# Dual-Frequency Patch Antennas

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### 1. Abstract

Dual-frequency patch antennas may provide an alternative to large-bandwidth planar antennas, in applications in which large bandwidth is really needed for operating at two separate transmit-receive bands. When the two operating frequencies are far apart, a dual-frequency patch structure can be conceived to avoid the use of separate antennas. In this paper, a critical overview of possible solutions for dual-frequency patch antennas is presented, and future perspectives are outlined.

### 2. Introduction

Patch antennas are popular for their well-known attractive features, such as a low profile, light weight, and compatibility with monolithic microwave integrated circuits (MMICs). Their main disadvantage is an intrinsic limitation in bandwidth, which is due to the resonant nature of the patch structure. On the other hand, modern communication systems, such as those for satellite links (GPS, vehicular, etc.), as well as emerging applications, such as wireless local networks (WLAN), often require antennas with compactness and low-cost, thus rendering planar technology useful, and sometimes unavoidable. Furthermore, thanks to their lightness, patch antennas are well suitable for systems to be mounted on airborne platforms, like synthetic-aperture radar (SAR) and scatterometers. From these applications, a new motivation is given for research on innovative solutions that overcome the bandwidth limitations of patch antennas. In applications in which the increased bandwidth is needed for operating at two separate sub-bands, a valid alternative to the broadening of total bandwidth is represented by dual-frequency patch antennas. Indeed, the optimal antenna for a specific application is one that ensures the matching of the bandwidth of the transmitted and/or the received signal. Dual-frequency antennas exhibit a dual-resonant behavior in a single radiating structure. Despite the convenience that they may provide in terms of space and cost, little attention has been given to dual-frequency patch antennas. This is probably due to the relative complexity of the feeding network which is required, in particular for array applications.

The need to operate at dual-frequency can arise in vehicular-satellite communication systems where low-cost antennas with an almost isotropic pattern over the upper hemisphere are required; this matches well the characteristics of patch antennas.

When the system requires operation at two frequencies too far apart, dual-frequency patch antennas may avoid the use of two different antennas; a typical case is that of SAR. As is well-known, the present SAR antennas employ different arrays for each band. The trend of SAR antennas of the future generation is to cover at least two of the three bands with a dual-frequency antenna. This would reduce weight and surface, thus improving the possibilities of accommodation under the launcher fairing. A dual-frequency patch antenna for SAR is very complicated to conceive, and represents an example in which all the possible critical issues involved in dual-frequency antennas coexist.

The first critical point concerns the design of the active part of the basic transmit-receive (T-R) module. Thc best solution would be to realize the same MMIC for the two bands, but this is often not practical, due to the large separation between the two frequencies that require different microwave components for each frequency (particularly for the receiving channel). The second issue is concerned with the design of a single feed network for the two frequencies. This is probably the most critical problem to solve, considering that dual-linear polarization is often required. This is strictly related to the architecture of the radiating part, which not only must ensure sufficient physical space for printing the microstrip feed lines, but also good isolation between the two frequencies, as well as between the two polarizations.

The two points mentioned above constitute open and challenging problems, and their discussion is beyond the purpose of this paper. Our attention will be focused on the radiating structures of dual-frequency patch antennas. In particular, in the next section, some solutions presented in the literature are reviewed. Next, some single-layer configurations that have been recently developed are presented. Finally, conclusions and future improvements will be discussed.

### 3. Dual-frequency techniques for patch antennas

In principle, dual-frequency planar antennas should operate with similar features, both in terms of radiation and impedance matching, at two separate frequencies. Obtaining these features by using planar technologies is not a straightforward matter, particularly when the intrinsic structural and technological simplicity typical of patch antennas is to be preserved.

