Microwave path design: The basics

Part 1 — This four-part primer on propagation-related basics, for those learning about microwave communications, begins with "line-of-sight" determinations and the evaluation of path clearances with regard to refractive effects.

By Jeff Ashley

Microwave communications path design poses many challenges. In addition to *static gain and loss* considerations, *terrain* and *propagation dynamics* can play a large role in determining whether a proposed path will have the required signal levels, clearances and reliability.

This four-part series will only scratch the surface with regard to these considerations, and it is not intended to cover *all* aspects of path design or any one in great depth. The primary focus of the series is to serve as a primer on some of the propagation-related basics for those who may be new to the microwave communications field.

Given this disclaimer, the following tasks are some of the fundamental components of microwave path design:

- ❑ determining whether a proposed path is "line-of-sight."
- ❑ evaluating path clearances with regard to refractive effects.
- ❑ evaluating path clearances with regard to Fresnel zones.
- ❑ considering path reflections.
- ❑ deriving a power budget and the fade margin.
- ❑ path reliability.

Part 1 of this series deals with the first two components; the others will be discussed in Parts 2, 3 and 4.

Determining line-of-sight

Although there are situations where microwave paths are designed to work without line-of-sight conditions, they would not be considered the "norm" for industrial microwave systems and therefore will not be dealt with here. In most situations, if a prospective path is not line-of-sight, then an alternate route is considered.

Determining whether a path is line-of-sight can be partially accomplished with the aid of a *topographical map.* This type of map will show the various elevations along the length of the path between proposed endpoints. Plotting these elevations at intervals will produce a path profile showing terrain relative to the antenna elevations, as shown in Figure 1 below left. This graphical representation aids in determining not only whether a line-of-site condition exists between endpoints but also in measuring clearances between the main beam (center) of the path and the surrounding terrain.

Topographical information for many areas throughout the world is now available on CD-ROM. Various software applications have been developed to design microwave paths based on this information. Since this option can be a bit "pricey," many opt to use inexpensive topographical maps, which may be used with the same success.

When evaluating a proposed path, the *path profile* should be developed first. This will identify path obstructions from terrain features. A field survey should follow, which offers the necessary visual confirmation that the height of foliage or of man-made objects (which are not indicated on a topographical map) will not be located in or too near the proposed path.

If visual confirmation is difficult because of the distance between path endpoints, a mirror can be used to reflect the sun's light from one path endpoint back toward the other. If the mirror's flash of reflected sunlight is seen back at the opposite endpoint, the path is line-of-sight. This is referred to as "flashing" a path. Alternate methods

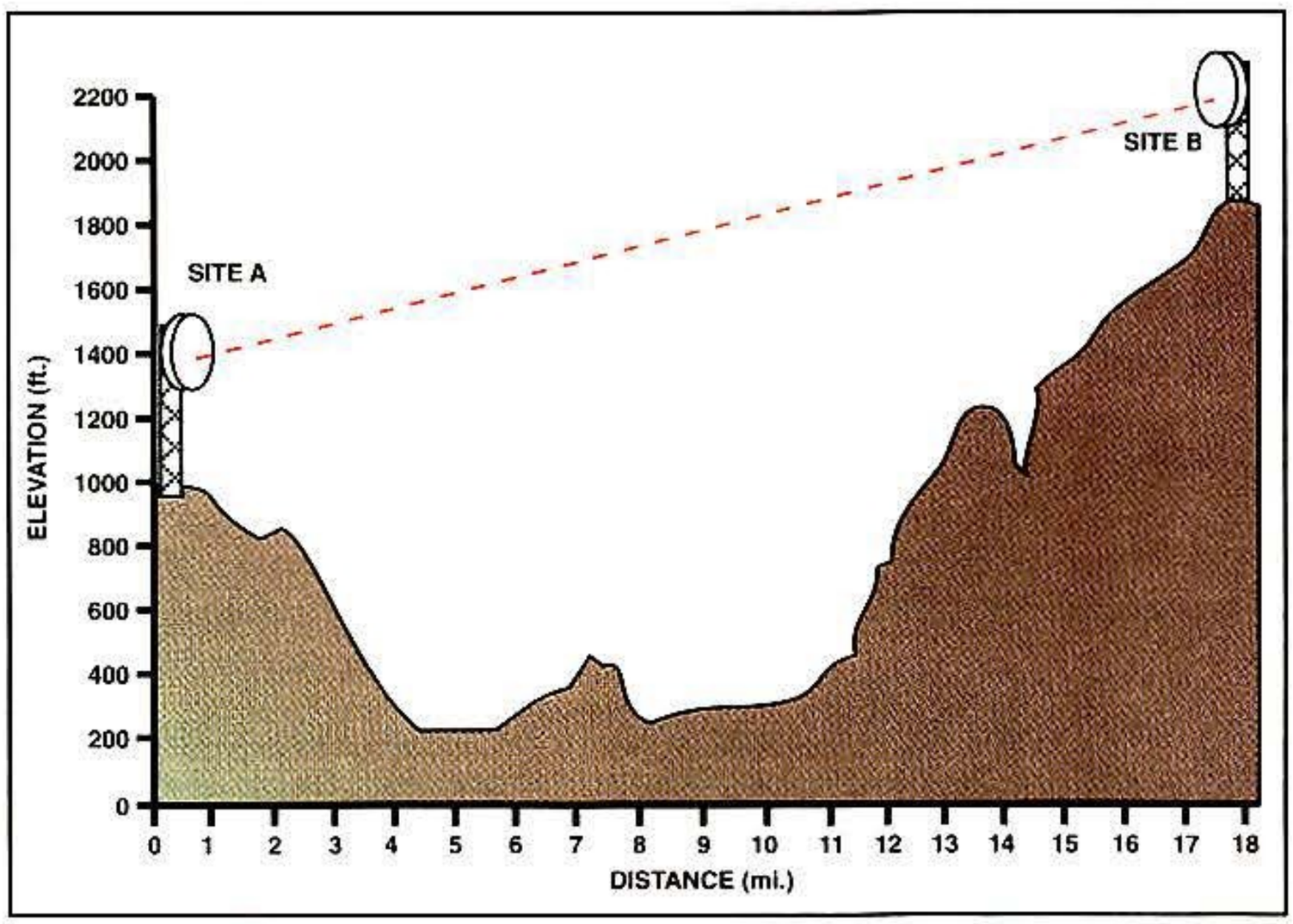

Figure 1. A typical path profile indicates terrain elevations relative to distance along the length of the path.

Ashley is a communications technician in Los Angeles.

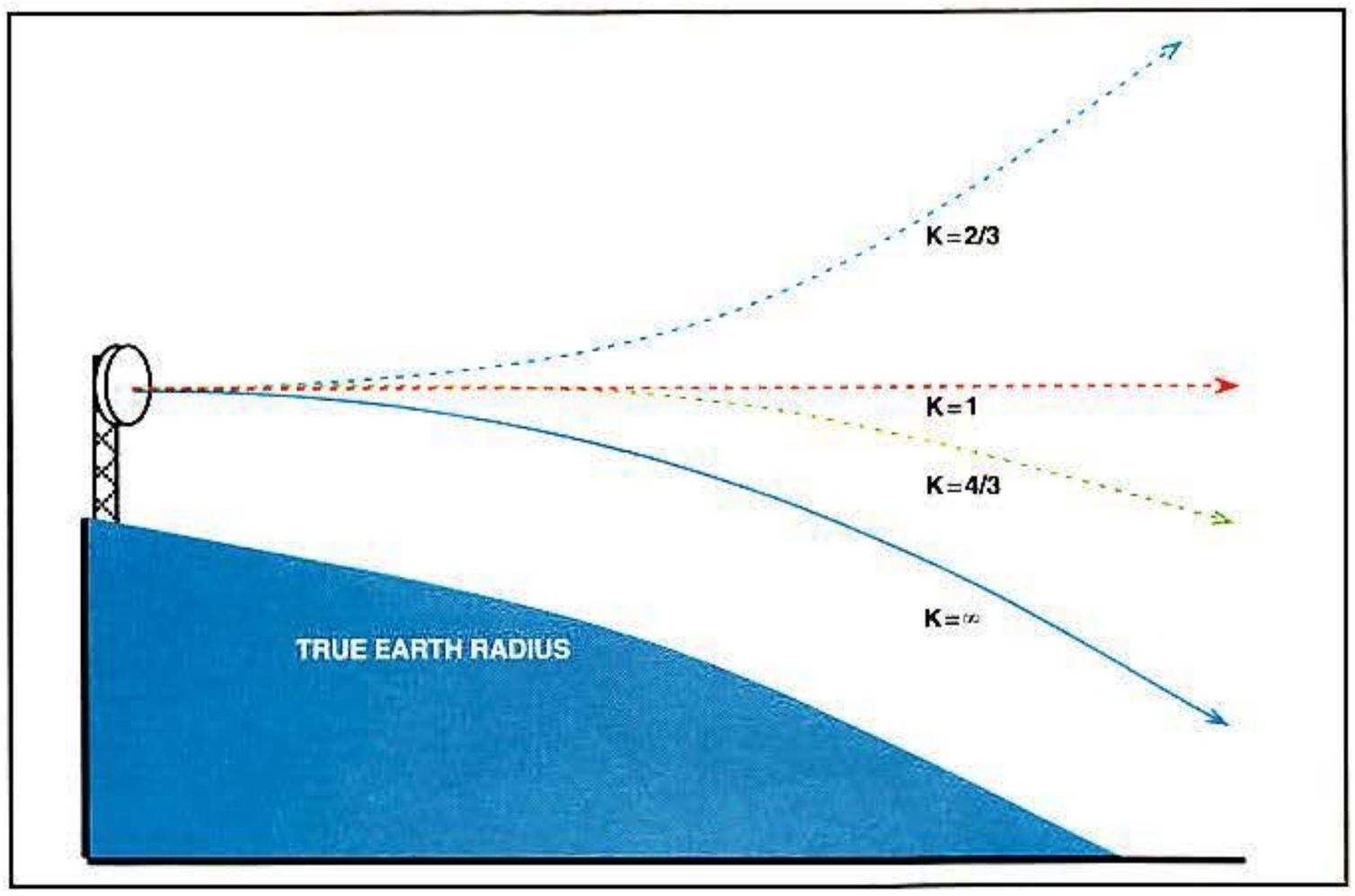

Figure 2. Where *K* equals infinity, the wavefront parallels the radius of the true earth. When no refractive effects are present (*K* = 1), no bending of the wavefront occurs.

might include the use of high-intensity search or strobe lights.

Evaluating path clearances—refraction

When the *dielectric constant* of the atmosphere changes with height-above-ground, the *refractive index* will also change. This refractive variation in turn causes the propagating wavefront to effectively "bend." If a condition exists where the dielectric constant of the atmosphere is constant with height-above-ground, no refraction will occur and the wavefront will travel in a straight line. This is not the "norm," however. Because of dielectric variations typically present, a wavefront usually will be refracted so that it will follow a path somewhere between a straight line and the true curvature of the earth (referred to as *R*). A wavefront propagating through the environment will be slowed by, and be effectively bent toward, an atmosphere that is more dense, such as one with a high water-vapor content (dense clouds or fog). That is why, rather than traveling in a straight line, the wavefront is normally bent earthward—which serves to extend its horizon.

