

Chapter 2

Leaky-Wave Antennas

Millimeter waveguides, such as coplanar lines, microstrip lines and dielectric guides, are open structures where energy leakage is bound to occur. This leakage can occur, for example, when there are discontinuities in the guiding structure, or perhaps when the guide is excited in an inappropriate mode. In terms of antenna design, this energy leakage may be advantageous and exploited. This is accomplished through intentionally creating discontinuities in the guiding structure in such a manner that the radiation produced is controlled by the antenna designer. These antennas are very suitable for integrated designs, seeing that they are compatible with the waveguides from which they are derived, essentially limiting the unwanted and imperfect transitions between guiding media. This property also eliminates the requirement for complex and lossy feed networks present in other types of planar structures.

Mathematically, a leaky wave is treated as a guided complex wave and the resulting radiation pattern is expressed in terms of the complex propagation constant. As a subset of traveling wave antennas, leaky-wave antennas are further divisible into two categories, namely one-dimensional and two-dimensional variants. These will be further reduced into subclasses and introduced as this chapter progresses. Similar to other traveling wave antennas, leaky-wave antennas radiate primarily in the endfire direction, but they are very suitable for frequency scanning and as such are often implemented for this purpose. Surface wave antennas and slot arrays are also part of the traveling wave family, and while they share some defining features, their performance expectations and design methodologies differ.

2.1 General Principles of Leaky Waves

The properties of leaky waves were originally derived in the pioneering work of Oliner and Tamir in the late 1950s and early 1960s [1–3]. This was followed by an extensive development of leaky-wave theory and application to antennas. However,

interest in the behavior and application of these antennas at millimeter-wavelengths only began several decades later.

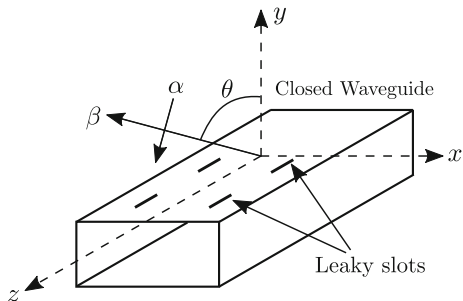
As mentioned earlier, a leaky wave arises from a guiding structure with some or other continuous or periodic discontinuity that facilitates energy leakage into the surrounding area. A simple example is a slotted waveguide antenna, where the waveguide is perturbed with periodic slots in the structure at a certain position. The energy leakage mechanism results in the waveguide having a propagation wave number that is a complex quantity. The phase constant is given by β and the attenuation constant is indicated by α . The attenuation constant varies in size depending on the leakage per unit length along the waveguide; a larger value for α means that a larger amount of energy leaks from the structure [4].

As seen from Fig. 2.1, the complex propagation parameters are dependent on the geometry of the leaky structure. Control over the beam shape and sidelobe levels may be obtained by using an appropriate aperture taper. This is accomplished in practice by slowly varying α along the length of the guide in a specific fashion, and simultaneously holding the phase constant at a fixed value, effectively adjusting the amplitude of the aperture distribution.

The entire waveguide in Fig. 2.1 is regarded as the effective aperture of the antenna, unless the energy leakage is so severe that the signal power fades away completely before reaching the end of the slotted section. A large value for α (which indicates a large leakage rate) in turn leads to a smaller effective aperture, and from antenna theory it is known that the smaller aperture would lead to an increase in beamwidth [5]. On the contrary, lower α values result in a much longer effective aperture and thus the antenna would radiate a narrow main beam, if the physical size of the structure permits it.

It should be noted that if the aperture is fixed beforehand and the leakage rate α is relatively small, the antenna pattern (particularly the beamwidth) is primarily influenced by the aperture itself, rather than the leakage rate. However, the radiation efficiency in such a case is strongly affected by the leakage rate. A reasonable design goal is to have approximately 90 % of the power in the waveguide radiated from the structure when the wave has propagated through the entire structure. In most practical cases, a matched load will be connected at the end of the waveguide in order to absorb the remainder of the signal power.

Fig. 2.1 Slotted rectangular waveguide leaky-wave antenna



Rectangular waveguides support either transverse electric (TE) or transverse magnetic (TM) modes of wave propagation, and for both cases the phase constant of the wave is frequency dependent [6]. As the frequency (and so also the phase constant β) changes, so does the direction in which the principal beam is pointing. Therefore, the antenna beam can be scanned by altering the excitation frequency.

2.2 Extension into Millimeter-Wavelengths

Leaky-wave antennas have been extensively studied in the last three to four decades, and their main attraction is high directivity, wide bandwidths, and their ability to scan with frequency. Earlier, leaky-wave antennas were based almost exclusively on closed waveguides and the leakage was introduced by cutting holes or slots into the guiding structure. Waveguide losses become exceedingly high at millimeter-wavelengths. As a result, millimeter waveguides are typically open structures, in an attempt to lower the attenuation constant that results from dielectric or metallic losses [7]. Examples of open waveguides are microstrip lines, groove guides, non-radiative dielectric guide and many other configurations of dielectric guides. Generally, the dominant modes that propagate in these types of waveguides are purely bound, which means that physical defects in the structure will not cause them to radiate. Instead, other techniques such as asymmetry or similar geometric alterations are often used.

Simple open waveguides support the propagation of slow waves, which do not radiate outward from the guide, and the periodic structure is chosen such that only the $n = -1$ space harmonic is able to radiate power. On the other hand, some millimeter waveguides can be closed structures, such as metal guides and finlines. The tapered antennas shown in Fig. 2.4, particularly, are examples of antennas that are easily integrated with the waveguides from which they are derived [8, 9].

