Microstrip Antennas

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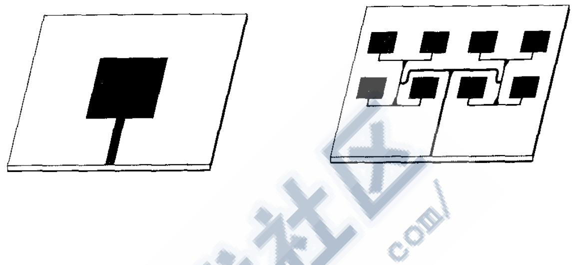
Invited Paper

Microstrip antennas have been one of the most innovative topics in antenna theory and design in recent years, and are increasingly finding application in a wide range of modern microwave systems. This paper begins with a brief overview of the basic characteristics of microstrip antennas, and then concentrates on the most significant developments in microstrip antenna technology that have been made in the last several years. Emphasis is on new antenna configurations for improved electrical performance and manufacturability, and advances in the analytical modeling of microstrip antennas and arrays.

I. Introduction

Although antenna engineering has a history of over 60 years, it remains, in the words of a recent review article [1], " . . . a vibrant field which is bursting with activity, and is likely to remain so in the foreseeable future." Within this field, if the numbers of journal articles, symposia papers, workshops, and short courses are any indication, it is probably safe to say that microstrip antennas form one of the most innovative areas of current antenna work. This paper will review some of the most significant advances in the design and modeling of microstrip antennas that have been made in the last several years.

The idea of the microstrip antenna dates back to the 1950's [2], [3], but it was not until the 1970's that serious attention was given to this element. As shown in Fig. 1(a), the basic configuration of a microstrip antenna is a metallic patch printed on a thin, grounded dielectric substrate. Originally, the element was fed with either a coaxial line through the bottom of the substrate, or by a coplanar microstrip line. This latter type of excitation allows feed networks and other circuitry to be fabricated on the same substrate as the antenna element, as in the corporate-fed microstrip array shown in Fig. 1(b). The microstrip antenna radiates a relatively broad beam broadside to the plane of the substrate. Thus the microstrip antenna has a very low profile, and can be fabricated using printed circuit (photolithographic) techniques. This implies that the antenna can be made conformable, and potentially at low cost. Other advantages include easy fabrication into



(a) (b)

Fig. 1. (a) A rectangular microstrip antenna with a microstrip line feed. (b) An eight-element linear array of microstrip patches with a coplanar microstrip feed network.

linear or planar arrays, and easy integration with microwave integrated circuits.

Since the original configuration was proposed, literally dozens of variations in patch shape, feeding techniques, substrate configurations, and array geometries have been developed by researchers throughout the world. The variety in design that is possible with microstrip antennas probably exceeds that of any other type of antenna element. Another interesting feature is that microstrip antennas can be fabricated rather easily in university or other research laboratories, which have often been a source of novel designs. Good reviews of much of this work, including the basic properties, analytical models, and design techniques for microstrip antennas, can be found in [4]–[8].

To a large extent, the development of microstrip antennas has been driven by systems requirements for antennas with low-profile, low-weight, low-cost, easy integrability into arrays or with microwave integrated circuits, or polarization diversity. Thus microstrip antennas have found application in both the military and the civil sectors, as shown in Table 1.

Disadvantages of the original microstrip antenna configurations include narrow bandwidth, spurious feed radiation, poor polarization purity, limited power capacity, and tolerance problems. Much of the development work in microstrip antennas has thus gone into trying to overcome these problems, in order to satisfy increasingly stringent systems requirements. This effort has involved the devel-

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Table 1 Some Applications of Microstrip Antennas

Platform	Systems
Aircraft	Radar, communications, navigation, altimeter, landing systems
Missiles	Radar, fuzing, telemetry
Satellites	Communications, direct broadcast TV, remote sensing radars and radiometers
Ships	Communications, radar, navigation
Land vehicles	Mobile satellite telephone, mobile radio
Other	Biomedical systems, intruder alarms

opment of novel microstrip antenna configurations, and the development of accurate and versatile analytical models for the understanding of the inherent limitations of microstrip antennas, as well as for their design and optimization.

In the following section some of the basic mechanisms and electrical characteristics of microstrip antennas are briefly reviewed; because of space constraints, further detail must be left to the references. The emphasis of the paper is on recent advances in microstrip antennas, as divided into two headings. Section III discusses new configurations of microstrip antennas that enhance their utility, such as alternative feeding techniques, geometries for enhanced bandwidth, and some new array concepts. Section IV discusses recent developments in the analysis and modeling of microstrip antennas and arrays, emphasizing full-wave solutions for a variety of geometries. Section V closes with a short discussion of future developments in microstrip antenna technology.

II. Basic Characteristics of Microstrip Antennas

In this section we will summarize the basic operation and electrical characteristics of microstrip antennas, as an introduction for the nonspecialist and to provide context for the next two sections.

Consider a basic rectangular microstrip antenna with a probe feed, as shown in cross section in Fig. 2(a). When operating in the transmitting mode, the antenna is driven with a voltage between the feed probe and the ground plane. This excites current on the patch, and a vertical electric field between the patch and the ground plane. The dielectric substrate is usually electrically thin ($d < 0.05\lambda_o$), so electric field components parallel to the ground plane must be very small throughout the substrate. The patch element resonates when its length is near $\lambda/2$, leading to relatively large current and field amplitudes. In view of the equivalence theorem, there are several ways to interpret the resulting radiation. The antenna can be viewed as a cavity with slot-type radiators at $x = 0$ and $x = L$, with equivalent magnetic currents, $\bar{M} = \bar{E} \times \hat{n}$, radiating in the presence of the grounded dielectric substrate. Alternatively, radiation can be considered as being generated by the induced surface