As is well-known, a simple rectangular patch can be regarded as a cavity with magnetic walls on the radiating edges. The first three modes with the same polarization can be indicated by $TM_{100}$, $TM_{200}$, and $TM_{300}$, where TM denotes the magnetic field transverse with respect to the interface normal. $TM_{100}$ is the mode typically used in practical applications; $TM_{200}$ and $TM_{300}$ are associated with a frequency approximately twice and triple of that of the $TM_{100}$ mode. This provides, in principle, the possibility to operate at multiple frequencies. In practice, the $TM_{200}$ and the $TM_{300}$ modes cannot be used. Indeed, owing to the behavior of the radiating currents, the $TM_{200}$ pattern has a broadside null, and the $TM_{300}$ pattern has grating lobes.

The simplest way to operate at dual frequencies is to use the first resonance of the two orthogonal dimensions of the rectangular patch, i.e., the $TM_{100}$ and the $TM_{010}$ modes. In this case, the frequency ratio is approximately equal to the ratio between the two orthogonal sides of the patch. The obvious limitation of this approach is that the two different frequencies excite two orthogonal polarizations. Anyway, this simple method is very useful in low-cost short-range applications, where polarization requirements are not pressing.

The above approach characterizes a first category of dual-frequency patch antennas, which will be identified as 1) *orthogonal-mode dual-frequency patch antennas*. This category can be extended to any kind of patch shape that offers two cross-polarized resonant modes. Most of the other dual-frequency patch antennas found in the literature can be subdivided into 2) *multi-patch dual-frequency antennas*, and 3) *reactively-loaded dual-frequency patch antennas*. In the following, a brief overview of these three types of dual-frequency antennas is presented. For the sake of convenience, a summary is included in Figure 1.

