing k by its equivalent $\frac{2\pi}{\lambda}$ and multiplying by δ_1 gives:

$$k\delta_1 = \frac{2\pi}{\lambda} u_1^2 \frac{d_1 + d_2}{d_1 d_2} = \frac{\pi}{2} u_1^2 \frac{2(d_1 + d_2)}{\lambda d_1 d_2} \quad (4.4)$$

and making the substitution:

$$v_1 = u_1 \sqrt{\frac{2(d_1 + d_2)}{\lambda d_1 d_2}} \quad (4.5)$$

we obtain:

$$k\delta_1 = \frac{\pi}{2} v_1^2 \quad (4.4a)$$

The quantity v_1 is usually described as the Fresnel parameter.

After the above substitutions and approximations the total field at R , due to a portion MN of the aperture above the crest, can be found by integrating equation 4.1 over the region MN :

$$E = \frac{A(u)}{f(r)} \int_{u_M}^{u_N} e^{-jkr} du = \frac{A(u)e^{-jkd_2}}{f(d_2)} \int_{u_M}^{u_N} e^{-j\frac{\pi}{2}v^2} du$$

(4.6)

This can be evaluated (see Jordan⁽⁸⁾ for example) and expressed relative to the free space field (E_0) to give the well established

knife-edge attenuation:

$$\frac{E}{E_0} = \frac{1+j}{2} \int_v^{\infty} e^{-j \frac{\pi}{2} v^2} dv \quad (4.7)$$

where

$$v = u \sqrt{\frac{2(d_1 + d_2)}{\lambda d_1 d_2}} \quad (4.7a)$$

The integral in equation 4.7 is the Fresnel Integral and is often evaluated by a construction known as Cornu's spiral. The magnitude of $\frac{E}{E_0}$ is plotted in figure 4.6. It is shown that the loss for a path which just grazes the top of the knife-edge is 6 dB. The field in the shadow decreases as the parameter v increases and can be regarded as arising only from a radiating line source situated along the crest of the diffracting obstacle, while the ripple in the illuminated region can be regarded as the result of the interference between the direct ray and the radiation from that crest line source. The Fresnel parameter v is positive when the knife-edge is above the line-of-sight path, and negative when it is below.

More rigorous treatments of the diffracted field have been attempted by many authors using different methods and more rigorous solutions, although with some degree of approximation, have been obtained. Derivation of the field diffracted by a perfectly-conducting knife-edge, for both vertical and horizontal polarisation, is also possible by application of those methods and it can be shown that this field behaves similarly to that diffracted by a perfectly-absorbing knife-edge.

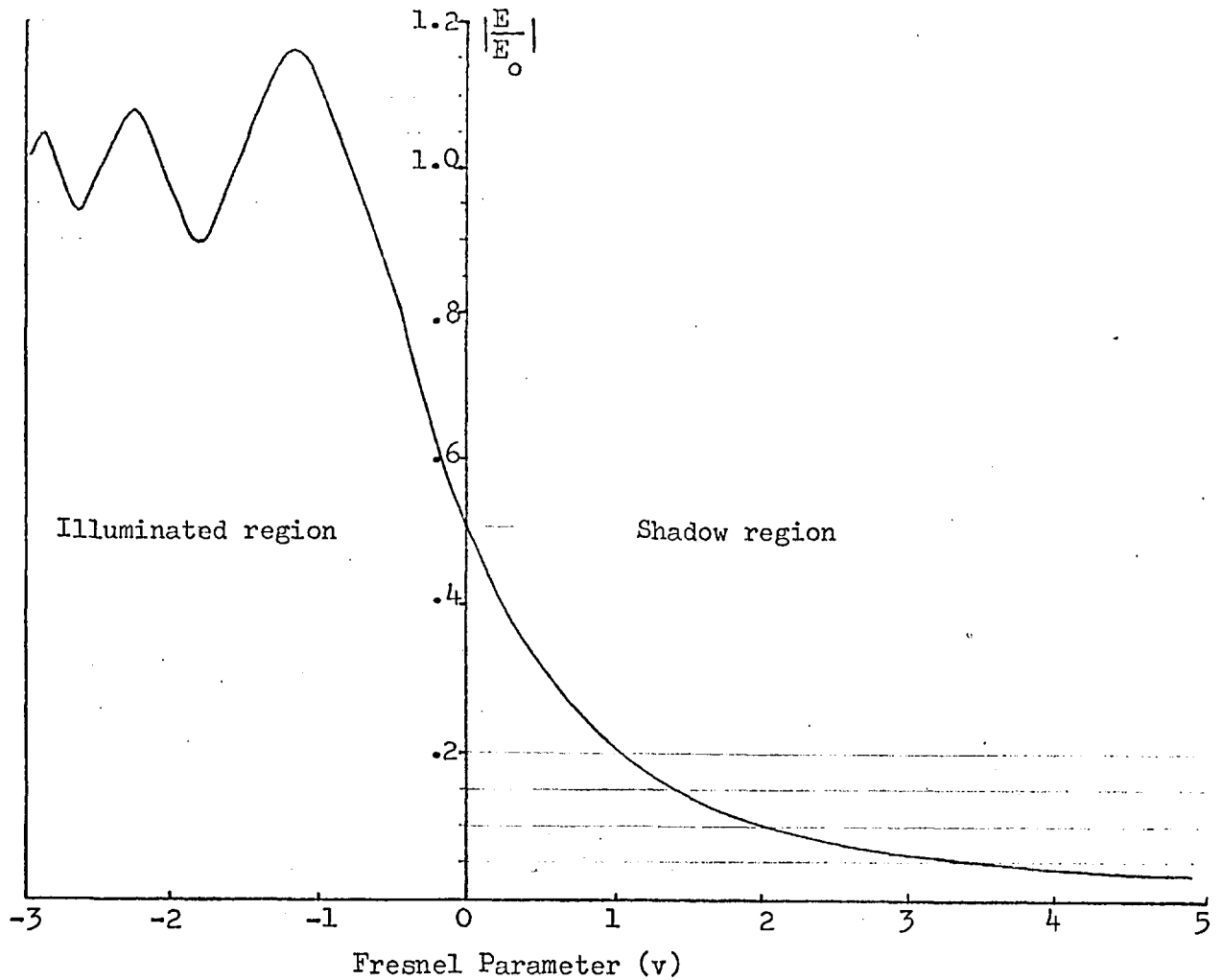


FIGURE 4.6

RELATIVE DIFFRACTED FIELD STRENGTH (CLASSICAL SOLUTION)

King⁽¹⁷⁾, using the plane wave angular spectrum representation of Clemmow⁽²⁰⁾, has obtained rigorous solutions for perfectly-absorbing and perfectly-conducting knife-edges, expressed in cylindrical coordinates. For the geometry shown in Figure 4.7 and for a plane wave, normally incident upon the diffracting knife-edge, the diffracted field $[E(r, \phi)]$ at R_1 , relative to the unobstructed one (E_0) at R_0 , is given by:

- a) for a perfectly-absorbing knife-edge and for either

where

$$v_1 = \sqrt{\frac{8r}{\lambda}} \sin \frac{1}{2} \phi = v_0 \quad (4.11)$$

$$v_2 = \sqrt{\frac{8r}{\lambda}} \cos \frac{1}{2} \phi \approx \sqrt{\frac{8r}{\lambda}} \quad (4.12)$$

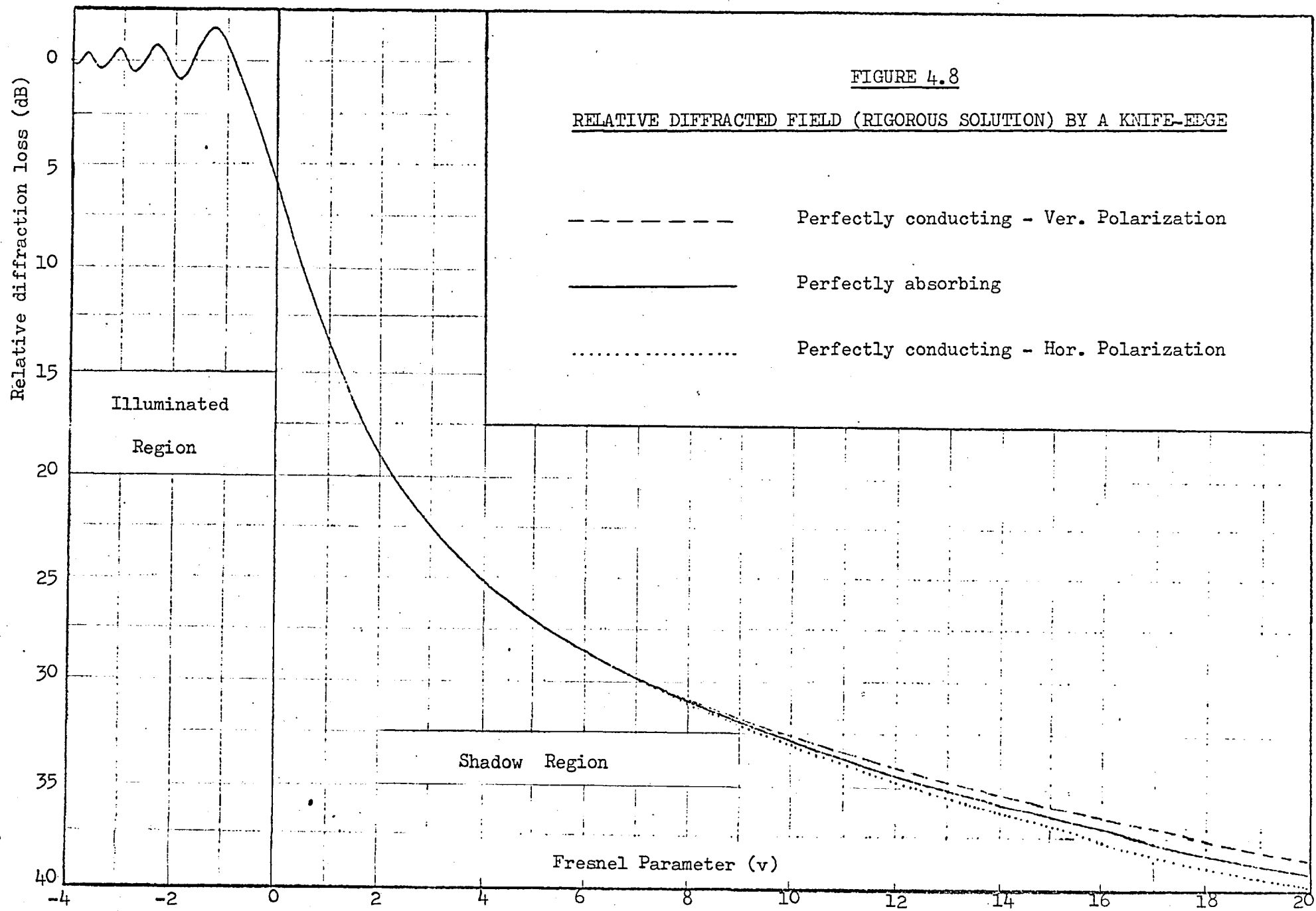
and the function:

$$G(v) = \exp \left\{ + j \frac{\pi}{2} v^2 \right\} F(v) \quad (4.13)$$

if $F(v)$ is the Fresnel Integral, as in equation 4.7.

The term $G(v_2)$ appearing in both perfectly-conducting cases is very small compared to $G(v_1) = G(v_0)$ for small diffraction angles ($\phi < 12^\circ$ say) and may be regarded as a second order term modifying the perfectly-absorbing knife-edge result in the case of perfectly-conducting edges.

Figure 4.8 illustrates graphically the diffracted field (relative to the unobstructed field at R_0) in all three cases described by equations 4.7 to 4.9. It is seen that even for relatively large diffraction angles (ϕ), corresponding to large values of (v), the difference between the fields for the above cases is very small. This difference decreases as the parameter (v) decreases. For very small or negative diffraction angles the term $G(v_2)$ becomes so small compared to $G(v_1)$ that its effect is practically zero. For large values of the Fresnel parameter, corresponding to the so-called "deep shadow" region, the diffracted field for the vertically polarized waves is greater than the field given by horizontally polarized waves. This result holds in all practical diffraction situations.

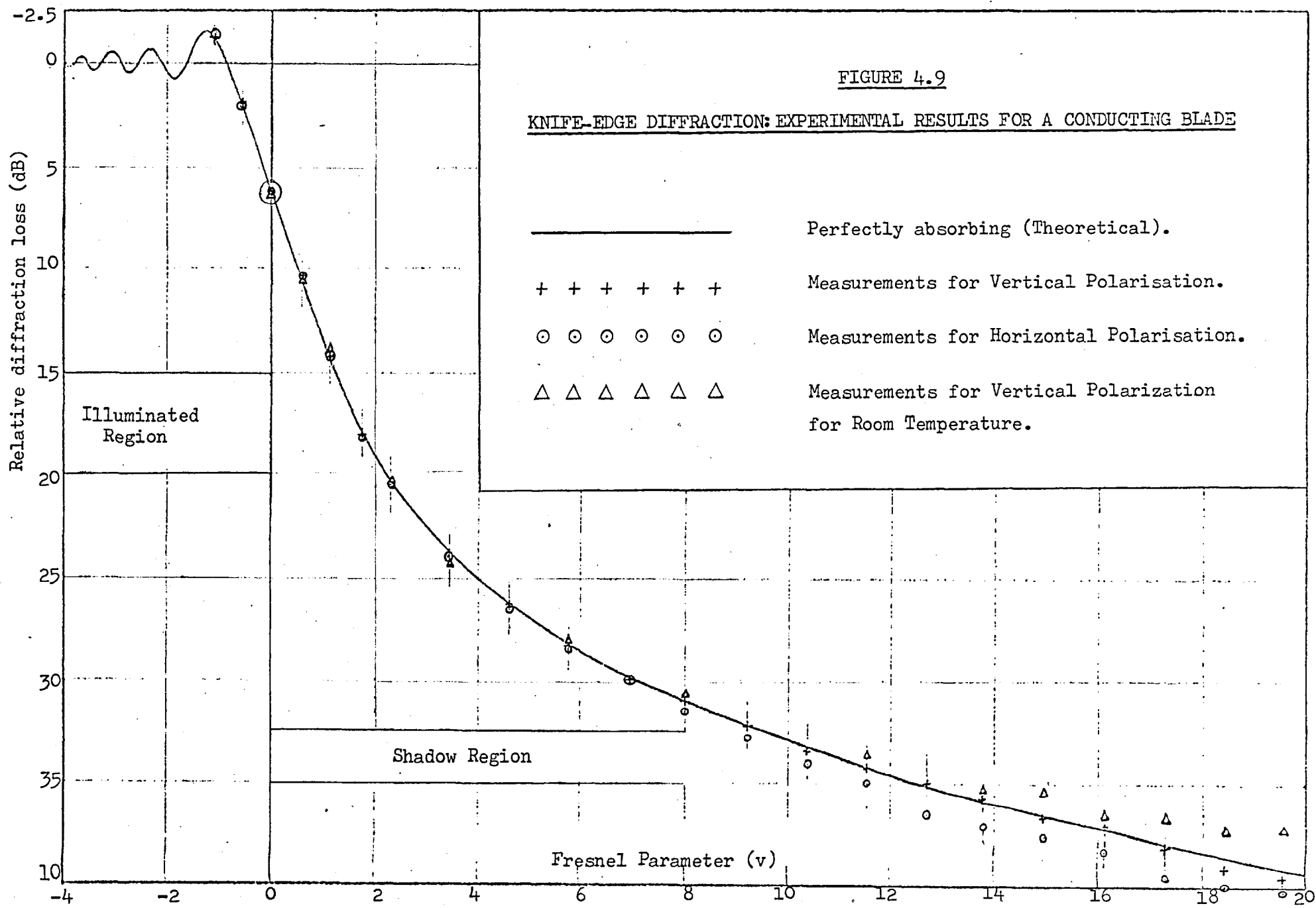


4.3.2. Experimental Results

Using the experimental system described in section 4.2, a series of measurements have been carried out and the results are given in figure 4.9. Before making any measurements all the components of the system were aligned so as to give a beam as parallel as possible to the bench. Neutral density filters were inserted as required to maintain the light level within the linear part of the photomultiplier characteristic, given in Figure 4.4, while the housing of the photomultiplier was filled with cardice. The 1mm aperture plate, locating the virtual source, was at a distance of 64 cm from the receiving aperture on the plane of which a beam, a little larger than 1 cm in diameter, was cast.

A new razor blade, darkened below its edge to avoid reflections, simulated the knife-edge. According to manufacturers data the curvature of the blade edge is about $1500 \text{ \AA}$ (Gillete laboratories), which corresponds to about $1/4$ of the laser wavelength, small enough for the blade to be practically a knife-edge. It is, however, reckoned that preparing the blade for the experiments (darkening, mounting etc.) had resulted in some bluntness of its edge; it is estimated that a curvature of about 4 - 6 wavelengths is a good approximation. The blade was at a distance of 23 mm from the receiving aperture with its edge vertical and normal to the beam. It was set on the beam axis by traversing it across the path until the output voltage (V_{ϕ}) was reduced to $1/4$ of the unobstructed axial value (V_0) as shown by theory (figure 4.6).

By scanning the receiving aperture across the path and normal to the blade edge the output voltage (V_{ϕ}) for various diffraction angles (ϕ) was recorded. These voltages, corresponding to intensities of the diffracted field for these particular angles (ϕ), were expressed relative to the unobstructed axial value (V_0), in decibels,



through the formula:

$$P(\phi) = 10 \log \frac{V_\phi}{V_0} \text{ dB} \quad (4.14)$$

The continuous curve in figure 4.9 is the theoretical perfectly absorbing knife-edge result, as given in figure 4.6, also in decibels. It is seen that for small diffraction angles the measurements for both kinds of polarization almost coincide with each other and with the perfectly-absorbing result, as expected from theory. As the diffraction angle increases the expected effect of polarization becomes apparent as the mean of the values obtained for the horizontally polarized waves lies below the mean of the values obtained when the waves were vertically polarized. The fact that the results do not lie symmetrically on both sides of the absorbing theoretical curve is mainly due to the finite curvature of the edge of the blade, as mentioned before. Comparison between theory (figure 4.8) and experiment (figure 4.9) indicate that apart from some experimental discrepancies and the finite curvature effect, the results are in very good agreement, which in turn indicates that the metal from which the razor blade is made is a very good conductor.