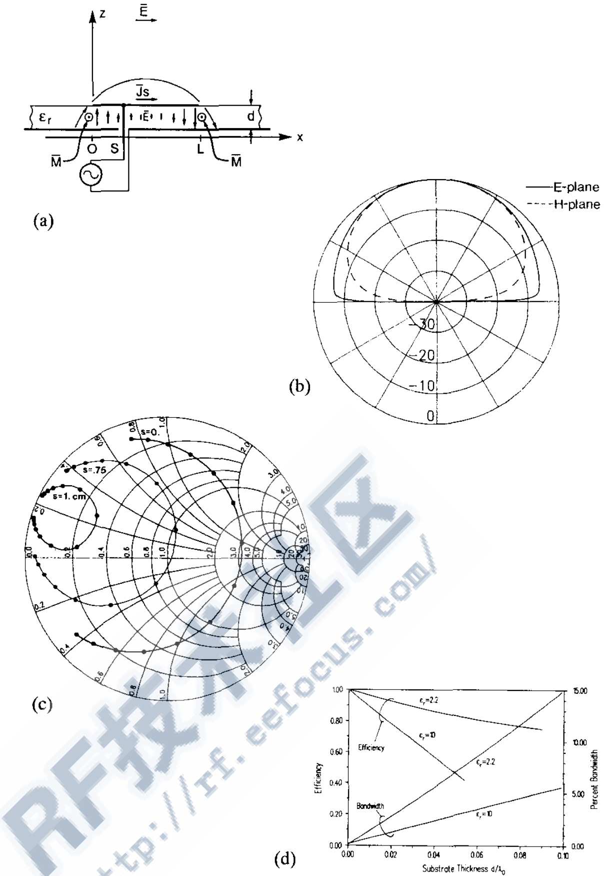


Fig. 2. Basic characteristics of a microstrip antenna. (a) Cross-section view of a probe-fed patch, showing the electric field lines below the patch, the surface current on the patch, and the equivalent magnetic currents at the radiating edges of the patch. (b) Principal plane radiation patterns of a microstrip antenna. (c) Input impedance loci of a probe-fed patch antenna versus probe position (substrate: $\epsilon_r = 2.2$, $d = 0.79$ mm; patch size: $L = 2.5$ cm, $W = 3.0$ cm; frequency span: 3600–4250 MHz, in steps of 50 MHz). (d) Efficiency and bandwidth of a microstrip antenna versus substrate thickness.

current density, $\bar{J}_s = \hat{n} \times \bar{H}$, on the patch element in the presence of the grounded dielectric substrate. In either case, the equivalent sources produce a broadside radiation pattern as shown in Fig. 2(b).

Early analytical models treated the microstrip antenna as a lossy cavity, or as a loaded transmission line resonator [4], [6]–[9]. These models involve several simplifying approximations, so they are easy to use and can provide useful information on antenna patterns, impedance, efficiency, and bandwidth. Although their accuracy and versatility are much less than more rigorous solutions (as discussed in Section IV), these simple models can provide an extremely useful intuitive view of the operation of a microstrip antenna. For example, in the cavity model the perimeter of the patch element is approximated as a magnetic wall, for which the electric and magnetic fields of the dominant resonant mode can be expressed as,

$$\text{vertical electric field} = E_z = E_o \cos \frac{\pi x}{L} \quad (1a)$$

$$\text{transverse magnetic field} = H_y = H_o \sin \frac{\pi x}{L}. \quad (1b)$$

Higher order resonant modes have more variations in the x- and/or y-directions. This result ignores the effect of fringing fields at the edges of the patch, so an ad hoc correction factor must be applied to the patch length to obtain accurate resonant frequencies. An effective dielectric loss tangent is used to account for power lost to radiation.

In terms of the transmission line model, the antenna is viewed as a length of open-circuited transmission line with light loading at the ends to account for fringing fields and radiation. The voltage and current on this equivalent transmission line can be approximated as,

$$\text{voltage} = V(x) = V_o \cos \frac{\pi x}{L} \quad (2a)$$

$$\text{current} = I(x) = \frac{V_o}{Z_o} \sin \frac{\pi x}{L}. \quad (2b)$$

This result gives a simple explanation for the input impedance ($Z_{in} = V/I$) variation of the microstrip antenna. For a feed point at a radiating edge ($x = 0$ or $x = L$), the voltage is a maximum and the current is a minimum, so the input impedance is a maximum. For a feed point at the center of the patch ($x = L/2$), the voltage is zero and the current is a maximum, so the input impedance is zero. Thus the input impedance can be controlled by adjusting the position of the feed point; typical input impedances at an edge of a resonant patch range from 150Ω to 300Ω. The impedance locus is that of a half-wave open-ended transmission line resonator, which can be modeled as a parallel *RLC* network. Typical examples are shown on a Smith chart in Fig. 2(c). The resonant patch element is sometimes made in a different shape (circular being the next most common), but the principles of operation are essentially the same as for the rectangular patch.

The magnetic wall approximation of the cavity model and the open-end approximation of the transmission line model become more realistic as the substrate becomes thinner. This implies that the Q of a patch antenna on a thin substrate is large, and that the bandwidth is small. From an alternative viewpoint, the current on the patch element is in very close proximity to its negative image caused by the presence of the ground plane. This causes near-cancellation of radiated fields, and relatively large stored energy below the patch. The impedance bandwidth of a microstrip element is shown versus substrate thickness in Fig. 2(d). Note two important trends: the bandwidth increases with substrate thickness, and decreases with an increase in substrate permittivity. Both of these effects are explained by the above arguments.

This behavior leads us to conclude that microstrip antennas operate best when the substrate is electrically thick with a low dielectric constant. On the other hand, a *thin* substrate with a *high* dielectric constant is preferred for microstrip transmission lines and microwave circuitry. Herein lies one of the paradoxes associated with the microstrip antenna concept, since it has often been claimed that the microstrip antenna can easily be integrated with a feed network and circuitry on the same substrate. If this is done, some compromise must be made between good antenna performance and good circuit performance. The root of this problem lies in the fact that antenna radiation and circuitry are distinct electrical functions, since antennas require loosely bound fields for radiation into space, while circuitry requires tightly bound fields to prevent undesired radiation or coupling. Several of the new antenna configurations discussed in Section III have been developed to overcome this fundamental problem.

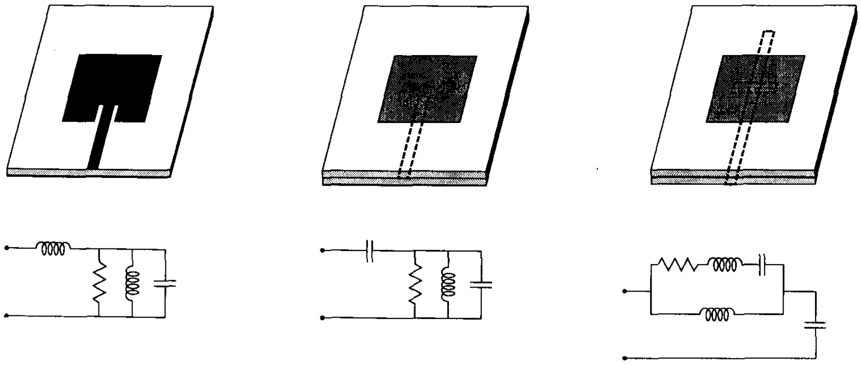
Losses in a microstrip antenna occur in three ways: from conductor loss, dielectric loss, and surface wave excitation [10]. Except for extremely thin substrates, conductor and dielectric losses for a microstrip element are quite small, usually accounting for no more than a few percent loss in radiation efficiency. Surface waves bound to the dielectric substrate can be excited by the antenna, and since it does not contribute to the primary radiation pattern of the antenna, surface wave power is generally considered as a loss mechanism. But surface wave power can also diffract from substrate edges or other discontinuities to degrade the antenna radiation pattern or polarization characteristics. Surface wave power generated by a single element increases with substrate thickness and dielectric constant, as shown in Fig. 2(d). This is another reason for preferring a substrate with a low dielectric constant. In large arrays of microstrip antennas, destructive interference of surface wave power can occur, raising the radiation efficiency, although at certain scan angles a constructive interference may be possible, leading to a scan blindness effect [11].