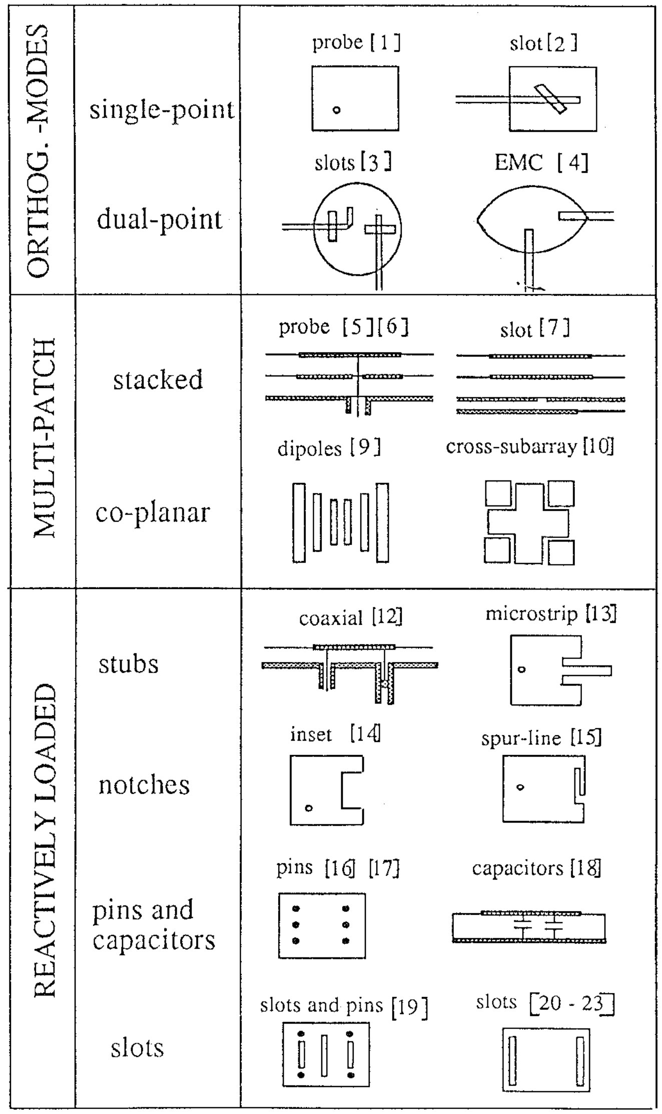


**Figure 1. Dual-frequency patch antennas.**

### 3.1 Orthogonal-mode dual-frequency patch antennas

As mentioned before, these antennas are characterized by two resonances with orthogonal polarizations. These may be obtained, in the simplest case, by a rectangular patch [1, 2]. An interesting feature of these antennas is their capability of simultaneous matching of the input impedance at the two frequencies with a single feed structure (denoted by "single-point" in Figure 1). This may be obtained with a probe-fed configuration, which is displaced from the two principal axes of the patch. As demonstrated in [1], the performance of this approach in terms of matching level and bandwidth is almost equal to that of the same patch fed separately on the two orthogonal principal axes. This provides the possibility of using the well-known design formula for standard feeds. It is also worth noting that the simultaneous matching level for structures that provide the same polarizations at the two frequencies is, in general, worse with respect to the case relevant to orthogonal polarization.

As suggested in [2], single-feed dual matching may be obtained by using slot coupling, in which the slot is inclined with respect to the microstrip feed line. The required slot length and inclination angle can be approximately obtained by projecting the slot onto the two orthogonal directions. The two projections can be thought of as the length of two equivalent slots that excite the patch at the two separate polarizations. The inclination of the slots may also be adjusted, in order to compensate for error introduced by the matching stub, which is designed to be a quarter of a wavelength for only one frequency.

Orthogonal modes may be excited by separated microstrips (see "dual-point feed" in Figure 1). In [3], a circular patch is used, in which two modes in the circular cavity are excited by two orthogonal slots. An isolation of 35 dB between the two ports can be obtained. This solution cannot provide flexibility in designing the frequency ratio. A different shape is suggested in [4], where the patch rim is composed of two intersecting portions of circles of the same radius, with their center displaced by a distance that is designed to have a given frequency ratio. Good isolation between the orthogonal ports can also be obtained with electromagnetically coupled microstrip feed lines (27 dB).

### 3.2 Multi-patch dual-frequency antennas

In these structures, the dual-frequency behavior is obtained by means of multiple radiating elements, each of them supporting strong currents and radiation at the resonance. This category includes multi-layer stacked patches (see "stacked" in Figure 1) that can use circular [5], annular [6], rectangular [7], and triangular [8] patches. These antennas operate with the same polarization at the two frequencies, as well as with a dual polarization. The same multilayer structures can also be used to broaden the bandwidth of a single-frequency antenna, when the two frequencies are forced to be closely spaced. In this latter case, the lower patch can be fed by a conventional arrangement, and the upper patch by proximity coupling with the lower patch [7]. In order to avoid disappearance of the upper resonance, the sizes of the two patches should be close, so that only a frequency ratio close to unity may be obtained. A direct probe feed for the upper patch may also be used [5, 6]. In this case, the probe passes through a clearance hole in the lower patch, and is electrically connected to the upper patch. This kind of configuration insures one more degree of freedom (the hole radius) in designing the optimum matching at the two frequencies, and allows a wider range of the frequency ratio with respect to the structure in which the upper patch is electromagnetically coupled. In comparison with the resonant frequencies of the two isolated patches, the frequency of the upper (smaller) patch increases, and the frequency of the lower (larger) patch decreases. In any case, due to the strong coupling between the two elements, simple design formulas cannot be found, so that a full-wave analysis is, in general, required in the first phase of the design.

Multi-frequency antennas can also be obtained by printing more resonators on the same substrate (see "co-planar" in Figure 1). Croq and Pozar [9] suggested parallel rectangular dipoles, coupled by a slot, that present a triple resonance, with frequency ratios of approximately 1.35. Radiation patterns have been shown to be consistent at all operating frequencies, and the antenna is attractive for its simplicity.