With a basic theoretical background on leaky-wave modes established, we will now move on to covering influential designs and concepts that have surfaced in recent years—in no particular order.

2.3 Leaky-Wave Antenna Classification

Depending on the geometry, principle of operation and the nature of structural perturbations, leaky-wave antennas can be divided into several categories [10]. As mentioned in the introduction, perhaps the most basic distinction is between one- and two-dimensional leaky-wave antennas. Furthermore, these can be identified as periodic, uniform or quasi-uniform structures. These configurations will be discussed separately in the succeeding sections, but it should be noted that some of the concepts relating to either the theory of operation or design principles overlap, sometimes with minor differences.

2.3.1 One-Dimensional Uniform Leaky-Wave Antennas

Theory

A one-dimensional structure is one that supports a wave traveling in a single, fixed direction. One such structure is a rectangular waveguide with a long uniform slit in one of its side walls, like the one shown in Fig. 2.2. The geometry of this guide can be considered uniform, seeing that it does not vary in the longitudinal direction (along the z -axis). The slot in Fig. 2.2 is modeled with a surrounding infinite ground plane that acts as a back-baffle, and its shape may be tapered if a specific beam shape is desired. For a narrow slot, the structure depicted in Fig. 2.2 becomes equivalent to a magnetic line current in the direction of the z -axis [4].

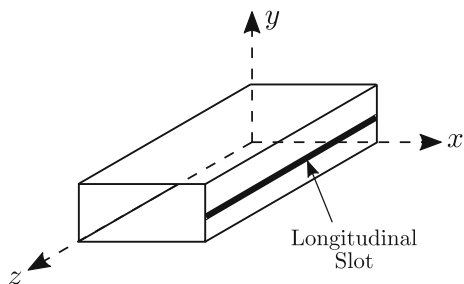
The radiation produced by this structure is limited to the $z > 0$ region, and the resulting beam is in the shape of a cone along the z -axis. As the angle approaches 90° (indicated by θ in Fig. 2.1)—in other words, as one approaches the broadside direction—the antenna pattern takes on a narrow donut-type shape.

It is difficult to achieve a broadside beam if the antenna is fed from only one end, since this equates to operating the waveguide at its cutoff. By feeding the waveguide with a frequency that is marginally higher than its cutoff, it may be possible to achieve a broadside beam. This may be accomplished by feeding it from both ends, or by placing the source in the middle of the guiding structure. From the examples provided here, it is clear that these antennas are axially significantly longer than laterally.

An extremely narrow pencil beam in both the azimuth and elevation planes can be obtained by placing the one-dimensional leaky-wave elements in a linear array [11]. The direction in which the beam points is then controlled by individual beam angles and the phase delta between elements. It is therefore possible to control the beam angle in the elevation plane by altering the excitation frequency and in the azimuth plane by altering the phase of the exciter.

The radiation pattern is obtained by taking the Fourier transform of the aperture distribution. In the case where the geometry is kept consistent along the length of the antenna, the field lines comprise a traveling wave with constant values for α and β . This then leads to an exponentially decaying amplitude distribution along the length of the guide. If the length of the antenna is modeled as being infinite, the resulting radiation pattern can be computed with acceptable accuracy as

Fig. 2.2 Rectangular waveguide with a longitudinal slot on its side wall



$$R(\vartheta) \approx \frac{\cos^2 \vartheta}{\left(\frac{\alpha}{k_0}\right)^2 + \left(\frac{\beta}{k_0} - \sin \vartheta\right)^2}, \quad (2.1)$$

where k_0 is the free-space wavenumber. Fundamentally, this pattern does not contain any sidelobes, but as the length is decreased the expression for $R(\vartheta)$ becomes more involved and sidelobes begin to appear. The sidelobes present in a leaky-wave antenna are often significant, and the tapering approach to its design is often the de facto route that designers follow in practice.

Design Principles

Once the values for α and β are known, the major parameters of the antenna—beamwidth, radiation efficiency, and scan angle—can be determined rapidly. The scan angle is given by

$$\sin \theta_{p,\max} \approx \frac{\beta}{k_0}, \quad (2.2)$$

where, as the subscript indicates, $\theta_{p,\max}$ represents the maximum achievable scan angle. For a waveguide with axial length L , the beamwidth can then be found as

$$\Delta\theta \approx \frac{1}{(L/\lambda_0) \cos \theta_{p,\max}}. \quad (2.3)$$

The unity factor in the numerator of Eq. (2.3) changes depending on the particular amplitude distribution. For example, for an aperture distribution that remains consistent over the length of the antenna, the numerator is replaced by 0.88, while a tapered distribution could have a numerator greater than 1.25 [4]. The physical length of the antenna L is chosen so that approximately 90 % of the power is radiated. The remainder is then absorbed by a matched load, and L is usually specified for a set value of α . Therefore, we can find L as

$$\frac{L}{\lambda_0} \approx \frac{0.18}{\alpha/k_0}, \quad (2.4)$$

where L and α are chosen independently of one another, the percentage of radiated power can differ significantly from the desired value of 90 %. The ratio of power that resides in the leaky mode at $z = L$ to the input power can then be written as

$$\frac{P(z = L)}{P(z = 0)} = \exp(-2\alpha L) = \exp(-4\pi(\alpha/k_0)(L/\lambda_0)), \quad (2.5)$$

where $P(z)$ indicates the power in the waveguide at a distance z along the length of the structure. The radiated power changes slightly when the beam is scanned with

frequency, given that α changes with frequency. However, the percentage of radiated power (indicated simply by P%) can be obtained easily by rewriting Eq. (2.5), assuming an exponential aperture distribution. The result is

$$P\% = 100\{1 - \exp[-4\pi(\alpha/k_0)(L/\lambda_0)]\}. \quad (2.6)$$

The aperture distribution will inevitably be changed in order to control sidelobes, but nonetheless, Eq. (2.6) remains a reasonable approximation.