Besides the inherent antenna element losses described above, one must often consider the effect of losses due to feed lines and associated feed circuitry. These can be the dominating loss contribution in a large array (see the article by Mailloux in this issue).

III. New Microstrip Antenna Configurations

A. Feeding Techniques

Early microstrip antennas used either a microstrip feed line or a coaxial probe feed [4], [6], [7]. These two feeding methods are very similar in operation, and offer essentially one degree of freedom (for a fixed patch size and substrate) in the design of the antenna element through the positioning of the feed point to adjust the input impedance level. For the case of a microstrip line feed, the patch can be notched to provide an inset feed point, as shown in Fig. 3(a). This figure also shows the equivalent circuit that applies to either of these feeds; the parallel *RLC* network represents the resonant patch, while the series inductor represents feed



(a) (b) (c)

Fig. 3. Three types of feeding methods for microstrip antennas, and their equivalent circuits. (a) Patch fed with an inset microstrip line feed. (b) Patch proximity coupled to a microstrip feed line. (c) Patch aperture coupled to a microstrip feed line.

inductance of the coaxial probe or the microstrip feed line. These feeds excite the patch through the coupling of the J_z feed current to the E_z field of the dominant patch mode:

$$\text{coupling} \sim \int_v E_z J_z dv \sim \cos\frac{\pi s}{L} \tag{3}$$

where L is the resonant length of the patch, and s is the offset of the feed point from the patch edge. This result shows that maximum coupling occurs for a feed at a radiating edge of the patch ($s = 0$ or L).

Such direct contacting feeding methods have the advantage of simplicity, but also have several disadvantages. First, these configurations suffer from the bandwidth/feed radiation trade-off discussed above, where an increase in substrate thickness for the purpose of increasing bandwidth leads to an increase in spurious feed radiation, increased surface wave power, and possibly increased feed inductance [12]. For practical purposes such antennas are thus limited in bandwidth to about 2%–5%. Coaxial probes can be used to feed patch elements through the ground plane from a parallel feed substrate [8], but in an array having thousands of elements such a large number of solder joints makes fabrication difficult and lowers reliability (an especially important consideration for space applications). Finally, although probe and microstrip line feeds primarily excite the dominant mode of the patch element, the inherent asymmetry of these feeds generates some higher-order modes which produce cross-polarized radiation.

In recent years a variety of noncontacting feeds have been developed for microstrip antennas. The proximity feed shown in Fig. 3(b) uses a two-layer substrate with a microstrip line on the lower substrate, terminating in an open stub below the patch which is printed on the upper substrate. This technique was originally developed for printed dipoles [13], [14], but has also been used for microstrip patches (proximity coupling is often referred to in the literature as "electromagnetic coupling," but this is an overly broad description). Proximity coupling has the advantage of allowing the patch to exist on a relatively thick substrate, for improved bandwidth, while the feed line sees an effectively thinner substrate, which reduces spurious radiation and coupling. Fabrication is a bit more difficult because of the requirement for reasonably accurate alignment between substrates, but soldering is eliminated.

The proximity coupled patch has at least two degrees of freedom: the length of the feeding stub and the patch width-to-line width ratio. The capacitive nature of this coupling method is reflected in the fact that the equivalent circuit, shown in Fig. 3(b), has a capacitor in series with the parallel *RLC* resonator that represents the patch. This can be used to impedance-match the antenna, as well as aiding in the tuning of the element for improved bandwidth. Bandwidths of 13% have been achieved in this manner [15].

Another type of noncontacting feed is the aperture coupled microstrip antenna [16], [17]. As shown in Fig. 3(c), this configuration uses two parallel substrates separated by a ground plane. A microstrip feed line on the bottom substrate is coupled through a small aperture (typically a narrow rectangular slot) in the ground plane to a microstrip patch on the top substrate. This arrangement allows a thin, high dielectric constant substrate to be used for the feed, and a thick, low dielectric constant substrate for the antenna element, thus allowing independent optimization of both the feed and the radiation functions. In addition, the ground plane eliminates spurious radiation from the feed from interfering with the antenna pattern or polarization purity. An important aspect of the aperture coupled patch is the fact that the aperture is usually smaller than resonant size, so the backlobe radiated by the slot is typically 15–20 dB below the forward main beam.

This geometry has at least four degrees of freedom: the slot size, its position, the feed substrate parameters, and the feed line width. Impedance matching is performed by adjusting the size (length) of the coupling slot, together with the width of the feed line, which is usually terminated in an open-circuited tuning stub. Coupling can occur via an equivalent electric or magnetic polarizability in the slot, but the magnetic case is the stronger mechanism, resulting in a coupling of the equivalent magnetic current M_y in the aperture to the dominant H_y field of the patch:

$$\text{coupling} \sim \int_v M_y H_y dv \sim \sin \frac{\pi s}{L}. \tag{4}$$

This result is the dual of the probe coupled case (3), and shows that maximum coupling occurs when the aperture is centered below the patch ($s = L/2$), where the magnetic field is maximum [16]. In this case the excitation of the patch is symmetric, which reduces the excitation of higher-order modes and leads to very good polarization purity [18]. The aperture coupled patch with a centered feed has no cross-polarization in the principle planes.

The equivalent circuit of an aperture coupled microstrip antenna is shown in Fig. 3(c). The patch resonator now appears as a series *RLC* network (since a series-type feed at the center of the patch is $\lambda/4$ away from a shunt-feed at the edge of the patch, resulting in an admittance inverter effect), with a shunt inductance representing the coupling slot. This network allows the possibility of double tuning for increased bandwidth, as will be discussed in Section III-B. A variation of this design uses a feed substrate that is oriented perpendicular to the antenna substrate [19].