Because atmospheric conditions are dynamic, the bending effect of the propagating wavefront will vary. To provide a simple model describing the path traveled by a wavefront for a particular refractive condition, engineers have developed a factor, *K*, where *KR* is considered the *"equivalent" earth radius*. This equivalent radius describes the bending of the wavefront relative to the *true earth radius*, as shown in Figure 2 on page 28.

The refractive index can change

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drastically with time. This means the microwave beam between path endpoints will "bow" to a greater or lesser extent. In fact, it may "bow" in an upward *or* a downward direction, depending on the value of K at the time. It is therefore important to evaluate clearances from path obstructions over a wide range of K in order to determine whether adequate path clearances are maintained under various degrees of refraction. Often, three values of K are used in the calculations. Two of the three values describe the *limits* or boundaries of refraction that might occur (although with anomalous conditions, values exist outside these limits), while the third describes what is considered "normal" or expected. The value of $K = \infty$, also known as "super-standard" atmosphere, is one extreme where the wavefront follows the true curvature of the earth. The other extreme value of K typically used is 2/3 and is termed "substandard" atmosphere. It is also a condition commonly referred to as "earth bulge." The median value of $K = 4/3$ is used to evaluate the path under "normal" atmospheric conditions in temperate climates.

There are two methods of graphically displaying the various *equivalent* earth radii with respect to the terrain between path endpoints: the "flat earth" method and the "curved earth" method. Figure 3 above is an example of a "flat earth" path profile. It is called "flat earth" because the horizontal axis (or baseline) has been kept flat and

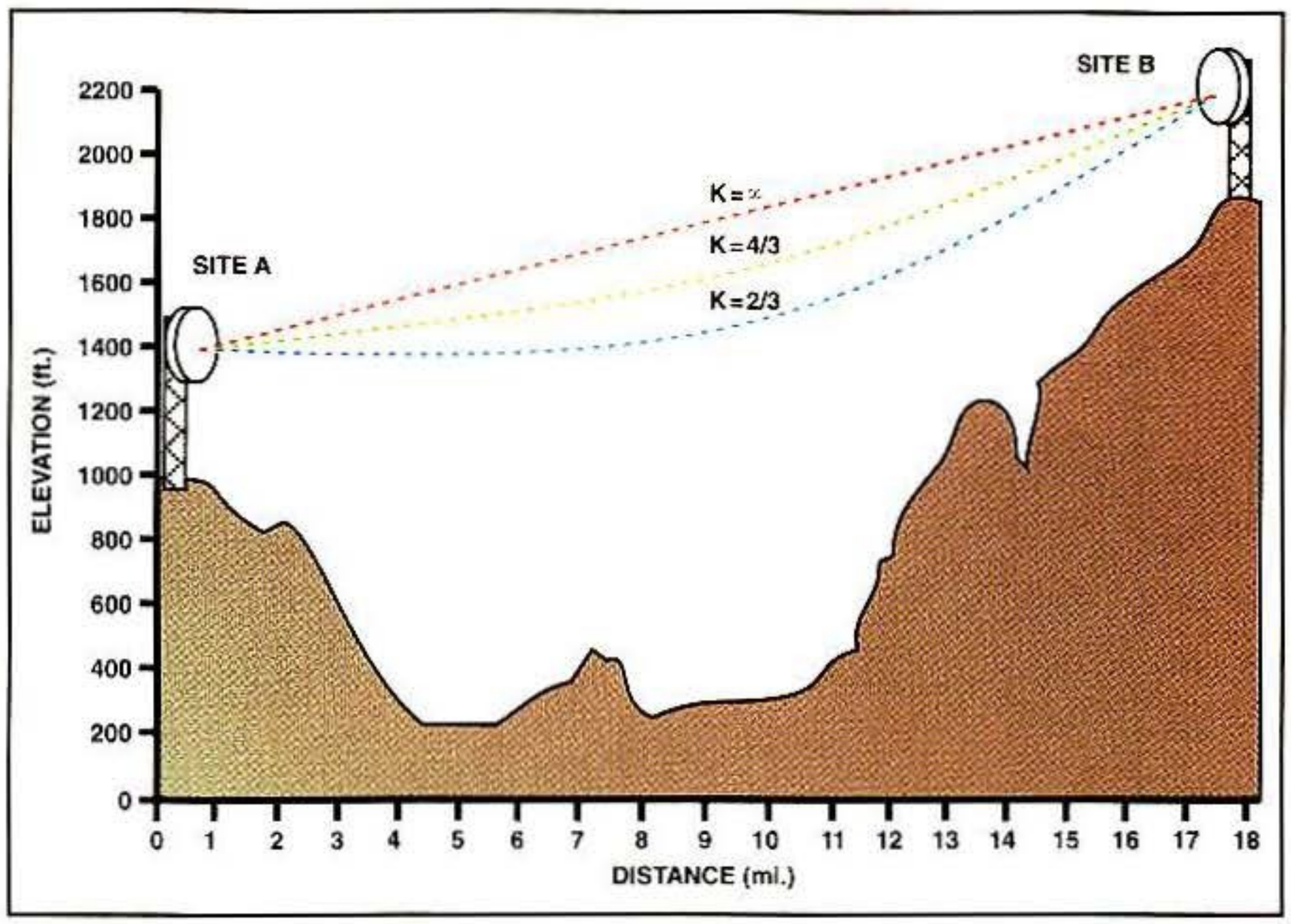

Figure 3. This is a 'flat-earth' profile. Terrain features become more of an obstruction consideration when $K = 2/3$.

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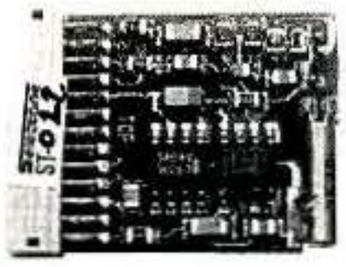

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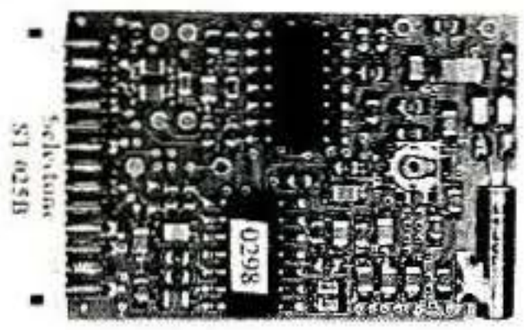

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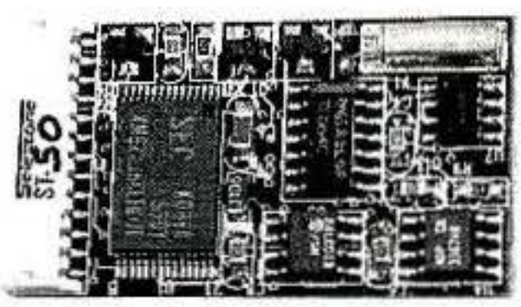

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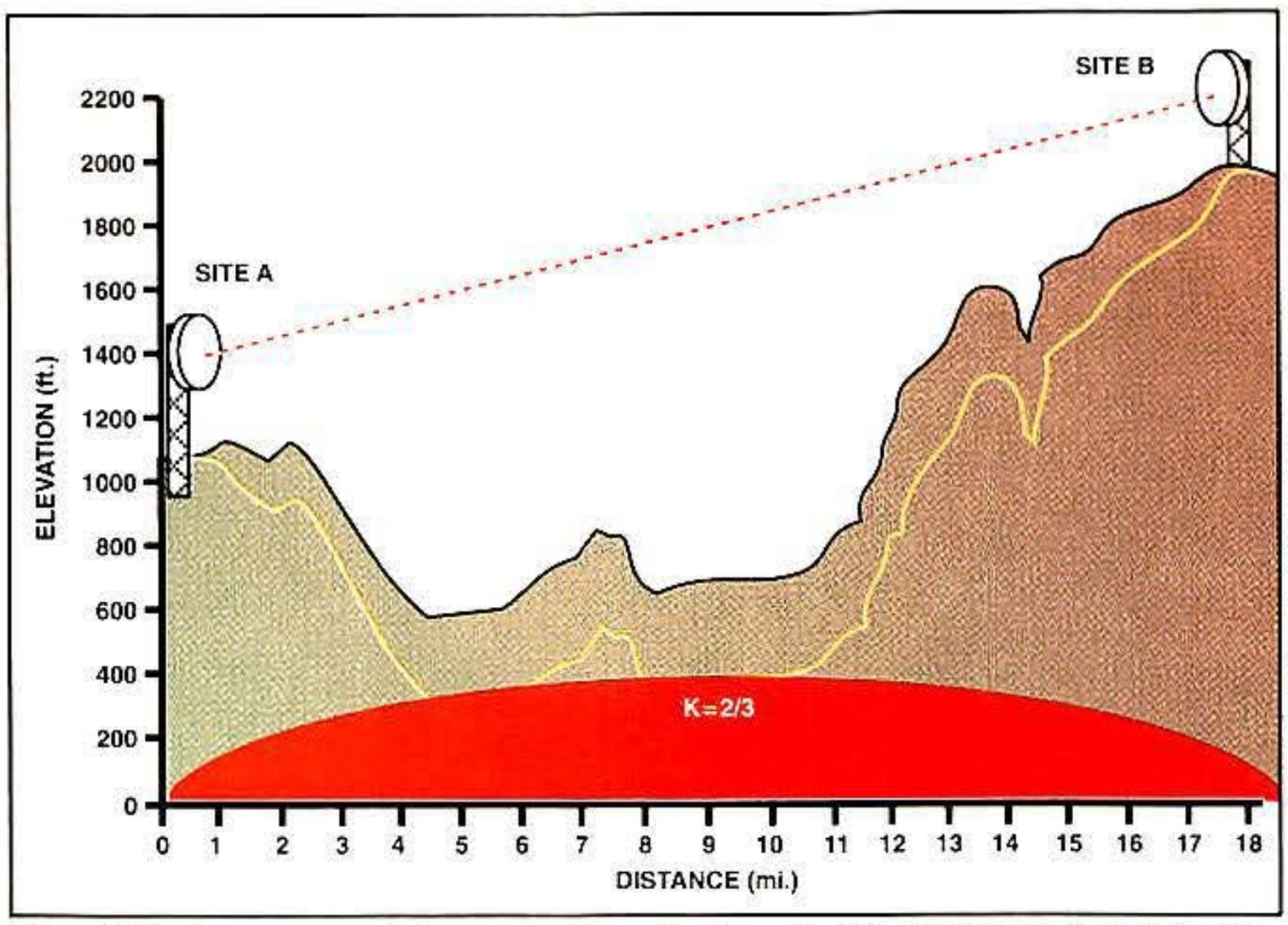