Some measurements have also been made without any cardice in the housing of the photomultiplier. The results are also given in the same figure 4.9 and are for room temperature and vertically polarized waves. The dark current effect is pronounced in the deep shadow region where the actual field is comparatively small.

4.4. Diffraction By Smooth Curved Obstacles

4.4.1. Theoretical Aspects

Although the distant diffracted field at points close to the

geometrical shadow is almost independent of the curvature of the diffracting crest and the properties of its surface, it can be demonstrated that the same does not apply for the field diffracted deeper into the shadow. Curvature and surface properties of the crest greatly modify the previously studied field diffracted by a knife-edge.

Huygens' principle may again be applied when investigating the problem of diffraction by rounded obstacles. Figure 4.10 illustrates the geometry of the problem.

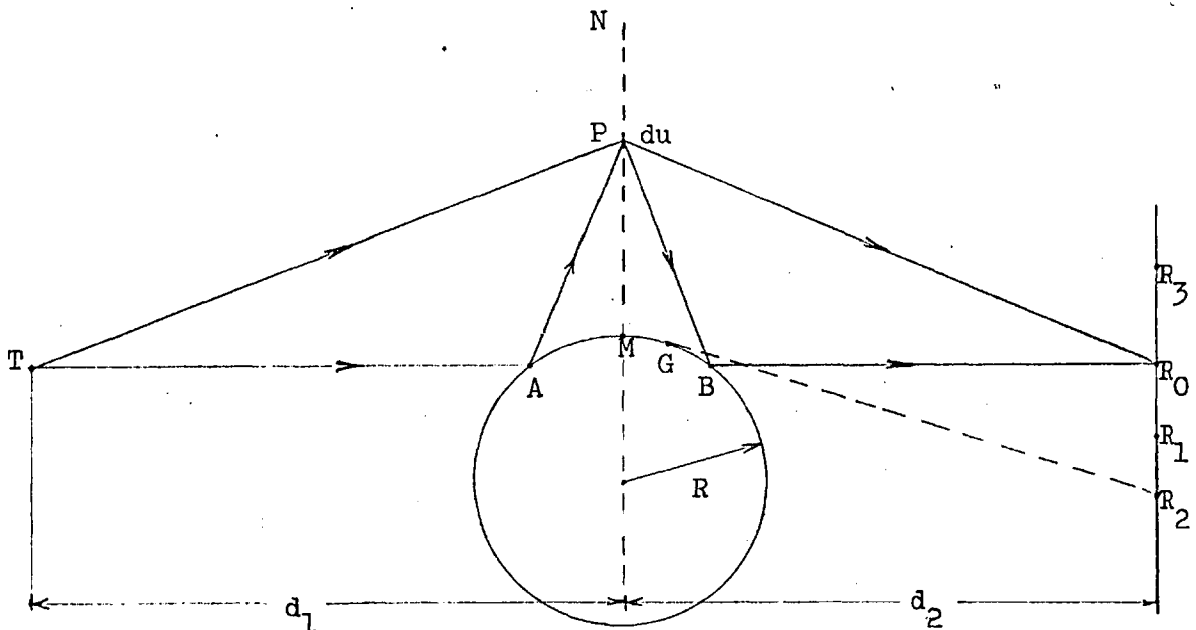


FIGURE 4.10

DIFFRACTION BY ROUNDED OBSTACLES

As in section 4.3.1 the field at R_0 may be calculated by assuming elemental sources over the aperture MN above the obstacle. However, these sources cannot be assumed unobstructed in this case because specular reflection from the curved surface to the left of M contributes to the field occupying the aperture of the hypothetical Huygens' sources. This contribution modifies the field due to the direct ray for many wavelengths above M and some Huygens' sources will be of zero magnitude if direct and reflected rays, undergoing phase reversal, are

of the same magnitude. This destructive interference will reduce the field in the aperture above M as compared to the knife-edge case and consequently, less field will appear at R_0 .

Moreover, the field at R_0 due to an elemental source of the aperture MN is the result of a similar mechanism, since specular reflection now takes place to the right of M which affects the situation at R_0 . Destructive interference will also be expected at the receiving aperture, the received field being further reduced.

The theory of diffraction by smooth curved surfaces has been widely treated and various solutions for the shadow field strength have been given for many obstacle configurations. A method based on Huygens' sources, produced and acting according to the mechanism described above, was used by Shkarofsky et al. ⁽²¹⁾. The field at R_0 in figure 4.10 is calculated as the result of four rays TPR_0 , $TAPR_0$, $TAPBR_0$ and $TPBR_0$ integrated over the aperture MN. Each ray can be treated separately to give a term containing a single Fresnel Integral of the form in equation 4.7.

Wait and Conda ⁽²²⁾, using a smooth circular cylinder as their mathematical model, have given a solution in an engineering form for cylindrical obstacles with a curvature much larger than the wavelength.

An engineering form for diffraction by a sphere was given by Domb and Pryce ⁽²³⁾ and can also be applied to obstacles of cylindrical shape. Cylinders of parabolic profile were considered by Rice ⁽²⁴⁾ about two decades ago.

A numerical comparison of the above results has been made by Hacking ⁽²⁵⁾ using a symmetrical arrangement of transmitter, obstacle and receiver ($d_1 = d_2$ in figure 4.10) and assuming that the diffraction angle is small (e.g. $< 12^\circ$). Figure 4.11 illustrates this comparison for horizontal polarization and for values of parameters such as:

$$\frac{d_1 d_2}{(d_1 + d_2) \left(\frac{R^2 \lambda}{\pi} \right)^{1/3}} = 25 \quad (4.15)$$

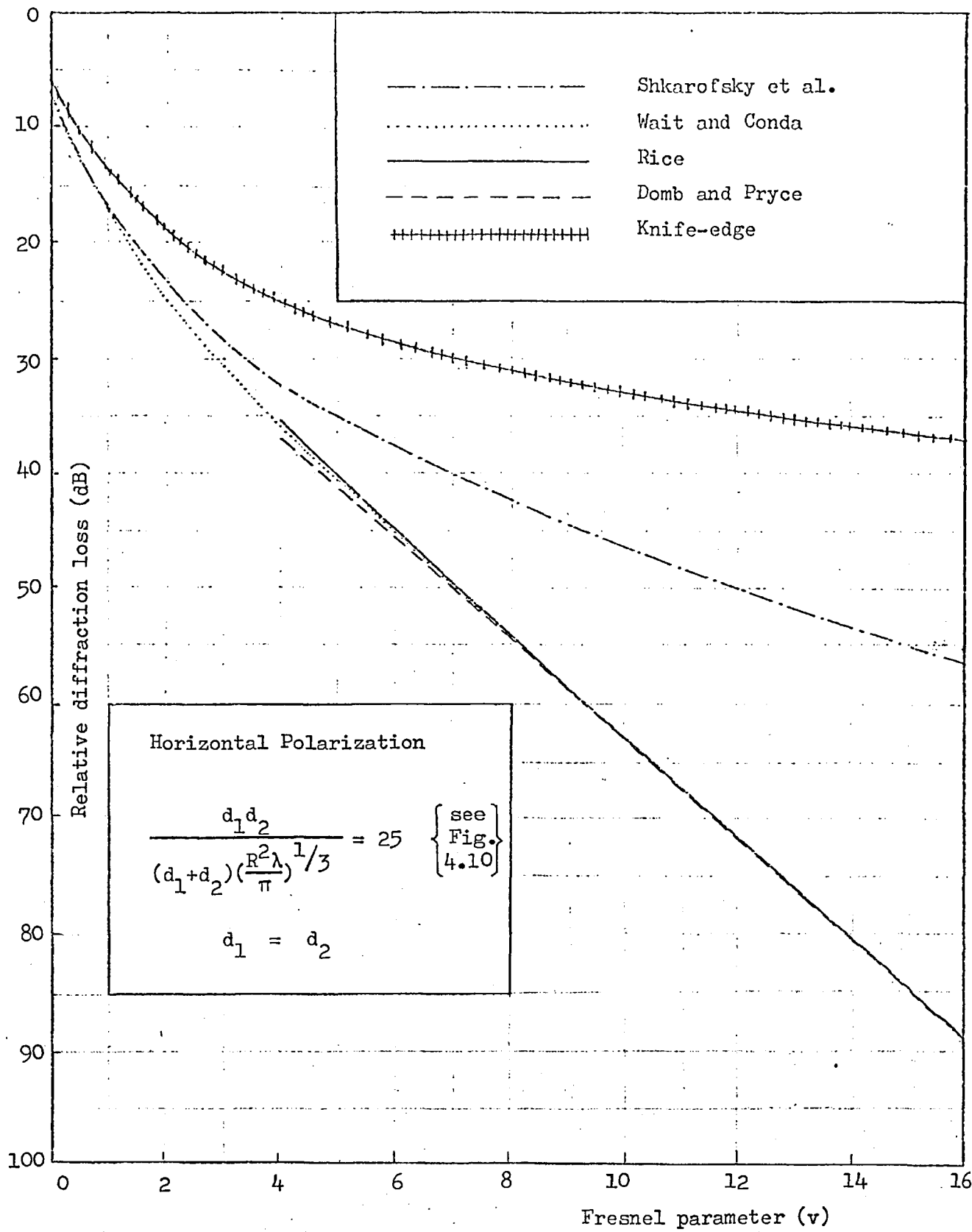


FIGURE 4.11

COMPARISON OF DIFFRACTION LOSS THEORETICAL SOLUTIONS
FOR CURVED OBSTACLES

It can be seen that, except for very small diffraction angles, the solution of Shkarofsky et al.⁽²¹⁾ considerably underestimates the diffraction loss obtained from the rest of the solutions which are in very good agreement. This underestimation, as Hacking suggested, is due to approximations involved in evaluating the Fresnel integrals. On the other hand, as Rice⁽²⁴⁾ has demonstrated, the "deep shadow" field is closely related to the distribution of surface current around the point (G) of the tangent to the cylinder from the receiver. If this is true, then the method of Shkarofsky et al., which does not take this current distribution into account precisely, is inadequate for comparatively large diffraction angles and consequently the underestimation of the loss may also be due to this inadequacy.

The knife-edge diffraction loss is also included in figure 4.11 for comparison. It can be seen that, as indicated by employing the Huygens' principle, the shadow region field diffracted by curved obstacles is considerably reduced compared with that which would appear if the curved obstacle were replaced by a knife-edge, the geometry remaining the same.

Hacking⁽²⁵⁾ has shown that the field diffracted by a circular cylinder can be expressed by means of the loss introduced by a knife-edge at the same position multiplied by a factor termed "excess loss" the magnitude of which is a function of the cylinder curvature (R) and path dimensions, assuming constant wavelength. This "excess-loss" can be derived from figure 4.11 and is given in figure 4.12 as a function of a parameter (γ) which is proportional to the cube root of the cylinder curvature, assuming constant wavelength and diffraction angle (ϕ), and given by:

$$\gamma = \left(\frac{\pi R}{\lambda} \right)^{\frac{1}{3}} \phi \quad (4.16)$$

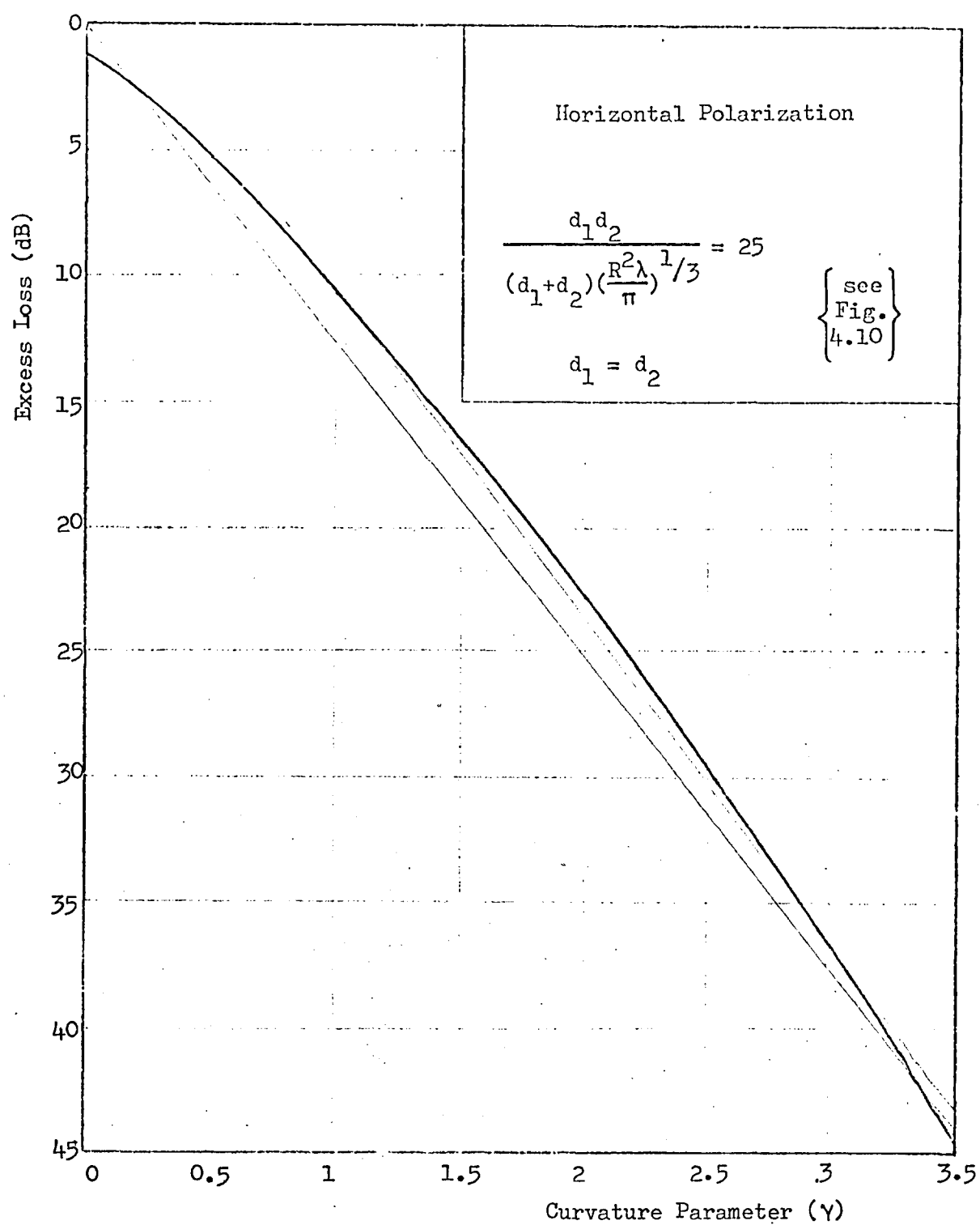


FIGURE 4.12

THEORETICAL EXCESS LOSS BY CURVED OBSTACLES

As shown in figure 4.12, the excess-loss curve is approximately straight except for very small diffraction angles or curvatures when in any case the excess loss itself is very small.

In terms of figure 4.11 and the "curvature parameter" (γ) the field E_R , diffracted by a rounded surface, relative to the unobstructed field (E_0) at R_0 in figure 4.10, is given by:

$$\frac{E_R}{E_0} = \sqrt{\frac{j}{2}} F(v) G(\gamma) \quad (4.17)$$

where v is the Fresnel parameter

$F(v)$ is the Fresnel integral

$G(\gamma)$ is the excess-loss factor, given by means of figure 4.12 through the formula:

$$G(\gamma) = \text{Antilog} \frac{\text{Excess-loss (dB)}}{20} \quad (4.18)$$

Although figures 4.11 and 4.12 refer to a particular combination of obstacle curvature and path dimensions, corresponding to a particular value (25) of the quantity given in equation 4.15, it holds sufficiently well as long as this quantity has a value greater than 10. For values below 10 a "close-distance correction" must be added to the excess loss. This correction is inversely proportional to both the quantity in equation 4.15 and the parameter (γ).

A further correction, which may be necessary when vertically polarized waves are investigated, is due to the finite permittivity of the diffracting crest, because the results in figures 2.11 and 2.12 refer to horizontal polarization for which the permittivity has little influence on the diffracted field strength. For a permittivity of large modulus a correction of about 1-3 dB must be subtracted from the excess-loss

when the radiation is vertically polarized. This correction is directly proportional to the diffraction angle and the radius of curvature.

King⁽¹⁷⁾ has shown that a segment of a cylinder gives the same loss as a complete cylinder provided the angle subtended at the axis of the cylinder is at least equal to α , where α is related to diffraction angle ϕ by:

$$\alpha = \phi + 2\left(\frac{4\lambda}{R}\right)^{\frac{1}{3}} \quad (4.19)$$

if R is the radius of the cylinder and λ is the wavelength of the diffracted beam. The position of the slab must be such that the bisector of angle (α) bisects also the angle subtended by the surface between the contact points of the tangent rays from the transmitter and the receiver to the obstacle.