The above feeding techniques use a single feed point to generate linear polarization. While it is possible to produce circular polarization with a single feed [4], [8], such techniques are usually very narrow-band. Thus it is more common to generate circular polarization by using two feed points to excite two orthogonal modes on the patch with a 90° phase difference between their excitations, as shown in Fig. 4(a). This technique has been extensively used with probe feeds and microstrip line feeds [4], [6]–[8], and can also be applied to proximity coupled patches and aperture coupled patches [20]. An alternative technique that works well for small subarrays is to sequentially phase the excitations of an array of patches that radiate orthogonal sets of linear (or circular) polarizations [21], [22]. An example of this type of arrangement is shown in Fig. 4(b). Yet another approach is to use sets of printed dipoles and slots proximity coupled to a microstrip feed line with the proper phasings to achieve circular polarization [23].

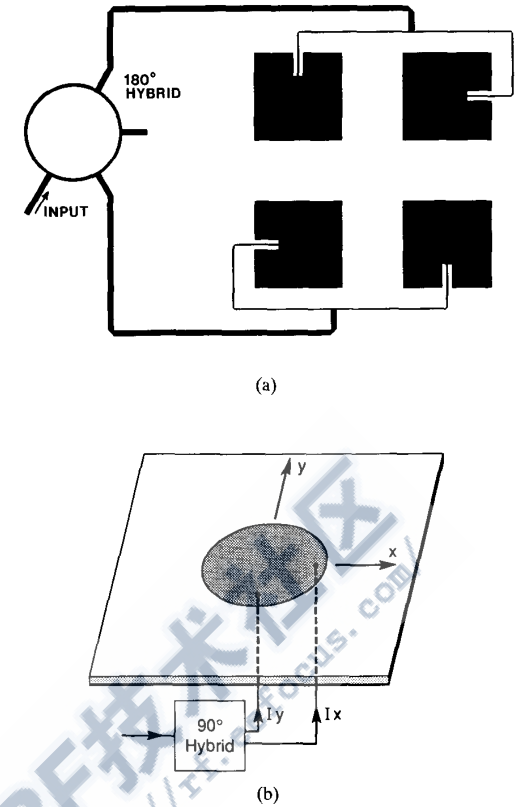


Fig. 4. Techniques for obtaining circular polarization. (a) Using two feed points for exciting two orthogonal modes on the patch with a 90 degree phase difference. (b) Using a subarray of orthogonally polarized elements with sequential phase rotations (the phasings at the four patches are 0°, 90°, 180°, and 270°).

B. Bandwidth Enhancement Techniques

In many cases the narrow bandwidth (2%–5%) of the traditional microstrip antenna element is its most serious disadvantage, preventing its use in many practical microwave applications. Thus a large amount of effort has gone toward the development of creative designs and techniques for improving the bandwidth of the microstrip antenna. A good review of many of these methods is given in [24]. All these techniques basically work to overcome the fundamental bandwidth limitation set by the small electrical volume occupied by the element [24], and most can be classified as either using an impedance matching network, or parasitic elements. In both cases a double-tuning effect is often exploited.

The most direct method of increasing the bandwidth of the microstrip element is to use a thick, low dielectric constant substrate. But as discussed above, this inevitably leads to unacceptable spurious feed radiation, surface wave generation, or feed inductance. Since the bandwidth of the element is usually dominated by the impedance variation (the pattern bandwidth is generally much better than the impedance bandwidth), it is often possible to design a planar impedance matching network to increase bandwidth. Bandwidths of 9%–12% [25] and 15% [26] have been obtained in this manner for probe-fed and microstrip line-fed elements. A bandwidth of 13% was achieved for a proximity coupled patch element with a stub tuning network [15]. Spurious radiation from the matching network may be a concern, however, if the matching network is coplanar with the antenna element.

If one is willing to use a more complicated element geometry, increased bandwidth can be obtained in a variety of ways by using parasitically coupled elements to produce a double-tuned resonance. One of the best ways to do this is with two stacked patches, as shown in Fig. 5. The top patch is proximity coupled to the bottom patch, which can be fed by any of the methods discussed in Section III-A. Bandwidths of 10%–20% have been achieved with probe-fed stacked patches [27]–[29], and 18%–23% bandwidths

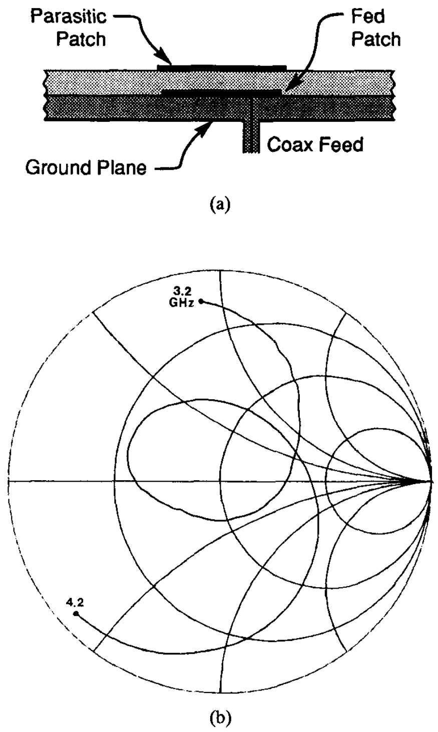


Fig. 5. Bandwidth enhancement using stacked patches. (a) Geometry of stacked patch antenna, with a probe-fed bottom patch. (b) Typical impedance locus showing characteristic double-tuned response (top and bottom substrates: $\epsilon_r = 2.2$, $d = 1.6$mm; top patch size: $L = W = 2.55$cm; bottom patch size: $L = W = 2.5$cm).

have been achieved for aperture coupled stacked patches [30], [31]. Figure 5 shows a typical impedance locus for a stacked patch design, showing the characteristic double-tuned response.

Double tuning can also be obtained with a single aperture coupled microstrip patch by lengthening the coupling slot so that it is near resonance. This simplifies construction, relative to using stacked patches, but the larger coupling slot may result in a higher backlobe level. Bandwidths in excess of 20% have been achieved for single elements and arrays using this technique [32], [33]. Figure 6 shows the geometry and impedance locus for an element of this type.

It is also possible to use parasitic elements coplanar with a driven element to produce double tuning [4], [8]. This configuration can be a disadvantage in array applications, however, because the lateral extent of such an element usually requires array spacings greater than $\lambda/2$; additionally, the phase center of this element will not be fixed over its operating bandwidth. Finally, the stagger-tuned resonator concept can be generalized to a log-periodic array of printed dipoles to achieve multioctave bandwidths [34].