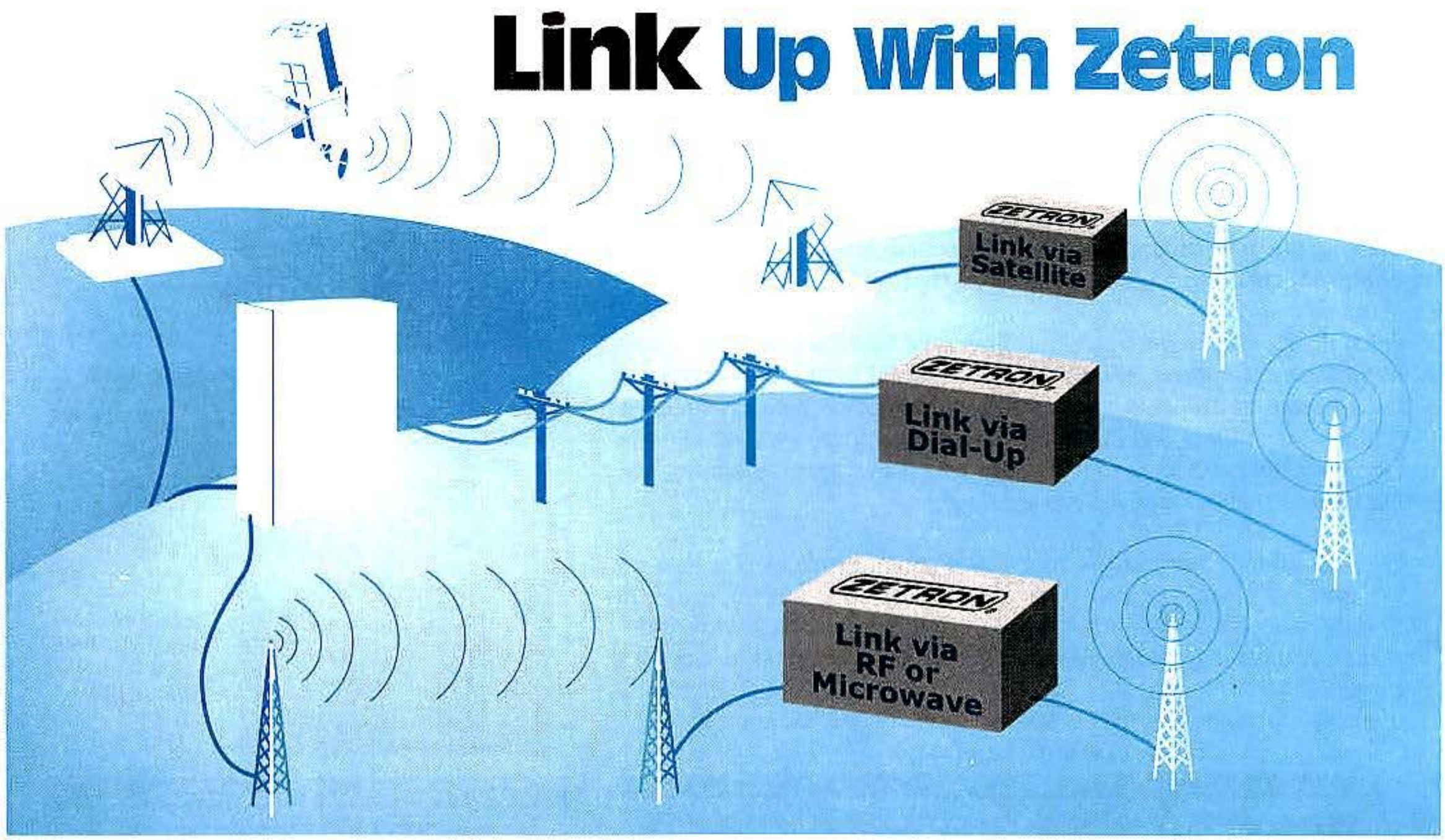
Figure 4. This is an exaggerated curved-earth profile where $K = 2/3$, with identical terrain features as shown in Figure 3. (Figure 3 elevations are represented by the yellow line.) Because the equivalent earth radius 'bulges up' in the center, the elevations of the terrain features increase, *reducing* path clearances. When $K = 4/3$, the earth bulge is less severe, resulting in greater path clearances. The greatest clearance is realized when $K = \infty$.

represents the earth's curvature as a flat line. The various equivalent earth radii represented by the different values of K are shown by bending the path beam line between antennas. The three lines shown in Figure 3 correspond to the refractive conditions where $K = \infty$, $K = 4/3$ and $K = 2/3$.

Elevations of terrain features between path endpoints are plotted using the flat baseline as a reference. The profile allows one to see if there is sufficient path clearance over the terrain for each value of K.

The curved earth approach, as shown in Figure 4 on the left, leaves the beam line between antennas straight and instead bows the horizontal baseline to correspond to one particular value of K. Elevation of all terrain features must be added to that of the bowed baseline reference. The straight beam line between antennas must now be checked for sufficient clearance from all terrain features for that value of K.

With the curved earth profile, the one limit of $K = \infty$ would be shown with the baseline flat. Here (as with the flat-earth approach) is the one "extreme" condition where the wavefront (shown as a straight line) follows



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the true curvature of the earth (also shown as a straight line). As the value of K gets smaller and reaches the other limit of K=2/3, the equivalent earth radius bulges upward more in the center, hence the term "earth bulge." As these profiles show, paths with values of $K = 4/3$ and $K = 2/3$ have less path clearance with respect to terrain features than when $K = \infty$.

The following equation can be used to calculate how much the horizontal axis (or baseline) of a "curved earth" profile will bow upward at any point along the path for a particular value of K:

$$h = \frac{d_1 \times d_2}{1.5 \times K}$$

where h is the vertical variation in height (ft.).

d_1 is the distance (mi.) from one end of the path to the point being considered.

d_2 is the distance (mi.) from the point considered to the other end of the path.

With the "flat earth" profile, h is subtracted from the direct ray ($K = \infty$) reference line (for positive values of K) at various distances along the path in order to develop a graphical representation of the amount of wavefront bowing that will occur for the value of K under investigation.

With the "flat earth" profile, the wavefront, rather than the baseline, is shown "bowed," and the amount of bowing changes as the value of K changes. The distance between the bowed wavefront and the terrain features can be measured graphically to determine path clearances for each value of K.

Whichever profiling method is used ("curved earth" or "flat earth"), it can be seen graphically that smaller values of K result in greater bulging or "bowing." This in turn results in less path clearance with regard to terrain features.

... Smaller values of K result in greater bulging or 'bowing.' This in turn results in less path clearance ...

Because the microwave path has *radial dimension*, the clearances considered are not only those above and below the center of the path, but also to either side.

Initially, these graphical methods allow path clearances to be measured as distances, *in feet*. Once this information has been obtained, it is however, typically converted to a more meaningful design measurement, called the *Fresnel zone*.

The next article in this series will discuss path clearances with regard to Fresnel zones.

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Microwave path design: The basics

Part 2 —This four-part primer on propagation-related basics, for those learning about microwave communications, continues with a discussion of path clearances and Fresnel zones.

By Jeff Ashley

In Part 1 of this series (*MRT* May), two of the basic propagation-related considerations in the microwave path design process were discussed:

❑ determining whether a proposed path is "line-of-sight."

❑ evaluating path clearances with regard to refractive effects.

Remaining fundamental design components to be discussed are:

❑ evaluating path clearances with regard to Fresnel zones.

❑ considering path reflections.

❑ deriving a power budget and the fade margin.

❑ path reliability.

This article will briefly cover item three, path clearances and Fresnel zones. The remaining items will be dealt with in the concluding parts of this series.

Evaluating path clearances—Fresnel zones

If one were to look directly into the center beam of a microwave path (gazing directly into the dish), the Fresnel zones could be defined as an infinite number of circles centered around the center beam, each having a larger radius than the previous one, as shown in Figure 1 below. The Fresnel zones get closer as they expand further away from the center. The radii of these zones reach a maximum at mid-path and diminish on either side of mid-path.

From a side view of the path, as shown in Figure 2 on page 30, one can see the creation of ellipsoids between path endpoints, centered around the main beam. Each ellipse represents the boundary of a particular Fresnel zone.

Fresnel zones are numbered from the center of the path out. The boundary of the first radius out from the center represents the first Fresnel zone, which contains all wave components whose distance traveled to the far-end is exactly 1/2 wavelength (λ)greater than the direct ray. The boundary of the second radius out from the center represents the second Fresnel zone, which contains all wave components whose distance traveled to the far-end is exactly 1λ greater than the direct ray. Each integer increase of Fresnel zone number represents an increase of 1/2λ to the distance traveled to reach the other end of the path.

The equation that determines the radius of the first Fresnel zone (in feet) at any particular distance along the path is:

$$F_1 = 72.1\sqrt{\frac{d_1 \times d_2}{f \times D}}$$

where

F_1 = the first Fresnel zone radius (ft.).

d_1 = distance (mi.) from one endpoint to the point being considered.

D = total path length (mi.).

$d_2 = D - d_1$.

f = frequency (GHz).

The radius of *any* Fresnel zone number can be calculated by the following equation once the radius of the *first* Fresnel zone has been determined:

$$F_n = F_1\sqrt{n}$$

where n = the Fresnel zone number being evaluated.

When examining path profiles, if one wishes to know what Fresnel zone is represented by a particular clearance (in feet), the