In terms of scan angle, uniform leaky-wave antennas can take on two forms. While the principles remain similar, the scan angle behavior of air-filled and dielectric-filled waveguides differ somewhat. Air-filled structures commonly encountered are rectangular waveguides and groove guides, both of which support dominant modes that are fast. On the other hand, partially dielectric-filled structures often used are the non-radiative dielectric guide and dielectric-loaded rectangular guide. To operate these as leaky-wave antennas, they should be excited with a fast wave ($\beta < k_0$).

Research Review

Considering uniform leaky-wave antennas, a lot of effort has been expended towards non-radiative dielectric guides, often referred to simply as NRD guides. Early work by Oliner et al. contained detailed theories and experimental results on rectangular NRD guides, although several other variations of leaky-wave NRD guides have been proposed for millimeter-wavelengths ever since. The first proposal on leaky-wave antennas derived from NRD guides appeared in 1981 and was authored by Yoneyama et al. [12].

The concept of asymmetrically perturbed planar leaky-wave antennas has been studied extensively by Gómez-Tornero et al. [7]. This work covered several important questions relating to planar leaky-wave antennas, and the frequencies of interest were around 40–65 GHz. For instance, modifying the leakage factor of the leaky-wave mode while keeping the phase constant stationary was investigated for a wide array of slot and strip designs.

Each of these designs used a different tapering topology and was specifically designed to be mechanically flexible and thus relatively simple to fabricate. One example of a tapered planar guide is shown in Fig. 2.3, where the width, amplitude

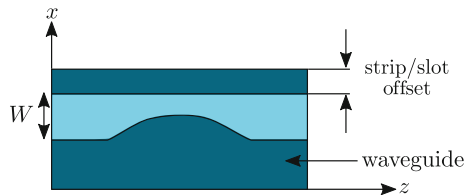


Fig. 2.3 Tapered slot leaky-wave antenna. The offset, shape, and maximum width of the taper are controlled to alter radiation characteristics

distribution, and the offset of the slot (or microstrip in some cases) affect the properties of the leaky-wave modes propagating in the guide.

Through the use of a hybrid-planar topology, flexibility in manufacturing became a substantial benefit when compared to conventional rectangular waveguide technology. This is especially important at millimeter-wavelengths since manufacturing tolerances play a key role in the performance of the antenna. The authors came to the conclusion that the slotted leaky-wave antenna proved to be better suited for offset tapering, while the stripline leaky-wave antennas were in turn better suited for asymmetric tapering around a zero offset.

The group published its analysis and design methods for the hybrid waveguide-planar technology later that year. It provided full-wave integral equations and design guidelines for this concept [13]. The derivations were based on the leaky-wave modes that propagate in laterally shielded, stub-loaded dielectric guides that are rectangular in shape. The perturbations in the guiding structures were considered as either rectangular strips or slots. The guidelines presented in this report allowed other designers to obtain leaky-wave dispersion curves easily. A demonstration of this was provided by developing a novel planar leaky-wave antenna. The configuration presented had the attractive property of being able to alter the stopband region and the leakage rates independently, because of relying on the principle of asymmetric radiation.

Gómez-Tornero et al. [14] continued work with one-dimensional leaky-wave antennas, reporting on a parallel-plate waveguide loaded on one end with a partially radiating surface and on the other end with a high impedance surface. This antenna is designed for operation at 15 GHz. By altering the dimensions of the dipoles etched on the two surfaces, independent control over the leakage rate and the leaky mode phase constant is possible. Using an air-filled waveguide also negates the dielectric loss, leading to radiation efficiencies of about 90 %.

Furthermore, the group also investigated conformal tapered microstrip leaky-wave antennas. These were based on using half-width microstrip sections at 15 GHz [15].

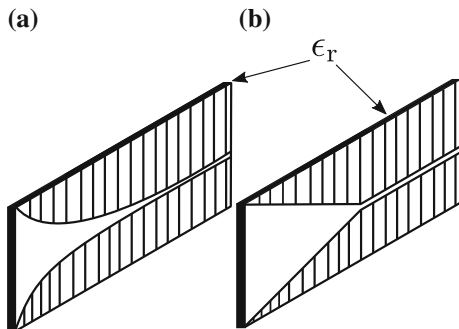
2.3.2 *One-Dimensional Periodic Leaky-Wave Antenna*

Theory

A one-dimensional periodic antenna comprises a uniform structure that supports the propagation of a non-radiating wave (i.e., $\beta > k_0$) and is modified periodically in the z-direction. This non-radiating wave is also referred to as a fast wave, since its dominant mode is fast relative to free-space velocity. In a periodic antenna, however, the dominant mode is a slow wave and therefore does not radiate by itself: the structure needs some kind of periodic modulation in order to produce the radiation.

An example is shown in Fig. 2.4, where slots are periodically inserted on a planar structure and the slot length is controlled through a taper function. This

Fig. 2.4 Tapered slot leaky-wave antennas, **a** using an exponential taper and **b** using a linear taper



configuration differs from the tapered slot antenna shown in Fig. 2.3, in that the slots are periodically inserted on the structure, although both are essentially planar configurations. The tapering is implemented differently and the effects on the antenna characteristics should become clear as the discussion progresses.