C. Microstrip Arrays

One of the best features of microstrip antennas is the ease with which they can be formed into arrays, and a wide variety of series-fed, corporate-fed, scanning, and

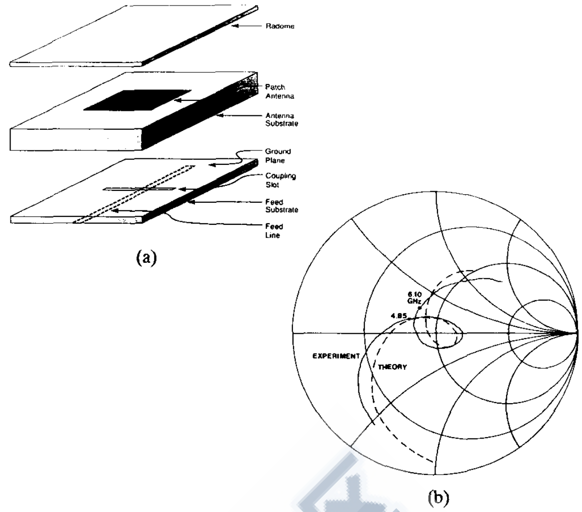


Fig. 6. Bandwidth enhancement using a double-tuned aperture coupled patch antenna. (a) Geometry of antenna element. (b) Resulting impedance locus (radome: $\epsilon_r = 2.2$, $d = 1.6$ mm; antenna substrate: $\epsilon_r = 1.01$, $d = 5.5$ mm; feed substrate: $\epsilon_r = 2.2$, $d = 0.762$ mm; patch size: $L = W = 17$ mm; slot size: $L = 15.4$mm, $W = 0.8$ mm; feed line width = 2.32 mm) Impedance bandwidth for SWR$\leq$ 1.5 is 22%. Data courtesy of F. Croq [33].

polarization-agile arrays have been designed using microstrip elements [5]–[8]. Here we will summarize some of the most recent advances in microstrip array design, including arrays using noncontacting feeds, the problem of spurious feed radiation, and the monolithic phased array. Further discussions of some of these topics can be found in other papers in this issue.

Series-fed arrays offer a very convenient form of array fabrication because both the feed network and the radiating elements can be made photolithographically, without any need for soldering to the elements. This technique is limited to fixed-beam or frequency-scanned arrays, but linear or planar arrays can be made, with single or dual polarization capability. Because of the fact that a change in the excitation of one element affects the excitations of all other elements in a series-fed array, it is very helpful to have an accurate CAD capability (including the effects of internal reflections, and possibly mutual coupling), for the design of such arrays. Progress in this area has been made for series-fed linear arrays of proximity coupled dipoles [35], [36], and aperture coupled patches [37], which have been designed using CAD models. Arrays using proximity coupled elements have the advantages of improved bandwidth and reduced spurious radiation over microstrip line-fed elements, but feed radiation is still high enough so that cross-polarization or side lobe levels better than about 20–25 dB are unlikely to be attained. This is not a problem with aperture coupled patch arrays, since the feed lines are shielded by the ground plane.

Corporate-fed arrays are generally used when more control is needed for the excitation of individual array elements, as in phased arrays, multibeam arrays, or shaped-beam arrays. Figure 7 shows a corporate-fed phased array antenna using MMIC modules for phase shifters and amplifiers with probe-fed microstrip elements. Proximity and aperture coupled elements are beginning to be considered for such arrays, and some progress has been made in the development of novel configurations and CAD models for arrays of this type [38].

The experience of a number of workers who have been designing and testing microstrip arrays in recent years has led to the conclusion that feedline radiation from series or corporate feeds is an unavoidable problem that sets a lower limit on cross-polarization and side lobe levels [39], [40]. Depending on the substrate thickness and the topology of the feed network, these levels may range from -15 to -25 dB. Achievement of better cross-polarization or side lobe levels requires that the feed network be isolated by a ground plane from the radiating face of the array, and coupled to the radiating elements with either feed-through probes or by aperture coupling. This and other factors were considered in a study of low side lobe microstrip arrays [41], where an array with a 40-dB relative (-35 dBi) side lobe level was demonstrated. Figure 8 shows the measured pattern of this array.

Microstrip antenna technology allows a greater level of integration with feed and control circuitry than ever before, and the culmination of this trend is the monolithic phased array antenna, which integrates radiating elements, feed circuitry, phase shifters, and amplifiers on a single monolithic chip. The design of such an antenna is one of the most challenging subjects in the antenna field today, involving a wide variety of interrelated issues such as antenna bandwidth, surface wave effects, feed losses, active circuit gain, active device yields, substrate selection, material properties, and heat dissipation, as well as the overall array architecture. These problems are formidable enough that no truly monolithic array has yet been produced, but the potential cost savings that would be realized relative to conventional array technology make this concept worth pursuing.

The specific substrate configuration of a monolithic array determines many of its electrical, thermal, mechanical, and fabricational characteristics. Thus there is merit in considering array geometries that use more than one substrate, possibly with noncontacting feeds, and there have been a variety of novel monolithic array architectures proposed along these lines [42], [43], each with its own benefits and trade-offs. Some of these different approaches are reflected in recent monolithic array work, which includes a Ka-band linear patch array on Duroid with coplanar MMIC phase shifter modules [44], a planar array design at 40 GHz that uses a stacked patch antenna element over a GaAs substrate for active circuits [45], and a "brick" construct using planar elements on a set of substrates mounted perpendicular to the array face [46]. This latter technique allows the use of separate substrates for the antenna elements and the MMIC's, and also provides a high degree of modularity and facilitates heat removal.