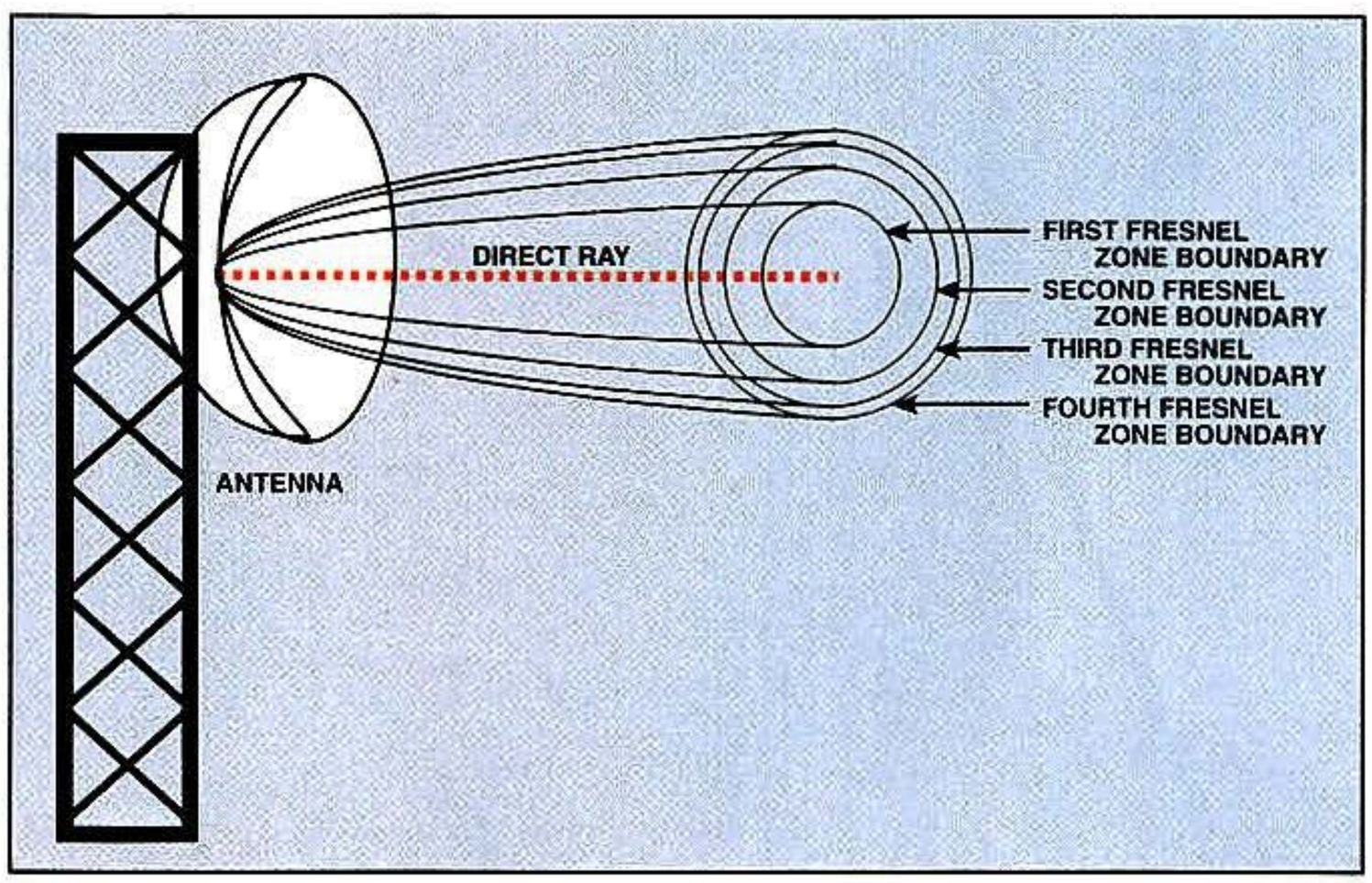

Figure 1. Fresnel zones could be viewed as an infinite number of circles centered around the direct ray, each circle having a larger radius than the previous one. Fresnel zones get closer together as they expand out further from the center.

Ashley is a communications technician in Los Angeles.

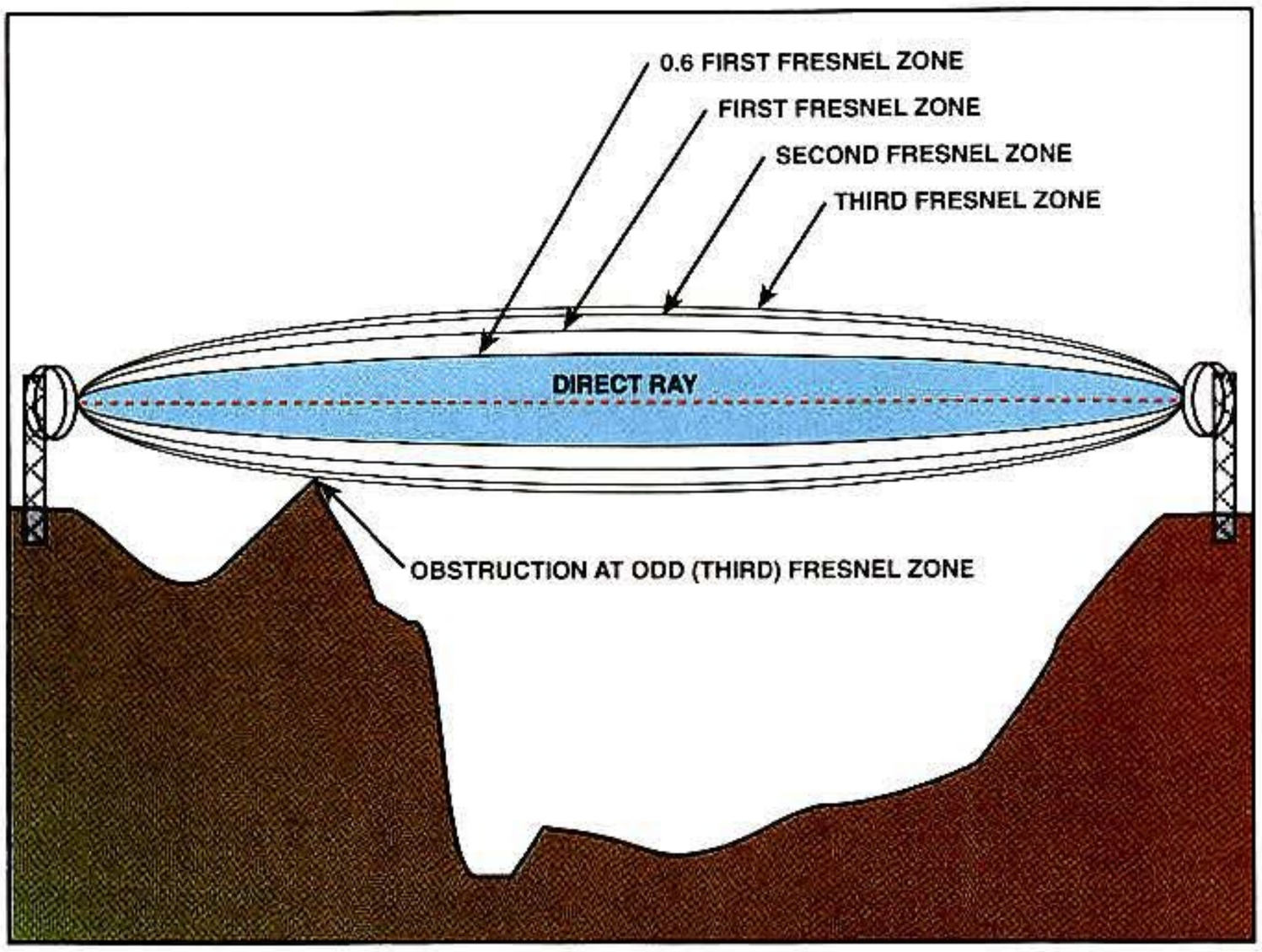

Figure 2. The Fresnel zone radii decrease with distance from mid-path. The Fresnel zone boundaries are defined by ellipsoids around the direct ray. Here, a path obstruction is shown with third Fresnel zone clearance. Because this is an odd Fresnel zone, signal energy reflected by the obstruction surface will be in phase and will add with the energy from the direct ray as it arrives at the far-end receiver. How much energy is reflected depends on the coefficient of reflectivity of the obstructing surface.

following form comes in handy:

$$\text{Fresnel zone \#} = \frac{\text{clearance}^2(\text{ft.}) \times f \times D}{72.1^2 \times d_1 \times d_2}$$

In general, if the path is line-of-sight with adequate clearance, the loss in RF signal between one end's transmitter and the other end's receiver will merely be that of free space, as described by the equation:

$$FSL\ (\text{dB}) = 96.6 + 20 \log_{10} F + 20 \log_{10} D$$

where
FSL (dB) = Free Space Loss (dB).
F = Frequency (GHz).
D= Distance (mi.).

As indicated in Figure 3 on page 32, if a path has less than 0.6 first Fresnel zone clearance, there will be more loss than merely that of free space, even though a line-of-sight condition may exist. With *odd* Fresnel zone clearances something *less* than free space losses may be realized, while with *even* Fresnel zone clearances losses may be quite significant. This is because of the phase relationship between the direct and reflected

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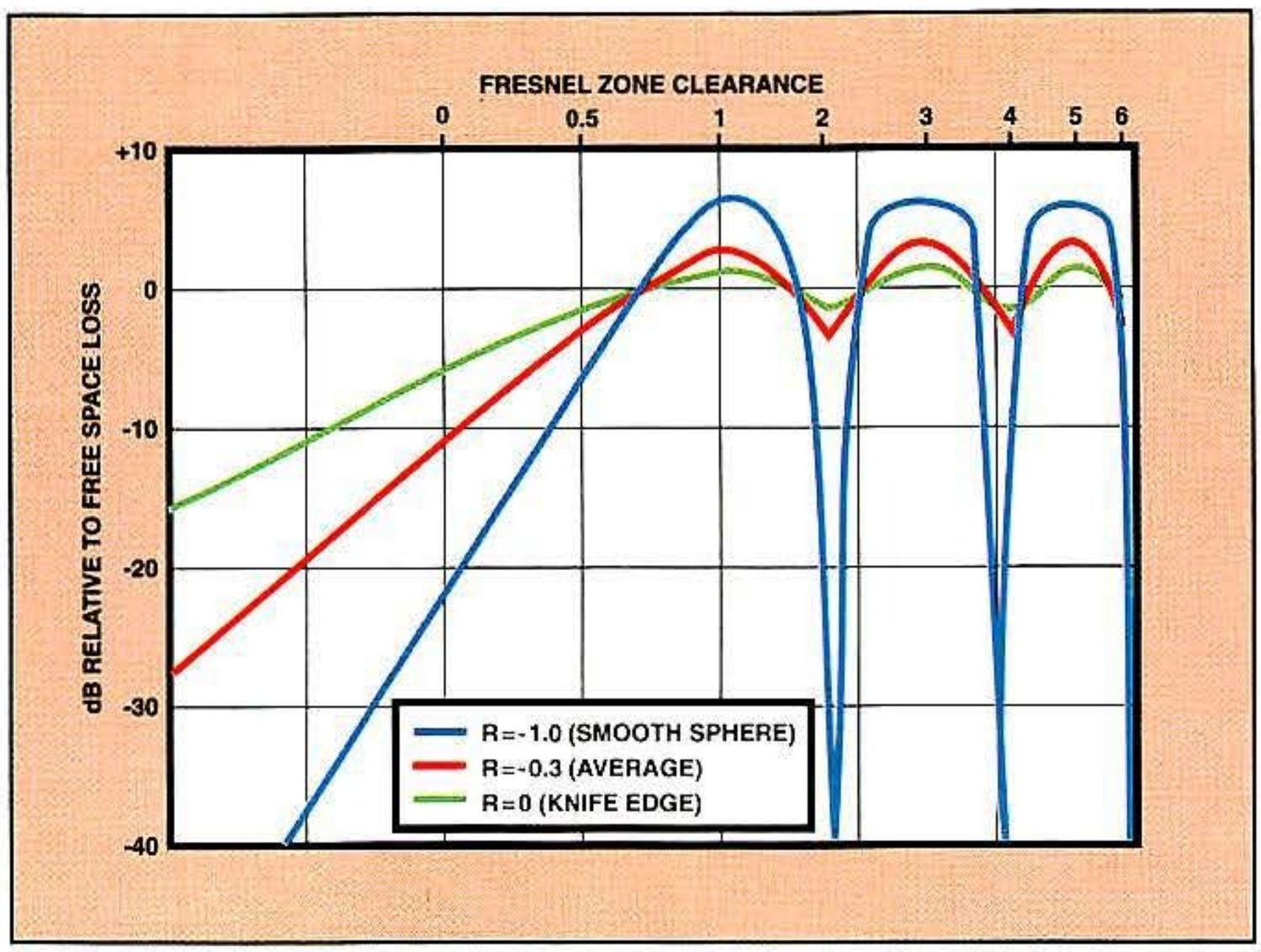

Figure 3. Deep fades result when out-of-phase reflected energy combines with the incident signal at the receiver, increasing the cancellation of the RF receive signal.

signal at the receiving antenna.