A primary advantage of periodic leaky-wave antennas is that the direction in which the beam points could be either forwards or backwards, depending on the phase constant used in the excitation.

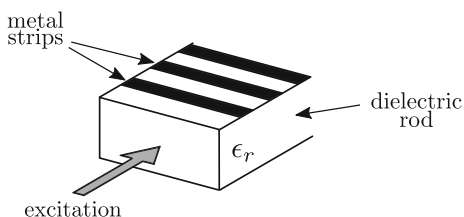
Although two planar structures are shown in Fig. 2.4, the concept applies to many other periodic configurations. For example, a dielectric waveguide can be periodically perturbed by small holes or slots and achieve similar radiation characteristics. The excitation of the waveguide is in the fundamental mode, and in order to prevent higher order mode propagation, the width is chosen to be relatively small.

The slots in Fig. 2.4 could also be replaced with a metal grating (i.e., periodic metal strips on the dielectric) and similarly a dielectric grating. In the latter case, the metal strips would instead be grooves in the dielectric surface, and the diffraction effect at this grating would then transform the excitation mode into a leaky wave [16].

Since the radiation mechanisms of uniform and periodic leaky-wave antennas result from different physical processes, these antennas differ in their achievable scan ranges. Consider the dielectric waveguide in Fig. 2.5.

The guide dimensions are chosen such that the dominant mode is the only mode above the cutoff frequency, and since $\beta > k_0$ (equivalently, $\beta_n/k_0 < 1$) for this particular mode, it is thus purely bound. After the dimensions have been set, the metal strips in Fig. 2.5 are added periodically, and it is this periodicity that in turn

Fig. 2.5 Dielectric rod, periodically loaded with metallic strips



creates an infinite number of space harmonics. These harmonics are characterized by a unique phase constant β_n and subsequent phase constants are then related by

$$\beta_n d = \beta_0 d + 2n\pi, \quad (2.7)$$

where d is the strip spacing and β_0 indicates the original phase constant of the dominant mode in a uniform guide [4]. The phase constant β_n can clearly assume a significant number of values, but if at least one of the space harmonics becomes fast, the whole mode will become leaky. Given that $\beta_n/k_0 < 1$ and $\beta_0/k_0 > 1$, the relationship in Eq. (2.7) can then be rewritten as

$$\frac{\beta_n}{k_0} = \frac{\beta_0}{k_0} + \frac{2n\pi}{k_0 d} \quad (2.8)$$

and since $k_0 = 2\pi/\lambda_0$, it is clear that it is indeed possible for $|\beta_n/k_0|$ to be less than unity, provided that λ_0/d is suitably selected and the harmonic number n is negative. For an antenna that is designed to have only one radiated beam, one can go ahead and choose $n = -1$. Following this approach, at low frequencies there will now be fast harmonics and the antenna will not radiate any beams. As the frequency increases and eventually reaches a critical point where the $n = -1$ harmonic becomes fast, a beam originates from backward endfire. Further increasing the frequency will result in the beam scanning from backward endfire, through broadside, up to forward endfire.

A periodic leaky-wave antenna therefore radiates as a result of the $n = -1$ space harmonic, which is the fundamental difference between uniform and periodic structures.

Design Principles

Tapering the antenna aperture according to the appropriate amplitude distribution leads to exceptional radiation characteristics and ultra-low sidelobes on paper. To obtain these properties in practice, careful design of the feeding method is required. When the antenna is derived from perturbing a closed waveguide, the taper will cause the aperture to radiate minimally at both ends. The discontinuity between the antenna and the closed feed waveguide is thus negligible, and considerations relating to the feed are nearly nonexistent.

This is not the case when the feed is an open waveguide. Surface waves are excited from a tapered transition out of a closed waveguide, and the transition creates spurious radiation. Wherever this type of transition is found in the feed network, the uncontrolled radiation may adversely affect the radiation pattern to the point where the initial design is invalidated. The influence that spurious radiation has on the radiation pattern is not an issue with most leaky-wave antennas, but given the effects on the antenna performance, it should not be overlooked in the design.

In order to ensure that the TE_{10} mode is a slow wave over the desired bandwidth, a lower limit on the permittivity of the substrate is given by

$$\epsilon_r > 1 + \left(\frac{\pi}{k_0 a} \right)^2, \quad (2.9)$$

where a indicates the greater of the two waveguide dimensions. Although there are many derivations of leaky-wave antennas based on rectangular waveguides, many properties are common and can be understood by considering this structure. The fundamental mode that exists in the waveguide is non-radiating, but the periodicity in the structure means that the modal field is in the form of a Floquet-wave expansion,

$$\mathbf{E}(x, y, z) = \sum_{n=-\infty}^{\infty} \mathbf{A}_n(x, y) \exp(-jk_{zn}z), \quad (2.10)$$

where

$$k_{zn} = k_{z0} + \frac{2n\pi}{p} \quad (2.11)$$

is the specific wave number for the n th Floquet mode—also often referred to as a space harmonic—recall the earlier discussion—and p indicates the perturbation period [10]. As mentioned earlier, leakage will occur if one of the space harmonics is a fast wave. This is usually the $n = -1$ harmonic, and the result is the requirement that $-k_0 < \beta_{-1} < k_0$. Altering the periodicity p in the appropriate manner can scan the main beam over to the desired angle and, typically, it can be scanned from backward endfire to forward endfire. Alternatively, beam scanning is achieved by changing the excitation frequency. Having a single beam over the scan range is an important design consideration. This occurs when the $n = -2$ harmonic remains as a slow backward wave ($\beta_{-2} < -k_0$), while the fundamental mode is kept as a slow forward wave ($\beta_0 > k_0$). At the highest frequency in the range of scan angles (that is, in the forward endfire direction), it is then required that

$$\epsilon_r > 9 + \left(\frac{p}{a} \right)^2, \quad (2.12)$$

which follows from $p/\lambda_0 < 0.5$. One thing to notice is that for uniform periodic leaky-wave antennas, a dielectric constant of greater than 9 is always required if a single beam is desired for the application. If the guiding structure does not support a fundamental mode, the constraint outlined in Eq. (2.12) changes. One example is a microstrip transmission line, which can support only a quasi-transverse electromagnetic (TEM) mode [6].