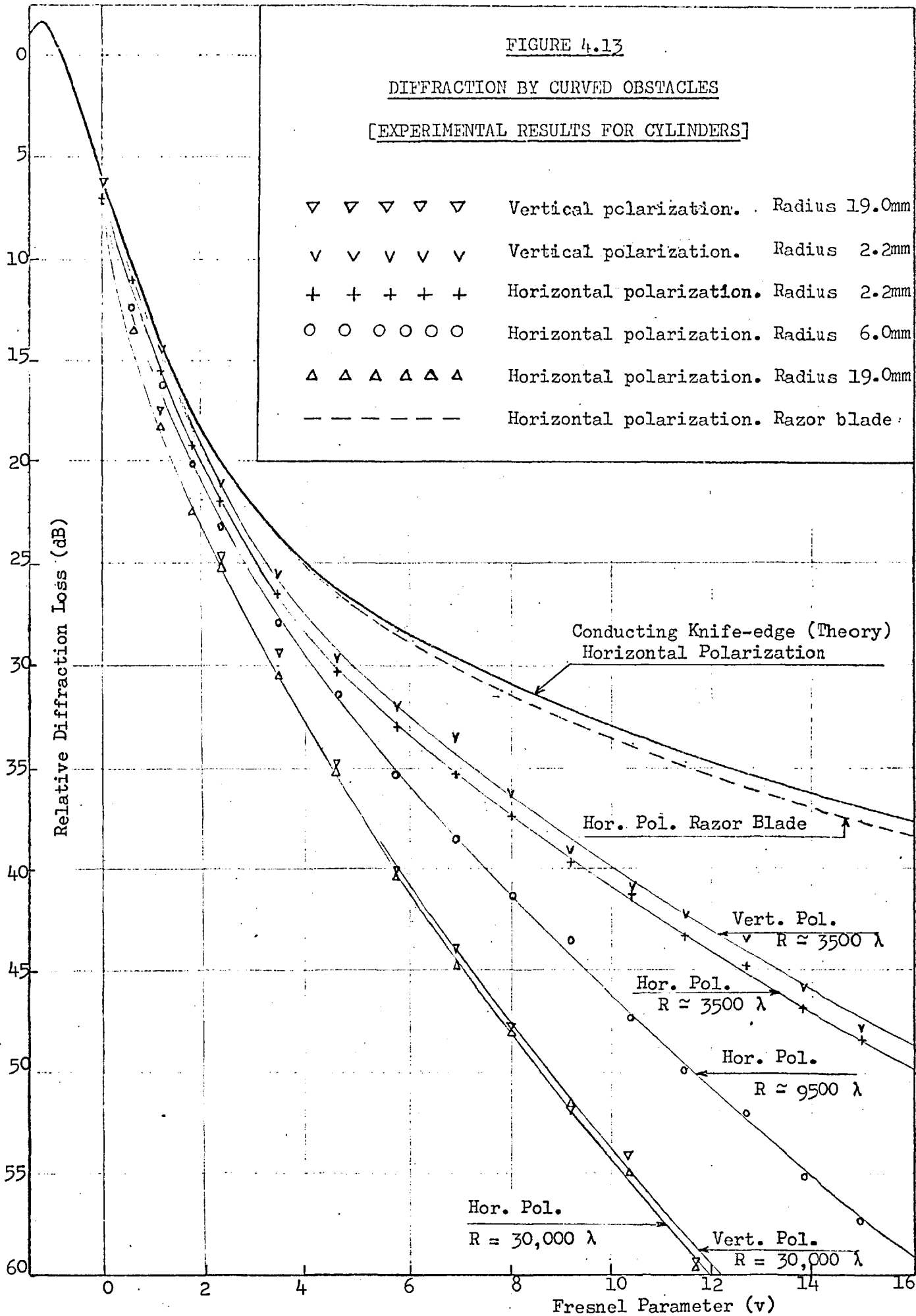
4.4.2. Experimental Results

The technique described in section 4.3.2. was used to carry out measurements using cylinders as diffracting obstacles whose radii of curvature were 19 mm ($\approx 30,000 \lambda$), 6 mm ($\approx 9500 \lambda$) and 2.2 mm ($\approx 3500 \lambda$). The geometry was the same as for the knife-edge measurements. The results obtained are given in figure 4.13. The theoretical result that, for large radius of curvature and horizontally polarized incident waves, the diffraction loss at grazing angles is about 1 dB greater than that of a knife-edge case (see figure 4.11) was taken into account; the larger cylinder ($r = 19$ mm) was moved across the path until the output voltage (V_{ϕ}) was reduced to $1/5$ of the unobstructed axial value (V_0). In between 6 dB and 7 dB values for axial loss were set when the other two cylinders were used. For

FIGURE 4.13

DIFFRACTION BY CURVED OBSTACLES

[EXPERIMENTAL RESULTS FOR CYLINDERS]



vertically polarized incidence the diffraction loss, at grazing angles, was obtained automatically when the polarization was changed from horizontal to vertical by rotating the polaroid sheet by 90° and the half wave plate for maximum output. It was almost equal to that in the case of the knife-edge in all three cases (i.e. 6 dB).

The cylinders were constructed from duralumin (6 mm and 9 mm) and brass (2.2 mm) and were burnished prior to their use. Considering that the modulus of the effective complex relative permittivity of aluminium at an optical wavelength is approximately 30 and the real part is the largest term, it is reasonable to assume that ordinary terrain at V.H.F. and metals at optical frequencies behave similarly; consequently the simulation experiments should be realistic for both vertically and horizontally polarized waves. Nevertheless, there is a deviation between the actual and the theoretical permittivity of the models used due, primarily, to the oxide film which is known to cover the surface of aluminium and brass soon after their exposure to the atmosphere. This deviation affects only the results for vertical polarization in the case of which, in general, the ohmic loss tends to reduce the resultant diffracted field. In contrast, for horizontally polarized incidence the permittivity variations have negligible effect⁽²⁶⁾ and in general the diffracted field remains unchanged.

For comparison reasons the theoretical conducting knife-edge diffraction curve for horizontal polarization is included in figure 4.13, as well as the razor blade results for the same case, both taken from figure 4.9. It can be seen that the decrease of the radius of curvature is accompanied by enormous increases in the field strength particularly in the deep shadow.

On the other hand the effect of polarization is indistinguishable for large values of radius of curvature and only just distinguishable for smaller ones. Moreover, greater dispersion of the measured values is

noticed for vertically polarized waves, which may be due to the greater sensitivity of this polarization to the angle of incidence, which for the model is grazing.

Finally one can conclude that the radius of curvature becomes less critical as it becomes larger, since a radius difference of about $20,000 \lambda$ ($30,000 \lambda - 9500 \lambda$) for large obstacles introduces almost as much loss as a radius difference of about $6,000 \lambda$ ($9500 \lambda - 3500 \lambda$) in the case of thinner obstacles.

4.5. Diffraction By Rough Curved Obstacles

4.5.1. Theoretical Aspects

The influence of the specularly reflected components, discussed in section 4.4.1, will be reduced if the surface of the obstacle is made rough, because the reflected ray will now be of a smaller magnitude than in the smooth obstacle case and will depend on the roughness. This case is the most realistic, since in most diffraction problems, rounded rough hills constitute the obstacles.

The mathematical calculation of the diffracted field behind a randomly rough obstacle is obviously impossible. Even if it were possible to represent the various minor obstacles, constituting the roughness of the diffracting obstacles, by profiles of constant curvature, the evaluation of the diffracted field strength would be complicated. However, several attempts have been made to evaluate approximately the field received behind a series of successive obstacles whose individual diffraction behaviour is supposed to be known.

One frequently mentioned approximate technique is the so-called multiple diffraction method, shown in figure 4.14, which can easily be extended to any number of obstacles of idealised configuration with known diffracting properties. It must be noted, however, that since no rigorous

solution of the diffracted field exists even for simple forms of obstacles this method is only an approximation. Nevertheless its simplicity and its ease of application to general configuration make it attractive for rapid prediction work.

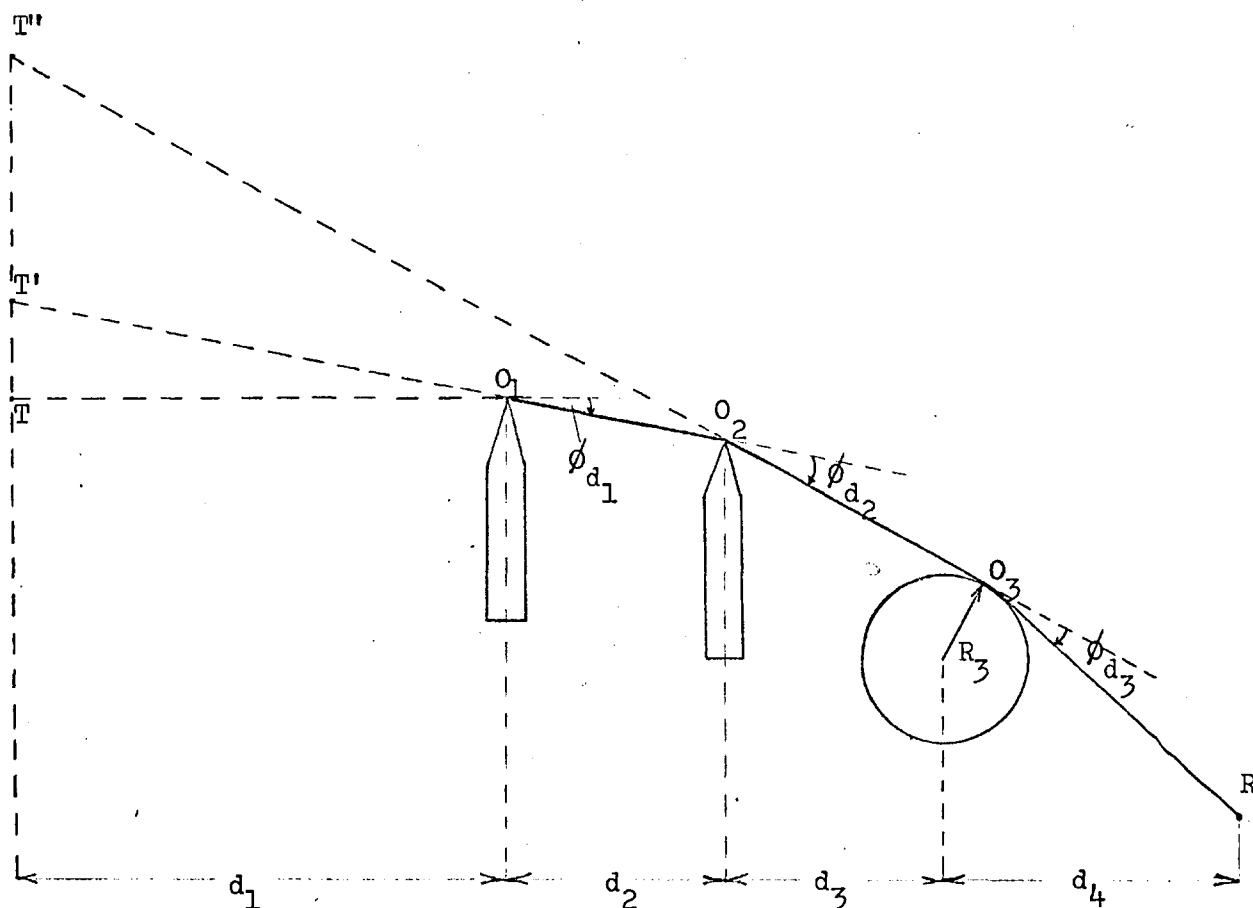


FIGURE 4.14

MULTIPLE DIFFRACTION METHOD

The basis of the method is the assumption, mentioned in section 4.3.1, that the diffracted field can be regarded as arising from a radiating line source situated along the crest of the diffracting obstacle.

With the transmitter at point T and the receiver at point R on both sides of a set of obstacles O_1 , O_2 , O_3 , as in figure 4.14, the relative diffracted field $|E_1|$ at O_2 due to the presence of O_1 is first evaluated. For a knife-edge at O_1 this field is given by equation 4.7 with a Fresnel parameter:

$$v_1 = \sqrt{\frac{2d_1 d_2}{\lambda(d_1 + d_2)}} \phi_{d_1} \quad (4.20)$$

Similarly, a field $|E_2|$ at O_3 can be evaluated by assuming obstacle O_1 absent and the source removed directly above T to a point T' on the line $O_2 O_1$. The Fresnel parameter is now given by:

$$v_2 = \sqrt{\frac{2(d_1 + d_2)d_3}{\lambda(d_1 + d_2 + d_3)}} \phi_{d_2} \quad (4.21)$$

Finally, a relative field $|E_3|$ at R, due to the source T at T' and the obstacle at O_3 , can be evaluated by assuming obstacles O_1 and O_2 removed and point T' at the intersection of the lines TT' and $O_3 O_2$. This field is now given by equation 4.17 with a Fresnel parameter:

$$v_3 = \sqrt{\frac{2(d_1 + d_2 + d_3)d_4}{\lambda(d_1 + d_2 + d_3 + d_4)}} \phi_{d_3} \quad (4.22)$$

and a curvature parameter:

$$\gamma = \left(\frac{\pi R_3}{\lambda}\right)^{\frac{1}{3}} \phi_{d_3} \quad (4.23)$$

where R_3 is the radius of curvature of obstacle O_3 and λ is the wavelength. Distances d_i and diffraction angles ϕ_{d_i} are as given in figure 4.14.

The relative field at R due to the source at T diffracted by the

set of obstacles O_1, O_2, O_3 will be given approximately by:

$$|E_R| = |E_1| \cdot |E_2| \cdot |E_3| \quad (4.24)$$

The method described is a good approximation when each obstacle is well into the shadow cast by the preceding obstacle; this is not the case when any two of the obstacles are relatively close together and empirical technique has been devised for dealing with these cases⁽¹⁷⁾.

4.5.2. Experimental results

A cylinder having a radius of curvature equal to 6 mm ($\approx 9500 \lambda$) was covered by conducting adhesive paper, whose surface was roughened, to simulate curved rough obstacles on the laser model. Three different roughnesses were used, being obtained by grinding the surface of the adhesive paper with carborundum granules and powders of various grades.

Surface profiles along the diffracting crest are given in figure 4.15. These profiles were measured with a Telysurf instrument, after the roughened papers were fixed on the cylinders, by dragging a fine diamond stylus across the surface and simultaneously amplifying and recording the surface height fluctuations. It is estimated that due to the comparatively large diameter of the stylus ($2.5 \mu \approx 4 \lambda$), the r.m.s. deviation at surface height and the mean width of the protuberances or "grain-size", as given in Table 4.1, are only approximate.

SURFACE ROUGHNESS	R.M.S. HEIGHT DEVIATION	AVERAGE WIDTH
Rgh I	18 λ	40 λ
Rgh II	13 λ	90 λ
Rgh III	23 λ	130 λ

TABLE 4.1 ROUGH SURFACE CHARACTERISTICS

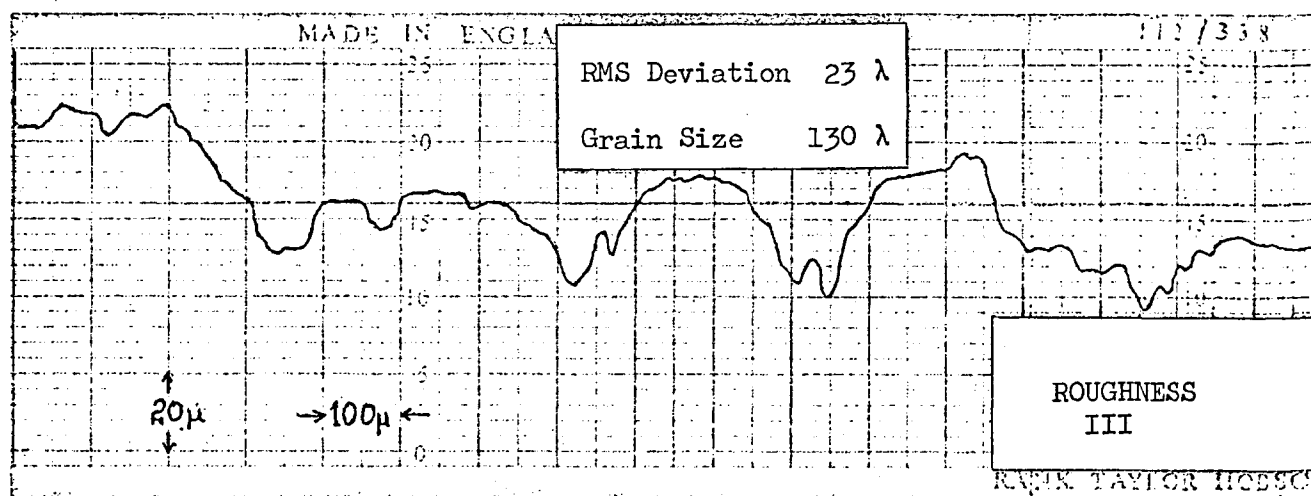
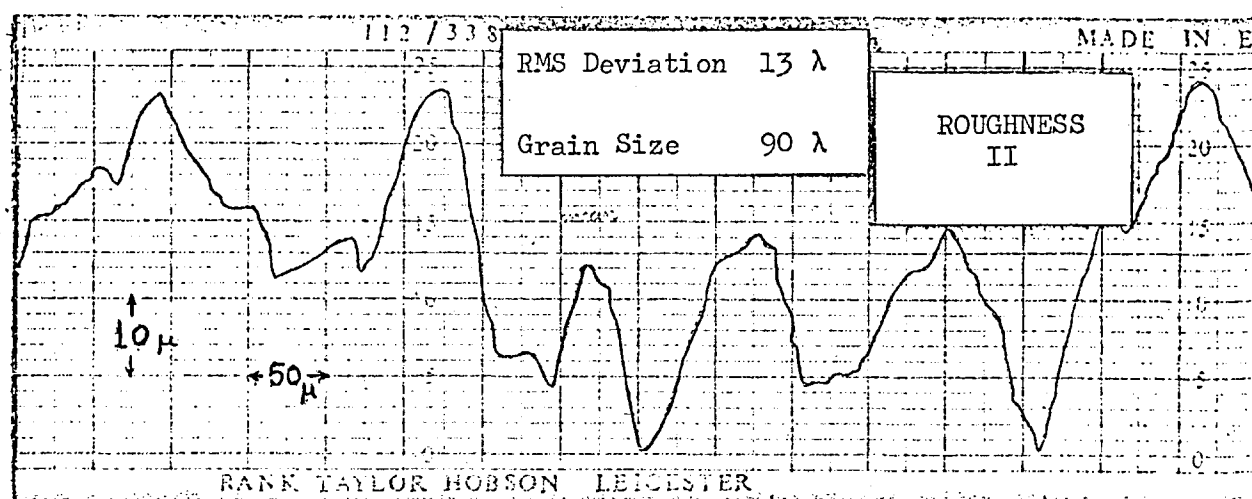
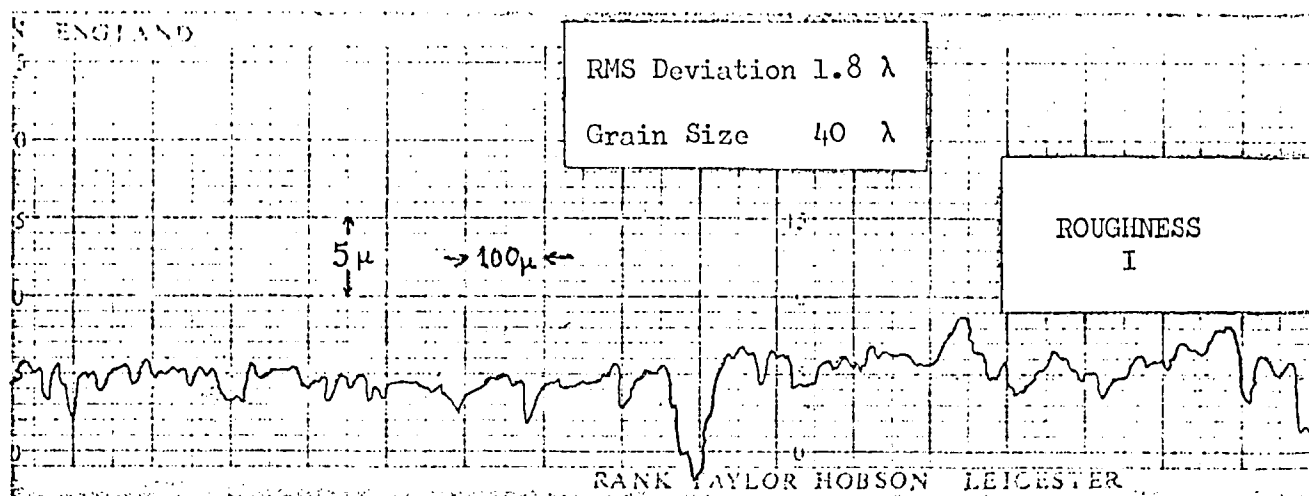