All of the multi-resonator antennas discussed above allow only a limited value of the frequency ratio, so that they are suitable for short-link transmit-receive modules or vehicular-satellite communications. Radar applications, such as SAR and multi-spectral scatterometers, often require a large separation between the frequencies, so that the multi-resonator structure must involve patches of very different sizes. A simple example of this concept is that presented in [10]. This consists of a cross-shaped patch for the lower frequency, and a subarray of four patches for the upper frequency. This structure will be discussed in more detail in the next section.

For a large separation between the two frequencies, the criteria of a drastic separation of the patch antenna associated with each frequency may be adopted. For example, the patches and the relevant feed networks for the two frequencies can be stacked on two different substrates, thus obtaining two almost-independent antennas. The structure presented in [11] is an array of dual-frequency modules with two superimposed two-layer structures, one for each frequency. The lower structure is a slot-coupled patch, resonating at 0.9 GHz; the upper structure is a four-patch sub-array, operating at 2.5 GHz. This sub-array is slot-fed by a Wilkinson divider network that is located between the finite slotted ground plane of the sub-array and the low-frequency patch. This arrangement has the advantage that the two sub-antennas can be designed almost independently, provided that orthogonal polarizations at the two frequencies are imposed. In this configuration, one should expect parallel-plate mode excitation between the slotted ground plane and the lower patch. This can produce spurious radiation and coupling between the array transmit-receive modules.

### 3.3 Reactively-loaded patch antennas

The most popular technique for obtaining a dual-frequency behavior is to introduce a reactive loading to a single patch. The simplest way is to connect a stub to one radiating edge, in such a way as to introduce a further resonant length that is responsible for the second operating frequency. This may easily be understood by resorting to the transmission-line model. As is shown in Figure 1, other kinds of loading can be used, including notches, pins and capacitors, and slots.

The reactive-loading approach was first used in [12], where an adjustable coaxial stub was employed. This structure may provide both tuning and design of the frequency ratio in a simple manner; on the other hand, it is encumbering and not well-suited for high frequencies. In [13], a more practical configuration is presented, in which the stub is constituted by a microstrip.

Loading the radiating edge with an inset [14] or a spur-line [15] ("notch loading") is an alternative way to introduce a dual-frequency behavior that creates the same effect as the microstrip-loading effect, with the advantage of reduced size. However, both with stubs and notches, the frequency ratio cannot be designed to be higher than 1.2 without introducing strong cross-polarization levels or pattern distortion at the additional frequency.

To obtain higher values of the frequency ratio, different approaches have been proposed. In particular, one can modify the resonant frequency of the $TM_{100}$ and/or of the $TM_{300}$ mode, by using shorting vias or lumped capacitors between the patch and the ground plane. As shown in [16], by locating the shorting pin where the current distribution of the $TM_{300}$ mode exhibits a minimum, a strong perturbation of its resonant frequency is obtained, while the $TM_{100}$ mode remains almost unperturbed. This permits a frequency ratio of from 2 to 3, by increasing the number of vias. The disadvantage is that the radiation pattern of the perturbed $TM_{300}$ mode is affected by spurious lobes. An exhaustive investigation of the use of shorting vias for changing the resonant frequency and the polarization is carried out in [17]. There, the use of pin diodes is proposed for changing the loading configuration, thus allowing frequency agility. Very high values of the frequency ratio (4-5) can be obtained by means of two lumped capacitors, connected from the patch to the ground plane [18].

Another kind of reactive loading can be introduced by etching slots on the patch. The slot loading allows for a strong modification of the resonant mode of a rectangular patch, particularly when the slots are oriented to cut the current lines of the unperturbed mode. In particular, as shown in [19], the simultaneous use of slots and short-circuit vias allows a frequency ratio of from 1.3 to 3, depending on the number of vias. Other kinds of slot-loaded patches have been independently introduced in [20] and [21], and consist of a rectangular patch with two narrow slots etched close to and parallel to the radiating edge. The same configuration has been investigated in [22], and extended to dual polarization in [23]. More details are given in the next section.