Research Review

The earliest forms of periodic one-dimensional leaky-wave antennas were in the form of dielectric rods (one such antenna was illustrated in Fig. 2.5). Most of these results were due to the pioneering work of Klohn et al. [11], and successful

millimeter-wave designs were demonstrated as early as 1978. Particular bands of interest were in the 55–100 GHz region, specifically around 60, 70, and 94 GHz, millimeter bands that are widely used today. Though discussed extensively in the introductory chapter of this text, it is worth mentioning once again that these developments were highly dependent on the advent of stable, high-power millimeter-wave sources, many of which were based on the IMPATT diode. The work of Klohn et al. also clearly demonstrated the manufacturing challenges associated with millimeter-wave designs. For example, it was found that a 0.1 mm change in the perturbation period led to a shift in scanning angle of approximately 8° . Nonetheless, these antennas were a crucial part of the development of millimeter-wave integrated circuits.

Investigation into the endfire tapered slot antennas shown in Fig. 2.4 was first reported by the Yngvesson group [8], a few years after the publication of the research done by Klohn's group. With applications such as remote sensing, radio astronomy, and satellite communications, there was an increasingly urgent requirement for multi-beam antennas and integrated systems. The Vivaldi antenna presented by Gibson in 1979 is an early example of a planar endfire antenna that exhibits acceptable gain and sidelobe levels. The Vivaldi antenna uses an exponential taper to control the amplitude distribution of the individual slots, and the configuration shown in Fig. 2.4a is thus a Vivaldi antenna. One substantial advantage offered by tapered slot antennas is their ability to produce symmetric E-plane and H-plane beams over large bandwidths, even with the planar implementation discussed here.

A group at the Valencia Polytechnic University has investigated the possibilities of constructing leaky-wave antennas that are based on the Goubau line [17].

High losses in the dielectric and ground plane of microstrip circuits offer a substantial challenge at millimeter-wavelengths, and other types of transmission lines are often required to solve this problem. At submillimeter wavelengths, single-wire waveguides have been demonstrated and validated for reasonably low-loss transmission; however, these lines are far too complex in structure to facilitate their practical use in array configurations. On the other hand, the Goubau line is an effective alternative to the single-wire waveguide. In fact, the variation discussed here is simply a planar implementation of the thin-wire waveguide on a thin substrate. This is shown in Fig. 2.6.

Similar to the traditional microstrip transmission line, the Goubau line does not use a ground plane (as illustrated in Fig. 2.6) and as a result, presents a lower attenuation as opposed to that of the microstrip line. The leaky-wave configuration that is based on the Goubau line is shown in Fig. 2.7.

The fundamental mode that propagates along the Goubau line is bounded. It can therefore be used as a leaky-wave antenna by adding dipole sources along the

Fig. 2.6 The Goubau transmission line in its planar form

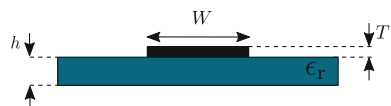
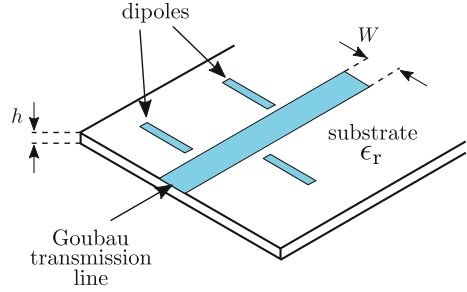


Fig. 2.7 Transverse dipole array with a Goubau transmission line. Note once again the absence of a ground plane



transmission line. The transverse planar dipole sources are periodic perturbations along the line and they give rise to an infinite number of space harmonics. If a single one of these harmonics is fast, it will cause the structure to radiate. The separation distance of the dipoles is chosen such that the main beam is not radiating in the broadside direction, since this significantly decreases the gain of the antenna. To ensure that the structure radiates a single main beam, the mode $n = -1$ is selected. The final configuration consists of a 16-element array (8×2) with a cosine distribution.

It was later discovered that the absence of a ground plane resulted in two main lobes, as opposed to the intended single-beam implementation. To realize the single-beam version, a metallic plate was added to the structure at an optimized distance from the antenna. This optimization was required so that the power radiated towards the metallic plate would be completely reflected as well as added in-phase to the generated antenna beam. Adding the two beams in-phase resulted in a maximum obtainable directivity. Furthermore, reduced losses could be achieved if the metallic plane and the antenna were separated by an air gap. Measurements were taken around 40 GHz, particularly in the frequency band of 37–41 GHz. The obtainable radiation efficiency was found to be 71 % (70 % without the reflecting metallic plate), and S_{11} remained below -10 dB over the specified operating bandwidth.

A later modification to the Goubau line approach presented by Sánchez-Escuderos et al. [17] involved the design of arrays that were circularly polarized. This was accomplished by printing crossed dipoles on the substrate with an electric contact between the elements. The measured H-plane radiation pattern obtained remained properly circularly polarized with and without the metallic reflector plane discussed in the earlier work.