FIGURE 4.15

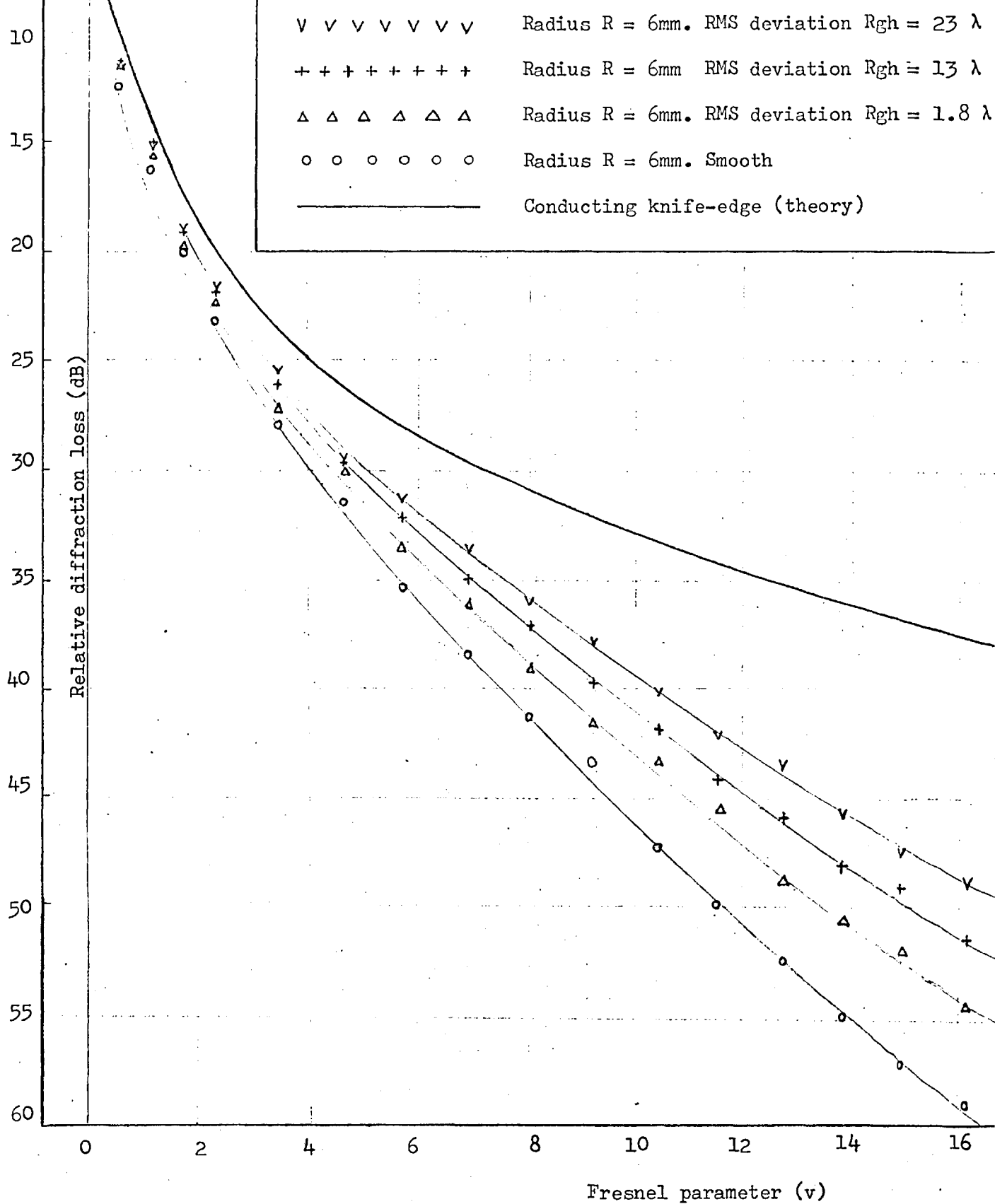
SURFACE PROFILES OF USED SAMPLES ON

A CYLINDER OF RADIUS $R = 6 \text{ mm}$

FIGURE 4.16

DIFFRACTION BY ROUGH CURVED OBSTACLES

(EXPERIMENTAL RESULTS FOR ROUGH CYLINDERS)

HORIZONTAL POLARIZATION

The diffracted field measurements were carried out with horizontally polarized incident waves and the geometry described for the knife-edge and smooth cylinder cases. The experimental results are given in figure 4.16 together with the theoretical conducting knife-edge diffraction curve and the measurements obtained for the smooth cylinder of the same radius of curvature (6 mm).

It can be seen that the results are in good agreement with the theoretical aspects described in the previous section.

It can also be seen that for even a small degree of roughness the diffracted field is increased by about 6 dB in the deep shadow.

However, the increase of diffracted field with r.m.s. deviation does not follow a linear law, the diffracted field becoming less sensitive as the r.m.s. deviation becomes large. In any case, provided the roughness remains very small compared with the cylinder radius, the diffracted field strength is relatively close to that of the smooth case.

CHAPTER FIVE

INVESTIGATION OF THE DIFFRACTED FIELD

PHASE DISTRIBUTION

5.1. General

An important difference between the optical model, described in section 4.2, and its radio frequency counterpart is the method employed in detecting the energy reaching the receiving point. At optical frequencies, photon energy is high enough to be detected by a photon detector, such as a photomultiplier; but even at microwave frequencies, photon energy is extremely low and the customary method of detection is by means of coupling to the electromagnetic wave with an antenna usually of parabolic dish or horn.

It is well-known that when using a microwave receiving antenna it is necessary to have a uniform phase front over the entire antenna aperture if the full theoretical gain is to be realized.

To keep the gain reduction within acceptable limits, a commonly-used criterion is that the phase errors should not exceed $\frac{\pi}{4}$. Phase errors greater than $\frac{\pi}{4}$ result in deterioration of the effective antenna gain and, in case of phase reversal at various points of the antenna aperture the gain may fall to a very low value.

Phase changes, as well as amplitude changes, may occur at various points of the receiving aperture plane due to perturbations of the Huygens' sources above the diffracting obstacle. Amplitude changes have the same effect on both methods of detection mentioned above, since they add arithmetically both when photon energy or electromagnetic energy is detected; but phase changes do not give the same result on both cases. The photomultiplier is sensitive to light intensity reaching the receiving aperture and conveyed by means of an optical guide; phase differences at

different points of the aperture are not taken into account, since only the photon energy is measured. On the other hand, phase differences at different points of a microwave antenna aperture are of great importance, since contributions from the various points add vectorially at the combiner and may cancel each other in the case of a difference of π .

Diffraction by randomly rough obstacles has not yet been well-established. Nevertheless, the planning of microwave links, taking obstacle diffraction into account, is heavily dependent upon the advance knowledge of the diffracted field, if correct predictions are to be made. This is particularly so in military applications where it is likely that, for security reasons, receiving antennas will most probably be screened by ground elevations, and the diffracted field prediction will usually be necessary. Furthermore, ground measurements are either impossible or inadequate due to time or security restrictions.

It is, therefore, very important to know the discrepancies, if any, between the optical model and its radio frequency counterpart, arising from irregular phase distribution on the receiving aperture. The purpose of this chapter is to study the possibility of investigating phase at various points of the receiving aperture which was employed in studying obstacle diffraction (Chapter four).

Firstly, the various methods by which this problem can be approached are examined and the results interpreted. Secondly, advantages and disadvantages of each method and the extent to which these methods can be employed to carry out meaningful experiments are discussed. Finally, an interpretation of some experimental results is attempted and some conclusions drawn.

5.2. Interferometry

It is well-known that interferometry, besides its many other applications, is a powerful method for relative phase measurements. Due to its simplicity

and economy of means which, however, give critically correct results, interferometry will be used in the methods described below. A brief description of its nature is therefore considered worthwhile.

When two light beams, both emanating from one point source, arrive at a screen after following different paths, interference will take place and their characteristics will combine to produce visible effects. If F_1 and F_2 (Fig. 5.1) represent two parallel uniform plane wavefronts, on arriving at screen S they will combine to produce a bright light when in phase and no light when 180° out of phase.

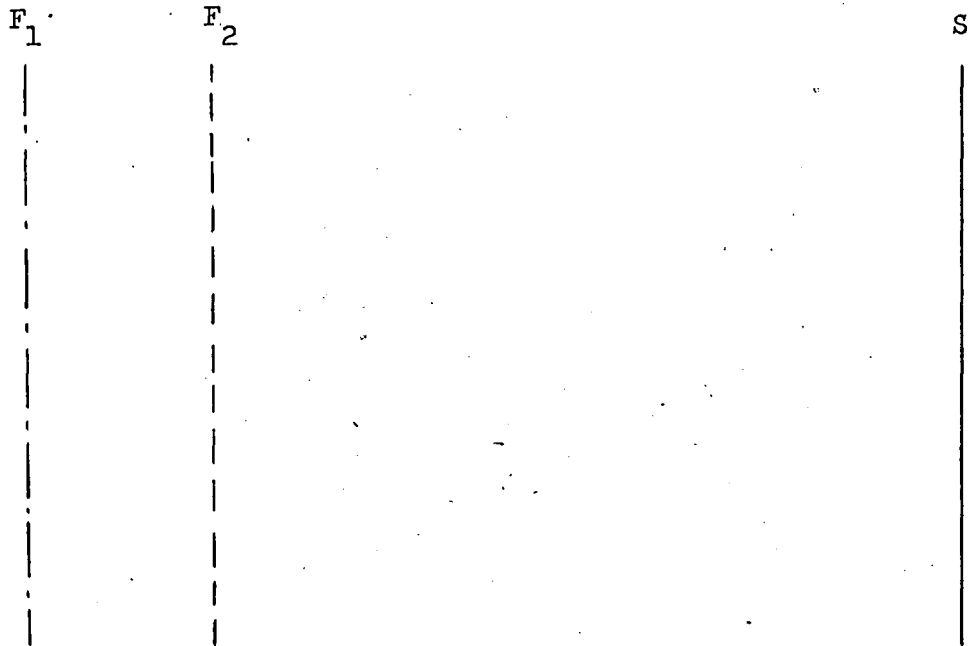


FIGURE 5.1

PARALLEL UNIFORM PLANE WAVEFRONTS

Intermediate brightness levels will be expected for phase differences between 0° and 180° . In every case the resultant light intensity (I) is given⁽²⁷⁾ by the equation:

$$I = I_1 + I_2 + 2\sqrt{I_1 I_2} \cos \delta \quad (5.1)$$

where I_1 , I_2 are the individual intensities of F_1 , F_2 respectively and δ is the phase difference between them due to the path difference.

When $I_1 = I_2$, equation 5.1 reduces to:

$$I = 2I_1(1 + \cos \delta) = 4I_1 \cos^2 \frac{\delta}{2} \quad (5.2)$$

Assuming an angle θ between the wavefronts F_1 and F_2 , as in figure 5.2(a), visible parallel fringes, as in figure 5.2(b), will appear on screen S , their separation (d) given by:

$$d = \frac{\lambda}{\sin \theta} \quad (5.3)$$

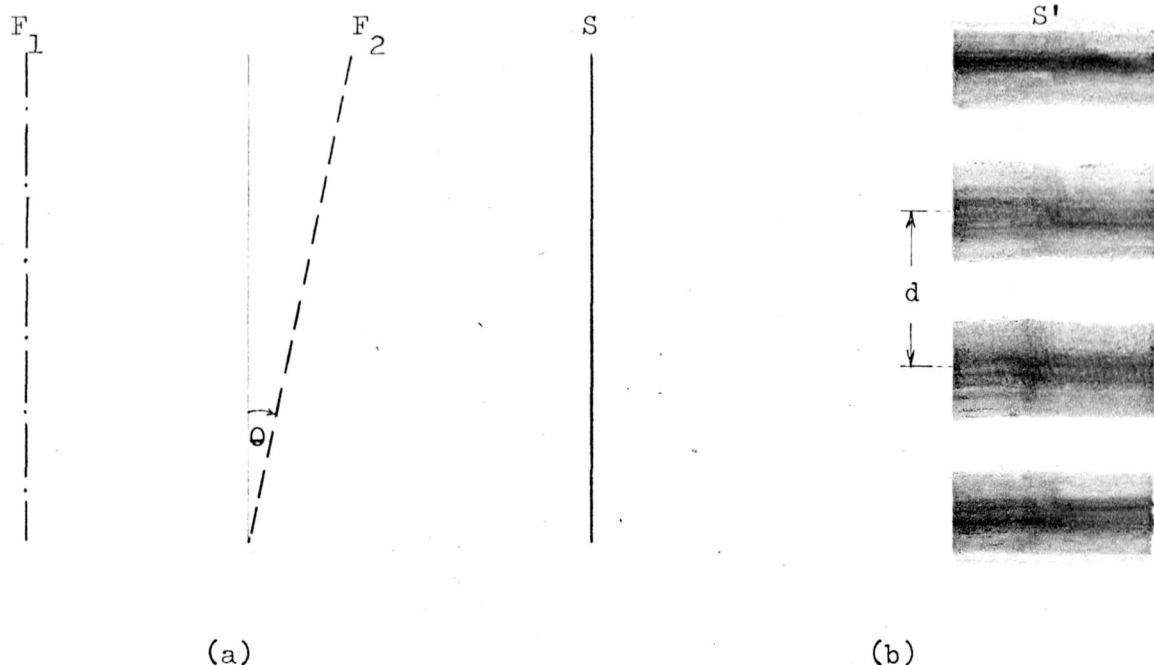


FIGURE 5.2

INTERFERENCE OF NON PARALLEL UNIFORM PLANE WAVES

where λ is the wavelength of the radiating source. Light intensity at each point is given by equation 5.1 or 5.2, which implies a gradual transition from very bright to dark as the phase difference δ changes from 0° to 180° due to the combined effects of path difference and angle θ .

A more complicated case, although unrealisable, is shown in

figure 5.3 where, besides angle θ between the wavefronts F_1 and F_2 , the second wavefront has a phase difference $\phi = 180^\circ$ between different points in the same plane perpendicular to the plane of propagation.