When the signal is reflected, it undergoes a 180° phase lag (equivalent to 1/2λ delay) referenced to the incident wavefront. (Although, as will be mentioned in Part 3 of this series, wave *polarization* and grazing angle can affect this). If the reflective surface is located at the first Fresnel zone, the reflected signal will arrive at the receiver delayed by another 1/2λ, for a total delay of 1λ. Signals reaching the receive antenna are therefore in phase and additive.

If the reflective surface is located at the boundary of the second Fresnel zone (which by definition, delays the signal by two 1/2 wavelengths) the wavefront, when reflected, reaches the receiving antenna delayed by a total of 1.5λ. This means it will be out of phase relative to the incident wavefront, and signals reaching the receiver will cancel to some extent.

This effect repeats for each odd and even Fresnel zone, which explains the slight rises and very deep nulls in signal level for the *even* and *odd* Fresnel zone clearances (as shown in Figure 3). Raising one or both antennas is one method of altering the Fresnel zone clearance. As the clearance increases to higher order Fresnel zone numbers, the intensity of this problem diminishes because the magnitude of the coefficient of reflection with regard to the reflecting surface diminishes.

Paths are usually designed to meet specific Fresnel zone clearance criteria for one or more values of *K*. "Heavy route"

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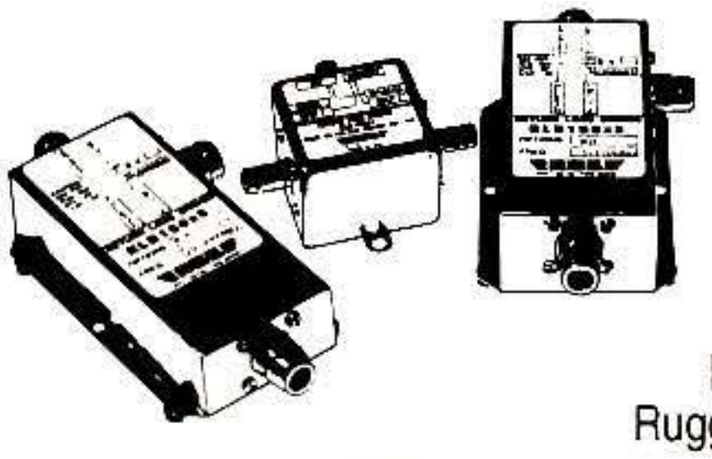

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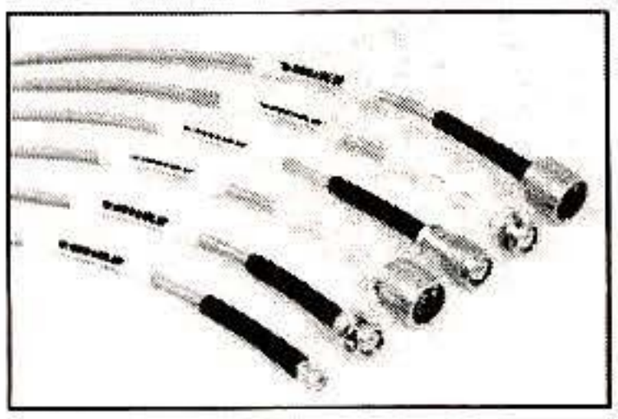

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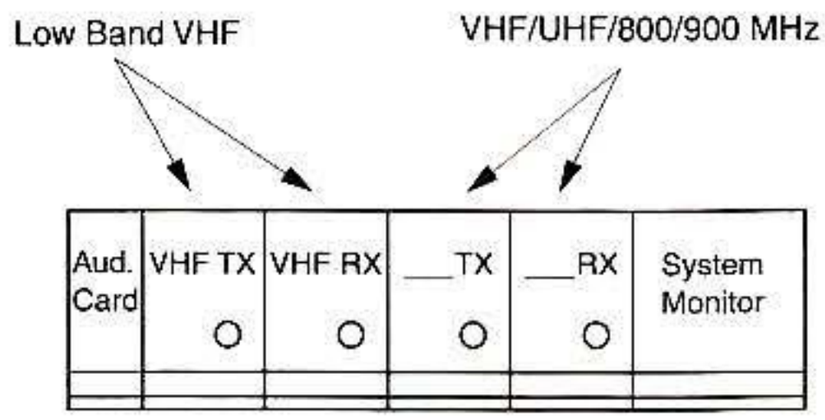

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- ☆ If you use a scrambler for privacy on a shared channel, sign off "in the clear" so other users know you are finished.
- ☆ Have your radio serviced periodically to be sure it is on frequency.
- ☆ Keep communications brief to promote harmony with other users who share the channel.
- ☆ Use the minimum transmitter power and antenna height you need to cover your area. That might be less than the power and antenna height specified on your license.

System operators:

- ☆ Post the name and 24-hour phone number of the site manager at the access point for each antenna site.

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requirements dictate that there must be at least 0.3 first Fresnel zone clearance when $K = 2/3$, and 1.0 first Fresnel zone clearance when $K = 4/3$, whichever is greater. "Light route" requirements dictate at least 0.6 first Fresnel zone clearance plus ten feet when $K = 1$.

Another important consideration here is what *type* of reflecting surface is being evaluated. In general, three categories of surfaces are defined, each with its own reflection coefficient: $R = 0$ (a sharp or "knife" edge), $R = -0.3$ ("average" roughness) and $R = -1.0$ (a smooth spherical surface). (The reflection coefficient will be described in more detail in the third article of this series.) The type of surface makes a big difference with regard to attenuation on the even Fresnel zone clearances because obstructions with larger, smoother surface areas reflect more energy. With more out-of-phase reflected energy combining with the incident signal at the receiver, greater cancellation of RF receive signal will be realized, resulting in deep fades as shown in Figure 3.

The next article in this series will briefly discuss path reflections, the power budget and the fade margin.

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Microwave path design series

Part 1: "Microwave path design: The Basics," May 1998.

Part 2: "Microwave path design: The Basics," June 1998.

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Microwave path design: The basics

Part 3—This series of four articles is a primer on some of the propagation-related basics of microwave path design. This month's installment explains the effect of path reflections on fading, how to derive a power budget and the fade margin.

By Jeff Ashley

In the first two parts of this series (*MRT* May and June), the microwave path design process was described in terms of:

- determining whether a proposed path is "line-of-sight."
- evaluating path clearances with regard to refractive effects.
- evaluating path clearances with regard to Fresnel zones.

This section covers path reflections and how to derive a power budget and fade margin. The concluding installment will examine path reliability.

Considering path reflections

A point of reflection will exist somewhere along the length of the path where (with regard to the reflecting plane and antennas at both ends) the angle of incidence equals the angle of reflection (see Figure 1 below). If antennas at each end of the path are at the same height above a flat, reflecting surface, the reflection point will be located halfway between the two. However, if antenna elevations are different, the reflection point will be closer to the lower antenna and will move along the length of the path, toward one endpoint or the other as K varies ("K" was defined in Part 1 of this series). If sufficient out-of-phase signal from the reflection point is received at the far-end receiver, catastrophic fading may occur.

In some cases the reflection point will be blocked naturally by terrain features. Manipulation of antenna heights is a method of moving the reflection point to a different spot along the length of the path so that it can be blocked by an obstruction or perhaps so that it may fall on a less reflective surface. Rough surfaces do not reflect as well as smooth, flat surfaces, such as water.

The magnitude of the reflection coefficient (amount of "reflectivity") also increases as the angle of incidence decreases. Because of this, short paths with endpoints at a considerable height above the reflecting surface will be affected less than those with endpoints barely higher than the reflective surface itself (a microwave link between two offshore oil platforms, for example). The magnitude of the reflection coefficient can vary from 0 to −1.0, the latter corresponding to 100% reflection. The negative sign refers to a 180° phase reversal at the point of reflection.

There is an advantage to using vertical polarization when the grazing angle on a reflective surface is a couple of degrees or greater. The magnitude of the reflection coefficient with vertical polarization is much less than that of horizontal polarization. Less reflected signal will be present at the receive antenna on the other end of the path. In fact, with a grazing angle of 6.5°, the magnitude of the reflection coefficient for a vertically polarized wave is only about −0.08 compared to about −1.0 for horizontal polarization. The magnitude rises as the grazing angle increases above 6.5° until it finally meets the same value of the horizontally polarized signal (at 90°).

The phase change (or lag) at the point of reflection for a vertically polarized signal is also significantly affected at low grazing angles as compared to the practically uniform 180° phase reversal for a horizontally polarized signal. As the angle increases from 1° to 30°, the phase change of a vertically polarized signal drops quickly from about 180° to 4°. With only a 4° phase lag, the reflected signal that reaches the receive antenna on the other end of the path would have more of an additive, rather than canceling effect.

The distance to the reflection point can be calculated using the equation

$$d_1 = nD \qquad \text{[Eq. 1]}$$

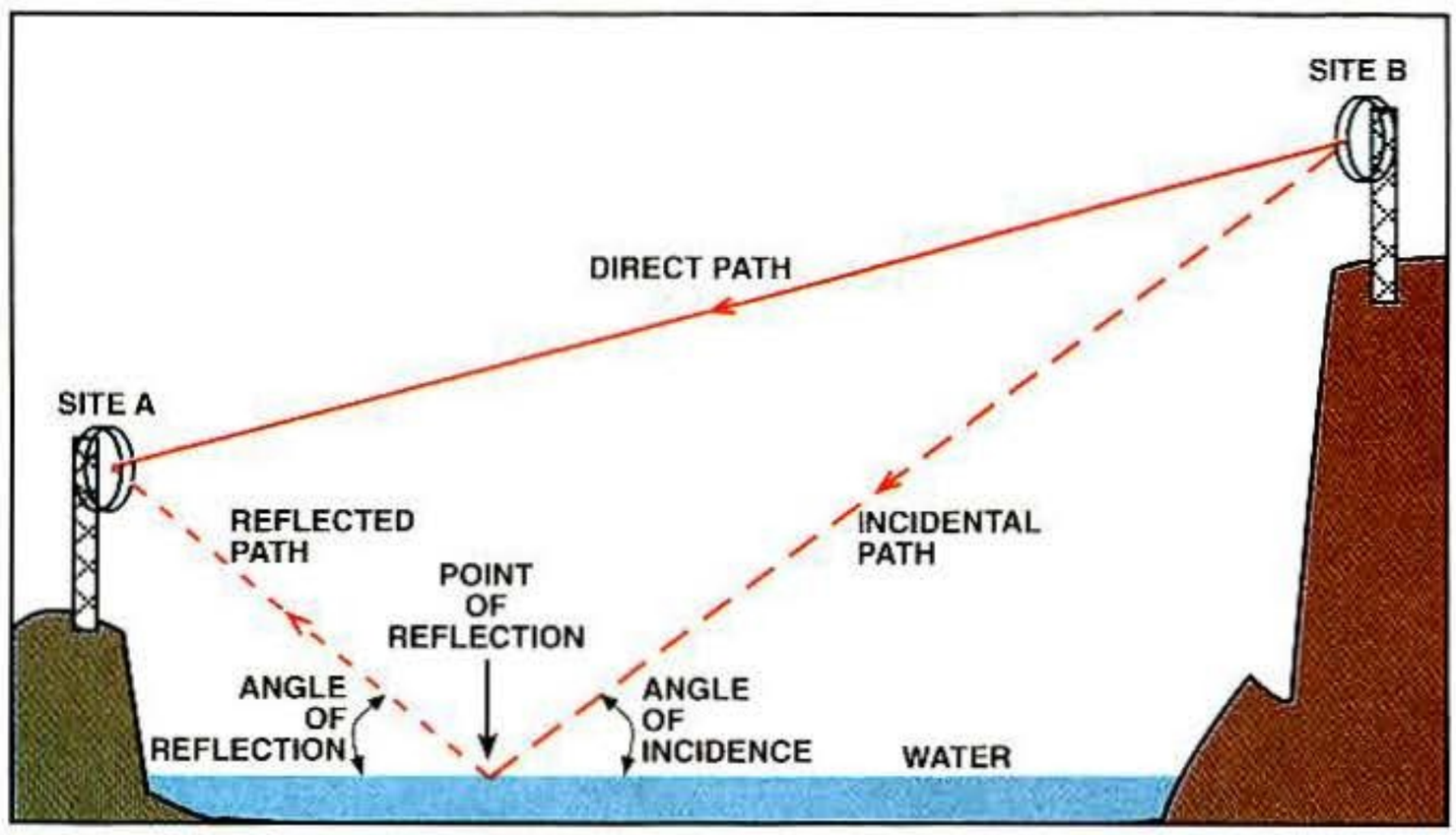

Figure 1. At the point of reflection, the angle of incidence equals the angle of reflection.