In a subsequent publication, Sánchez-Escuderos et al. [18] reported detailed measurements on the circularly polarized variation of the Goubau line leaky-wave antenna. The measured results obtained from the prototype antenna were promising, with a 15 % impedance bandwidth (for $S_{11} < -10$ dB) and a 3-dB axial ratio bandwidth of 7.6 %. Over the specified bandwidth, the radiation efficiency was over 90 % and the measured gain of the antenna was found to be 15.6 dBi.

A recent investigation on periodic leaky-wave antennas by Nechaev et al. [19] compared two types of center-fed configurations for millimeter-wave operation.

The first of these consisted of a periodically grating grounded dielectric waveguide. The grating comprised two arrays of parallel metallic strips, and the design frequency was centered at 81 GHz.

2.3.3 Two-Dimensional Leaky-Wave Antenna

Theory

A leaky wave that originates from a two-dimensional guiding structure propagates radially from the feed point. This configuration provides an easy method of obtaining a directive beam at broadside, requiring only a simple source. The general form of such a structure is a partially reflecting surface above a ground substrate. This is illustrated in Fig. 2.8.

The excitation source shown in Fig. 2.8 is a simple horizontal dipole placed within the substrate, at a set distance above the ground plane. The antenna pattern, however, depends on the structure and not on the excitation. The substrate/superstrate structure could also be extended to include multiple layers of the dielectric material, with the advantage of narrowing the beamwidth [10]. Increasing the permittivity of the superstrate layer results in a narrower beamwidth.

Another example of a partially reflecting surface is shown in Fig. 2.9.

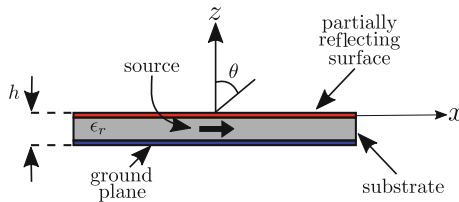


Fig. 2.8 Substrate/superstrate structure of a two-dimensional leaky-wave antenna

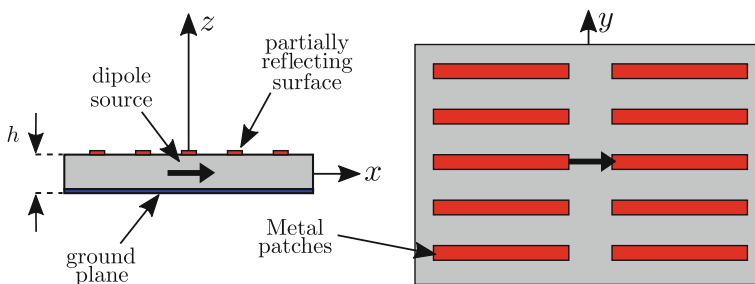


Fig. 2.9 Construction of a partially reflecting surface using metal patches

Larger rectangular microstrip elements will result in a narrower beamwidth. This structure is very suitable for photolithographic fabrication. A similar configuration to the one shown in Fig. 2.9 is to replace the patches with slots. Conversely, smaller slots will result in a narrower beamwidth, and it should be noted that the surface wave feeding the antenna is directed in the y -direction (rather than in the x -direction as with the microstrip array).

As the excitation frequency is increased, the pencil beam originating from the antenna will assume a conical shape, effectively losing gain in the broadside direction [20]. Generally, the guiding structure in a two-dimensional leaky-wave antenna is uniform (or at least quasi-uniform) and the excited wave is a fast wave.

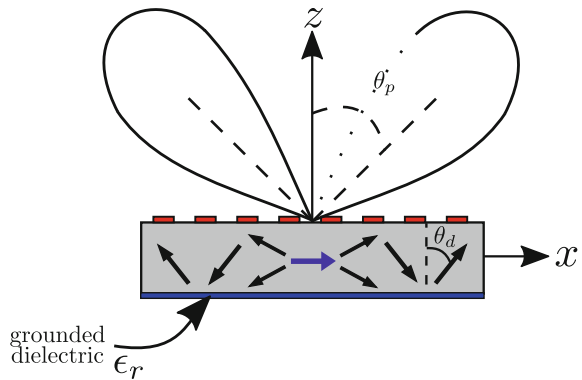
Design Principles

Two popular methods of producing frequency scanning pencil beams from a leaky-wave antenna is using an array of linear or uniform antennas, or by using a two-dimensional structure. While either of these approaches will produce highly directive beams, it is often required to stack multiple layers of dielectrics or use layers with exceedingly high permittivity. In order to improve on this technique, the work reported by Zhao et al. [21] describes a two-dimensional structure that consists of a periodic array of patch elements on a grounded dielectric substrate. Such a structure was illustrated in Fig. 2.9, and the design of such an antenna will be discussed in greater detail here.

As mentioned previously in this chapter, the partially radiating surface causes the antenna to behave like a leaky parallel-plate waveguide operating in the $n = 1$ mode. The array of patch elements acts as a leaky conducting plate on top of the dielectric surface, facilitating radiation at a particular scan angle θ_p . If the parallel-plate waveguide approach is then continued, the substrate thickness required for a particular scan angle can be computed as

$$k_{z1}h = (k_1 \cos \theta_d)h = n\pi \quad (2.13)$$

Fig. 2.10 Propagation of a leaky wave inside the substrate of a 2-D antenna



where $n = 1$, and θ_d indicates the angle at which the waves contained in the dielectric propagate. These are related to the radiation angle described by Snell's law [22] and this effect is detailed in Fig. 2.10.