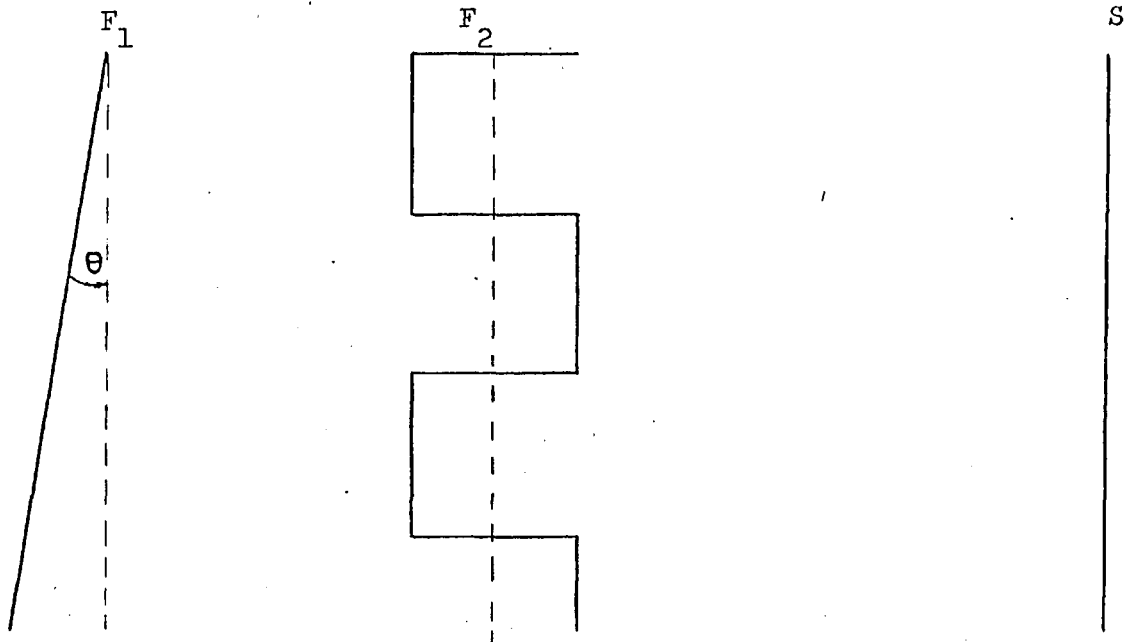


FIGURE 5.3

INTERFERENCE BETWEEN TWO DIFFERENT WAVES

Assuming for convenience, an angle $\theta = 14^\circ 29'$ between the wavefronts, interference fringes as in figure 5.4 will appear on the screen due to the interaction explained below in connection with figure 5.4. Light intensity is calculated from equation 5.2 because equal intensity for both beams is assumed.

Due to the angle $\theta = 14^\circ 29'$ between the wavefront they are in phase just below points A and J and just above point E, while a phase difference of 180° occurs just above points A, J and just below point E. A phase difference of 90° at points C and G results in a $2I_1$ intensity since $\cos \delta = 0$. A value of $.6I_1$ is due to a phase difference of 135° (or 225°) at one side of points B, D, F and H. A value of $3.4 I_1$ is

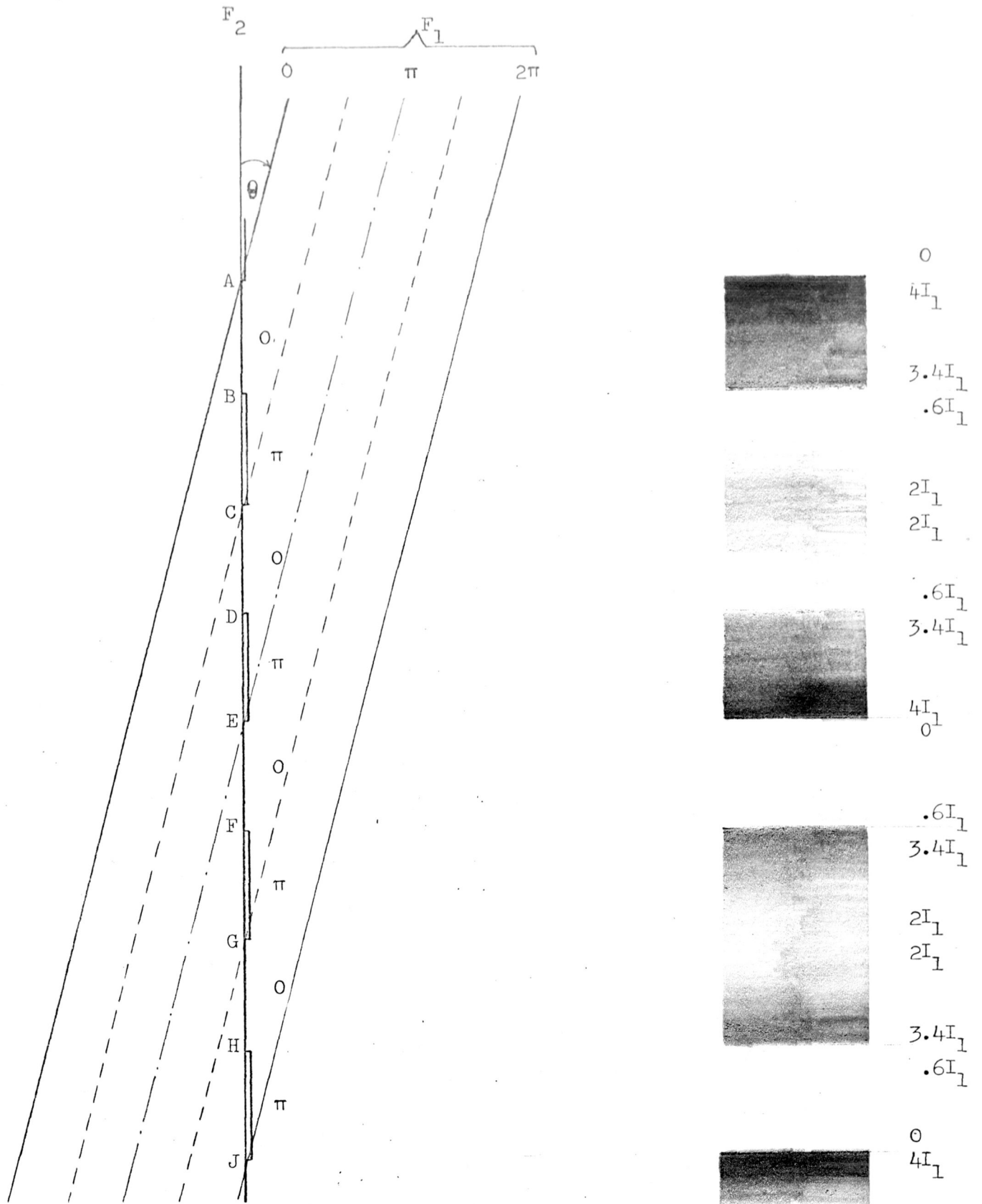


FIGURE 5.4

INTERFERENCE FRINGES CAUSED BY THE WAVES IN FIGURE 5.3

due to a phase difference of 45° (or 315°) at the other side of points B, D, F and H.

Interference patterns similar to, and even more complicated than, those mentioned above will be the object of observation in the following methods of investigating the phase distribution of the field. Their behaviour, when either of the interfering beams is disturbed by the diffracting obstacle, is expected to give the necessary information about phase distribution changes due to the presence of the obstacle.

5.3. Methods of Phase Investigation

5.3.1. A Method Using Michelson's Interferometer

The idea of this method is based on the assumption that, when obstacle diffraction takes place, the resultant field at the receiving point is related to the amplitude and relative phase of the hypothetical Huygens' sources over the aperture plane above the obstacle. In the simple case of knife-edge diffraction this relationship is given by the known equation (See section 4.3.1):

$$\frac{E}{E_0} = \frac{1+j}{2} \int_v^\infty e^{-j \frac{\pi}{2} v^2} dv \quad (5.4)$$

where v is the Fresnel parameter.

Furthermore, according to Fresnel, the hypothetical sources included in only a few Fresnel zones above the obstacle need to be considered. The height of such a zone can be calculated from the geometry in figure 5.5, assuming $d_1, d_2 \gg \lambda$ and ϕ confined to small angles. This height is given by

$$H_F \approx \frac{d_1 d_2}{d_1 + d_2} \left(\sqrt{\phi^2 + \frac{\lambda(d_1 + d_2)}{d_1 d_2}} - \phi \right) \quad (5.5)$$

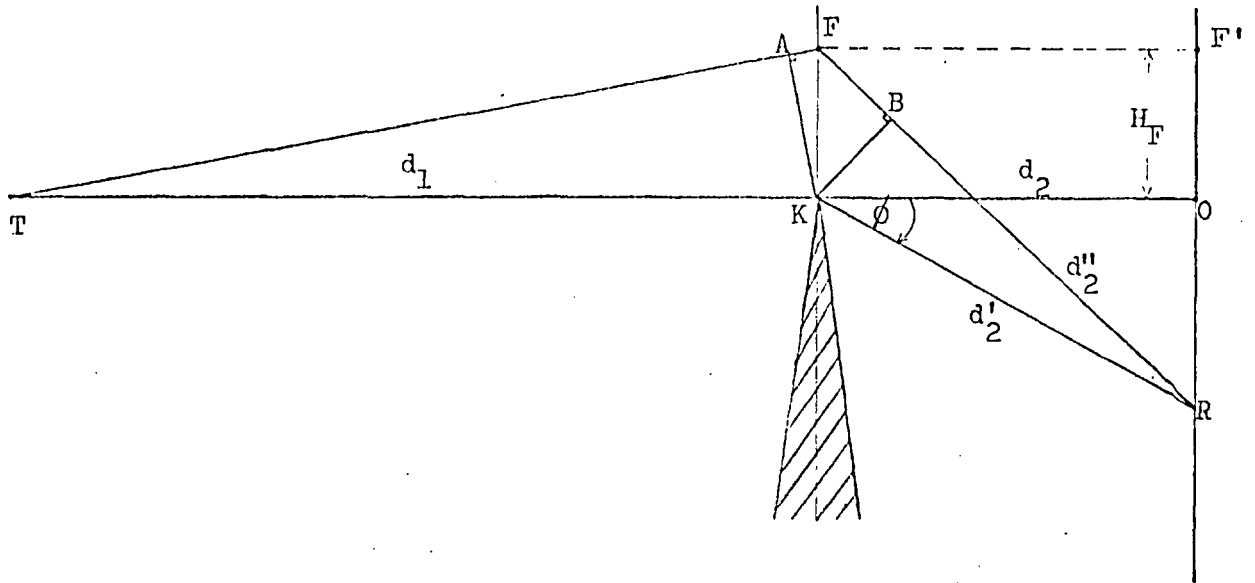


FIGURE 5.5

THE HEIGHT OF A FRESNEL ZONE ABOVE A KNIFE-EDGE

where ϕ is the diffraction angle in rad.. For the usual case of constant d_1 , d_2 and λ , H_F is dependent upon ϕ and is largest ($\approx 150 \lambda$) at grazing angles while it is only 5λ for a diffraction angle of about $5^\circ - 6^\circ$.

In any case, the important height above the abstacle is included in not more than a fraction of a millimetre, and this is the area where the investigation should be directed.

The arrangement for this method is shown in figure 5.6.

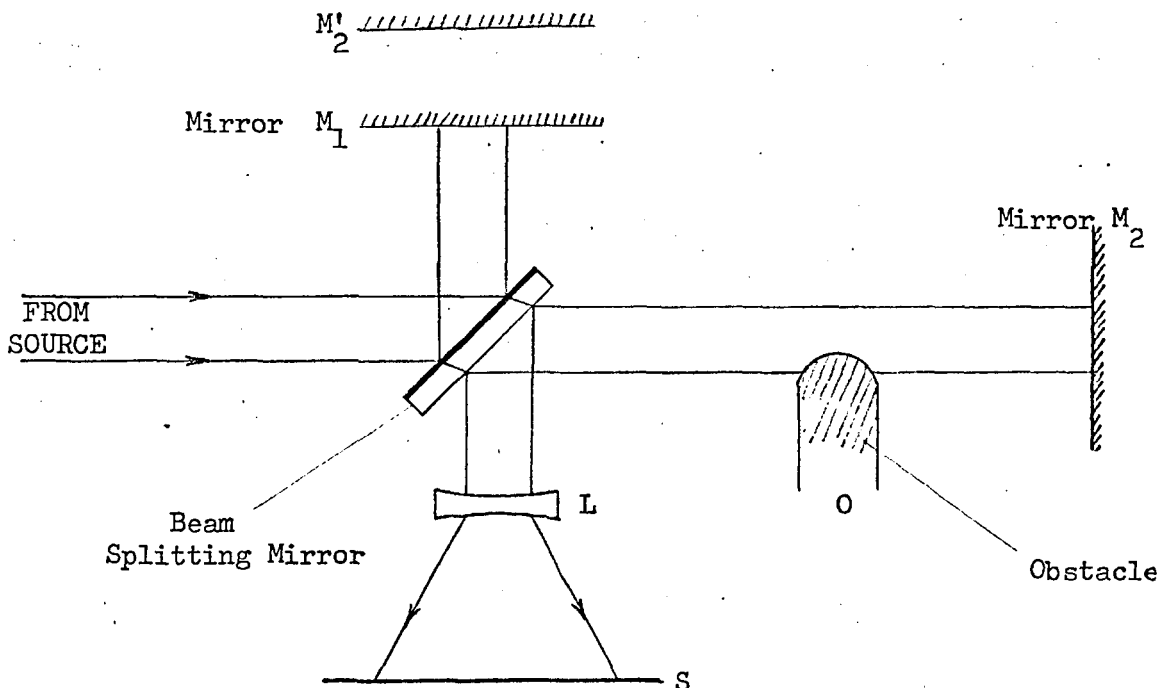


FIGURE 5.6

A METHOD USING MICHELSON'S INTERFEROMETER

Optically the system, as viewed from S , consists of two sources M_1 and M_2' in general behind each other with their planes inclined in optical terms. The interference fringes are, according to theory, straight lines parallel to the line of intersection between the planes of the sources and are projected on screen S , after passing through lens L in order to be magnified for better investigation.

By inserting the obstacle O between the splitter and mirror M_2 , one of the interfering beams is obstructed in proportion to the insertion of the obstacle. Any perturbation in the beam will be reflected by the mirror and will affect the interference pattern on the screen.

It is realized that, because of the diffraction, the wave-front distribution at the mirror is not the same as that above the obstacle. Nevertheless, the idea of the method is that after the interference pattern for one position of the obstacle has been established, any subsequent disturbance following the insertion or withdrawal of the obstacle will be due to this and only this movement, if all other factors are unchanged. A relation between the aperture distribution above the obstacle (in its new position) and the resulting interference pattern can then be sought.

The position of the obstacle relative to the beam and the interference pattern can be estimated by diffraction techniques in connection with the micrometric head monitoring the movements of the obstacle. (This technique is described in the next section dealing with the second method).

Although this method has the advantages of simplicity and economy of means it is obvious that the possible modification of the interference pattern is not only due to the obstruction as the beam travels towards mirror M_2 , but it is also due to the obstruction by the other side of the obstacle as the beam is reflected back. Unless no disturbance

occurs, due to either side of the obstacle, this method seems to be insufficient since neither the obstacle shape on both sides is ideally the same nor can its effect be assumed identical when obstructing the transmitted and reflected beams.

Another disadvantage associated with this method, as well as with the next, is the appearance of some undesirable fringes caused by interference between rays reflected from the front and rear faces of the splitter. These fringes cannot be removed unless a specially prepared beam splitter is used. Their existence introduces some problems when studying the interference pattern, although their shape and position are stable and unaffected by the movements of the obstacle. This disadvantage is treated in more detail in the next method where the use of two splitters simultaneously makes it more pronounced.

Finally, dust or other particles, casually appearing in the beam or on any optical item used, may significantly disturb the interference pattern. This possibility is another disadvantage of this and the next method.

5.3.2. A Method Using a Modification of Mach-Zehnder Interferometer

In order to avoid the major disadvantage of the previous method, i.e. the interposition of the obstacle in both the transmitted and reflected beams, a set up whose important parts are shown in figure 5.7 can be used. This is a modification of the Mach-Zehnder⁽²⁸⁾ interferometer the main dissimilarity being the great path difference inherent in the present set up. The interference pattern can be magnified as described in section 5.3.1.

Apart from the disadvantage mentioned above all others associated with the previous method are also encountered in this method. Further disadvantages are difficulty in alignment and the necessity for more components.