Ashley is a communications technician in Los Angeles.

where (for $K = \infty$ *only*):

$$n = \frac{h_1}{h_1 + h_2}$$

and

h_1 = the elevation of the lower antenna above the reflecting surface.

h_2 = the elevation of the higher antenna above the reflecting surface.

d_1 = the distance (mi.) to the reflection point from the h_1 end.

D = the total path length.

Because calculations for n, where $K = 2/3$ and $K = 4/3$, get a bit involved, estimated values can be obtained from graphs offered in R.F. White *(see references)*.

As refractive conditions change, the degree of bowing that occurs on the wavefront will change. As the amount of bowing changes, the location of the reflection point will vary in relative distance between the two path endpoints. In designing a path, it is useful to analyze the overall range of movement (or distance of travel) that can occur for this reflection point. The design engineer must determine whether (in its range of movement) it may wander onto a highly reflective surface (such as a body of water) that lies between path endpoints. In such an undesirable situation, measures might be taken to move the reflection point elsewhere by changing antenna heights, or by attempting to block the reflection.

To determine the practical limits of travel for the reflection point, two values of K are typically used that would describe limits of refractive conditions. One limit is when refractive conditions correspond to a value of $K = \infty$. The other extreme limit will be the value of K that would bend the path downward to the point of grazing the ground. Beyond this grazing point, no reflection can exist because the path itself is being obstructed by the earth.

The value of K that corresponds to "grazing" can be calculated from the equation:

$$K = \frac{1}{1.5\left(X + Y + 2\sqrt{XY}\right)} \quad \text{[Eq. 2]}$$

where

$$X = \frac{h_1}{D^2} \quad \text{and} \quad Y = \frac{h_2}{D^2}$$

(h_1, h_2 and D are already defined in Equation 1).

The distance to the grazing point is calculated from Equation 1, where:

$$n = \frac{1}{1 + \sqrt{\frac{Y}{X}}}$$

Deriving a power budget

A path power budget is nothing but an itemized list of all system losses and gains (in decibels) from the transmitter on one end of the path to the receiver on the other end, and everything in between.

For example, if starting with the transmitter on one end of the path, one would note transmitter output power (+dBm), then subtract all the losses in subsequent connectors and circulators external to the radio assembly, along with waveguide losses (−dB) until the antenna is reached. The gain of the antenna would be added to that value (minus the loss in the radome, if used) to give the resultant effective radiated power (ERP.). This is the available RF signal that will be propagated through the atmosphere to the other

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end of the link. As the signal makes its way through the atmosphere to the receiving site, free space loss (−dB) (the formula for this was given in Part 2, *MRT* June), in addition to atmospheric losses (−dB), would then be subtracted from the ERP value.

The term *atmospheric loss* or "absorption" pertains to effects that atmospheric gases, primarily oxygen and water vapor, have on microwave propagation. These effects are calculated separately from free space losses. Attenuation figures are directly proportional to path length and increase with frequency (0.0002 dB/mi. at 2GHz, 0.002 dB/mi. at 8GHz and almost 0.2dB/mi. at 20GHz).

At the receiving end of the link, the considerations are the same as on the transmitter end, only in reverse order (radome loss; antenna gain, waveguide, connector and hybrid/coupler losses), with the remaining signal level being applied to the receiver (−dBm).

Fade margin

The receiver manufacturer will offer a *noise threshold specification* (expressed in −dBm) which, simply stated, is the sensitivity of the receiver. It is considered to be the lowest receive signal level at which the receiver will still be considered "unusable." There are variations in the definition of the term "unusable," which means the use of different standards in determining how a particular receiver threshold specification will be obtained. For example, on most industrial analog microwave systems the practical threshold of the receiver is considered the absolute RF signal level (−dBm) injected into the receiver that will establish a 30dB (flat) signal-to-noise ratio (58dBrnc0) as measured in a discriminated baseband noise slot. The telephone companies, however, have used a 33dB (flat) signal-to-noise ratio (55dBrnc0). (The baseband frequency of this noise slot will typically be specified by the equipment manufacturer and is dependent on the particular channel loading of that equipment. It is usually assigned somewhere above the highest occupied frequency slot on the baseband. Because noise effects are worst on the high end of the baseband, this slot would offer a worst case scenario for the signal-to-noise measurement.)

The threshold specification on digital microwave receivers makes reference to the bit-error rate (BER) of the recovered data rather than to a discriminated analog noise measurement. The old Bell standard for digital systems is the RF signal level injected into the receiver that will produce a BER that exceeds 10^{-3}. This is typically referred to as the "outage" threshold. Often, however, a BER of 10^{-6} (referred to as the operating point threshold) is used because this is a more common T-1 level standard. (As a note, the difference between these two bit-error rates relates to a change in RF receive level of only 2dB–3dB. This is because BER performance on digital radios degrades rapidly after the 10^{-6} point.) As long as the BER is defined along with the threshold specification, no confusion over terms should exist.

On analog microwave systems, the difference between the receiver threshold value (−dBm) and the receive signal level (RSL) being applied to the receiver (−dBm) under normal path conditions is referred to as the *fade margin* (expressed in dB). This is a "safety margin" of excess signal that the path can fade before the receiver becomes "unusable" due to

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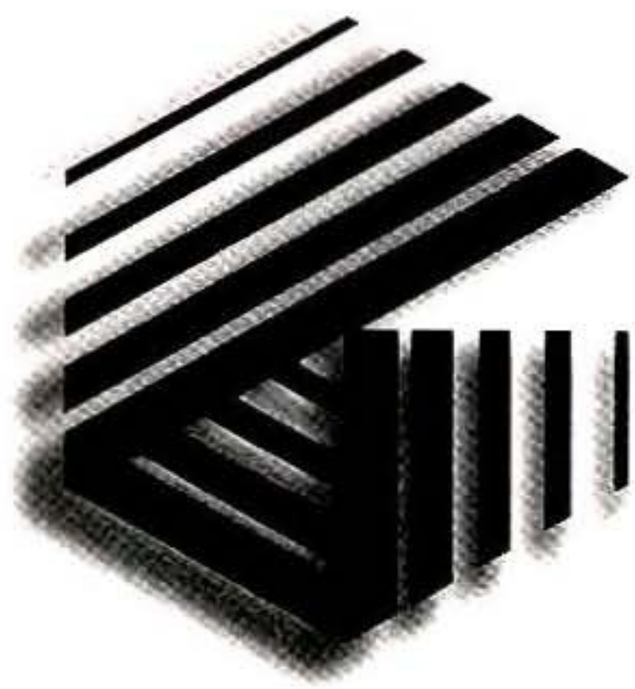

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noise. The same general definition could be used for digital microwave systems, although in this case it might be termed "flat" or *thermal fade margin* (TFM). Other terms applicable to digital systems are *dispersive fade margin* (DFM), *adjacent channel interference fade margin* (AIFM), and *external interference fade margin* (EIFM). *Composite fade margin* (CFM) is a term used to describe the composite affects of TFM, DFM, AIFM and EIFM. Typical receiver threshold levels for digital or analog commercial microwave receivers are usually in the ballpark area of −70dBm to −85dBm.

It should be noted that microwave receivers have a specified range of RF input signal level over which they should properly detect and decode modulated information. This means not only that there is a *minimum* signal level (receiver "threshold") but also a *maximum* signal level that can be applied to the receiver input without expecting distortion or data errors in the recovered signal. This restricts the path designer from creating a huge fade margin by applying excessively high RF levels to the receiver under normal path conditions. Additionally, a path phenomenon referred to as "up-fading" can occur (particularly around the end of a period of severe multipath activity) where the microwave link receivers realize an extreme increase in RF signal level (relative to normal RSL). Up-fading is thought to be the result of an anomalous "lens" effect in the atmosphere. For such reasons, the design engineer should leave a margin of at least 10dB between normal path RSL and the receiver's upper RF input limit. The typical design RSL for a microwave receiver is about −35dBm.

The next and final article in this four-part series will discuss microwave path reliability.