From the figure, it should be clear that a broadside beam would correspond to $\theta_p = 0^\circ$, and a main beam in the endfire direction corresponds to $\theta_p = 90^\circ$.

The thickness of the substrate is then given by

$$\frac{h}{\lambda_0} = \frac{0.5}{\sqrt{\epsilon_r - \sin^2 \theta_p}}. \quad (2.14)$$

The spacing of patch elements should be small enough to result in a reflection coefficient of -1 from the radiating surface. Once the substrate thickness is found, the required permittivity can be computed.

A wide range of substrate permittivity values will result in a pencil beam and it is therefore a relatively arbitrary parameter. However, the permittivity heavily influences the scanning ability of the antenna. In a case where the substrate permittivity is chosen too low, it is very likely that a critical angle will exist at which the structure will produce secondary beams due to the excitation of higher order modes. This is highly undesirable, and in order for the beam to be able to scan at any angle between zero and endfire (i.e., $0 < \theta < 90^\circ$), the $n = 1$ mode must scan over the whole range of interest before the next higher order mode appears at broadside.

For the principal higher order mode ($n = 1$) to produce endfire radiation, the substrate thickness can be found by entering the desired angle into Eq. (2.14), more specifically,

$$\frac{h}{\lambda_0} = \frac{0.5}{\sqrt{\epsilon_r - 1}}. \quad (2.15)$$

Furthermore, for the $n = 2$ mode to produce broadside radiation, the required substrate thickness can be found as

$$\frac{h}{\lambda_0} = \frac{1}{\sqrt{\epsilon_r}}. \quad (2.16)$$

Equating Eqs. (2.15) and (2.16), a value for the permittivity is found as $\epsilon_r = 4/3$, which is effectively the lower limit that will result in a single beam that is achievable at any given scan angle. The maximum achievable scan angle can then be determined by first finding the required substrate thickness when the $n = 2$ mode is excited, and then proceeding to evaluate Eq. (2.14) to find the scan angle. The maximum achievable scan angle (without exciting higher order modes) is then given by

$$\theta_p = \sin^{-1} \frac{\sqrt{3\epsilon_r}}{2}. \quad (2.17)$$

The last remaining design parameter is the spacing between patch elements. In order to prevent grating lobes existing at any given scan angle, the element spacing should be constrained. This value can be found by setting the $n = -1$ mode at backward endfire, while the main $n = 0$ mode is at forward endfire. The $n = 0$ mode will thus be at the forward endfire position when

$$\sqrt{k_1^2 - \left(\frac{\pi}{h}\right)^2} = k_0. \quad (2.18)$$

Furthermore, the $n = -1$ mode radiates in the backward endfire direction when

$$\sqrt{k_1^2 - \left(\frac{\pi}{h}\right)^2} - \frac{2\pi}{a} = -k_0, \quad (2.19)$$

where a indicates the element periodicity. Once again enforcing these two equations, a limit on the periodicity that will prevent grating lobes is established as $a/\lambda_0 < 0.5$. This relationship is commonly found in array design.

Research Review

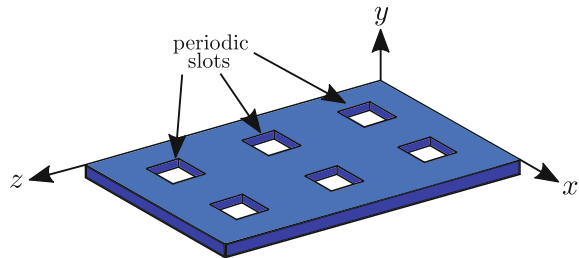
The proposed results were obtained around 12 GHz, but nonetheless the metal patch array leaky-wave antenna proposed and investigated by Zhao et al. has inspired several other designers in the millimeter region to approach leaky-wave antennas in a different manner. This approach is essentially an alternative to dielectric layer leaky-wave antennas [23, 24]. The main advantage offered by the metal patch approach is in terms of directivity. For the dielectric layer antenna to achieve very high directivity, the substrate permittivity needs to be exceedingly high. On the other hand, it is far simpler to obtain high directivity with the patch method.

Slot arrays are an important class of millimeter-wave antennas, and the structure is based on cutting slots with different shapes and rotations into the walls of a waveguide [25]. This technique can be implemented with rectangular waveguides, or with substrate integrated waveguide technology.

One such slot array has been proposed by Zhao et al. [26]. Both the metal patch and slot array configurations were able to achieve a scan range of approximately 60° , if an air substrate was used. However, upon substituting air with a dielectric substrate, scanning to the horizon was made possible in both the E- and H-planes. For practical dimensions, it is simpler to generate narrower beamwidths by using a slot array, since there is effectively less leakage.

Originally investigated by Xu et al. [27], the concept of deriving leaky-wave antennas from substrate integrated waveguides is essentially a continuation of the earlier work. A substrate integrated waveguide is an extremely popular transmission

Fig. 2.11 Leaky-wave antenna derived from a periodically slotted substrate integrated waveguide



line at microwave frequencies and by constructing such an open periodic waveguide with large spacing between vias (or slots) can be used for leaky-wave antenna design. A novel design and simulation approach was proposed, and the measured return loss was lower than -15 dB over the 28–34 GHz band (a 6 GHz bandwidth). An example of such a slotted array built on a substrate integrated waveguide is shown in Fig. 2.11.