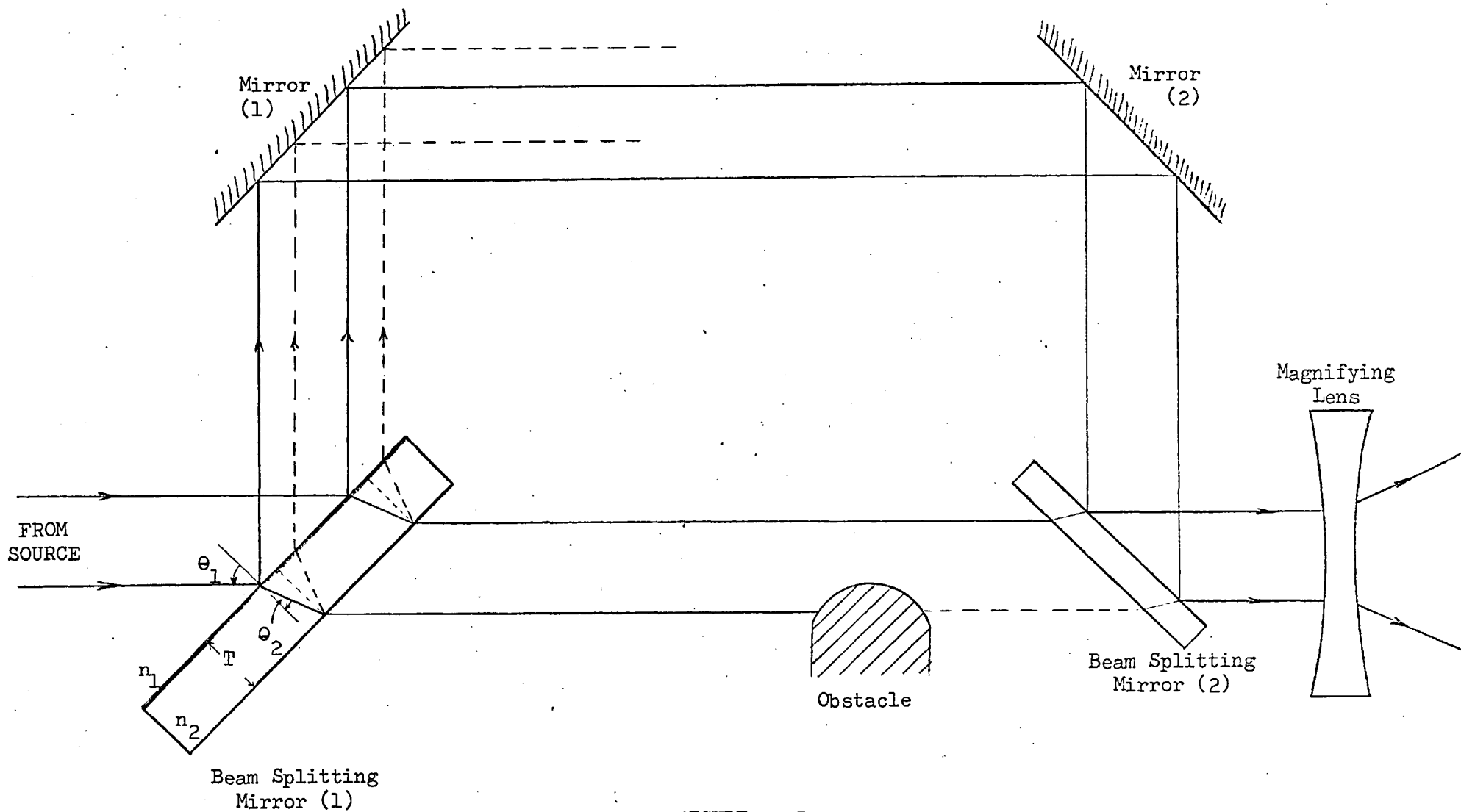


FIGURE 5.7

A METHOD USING THE MACH-ZEHNDER INTERFEROMETER

A problem arising in both methods is the difficulty of precisely determining the position of the obstacle relative to the beam and comparing it with the relative position shown on the screen. A way round this problem is to use diffraction techniques as follows:

After determining the magnification of the interference pattern and its exact position on the steady screen, the centre point of the object beam is found by scanning the receiving aperture vertically and horizontally, while monitoring its movements by the micrometric heads and observing the photomultiplier output. With the receiving aperture at this centre point, the obstacle is traversed until the photomultiplier output is reduced to $1/4$ of its unobstructed value. By removing the receiving aperture and its mount, the interference pattern, including the obstacle, will be cast on the screen magnified and the relative pattern-obstacle position can be determined for the axial position of the obstacle. Any subsequent movement of the obstacle can be controlled and monitored by the micrometric head and determined on the screen by the magnification.

Because two beam splitters are used in this method the undesirable fringes arising from them, as mentioned in section 5.3.1, have a greater effect on the clarity of the interference patterns and the differences due to obstacle blocking. The mechanism by which these fringes are produced is shown in figure 5.7.

With an angle of incidence θ_1 on the first face of the splitter the angle of refraction θ_2 is given by Snell's law:

$$n_1 \sin \theta_1 = n_2 \sin \theta_2 \quad (5.6)$$

where n_1 and n_2 are the refractive indices of the air and the splitter. Unless the second face is given an ideal $\lambda/4$ coating, part of the transmitted wave will be reflected at this boundary. A large proportion

of this reflected beam will be refracted at the first face and emerge at the same angle as the main reflected beam, as shown in the figure. Undesirable fringes are produced when these two beams interfere. At the same time, similar fringes, but far less clear, are obtained from the transmitted beam and the portion of the secondary beam reflected internally at the first face.

Considering the second face of the splitter, the reflection coefficient (R) for horizontally polarized waves is given by:

$$R_H = \frac{\cos \theta_2 - \sqrt{1/n^2 - \sin^2 \theta_2}}{\cos \theta_2 + \sqrt{1/n^2 - \sin^2 \theta_2}} \quad (5.7)$$

and for vertically polarized waves is given by:

$$R_V = \frac{1/n^2 \cos \theta_2 - \sqrt{1/n^2 - \sin^2 \theta_2}}{1/n^2 \cos \theta_2 + \sqrt{1/n^2 - \sin^2 \theta_2}} \quad (5.8)$$

where n is the refractive index of the splitter.

The refractive index of the material from which these beam splitters are usually made is approximately 1.52. Equal transmission and reflection by the beam splitters used with the model is achieved for horizontally polarized waves incident at an angle of 45° . This, in connection with equation 5.6, gives:

$$\theta_2 = \arcsin \frac{\sqrt{2}}{3.04} = 27^\circ 43' \quad (5.9)$$

Substitution in equations 5.7 and 5.8 gives:

$$R_H = \frac{.8853 - \sqrt{1/1.52^2 - .4652^2}}{.8853 + \sqrt{1/1.52^2 - .4652^2}} = -0.36 \quad (5.10)$$

$$R_V = \frac{1/1.52^2 \times .8853 - \sqrt{1/1.52^2 - .4652^2}}{1/1.52^2 \times .8853 + \sqrt{1/1.52^2 - .4652^2}} = 0.042 \quad (5.11)$$

The undesirable effects mentioned above may be reduced by using vertical polarization arranging for the angle of incidence on the second surface to be equal to the Brewster angle. This, assuming a perfect beam splitter, makes the reflection coefficient in equation 5.11 equal to zero and gives a Brewster angle

$$\theta_2 = \arctan \frac{1}{1.52} \approx 33^\circ 20' \quad (5.12)$$

Substitution in equation 5.6 gives the necessary angle of incidence

$$\theta_1 = 56^\circ 39' \quad (5.12a)$$

In practice, due to the presence of two beam splitters, critically aligned to cause interference of the beams, it is impossible to achieve the required Brewster angles simultaneously for both beam splitters. Moreover, the use of vertical polarization significantly reduces the reflection-transmission ratio of the splitter. However, some improvement is always possible by using this technique, particularly on the first beam splitter.

Another technique to avoid the undesirable fringes is to use beam splitters thick enough to achieve a separation between the reflected beams. This is shown in figure 5.8 where the reflected beam B_r and B'_r are now adjacent.

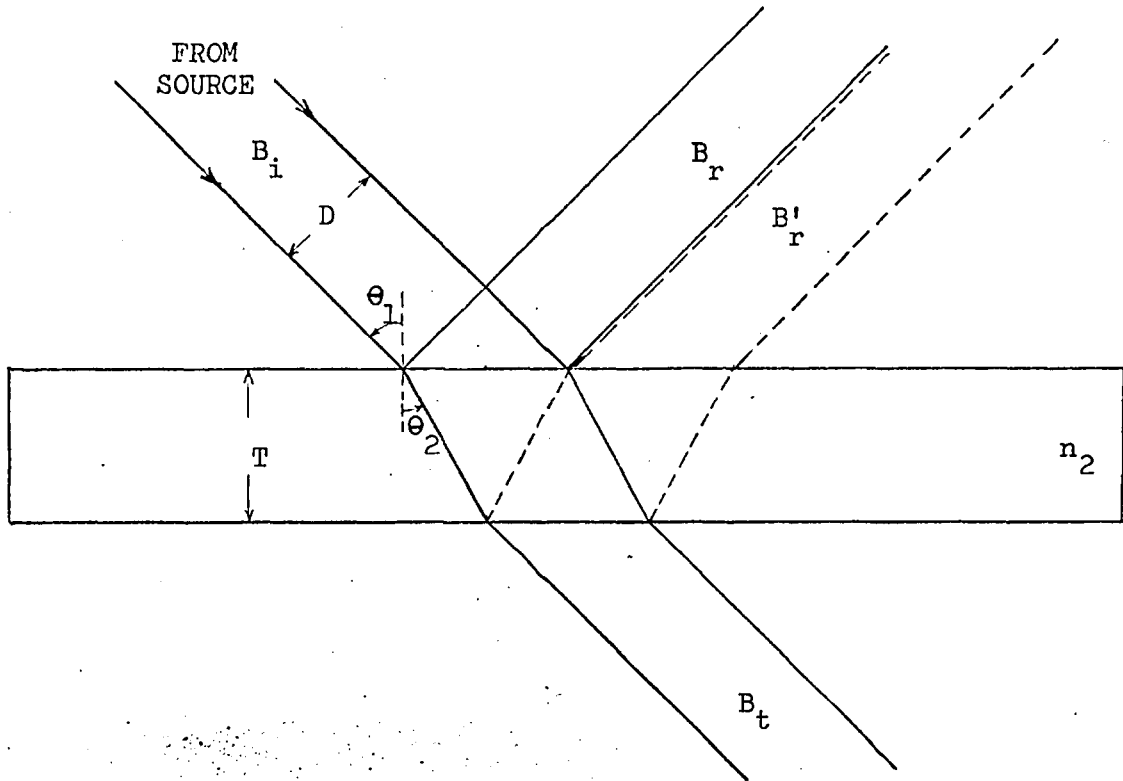


FIGURE 5.8

AVOIDING INTERFERENCE BY USING A THICK BEAM SPLITTER

Depending upon the angle of incidence (θ_1) on the front face, the refractive index (n_2) of the material of the splitter and the diameter (D) of the incident beam, the thickness (T) of the beam splitter can be calculated through the equation:

$$T = \frac{D}{2 \cos \theta_1 \tan \theta_2} \quad (5.13)$$

where the refraction angle θ_2 is given by equation 5.6.

Assuming a refractive index $n_2 \approx 1.52$ (glass BK7) and an angle of incidence $\theta_1 = 45^\circ$, the refraction angle is given by equation 5.9

($\theta_2 = 27^\circ 43'$) and the necessary thickness for a beam diameter $D = 15$ mm is:

$$T = \frac{15 \text{ mm}}{2 \cos 45^\circ \tan 27^\circ 43'} \approx 20 \text{ mm} \quad (5.14)$$

It is advisable to make the splitters a little thicker in order to account for the possible variations in the values of the parameters.

5.3.3. Experimental Results

The optical model used for diffracted field measurements (see chapter four) was appropriately modified to carry out the experiments described below. A close-up photograph of the modification is given in figure 5.9. The schematic diagram is given in figure 5.7.

By means of the polarization plate and a number of neutral density filters the shape, separation and direction of the interference fringes were appropriately modified in order to facilitate the investigation.

The object and the reference beams are given in figures 5.10 and 5.11 respectively, while their interference pattern is shown in figure 5.12. The undesirable fringes caused by the beam splitters can be clearly seen on the reference beam which undergoes reflections by both beam splitters. The object beam, transmitted through the splitters, is far less affected by the fringes although they can still be distinguished on the picture. The effect of these unwanted fringes on the interference pattern is shown in figure 5.12. It seems that their intensity lies somewhere between that of the reference and the object beam cases, as expected.