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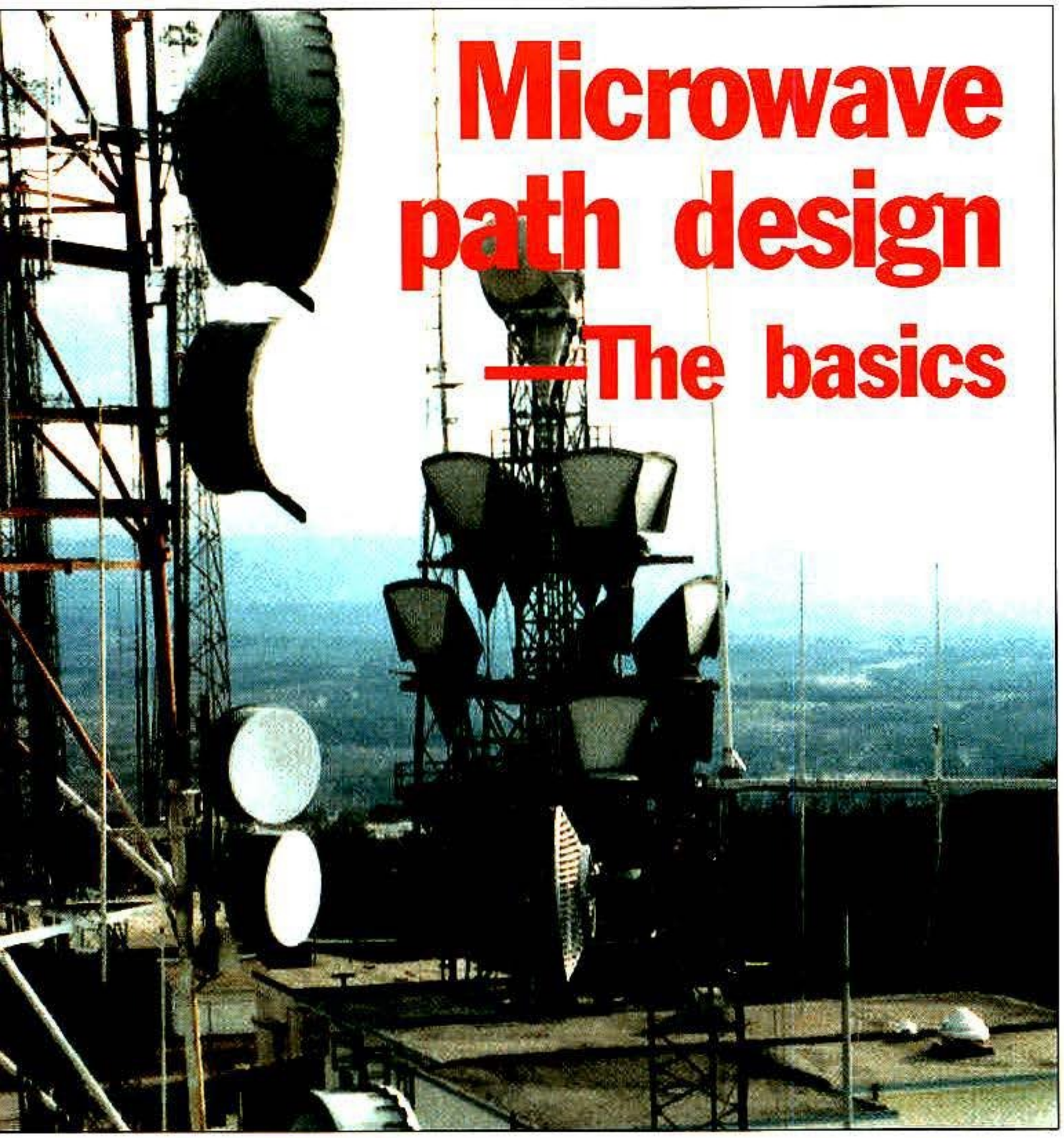

Microwave path design—The basics

Part 4—This series of four articles is a primer on some of the propagation-related basics of microwave path design. This last installment discusses path reliability.

By Jeff Ashley

Preceding installments in this series have described basic considerations:

- Determining whether a proposed path is "line-of-sight."
- Evaluating path clearances with regard to refractive effects.
- Evaluating path clearances with regard to Fresnel zones.
- Considering path reflections.
- Deriving a power budget and fade margin.

This concluding article covers path reliability.

Path reliability

Usually, microwave links are designed to meet a specific reliability factor. Reliability may also be termed "availability" and is expressed as a percentage. It represents the percentage of time the link is expected to operate without an outage caused by propagation conditions. The Bell standard for short-haul propagation reliability is 99.995% (minimum), while requirements for high-capacity, long-haul may be 99.9999%. "Unavailability" or the probability of an *outage* (because of propagation conditions) is often referred to and, if expressed in percentage, the value is determined by subtracting the availability (expressed as a percentage) from 100. For example, a short-haul reliability (availability) factor of 99.995% can also be expressed as an *unavailability* factor of 0.005%.

Availability and unavailability are referenced to a year. In other words, if the unavailability is 0.001%, it could also be stated there will be:

$$365.25\,(\text{days / yr}) \times 24\,(\text{hr / day}) \times 60\,(\text{min / hr}) \times \frac{0.001}{100}$$

= 5.26 minutes of unavailability (outage) *per year.*

The propagation-related unavailability is derived from Equation 1 below. This equation is used as an approximation for the one-direction effects of multipath fading with various terrain and atmospheric conditions.

$$P_m = a \times b \times 2.5 \times 10^{-6} \times f \times D^3 \times 10^{-F/10} \qquad \text{[Eq. 1]}$$

where

P_m = Outage probability (unavailability) due to multipath fading.
D = path length (in miles).
f = frequency (in GHz).
F = fade margin of the path (in dB).
a = one of the following generalized terrain factors:
  4: very smooth (including water).
  1: average, some roughness.
  1/4: very rough.
b = one of the following generalized climate factors:
  1/2: hot, humid.
  1/4: normal, temperate.
  1/8: very dry.

(Note that the result of this equation must be multiplied by 100 in order to be expressed as a percentage.)

Another factor that can affect path reliability is rain attenuation, particularly past 10GHz. At these higher frequencies, the wavelength becomes short enough that rain droplets (depending on their size and shape) can present themselves as obstructions to the path. This occurs when the diameter of the droplets approach 1/4 wavelength of the frequency being used. The maximum instantaneous intensity of the rainfall and the percentage of the total path within the rain cell are other factors to be considered.

Because rain rates differ by geographic location, statistics for the particular area of interest must be obtained. Commercial path design software typically has this information in tables, requiring the user simply to select the geographic location of the proposed path. The values for that area are then accessed. The National Weather Bureau also maintains this information for various cities throughout the United States.

Increasing path reliability

One method of increasing path reliability is to increase the fade margin. If the

Ashley is a communications technician in Los Angeles.

design engineer dictates a particular fade margin requirement from the start, the power budget can be determined in a particular sequence. Knowing the noise threshold of the receiver being used (-dBm) and the required fade margin, the nominal signal level into the receiver can be determined (noise threshold [-dBm] plus fade margin [+ dB] = nominal receive level [-dBm]). The difference in decibels between the transmitter RF output level (+ dBm) on one end of the path and the desired receive signal level (RSL) on the other end of the path will be the maximum allowable net path loss. Because path distance and antenna heights are known, free space and atmospheric losses, along with all hardware losses on both ends, can be determined and will exceed this allowable net path loss by some amount (measured in decibels). The amount by which it is exceeded will be the necessary combined gain of the antennas on each end of the path. To achieve this required gain, proper antenna reflector diameters must be selected.

Parabolic ("dish") antennas typically have gain efficiencies of 55% to 65%. Using an efficiency of 55%, the gain of a parabolic antenna can be calculated as follows:

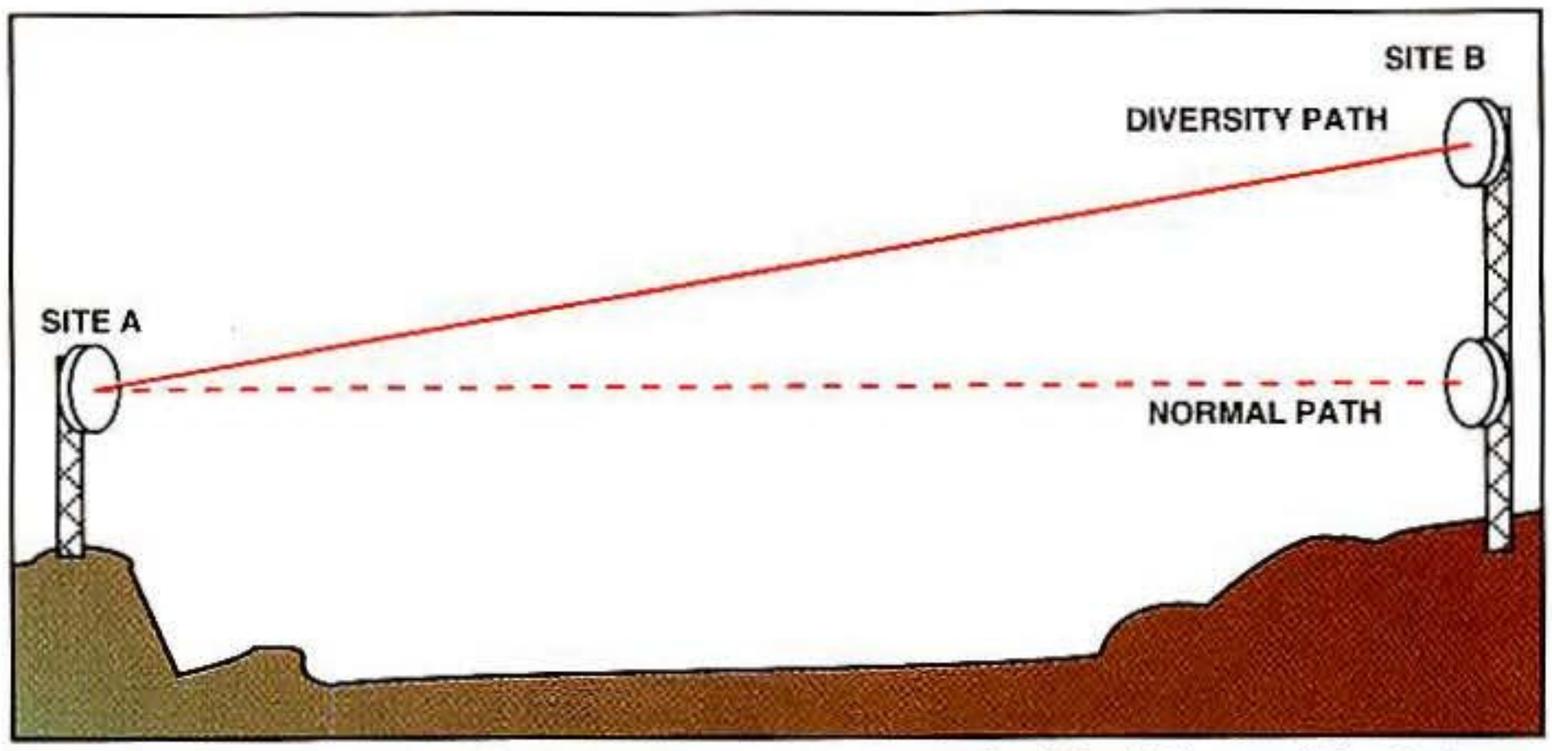

Figure 1. With adequate spacing between antennas, there should be little correlation between paths with regard to fading.

$$G = 20 \log_{10} B + 20 \log_{10} F + 7.5 \qquad \text{[Eq. 2]}$$

where
G = gain over an isotropic antenna.
B = parabola diameter in feet.
F = frequency in GHz.

After a proposed microwave path has been evaluated with regard to reliability, the calculations may show that it will not meet the required standards. In such situations, additional techniques known as *diversity* methods may be employed.

One such technique is *space diversity*, which involves the addition of another receive antenna, separated in distance from the first, as shown in Figure 1 above. An additional receiver and associated switching/combining equipment will also be needed. The additional "diversity" antenna is usually mounted on the same tower structure, yet vertically separated from the site's original receive antenna. This diversity antenna is connected to the diversity receiver. This additional equipment essentially creates another microwave path between the two sites. Receiver switching and combining equipment then senses which receiver has the troubled signal (if fades occur) and switches that receiver "off-line."

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One drawback of space diversity is increased cost. Tower heights must be greater to achieve sufficient vertical antenna separation. In some cases, the existing tower structure might have to be retrofitted to handle the additional wind loading created by the diversity antenna. The diversity antenna, receiver and combining and switching equipment create additional cost.

When spacing is adequate between antennas, there should be little correlation between the two paths with regard to propagation characteristics. Improvement in reliability comes from the reduced probability that both paths will be adversely affected by fading at the same time.