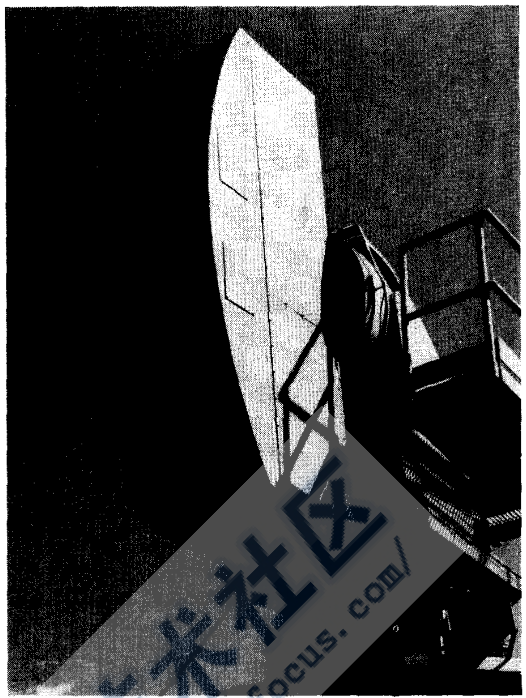


Fig. 7. Photograph of a 64-element S band conformal microstrip phased array antenna. Each element is driven with a $GaAs$ MMIC T/R module containing a 5-bit phase shifter, a low-noise amplifier, and a power amplifier. The array scans $\pm 60°$ in azimuth. It was developed by Raytheon Co. for the Rome Air Development Center, Hanscom AFB, MA. Photograph courtesy of Jerome D. Hanfling, Raytheon Co., Equipment Division.

IV. Advances in the Analysis and CAD of Microstrip Antennas

Antenna analysis is important for several reasons. It can aid the design process by reducing the amount of costly and time-consuming cut-and-try cycles, which is especially useful when optimizing an antenna design over one or more design parameters. Analysis can also provide an understanding of the operating principles of an antenna, which is useful in design as well as in the determination of the limitations in the performance of the antenna, and in the development of new antenna configurations or modifications to an existing antenna design. A good antenna model thus has the following characteristics.

- It can be used to calculate all the necessary parameters for the antenna under consideration (impedance, patterns, etc.).
- Its results are accurate enough for the intended purposes (increased accuracy generally requires a more complex solution).
- It is as simple as possible, while providing the above features (the simplest solution will generally be the easiest to implement and the easiest to use).
- It lends itself to interpretation in terms of intuitive

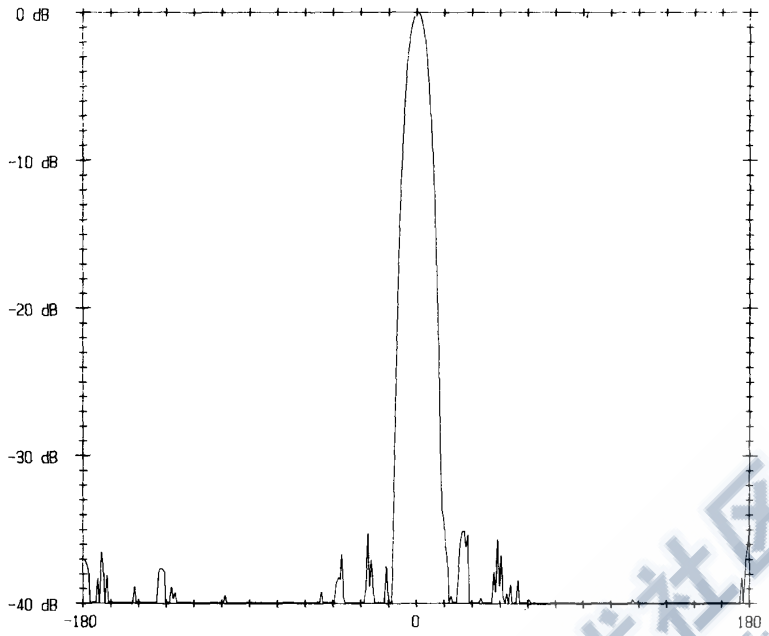


Fig. 8. Measured pattern of a 16-element C-band low side lobe microstrip array [41]. (substrate: $\epsilon_r = 2.22$, $d = 1.6$ mm; patch size: $L = W = 1.85$ cm; element spacing = 3.0 cm; frequency = 5.0 GHz) The patches are probe-fed, with a 40-dB Chebyshev amplitude taper.

models. (the simplest solution often provides the most physical insight).

Because it generally requires the full solution of Maxwell's equations, antenna modeling is a challenging problem, and one that is still undergoing a significant amount of development. Before computers were as pervasive as they are today, antenna analysis was limited to geometries that were amenable to solution by boundary value or variational techniques. Computer solutions using moment methods, finite difference techniques, and asymptotic techniques have greatly expanded the set of antenna geometries can be accurately and efficiently modeled. Thus there exists today software and CAD codes for fairly general analysis of wire antennas, reflector antennas, horn antennas, and waveguide slot antennas. Microstrip antenna CAD has not progressed as far as these other types of antenna codes, however, primarily because microstrip antennas are relatively new, and are somewhat more difficult to model.

Relative to other types of antennas, microstrip antenna analysis is complicated by the presence of a dielectric inhomogeneity, narrow-band electrical characteristics, and a wide variety of patch, feed, and substrate configurations. Thus existing microstrip antenna models invariably compromise one or more of the features listed above. Most theories to date can be categorized as either simplified, or reduced, analyses that maintain simplicity at the expense of accuracy or versatility, or full-wave models that maintain rigor and accuracy at the expense of computational simplicity. We will review some of the most recent work in both of these areas in this section.

A. Reduced Analyses

Reduced analyses refer to microstrip antenna models that introduce one or more significant (but reasonable) approximations to simplify the problem. These include the cavity model [4], [8], [9], which uses a magnetic wall boundary condition approximation for the periphery of the patch, the transmission line model [4], [8], which models the antenna as a transmission line section with lumped loads, and the multiport network model [8] which generalizes the cavity model. These models were the first to be developed for microstrip antennas, and have been extremely useful for practical design as well as providing a good intuitive explanation of the operation of the microstrip antenna. Drawbacks of these models have been limited accuracy for resonant frequency and input impedance for substrates that are not very thin [12], and a limited capacity to handle related problems such as mutual coupling, large arrays, surface wave effects, and different substrate configurations. Although many of these limitations are inherent in these modeling techniques, there have been notable advances in the accuracy of these models, and in their application to a broader range of microstrip antenna problems.

One such development has been the extension of the cavity model to mutual coupling calculations [47]. The basic cavity model with magnetic sidewalls precludes the effect of anything external to the cavity, such as nearby elements or radome layers, but by considering the interaction of equivalent magnetic slot radiators coupling between nearby patch elements can be calculated with reasonable accuracy in a very efficient manner. The cavity model has also been extended to the analysis of aperture coupled patch antennas [48], [49], with good results for the input impedance of antennas on thin substrates. A further generalization of the cavity model breaks the periphery of the patch into a set of ports, each with an effective load admittance modeling the edge of the patch as well as the feed point. This multiport network model [8], [50] has the advantage of being capable of handling fairly arbitrary patch shapes and multiport elements. It can also be used to model microstrip feeds, parasitic elements, and mutual coupling in arrays.