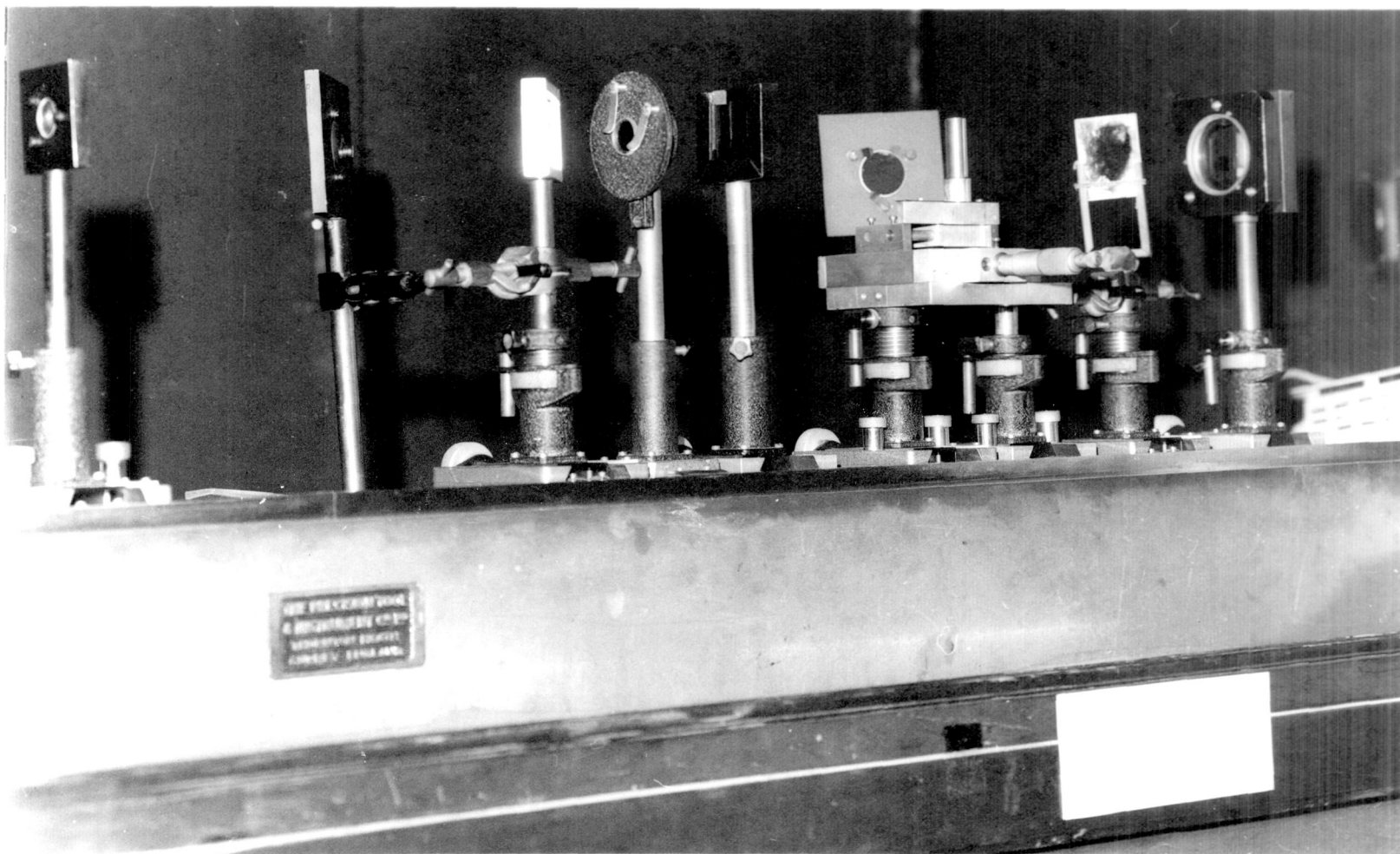


FIGURE 5.9

THE OPTICAL MODEL MODIFICATION FOR PHASE INVESTIGATION

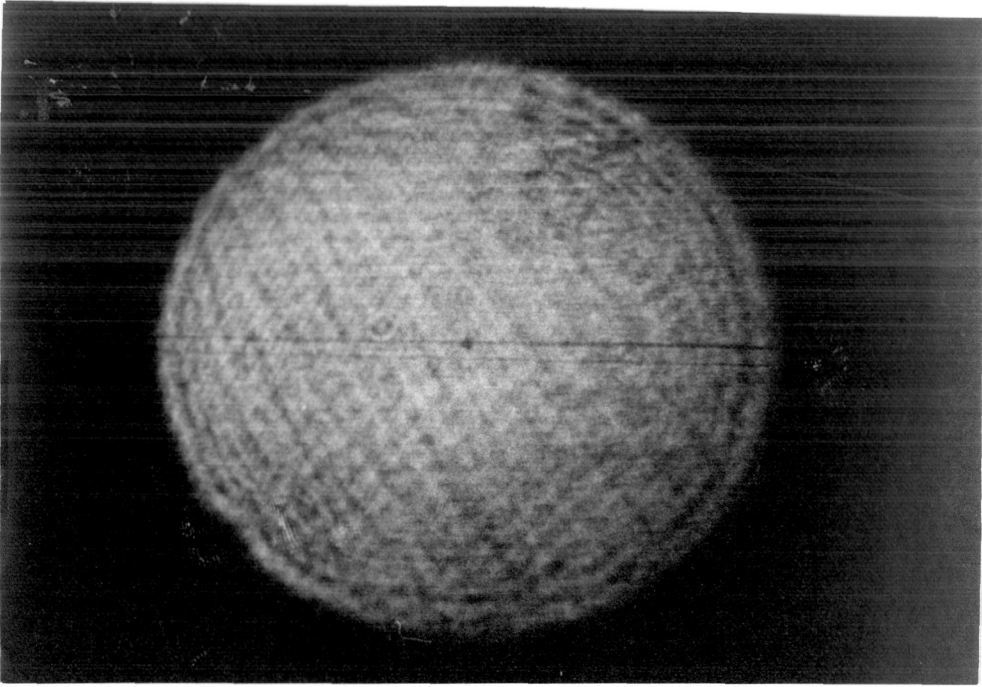


FIGURE 5.10

THE OBJECT BEAM

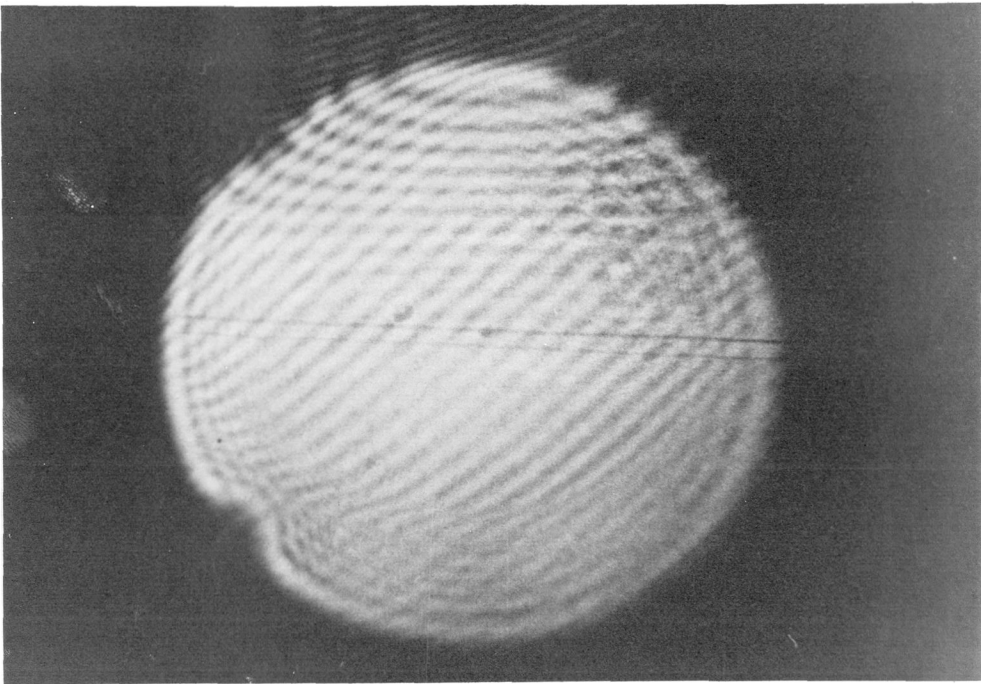


FIGURE 5.11

THE REFERENCE BEAM

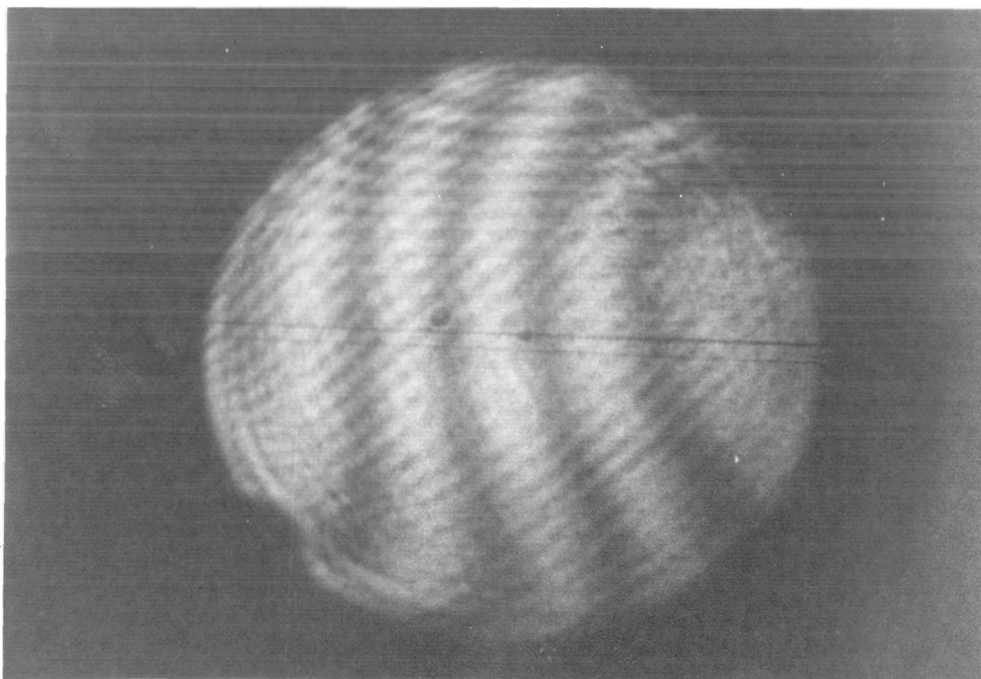


FIGURE 5.12

THE INTERFERENCE PATTERN UNOBSTRUCTED

The disturbance by dust or other particles, as mentioned in section 5.3.1, can be clearly seen on any of the pictures, particularly in figure 5.12.

The object beam obstructed by a knife-edge is given in figure 5.13. The knife-edge was placed in the middle of the beam (axial position). The diffracted field can be distinguished above the black line, indicating the position of the knife-edge in this case. The ripple on the illuminated region field can also be seen above the knife-edge.

A magnification of 5:1 was used for figures 5.10 to 5.13.

Figure 5.14 illustrates the effect of a cylinder obstructing the object beam. The right hand half of the beam is obstructed by a smooth cylinder while the other half is obstructed by a rough one. The transition from the rough to smooth cylinder cases is shown by a step due to the relatively greater radius of the rough cylinder, which was

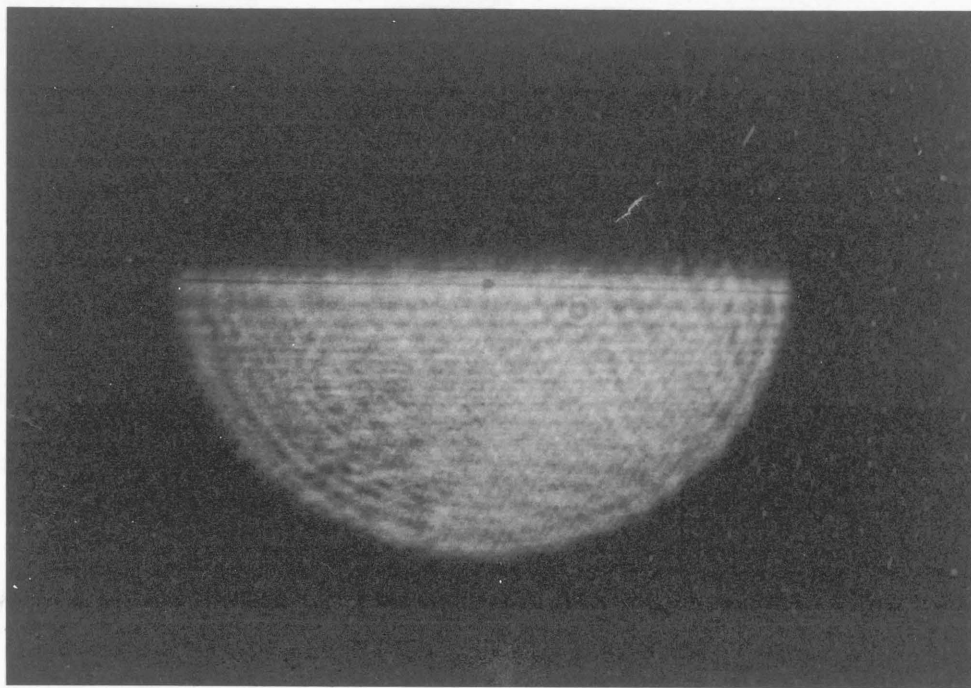


FIGURE 5.13

THE OBJECT BEAM OBSTRUCTED BY A KNIFE-EDGE

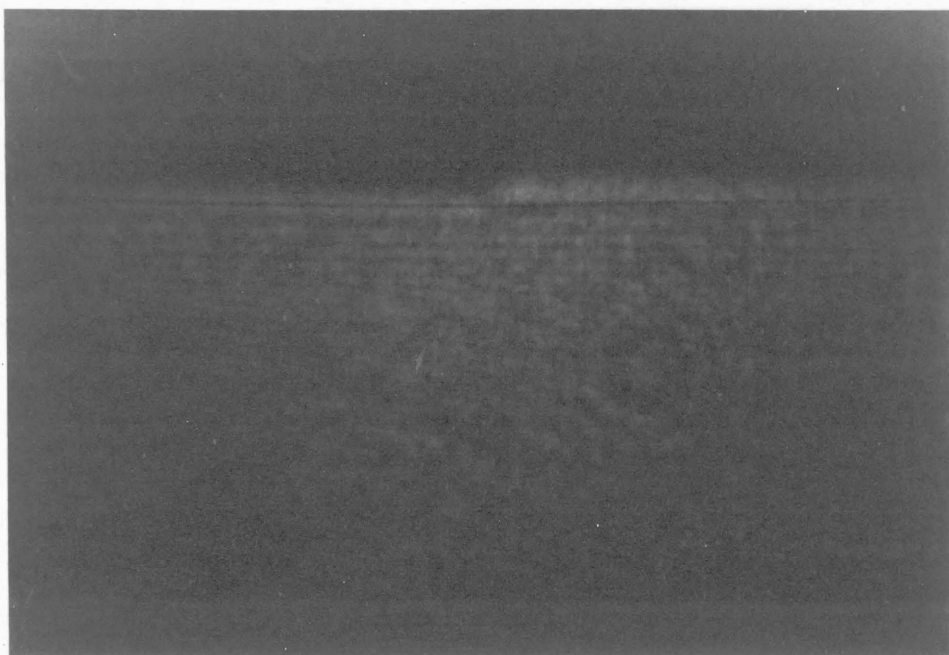


FIGURE 5.14

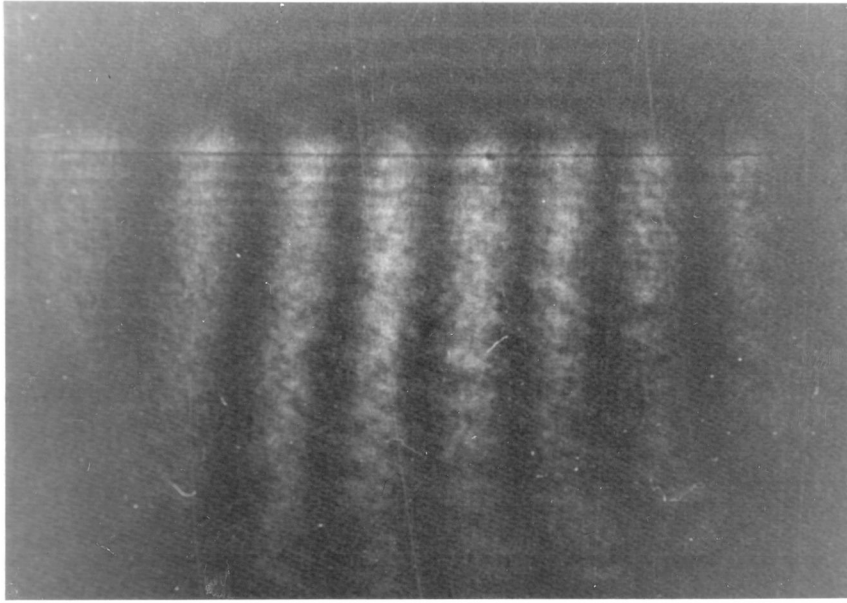
THE OBJECT BEAM OBSTRUCTED BY ROUGH AND SMOOTH CYLINDERS

obtained by covering part of a smooth cylinder by roughened aluminium tape. This is why the diffracted field seems to be higher in the smooth case.

The effect of a knife-edge on the interference pattern is shown in figures 5.15(a) and 5.15(b) for two slightly different positions of the knife-edge. In the first case the knife-edge was in the axial position, while in the second it was inserted a little further into the beam. It can be seen that apart from the diffracted field ripple, already noticed in figure 5.13, no significant modification of the fringes either horizontally or vertically appears in any of the pictures.

Similar results are obtained when the cylinder cases are investigated as can be seen in figures 5.16(a) and 5.16(b). Here the obstacle was inserted by a greater amount than in the knife-edge case before the second picture was taken.

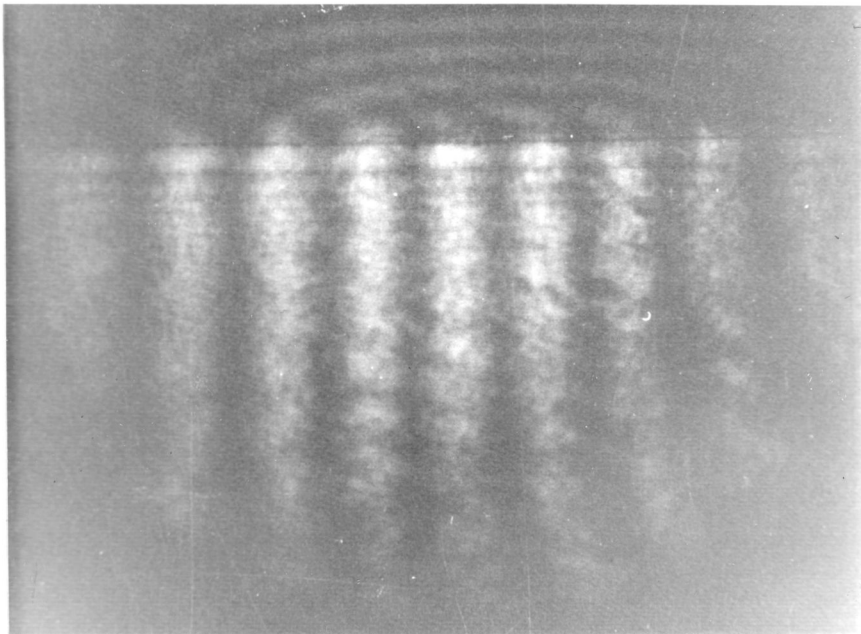
A magnification of 8:1 was used in figures 5.14 to 5.16(b), since only part of the pattern is needed. This means that if a change in the interference pattern had occurred the area of interest in figures 5.15 - 5.16 would have been that within about 1 mm from the dark area. Although the photographs are not as clear as we would wish (due to the unwanted fringes mentioned earlier) they do indicate that the disturbance of the phase just above the obstacle, i.e. below the black line, is small. The fringes in the shadow region, i.e. above the black line, are due to the beam splitters.



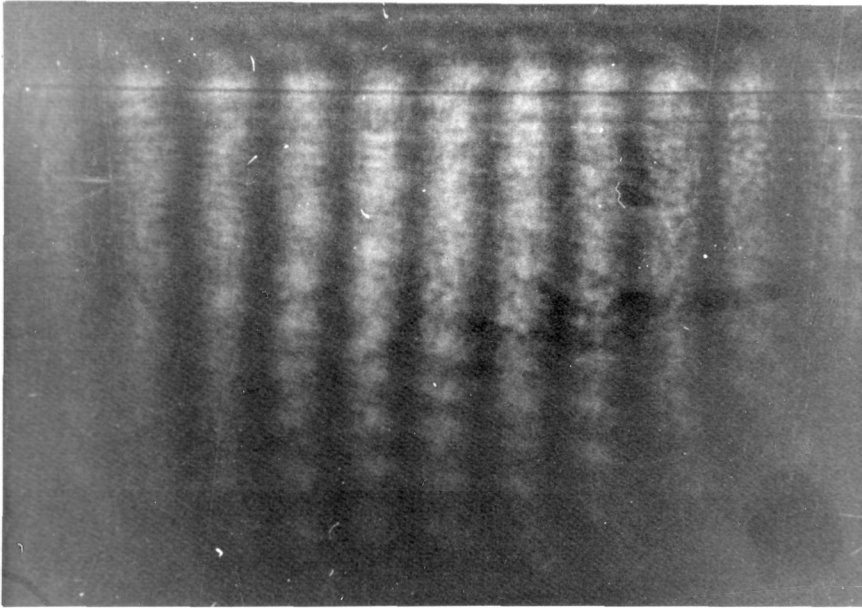
(a)

FIGURE 5.15

EFFECT OF KNIFE-EDGE ON THE INTERFERENCE PATTERN

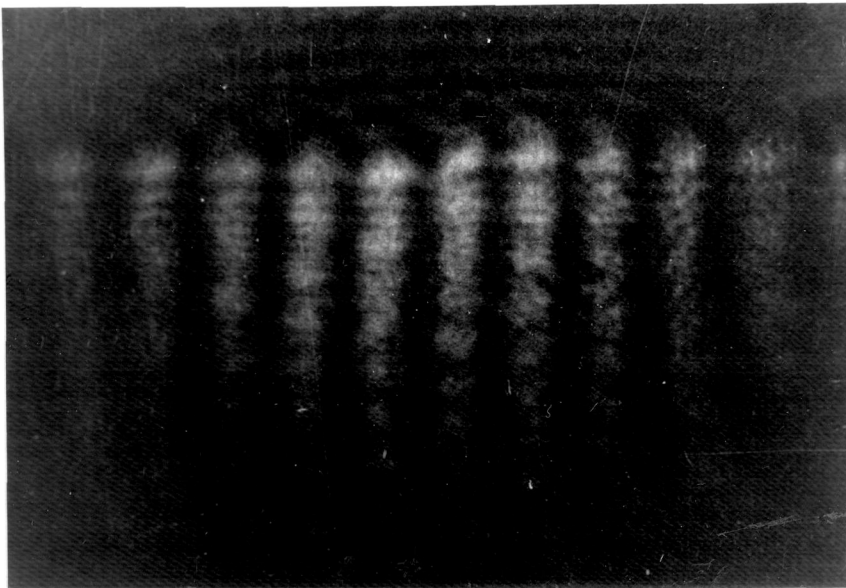


(b)



(a)

FIGURE 5.16
PARTLY PARTLY
EFFECT OF ROUGH AND SMOOTH CYLINDERS ON THE INTERFERENCE
PATTERN



(b)

5.3.4. Conclusions

Due to the undesirable fringes, caused by the splitters, and the other perturbations present in either of the interfering fringes, when observed individually, the results obtained may not be reliable and further investigation is needed. Some experiments with specially designed beam splitters and other optical components, carried out under better conditions, probably in a vacuum, would give better and more reliable results.

Furthermore, it is realised that by taking measurements at a plane other than that of the Huygens' sources aperture above the obstacle, an uncertainty is introduced. This uncertainty would be eliminated if a network, capable of transferring the Huygens' sources aperture unchanged to another plane, could be used. This seems impossible because such a network would be described by an equation containing the distribution of the aperture, which is to be studied.

5.3.5. A Method Using Holography

One way of investigating the phase distribution directly on the receiving aperture of the model is by means of holographic techniques. Gabor⁽²⁹⁾ has shown that the amplitude and phase information contained in a wavefront can be recorded on a photographic plate which after being developed, can be used to reveal that information. Such a record is called a hologram. The main differences between holography and photography lie in the information recorded and in the techniques employed in revealing that information.

A hologram is a photograph, on a transparent plate, of the interference pattern between the wavefront to be recorded and a coherent plane wave, the so-called reference beam. This pattern is a complex system of fringes unique for every recorded wavefront. When the hologram is illuminated with a beam of light, similar to the original reference beam, light

passing through the pattern undergoes a specific treatment at each point depending upon the distribution of the fringes; as a result, one transmitted wave duplicates exactly the recorded wavefront.