In general, more vertical spacing between antennas offers less path correlation and better path reliability. The rule of thumb for antenna spacing is as follows: 60–80 feet at 2GHz, 45–50 feet at 4GHz, 30-40 feet at 6GHz and 25–30 feet at 11GHz.

The following modified form of Arvids Vigants' fade reduction factor equation can be used to estimate the improvement offered by space diversity:

$$I_{sd} = \frac{7.0 \times 10^{-5} \times f \times s^2 \times 10^{F \div 10}}{D}$$

[Eq. 3]

where

I_{sd} = improvement factor (space diversity).

f = frequency in GHz.

s = vertical antenna spacing, in feet, between centers.

D = path length in miles.

F = path fade margin (if one antenna's path has a smaller fade margin, use that figure).

Frequency diversity is another method used to increase path reliability. Microwave transmitters operating on two frequencies (with a typical in-band diversity spacing of about 2%), and sometimes in two frequency bands (called crossband diversity), are used to transmit the same information to separate receivers at the other end of the path.

Reliability improvement comes from the reduced chances of fading occurring on both frequencies (or frequency bands) at the same time.

Frequency diversity does not require the vertical antenna spacing used in space diversity, and therefore reduces tower cost. It does, however, require the use of more spectrum because it uses two sets of frequencies.

Because available frequencies are in limited supply, the FCC disallows the use of frequency diversity except under certain conditions. Test comparisons between space and frequency diversity indicate that improvements realized with space diversity are generally greater than those achieved with frequency diversity.

A general formula to calculate the improvement factor given by frequency diversity is as follows:

$$I_{fd} = K \times \left(\frac{\Delta f}{f}\right) \times 10^{F \div 10}$$

[Eq. 4]

where

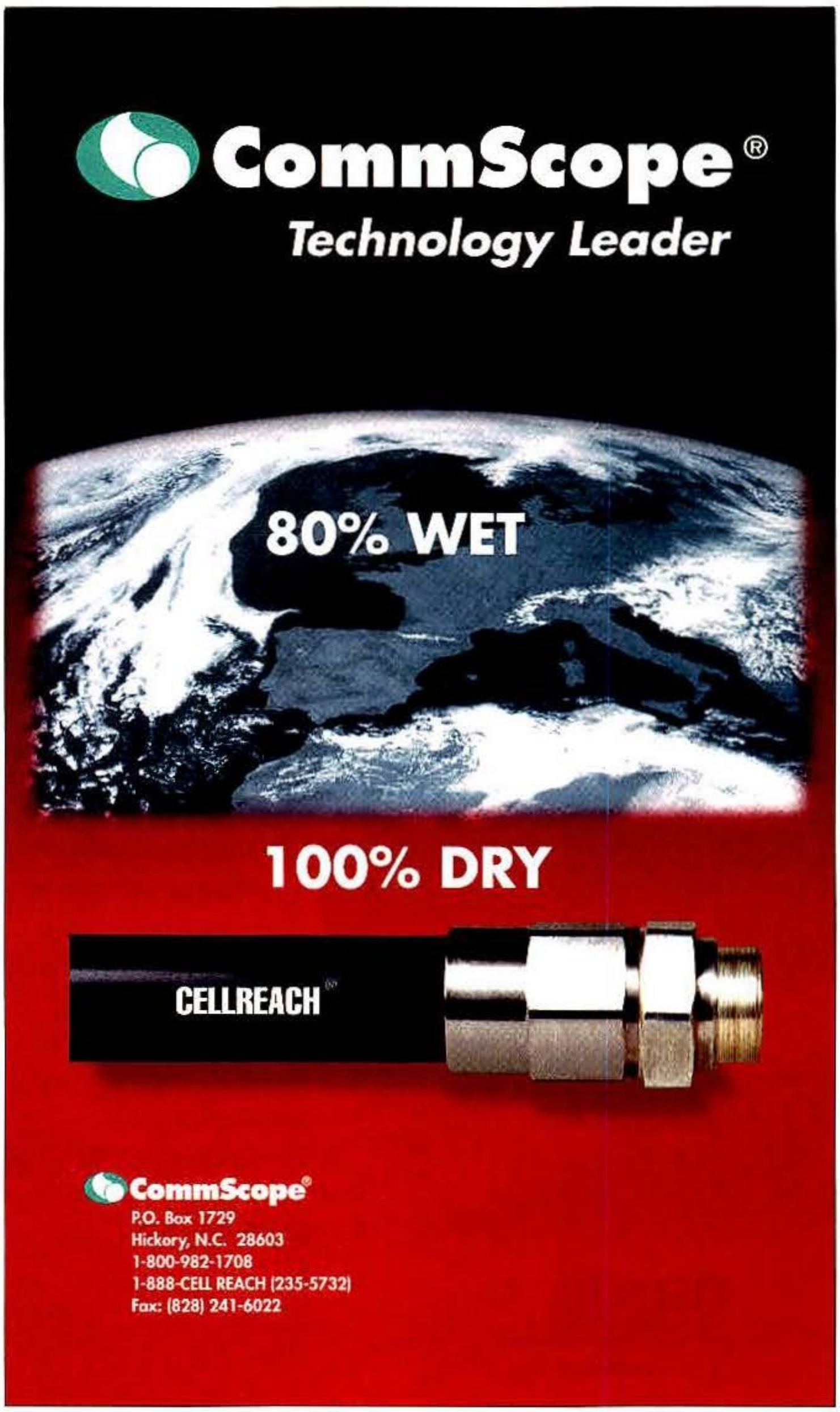
I_{fd} = frequency diversity improvement factor.

$\frac{\Delta f}{f}$ = frequency diversity spacing (*percent* ÷ 100).

F = fade margin.

K = 1/2 for 4 GHz band.
1/4 for 6 GHz band.
1/8 for 8 GHz band.
1/12 for 12 GHz band.

(The values of K shown above for the 8GHz and 12GHz bands are GTE-Lenkurt engineering estimates based on work from



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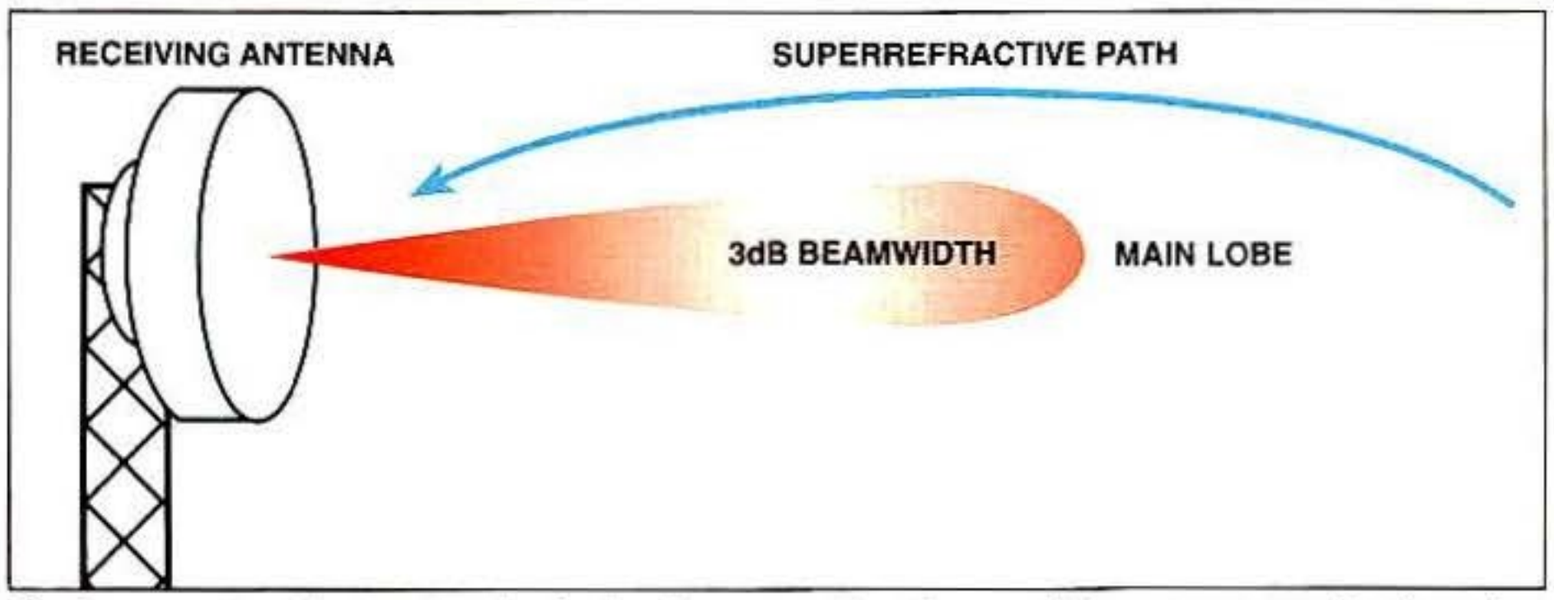

The increased vertical angle of arrival under superrefractive conditions may cause the incoming wavefront to fall outside the beamwidth of the antenna. This is called 'antenna decoupling' and can cause a significant reduction in receive levels.

W.T. Barnett that dealt strictly with the 4GHz and 6GHz bands.)

Neither space diversity nor in-band frequency diversity provide improvement against rain attenuation, although out-of-band frequency diversity (such as 6GHz/12GHz) does.

Under superrefractive atmospheric conditions, negative values of K can exist where the path bows skyward at mid-path and comes downward at the ends. This will create a wavefront that will arrive at the antenna with a greater vertical angle than normal (see Figure 2 above). Because microwave antenna beamwidths can be small, this increased vertical arrival angle may be outside the antenna's half-power beamwidth. The received signal level realized may be sufficiently low to cause noise problems or even an outage. This has been termed "antenna decoupling." A method of reducing this effect is to use a special *angle diversity* feed horn arrangement that essentially creates two vertically separated main lobes, so that if the signal level from the lower (normal) main lobe diminishes sufficiently because of the wavefront's increased arrival angle, the other (upper) lobe will be better oriented. Because most fades are due to reflections and multipath effects rather than this anomalous superrefractive condition, angle diversity offers only a limited improvement factor over that of a non-diversity path under typical path conditions.

Hybrid diversity is where both frequency *and* space diversity are employed simultaneously. However, most of the improvement comes from the effect of the space diversity.

The following equation is used when calculating fade probabilities with various methods of diversity:

$$U_{div} = \frac{P_m}{I} \qquad \text{[Eq. 5]}$$

where

U_{div} = unavailability or probability of outage (using some diversity method).

P_m = annual unavailability (non-diversity) due to multipath (Eq. 1).

I = diversity improvement factor (Eq. 3 and Eq. 4).

In multiple-hop microwave systems, the overall reliability figure (expressed as a percentage) is obtained by taking the sum of the unavailability figures (as a percentage) for each hop, and subtracting that figure from 100.

Conclusion

Many other considerations go into the complete design of a microwave path and communications link. Additionally, each of the points discussed can be taken in much greater depth. Volumes of information are available to the interested reader from many of the large communications companies, both past and present.

This series of articles has offered a glimpse into some of the more fundamental propagation-related design considerations and will be of value to those just getting their feet wet in the microwave communications field. ■

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