The transmission line model is one of the most intuitively appealing models for a microstrip antenna, but it suffers from limited accuracy and the need for several ad hoc correction factors. Improving the accuracy of the equivalent load admittances that represent the open ends of the patch, including the effect of radiation from the sides of the patch, and the effect of mutual coupling between the ends of the patch, has led to a much improved transmission line model [51], [8]. This technique has also been used for mutual coupling between two patches, in a manner similar to the cavity model.

Surface wave effects are important from the viewpoint of losses, and also play an important role in mutual coupling effects, but are not directly included in these models. Because of the fact that surface wave excitation is relatively independent of antenna element size or shape, it is possible to obtain general results for surface wave power from a full-wave analysis of an arbitrary microstrip element. Such results are simple enough to be reduced to closed-form expressions [8], [52] that can be used to augment cavity or transmission line models.

B. Full-Wave Analyses

Microstrip antenna models that account for the dielectric substrate in a rigorous manner are referred to as full-wave solutions. These models usually assume that the substrate is infinite in extent in the lateral dimensions, and enforce the proper boundary conditions at the air-dielectric interface. This is most commonly done by using the exact Green's function for the dielectric substrate, which allows space wave radiation, surface wave modes, dielectric loss, and coupling to external elements to be included in the model. Using the Green's function in a moment method solution results in a model that is accurate and extremely versatile, although the computational cost is high. Features of full-wave solutions include the following.

- Accuracy: full-wave analysis techniques generally provide the most accurate results for input impedance, mutual coupling, radar cross-section, etc.
- Completeness: full-wave solutions include the effects of surface waves, space wave radiation, and external coupling.
- Versatility: full-wave solutions can be implemented for arbitrary microstrip elements and arrays, various types of feeding techniques, multilayer geometries, and for anisotropic substrates.
- Computational complexity: full-wave solutions are numerically intensive, and require careful programming in order to be computationally efficient.

Green's function moment method solutions for printed antennas generally employ the electric field integral equation to solve for the unknown currents on antenna elements and feeds. This is done by expanding the unknown electric and/or magnetic currents in a set of expansion modes, then using a set of weighting modes to discretize the integral equation. The key step in this process is the evaluation of impedance matrix elements that involve the integration of the fields due to an expansion mode multiplied by a weighting mode. The form of this integration is as follows:

$$\int_x \int_y \int_{x_o} \int_{y_o} \oint_{k_x} \oint_{k_y} J_t(x,y) G(k_x,k_y) J_e(x_o,y_o) \cdot e^{-jk_x(x-x_o)} e^{-jk_y(y-y_o)} dk_y \, dk_x \, dy_o \, dx_o \, dy \, dx \tag{5}$$

where J_e is the expansion mode, G is the Fourier transform of the Green's function for the dielectric substrate, and J_t is the weighting (or testing) mode. This six-fold integration involves surface wave poles, a slowly converging integrand, and possibly near-singularities associated with the source point, thus requiring considerable care in its evaluation. It is also the most time-consuming part of the solution.

The order of integration in (5) is commonly chosen in one of two ways, leading to solutions with somewhat different characteristics. In the full-wave space domain approach [53]–[55], the transform variables k_x and k_y are converted to polar coordinates α and β, where $k_x = \beta \cos\alpha$, $k_y = \beta \sin\alpha$. Then the α-integration can be done in closed-form, so that (5) reduces to,

$$\int_x \int_y \int_{x_o} \int_{y_o} \oint_{\beta} J_t(x,y) H(x,y,x_o,y_o,\beta) \cdot J_e(x_o,y_o) d\beta \, dy_o \, dx_o \, dy \, dx. \tag{6}$$

Integrating H with respect to β gives the field at the point x, y due to a source at the point x_0, y_0, and so this approach is conceptually similar to standard moment method solutions for wire antennas, and the techniques that have been developed for those solutions can be applied. The remaining six integrations must be done numerically, but in many cases of practical interest Toeplitz-type symmetries can be exploited to eliminate one or two integrations. A problem that generally remains, however, is the near-singularity that occurs when the source point gets close to the field point during the course of the integration.

In the full-wave spectral domain approach [56], [57], the space integrations in (5) are done in closed-form. Because of the exponential kernels, this amounts to taking the Fourier transforms of J_t and J_e, which usually result in simple closed-form expressions. Then (5) reduces to,

$$\oint_{k_x} \oint_{k_y} F_t(k_x,k_y) G(k_x,k_y) F_e(k_x,k_y) dk_y \, dk_x \tag{7}$$

where F_e and F_t are the Fourier transforms of the expansion and weighting modes, respectively. The remaining inverse transform integrations must be done numerically. This can be a time-consuming process, because the integrand in (7) is slowly convergent. This slow convergence is a direct result of the fact that the source singularity that occurs in the space domain approach has been eliminated, but spread out, in the spectral domain. The spectral domain approach also has the advantage of being easily extended to the infinite array case; this basically amounts to replacing the double integration in (7) with a double summation. (It is interesting to note that the solution for an infinite array of elements is computationally much faster than the solution for a single element of the same type.)

Because of their appealing features, space domain and spectral domain full-wave solutions have dominated microstrip analysis in recent years. One geometry that has received significant attention in this regard is the basic probe-fed microstrip antenna. While this problem is relatively easy to treat in an approximate manner, its rigorous modeling is rather difficult because of the requirement for an accurate treatment of the rapidly varying fields and currents near the probe-to-patch attachment point. This was done for infinite arrays of probe-fed rectangular patches using a decomposition that separated the probe feed from the rest of the patch [58], and more recently using a special attachment mode to model the current flow from the probe to the patch [59]. This latter technique has also been applied to circular patches [60], [61]. The attachment mode approach used in [59] and [60] has also been applied to isolated rectangular and circular patches [62]. All these solutions are spectral domain solutions that enforce current continuity from the feed probe to the patch, and account for the probe self-reactance; calculated input impedance results are in good agreement with theory, for a wide range of substrate thicknesses and dielectric constants. Figure 9 shows a typical result, comparing calculated input impedance with measured data.

A related problem is the microstrip element with a microstrip line feed, which has been rigorously modeled by expanding the non-TEM microstrip line currents near the patch in a set of modes, and enforcing current continuity from the feed line to the patch [63]. An infinite array of proximity coupled patches was analyzed using a similar technique [64], and the model was used to optimize the antenna bandwidth over a 60° scanning range. Figure 10 shows an example from this work. Stacked patches with a probe-feed have been analyzed using the space domain method [65], using an idealized model for the probe-feed that is valid for thin substrates.