In the simple case of recording a plane wavefront, such as F_1 in figure 5.2(a), using a reference beam with a wave front such as F_2 , the resulting hologram will consist of a series of interference fringes identical to those given in figure 5.2(b). Illuminating the hologram, after developing, with a beam similar to F_2 will produce one transmitted wave identical to F_1 .

What is important in our case is the phase distribution of the diffracted wavefront at a distance beyond the obstacle i.e. the phase information contained in a specific area of a particular plane - the plane of the receiving aperture. If this distribution can be recorded for various positions of the diffracting obstacle, as well as without the obstacle, useful information can be extracted by comparison. Qualitative and quantitative relationships between the various cases can then be formed. Holography seems to be a very powerful and promising technique in this field, if some practical problems can be solved. One of these problems is the difficulty of visually detecting small intensity differences. According to the Rayleigh criterion, intensity differences of less than 20% (28) cannot be visually detected.

A measure of the fringe clarity is a quantity introduced by Michelson (see Born and Wolf⁽²⁷⁾, for example) and called visibility (V). This is given by:

$$V = \frac{I_{\max} - I_{\min}}{I_{\max} + I_{\min}} \quad (5.15)$$

where $I_{\max}$, $I_{\min}$ are the intensities of maxima and minima respectively.

Taking into account the Rayleigh criterion (20%) we obtained

$I_{\min} = 0.8 I_{\max}$ and a lowest visibility of $V_{\min} = 0.11$. In the case of the interference between two beams with $\frac{\pi}{4}$ phase difference, as just above B in figure 5.4, a visibility $V = 0.08$ is obtained since $I_{\min} = 0.85 I_{\max}$. This visibility is lower than the minimum. On the other hand the minimum $V_{\min} = 0.11$ corresponds to a phase difference of 53° . This means that this method is reliable up to 53° as far as visibility is concerned. Considering the commonly used criterion of $\frac{\pi}{4}$ for reasonable phase errors at the antenna aperture, this value of 53° is reasonable. On the other hand, it is only a possibility which in any case will reduce the antenna gain by a small amount.

Another disadvantage of this method lies in the size and position of the receiving aperture of the model. It is too small to satisfactorily interfere with the reference beam and give a reasonably sized hologram, while its position close to the diffracting obstacle makes it impossible to direct the reference beam there, at a reasonable angle, without being obstructed by the obstacle. Both of these difficulties can be satisfactorily overcome by placing an objective lens behind the aperture which will be magnified and at the same time relayed to a remote plane where the holographic procedure can take place. This is shown in figure 5.19 given in section 5.3.7.

5.3.6. Holographic Interferometry

A very important application of holography, because of its uniqueness in dealing with some problems, is holographic interferometry⁽³⁰⁾. This application will be used in this method and, therefore, will be briefly described here.

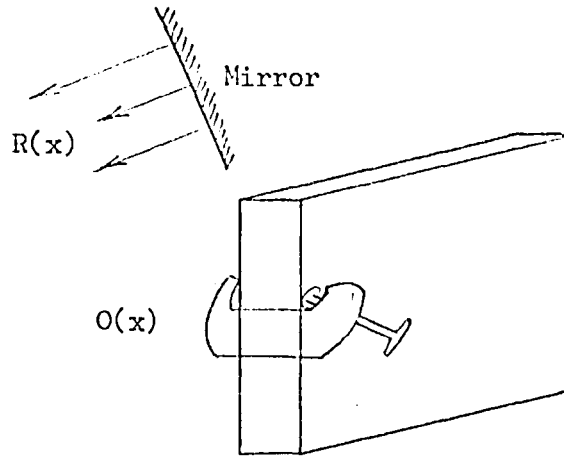
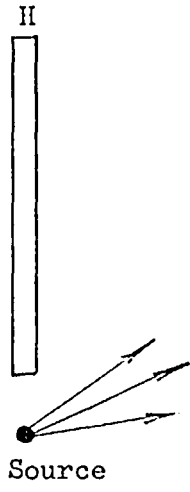
Use of interferometry, as described in section 5.2, is restricted to relatively simple cases where information about two dimensional

objects is needed. In addition the shape of the objects must remain unchanged with time. With holographic interferometry neither the shape nor the dimensions of the obstacle have to be specified because of the three-dimensional nature of the hologram. Complex cases can be examined from many different perspectives because a single interferometric hologram is equivalent to many views by conventional interferometry.

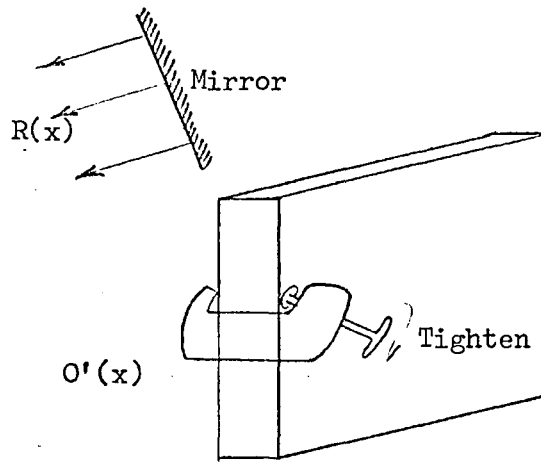
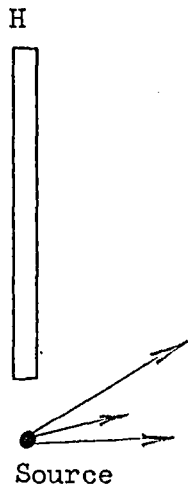
The most interesting difference between conventional and holographic interferometry, particularly in our case, is the possibility of detecting any changes on the object over a period of time. Using holographic interferometry we can compare the present object with its previous state. Comparing the picture of the receiving aperture when illuminated by an unobstructed beam with its picture when illuminated by the diffracted field will reveal the probable phase differences over the aperture in the second case, provided that the unobstructed beam is a plane wave.

There are many methods of holographic interferometry depending upon the technique employed in combining the reference beam with the wavefronts under investigation. Single-exposure and double-exposure methods are the simplest and commonest.

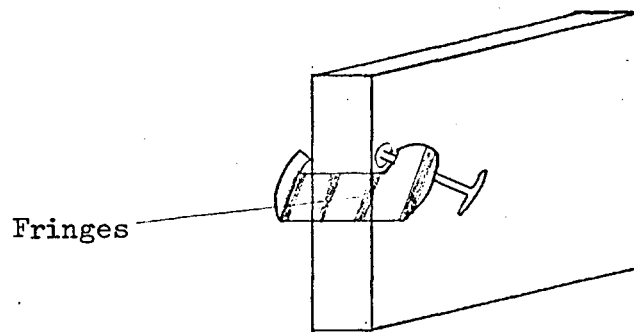
A simple example of the single-exposure method is given in figure 5.17. Wave $O(x)$, scattered by the illuminated object, interferes with the reference wave $R(x)$ at the hologram plane H . After development the hologram is placed in exactly the same position it occupied during exposure. By allowing for slight deformation in the object, as in figure 5.17(b) scattered wave $O(x)$ will be changed to wave $O'(x)$. These two waves have the same amplitude but slightly different phase distribution. Illuminating the hologram with the new object wave $O'(x)$, called the test wave, and the reference wave $R(x)$ simultaneously,



(a)



(b)



(c)

FIGURE 5.17

SINGLE EXPOSURE HOLOGRAPHIC INTERFEROMETRY

interference will take place between the reconstructed wavefront $O(x)$ and the directly transmitted $O'(x)$ to give fringes as in figure 5.17(c). These fringes are related to the deformation of the object surface. It is obvious that with this method many degrees of deformation can be investigated, since every one will produce a new object wave $O'(x)$ and accordingly a new set of fringes.

The double-exposure method is similar to the single-exposure one in most respects. However, placing the hologram in the exact position is no longer required. The main difference between these two methods is that, in the latter case, the shape of the object before and after the deformation is permanently recorded on the hologram, which is twice exposed before developing. The first exposure is made before the deformation of the object and the second one is made after the deformation. This hologram can be illuminated even on a separate arrangement for the reconstruction of the two permanently recorded wavefronts, which interfere automatically to reveal, in a form of fringes, any phase difference between them. The first of the two holograms, i.e. the first exposure, yields a primary image which constitutes the comparison wave, just as the wave $O(x)$ in the single-exposure method. The second hologram reconstructs the image of the changed object constituting the test wave $O'(x)$.

The position, shape and intensity of the fringes are related to the changes in the object over the period of time between the two exposures and must give a regular picture for a uniformly deformed object. Figure 5.18 shows the picture of a double-exposure hologram of a motor car tyre in which the air pressure in the tyre has been changed between the two exposures. The irregularity of the interference pattern, produced upon the reconstruction and shown by the arrows, reveal two flaws in the tyre, which did not expand or shrink uniformly when the air pressure was changed.

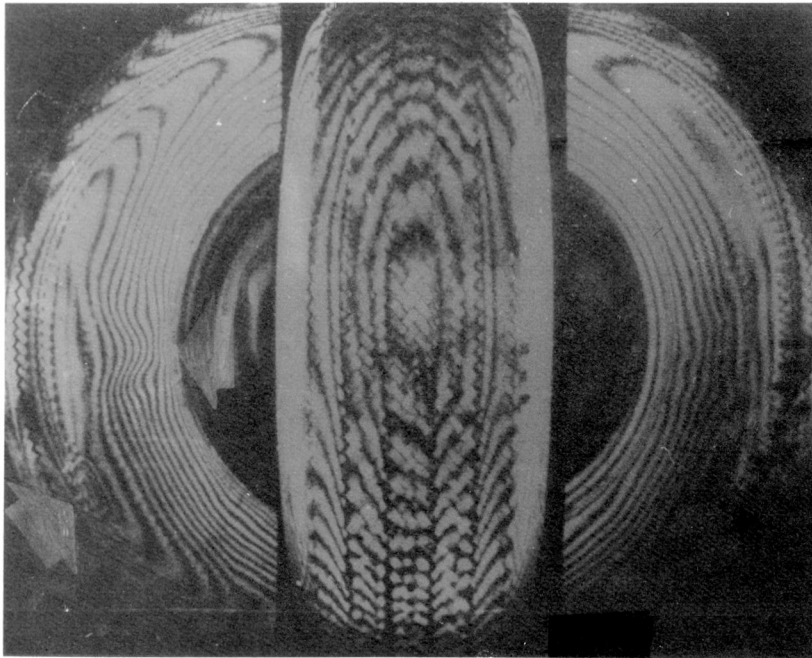


FIGURE 5.18

THE RECONSTRUCTION OF A DOUBLE-EXPOSURE HOLOGRAM

REVEALING TWO FLAWS IN A TYRE

[G.C. Optonics Inc., Ann Arbor, Michigan, U.S.A.]

5.3.7. Experimental Results

The optical model described in section 4.2 was modified as in figure 5.19 to carry out experiments using double exposure holographic interferometry.

Beam amplitude division was employed to obtain the reference beam, which was directed to the hologram plate at an angle of 10° due to the presence of the receiving aperture mount. A magnification of 60 was achieved for the receiving aperture at the hologram plate by an objective lens.

The first exposure hologram was made with the object beam

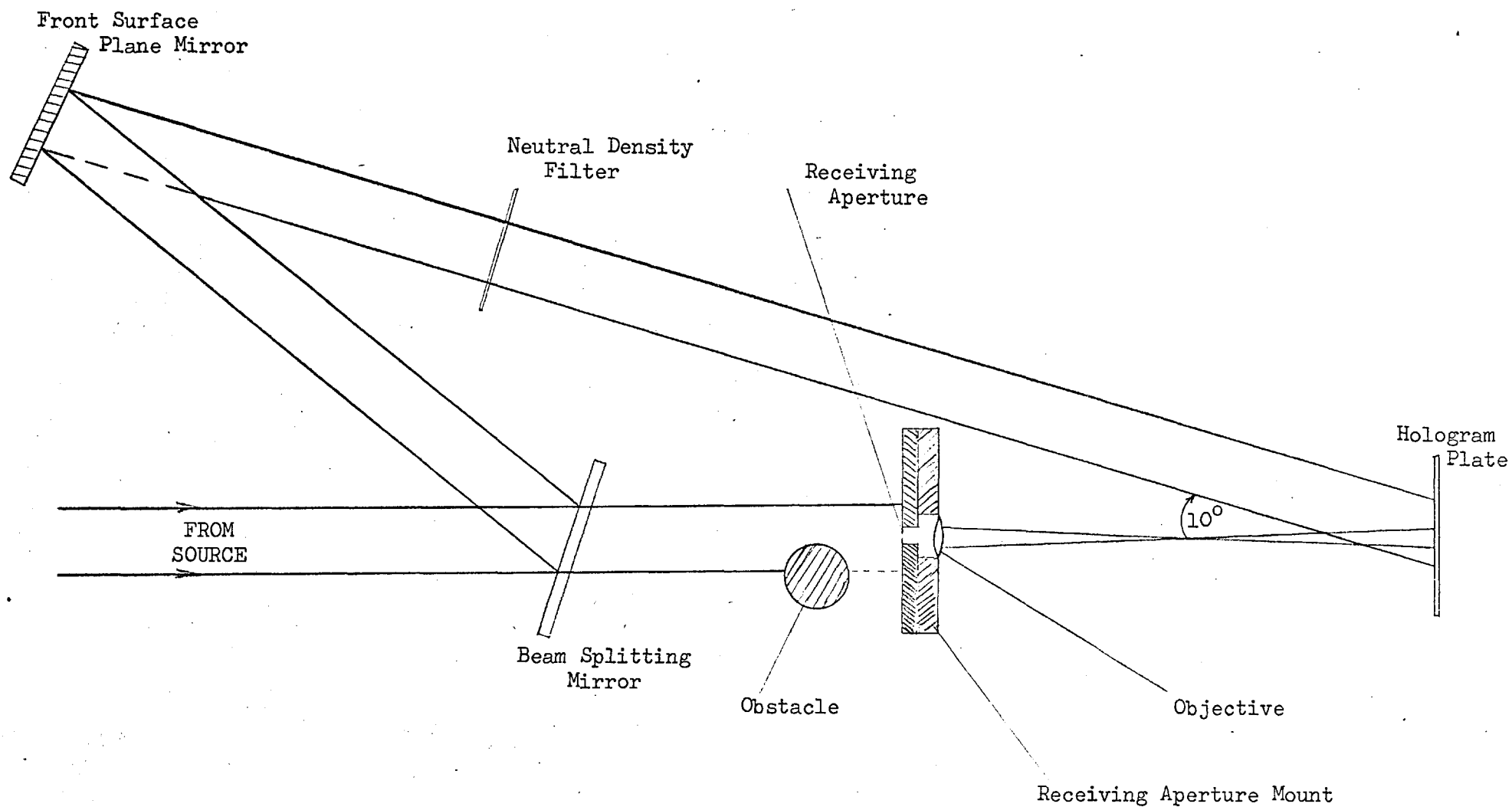


FIGURE 5.19

OPTICAL BENCH EQUIPMENT FOR PHASE CHANGES INVESTIGATION USING DOUBLE-EXPOSURE HOLOGRAPHIC

INTERFEROMETRY

unobstructed and the second one with the beam obstructed by a rough cylinder positioned 300 μm beyond its axial position. The reconstructed interference pattern was photographed through a microscope and is given in figure 5.20.

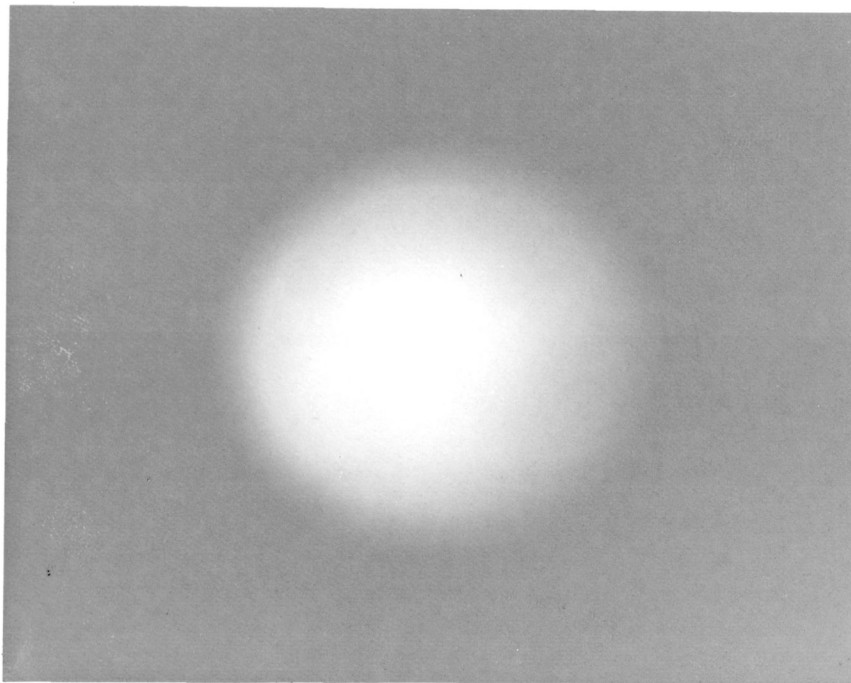


FIGURE 5.20

A DOUBLE EXPOSURE HOLOGRAM OF THE PHASE

DISTRIBUTION ON THE RECEIVING APERTURE OF THE MODEL

It can be seen that no visible fringes or intensity differences exist on the reconstructed interference pattern. Assuming that the unobstructed beam gives a uniform phase distribution (i.e. is a plane wave), it is therefore concluded that the phase distribution of a diffracted field on the receiving aperture is uniform within detectable limits.

The higher intensity in the centre is due to the amplitude taper of the transmitted beam.

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