The work also illustrated the differences between the leaky-wave properties of the TE_{20} and TE_{10} modes, concluding that the TE_{20} mode exhibits improved radiation properties compared to the TE_{10} mode. Discussed earlier in this chapter, the design highlighted the importance of substrate permittivity and slot periodicity with regard to radiation efficiency, directivity, and frequency scanning capability.

Several other reports on leaky-wave antennas based on substrate integrated waveguides have surfaced in recent years. A slot array that uses a plastic substrate, fabricated using micromachining and injection molding techniques, has been demonstrated by Fuh et al. [28]. The input flange is integrated directly into the waveguide through a one-shot molding process, significantly increasing the consistency and robustness of the manufacturing process. The design of this antenna explores many interesting manufacturing techniques. First, the waveguide is molded from a polymeric material through an injection molding process, whereafter a metallic layer is deposited onto the waveguide through a sputtering and electroplating process. A self-alignment mechanism is introduced between the top metallic plate and the plastic waveguide by using precision alignment pins. Finally, the feed connectors are constructed from an integrated flange that is molded together with the plastic waveguide. All of these features serve to reduce manufacturing costs and this antenna provides a flexible architecture for radar designers.

Measurements of this antenna were taken in the W-band region (73–79 GHz). To investigate the effects of surface roughness on the signal attenuation, a surface profiler was used to scan the waveguide area. An RMS surface roughness of 1.0 ± 0.3 μm was found, which is higher than that of most commercially available metallic waveguides, which generally exhibit a surface roughness of around 0.8 μm on copper. The -10 dB bandwidth was measured as 9.3 GHz, with a 9.6 dB gain and -13.5 dB sidelobe levels.

Another radar antenna based on substrate integrated technology was recently proposed by Cheng et al. [29]. The system consists of a monopulse comparator, a 16-way power divider network and a 32×32 slot array, all integrated on a single

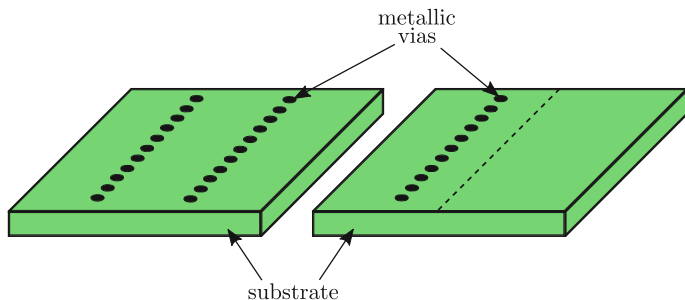


Fig. 2.12 Substrate integrated waveguide; the regular configuration is shown on the *left*, with the half-mode variant on the *right*

substrate. Monopulse antennas are particularly attractive for high-resolution tracking applications, and this antenna provides a highly directive beam and operates around 93–96 GHz. The system is etched on a single Rogers RT/Duroid 5880 substrate with a thickness of 0.508 mm and a dielectric constant of $\epsilon_r = 2.2$. The final assembly is 130×125 mm in size.

As expected from such a large array, the antenna produces a narrow beamwidth of 2° – 3° in the E-plane and 3° – 4° in the H-plane, when excited at the sum port. The minimum and maximum values of the measured gain vary between 21.29 dBi at 96 GHz and 25.75 dBi at 93.6 GHz. Sidelobe levels varied between 6 and 20 dB in the H-plane and 10–12 dB in the E-plane.

When excited at the first difference port (in this case, the H-plane difference port), the optimal amplitude balance between the two difference beams was measured as 2.12 dB at 95.2 GHz. Moreover, the lowest null depth was measured as -45.81 dB at 93.8 GHz. The second (E-plane) difference port excitation produced a maximum null depth of -37.71 dB at 95.2 GHz, with an optimal amplitude balance of 0.4 dB achieved at 93.6 GHz. While the slot array is by definition a leaky-wave antenna, the substrate integration with the combiner and coupler network creates an integrated antenna system. Such antennas are the focal point of the next chapter.

The half-mode substrate integrated waveguide has been used as a feed method for transverse slot arrays, and reported by Wu et al. [30]. Figure 2.12 depicts a half-mode substrate integrated waveguide.

This guide was first proposed by Wu et al.—the antenna work followed soon after—and it is advantageous because it retains all the characteristics of the full-mode substrate integrated waveguide, while occupying half the space [31]. For their experimentation process, two transverse slot array antennas were built for two different frequencies. The first design was for a center frequency of 9.8 GHz, while the second was for 31.4 GHz. Both arrays consisted of a total of eight slots and were fabricated on a low-loss 0.508 mm thick substrate. The antennas were fairly narrowband, and measurements of the first antenna revealed a 2.8 % impedance bandwidth, and 2.7 % for the second. The measured 3 dB beamwidth in the

E-plane also remained similar between the two, with 23.3° for the low-frequency version and 21.0° for the high-frequency version.

Work done by Lai et al. further demonstrated the use of the half-mode substrate integrated waveguide, in this instance to feed a dielectric resonator antenna [32]. The energy is coupled from the waveguide to the dielectric resonator through an aperture that separates the two. Experimentation revealed an impedance bandwidth of 24.2 % (for $S_{11} < -10$ dB) around 60 GHz, and maximum radiation efficiency of 92 %. Therefore, this antenna covers the entire ISM band at 60 GHz; the measured gain was around 5.5 dB.

2.4 Closing Remarks

As is evident throughout this chapter, despite the state of development of leaky-wave antennas, there is still a lot of room for innovation in terms of extending these antennas into the millimeter-wavelength region. Trends continue to follow substrate integrated techniques, as opposed to conventional waveguides, as well as antenna on-chip systems.

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