Aperture coupled patch antennas have also been modeled using the spectral domain technique. In [66] a rectangular aperture coupled element was modeled using a set of expansion modes on the feed line, while in [67] a reciprocity technique was used to handle the feed line. In both cases, the equivalent magnetic current in the coupling slot and the electric current on the patch are expanded in a set of expansion modes, and solved for using the electric

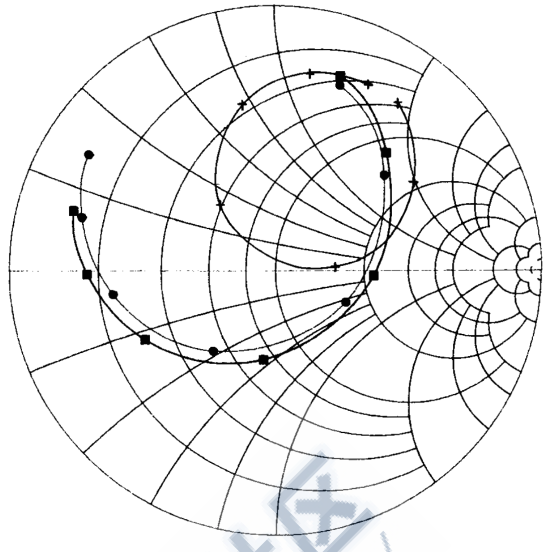


Fig. 9. Measured and calculated input impedance of a probe-fed rectangular microstrip antenna. * — experiment; • — theory using rigorous feed model [62]; ■ — theory using idealized feed model [56]; + — theory using cavity model. (substrate: $\epsilon_r = 2.20$, $d = 0.79$ mm; patch size: $L = 1.25$ cm, $W = 2.0$ cm; feed probe is 0.4 cm from a radiating edge). The frequency range is from 7.3 to 8.5 GHz, in steps of 0.2 GHz.

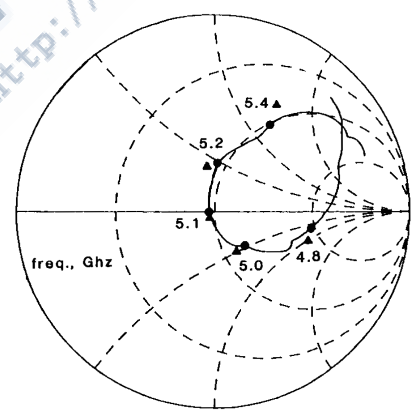


Fig. 10. Measured and calculated input impedance of an element in an infinite array of proximity coupled rectangular microstrip antennas (see Fig. 3(b)). ▲ — theory; • — experiment. (top and bottom substrates: $\epsilon_r = 2.33$, $d = 1.6$ mm; patch size: $L = W = 1.7$ cm; feed line width = 0.4 cm; E plane spacing = 3.35 cm; H plane spacing = 3.60 cm) The measurements were obtained using a waveguide simulator. (Data courtesy of J. Herd [64].)

field integral equation and the condition that the tangential magnetic field must be continuous in the slot. An infinite array of aperture coupled patches has also been treated in this manner [68].

The present state of affairs in microstrip antenna modeling is that there are a large number of separate codes for the solution of specific antenna geometries, with very little commonality or generalization. Making a fairly minor modification to the basic geometry (for example, adding a radome layer), can involve a fairly substantial effort, and usually engenders the development of a new solution and computer code. Besides being cumbersome, time-consuming, and expensive, this process reduces the likelihood of achieving a thorough or innovative design. Some progress has been made along these lines with the development of a full-wave model that can handle fairly arbitrary multilayer microstrip antenna geometries [69]. This model can treat an arbitrary multilayer substrate geometry, with proximity or aperture coupled rectangular patch or dipole elements. Stacked patches can be treated, and a general set of feed lines can be used, including microstrip, slotline, and stripline.

There have also been advances in alternative microstrip antenna modeling techniques which can be considered full-wave, but do not use a dielectric slab Green's function. Such solutions generally assume that the dielectric substrate is finite in size, and solve for the equivalent polarization currents in the dielectric, as well as the currents on the patch elements and feeds. This approach can thus account for finite substrate effects (such as space and/or surface wave diffractions from substrate edges), and has the advantage of a formulation that is less complex than a Green's function moment method approach. It is more of a "brute-force" technique, however, generally requiring the inversion of very large matrices. For geometries having rotational symmetry, considerable simplification can be achieved by formulating the solution as a body of revolution problem. This has been done for a circular microstrip antenna on a circular substrate [70].

V. Conclusions and Future Developments

This paper has emphasized two major aspects of the recent developments in microstrip antenna technology: new configurations of microstrip antennas for improved electrical performance and manufacturability, and recent advances in the analytical modeling of microstrip antennas and arrays. An extensive list of references has been given, but these represent only a fraction of the large volume of work on microstrip antennas performed by researchers throughout the world.

Microstrip antennas are one of the most innovative topics in antenna technology today, and this trend is likely to continue because the characteristics of microstrip antennas make them very appealing from a systems perspective. As discussed above, it is not the electrical characteristics of the basic microstrip antenna that has lead to this appeal, but rather its mechanical and fabricational features such as low-cost, lightweight, conformability, and easy integration with MIC's. Thus one of the main topics in this review has been the research and development in recent years that has gone toward improving the electrical characteristics of the microstrip antenna, such as increased bandwidth, novel feeding techniques, and control of spurious feed radiation, and it is expected that development will continue along these lines. The modeling of microstrip antennas and arrays is another area where much more work is needed, if the goal of providing versatile and accurate CAD software for microstrip antenna design is to be achieved.

Besides these topics, it is likely that microstrip antenna technology will form a synergistic link with one or more technologies that are just now emerging. High temperature superconductors are one such possibility, since these materials can be easily integrated with planar antennas. Material developments, in general, offer a variety of possibilities for improving the performance of microstrip antennas, in terms of low-loss substrates and improved thermal and mechanical characteristics. Materials with electric or magnetic anisotropy should also find useful applications in microstrip antennas and arrays. A likely possibility here is the use of ferrite materials, which have received little attention to date for microstrip antenna applications, but could potentially be used for frequency agility, polarization agility, or beam steering. Another area which almost certainly will lead to a mutually beneficial result is the application of high-speed analog electro-optic links to microstrip phased array antennas. Electro-optic technology with fiber optic lines can be used to distribute RF and/or control signals to radiating modules in the array, thus simplifying the layout and construction of large array antennas.

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