Before proceeding further, it is worth noting that some original dual-frequency structures, such as those proposed in [24] and [25], cannot be easily framed into the categories that we have mentioned.

## 4. Some geometries for single and dual linear polarization

Examples of dual-frequency patch antennas are shown in Figures 2-4 [20-23]. These belong to the category of the reactively-loaded patch antennas, in which the loading is obtained by using slots on the patches. The first configuration is used for linear polarization, while the second and the third are suitable, with different features, for dual-linear polarization. For all of these antennas, the frequency ratio may be easily adjusted from 1.6 to 1.9. For a frequency ratio of from 2 to 3.5, the multi-resonator structure shown in Figure 5 can be successfully used. In the following, the above antennas are discussed.

### 4.1 Slotted rectangular-patch antenna

The basic geometry is a slotted rectangular-patch antenna, in which two narrow slots, with dimensions $L_s$ and $d$, are etched on the patch close to and parallel to the radiating edges. The location of the slots with respect to the patch is defined by the quantities $w$ and $l$, which are very small with respect to the dimensions $L$ and $W$ of the patch. The antenna may be fed with either an aperture [20-21] or a probe feed [22].

The dual-frequency operation in the slotted structure can be interpreted as that associated with two modes that arise from the

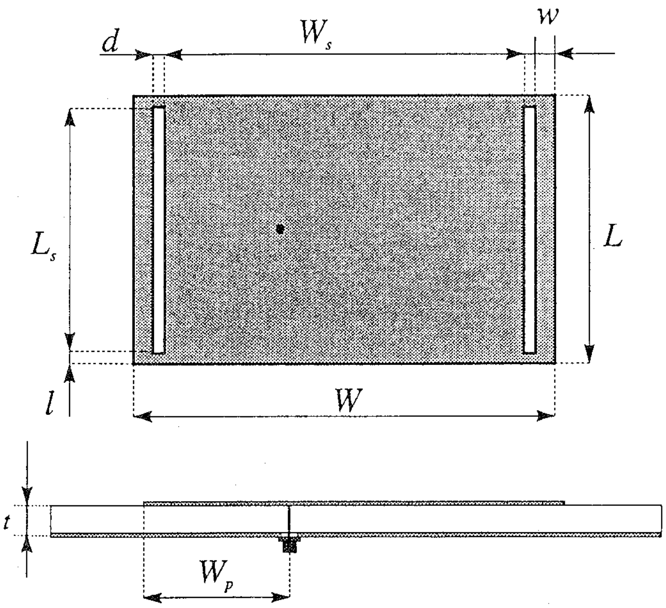


**Figure 2. The geometry of the slotted rectangular-patch antenna.**

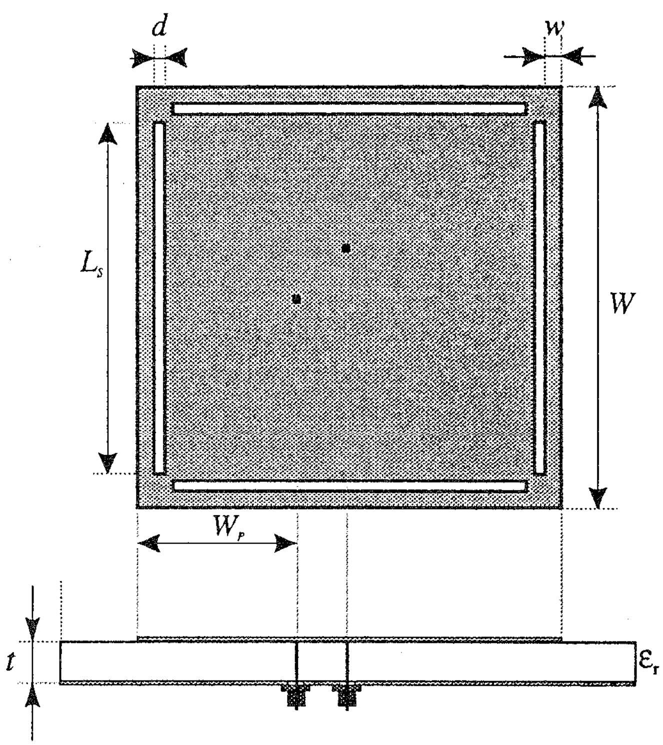


**Figure 3. The geometry of the slotted square-patch antenna.**

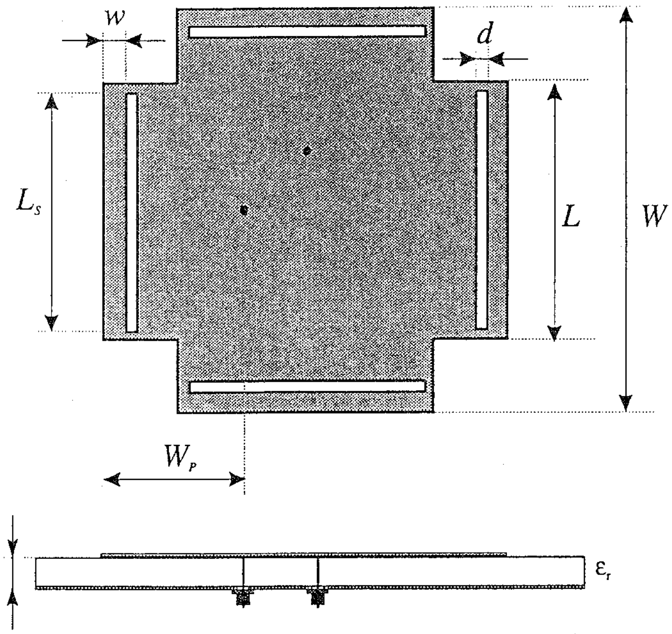


**Figure 4. The geometry of the slotted cross-patch antenna.**

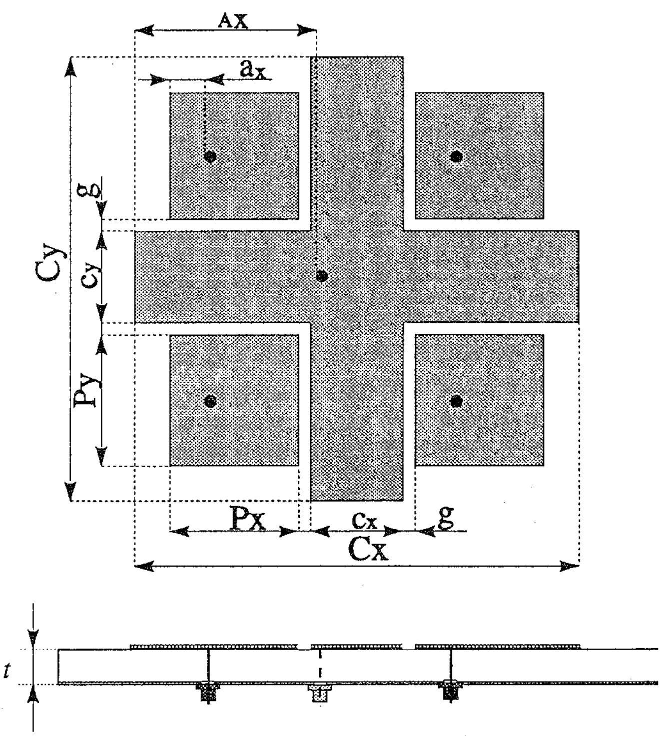


**Figure 5. The geometry of the cross-subarray patch antenna.**

perturbation of the $TM_{100}$ and of the $TM_{300}$ mode. In particular, since the narrow slots are etched close to the radiating edges, they are located close to the current minima, so that minor perturbations of the $TM_{100}$ mode are expected. The radiative mechanism associated with this first mode is essentially the same as that of a patch without slots. As a consequence, its resonant frequency is only slightly different from that of a standard patch.

On the other hand, the current distribution of the $TM_{300}$ mode is strongly modified, since the slots are located where the currents of the unperturbed $TM_{300}$ mode should be significant. The currents circulate around the slots, and find a resonant condition with nulls close to the two edges of each slot. This condition forces the central section of the current distribution to be broader than that corresponding to the unperturbed $TM_{300}$ mode, thus becoming similar to that of the $TM_{100}$ mode. The resonant frequency decreases, due to the increase in the current-line path length.

Regarding the radiation pattern, in [22] it is shown that when the slots are very short, the radiation pattern exhibits three lobes, as does that of the unperturbed $TM_{300}$ mode. As the slot length increases, the central lobe of the pattern first reduces, and then disappears. A condition may be found in which the radiation pattern exhibits a null at broadside. When the slot length ($L_s$) further increases, the field contribution from the currents located in the central part of the patch becomes more significant, and this fills up the pattern in the broadside direction. For $L_s$ comparable to $W$, the radiation pattern is very similar to that of the $TM_{100}$ mode of a standard patch, so that the two modes exhibit radiation properties very similar to each other, as desirable for a dual-frequency antenna. This consideration has to be accounted for when designing the frequency ratio of the antenna. Indeed, $L_s$ is the most important parameter for managing the upper frequency, so that the regularity of the pattern at the upper frequency imposes constraints on the frequency-ratio range.

Let us denote by $f_{100}$ and $f_{300}$ the resonant frequencies associated with the modified $TM_{100}$ and $TM_{300}$ modes, respectively. In order to design the two frequencies, simple semi-empirical formulas, based on physical models, have been found very useful. Since the first resonance is not much affected by slot loading, the first frequency can be predicted by slightly modifying the well-established formula for rectangular unslotted patches, i.e.,

$$f_{100} = \frac{c}{2(W+\Delta W)\sqrt{\varepsilon_e(L/t,\varepsilon_r)}}G, \qquad (1)$$

where $c$ is the free-space speed of light,

$$\varepsilon_e(x,y) = \frac{y+1}{2} + \frac{y-1}{2}\left[1+\frac{10}{x}\right]^{-1/2}, \qquad (2)$$

$$\Delta W = \frac{t}{\pi}\frac{L/t+0.336}{L/t+0.556}\left\{0.28+\frac{\varepsilon_r+1}{\varepsilon_r}\left[0.274+\ln\left(\frac{L}{t}+2.518\right)\right]\right\}, \qquad (3)$$

and

$$G = \left[1+\frac{\left(1.5\frac{w}{W}-0.4\frac{l}{L}\right)}{1+\Delta W/W}\right]^{-1} \qquad (4)$$

is a correction factor that accounts for the slot-loading effect. The parameters that occur in the above formulas are defined in Figure 2.

The upper resonant frequency can be predicted according to a simple transmission-line model, which is derived by observing the behavior of the current distribution at the upper resonance. Since the currents find a resonant condition by circulating around the slots, the narrow portion of the conductor that encircles the slot behaves like two half-wavelength open-circuit stubs. The effective dielectric constant of these equivalent stubs can be determined by assuming a $w$-width microstrip-line model to describe the current distribution around the slot. This allows predicting the second resonant frequency according to

$$f_{300} = \frac{c}{2(L-2l+d)\sqrt{\varepsilon_e(w/t,\varepsilon_r)}}. \qquad (5)$$

The ranges of validity and the accuracy of Equations (1) and (5), with respect to results from a full-wave analysis, are given in [22]. From these formulas, the frequency ratio can be well controlled in a range from 1.6 to 1.9, by changing the patch dimensions and the slot lengths. In order to extend this range, two resonating microstrip stubs can be printed on a back substrate, and connected at a point between the slots and the radiating edges of the patch by two vias through the ground plane [22]. This arrangement permits a frequency ratio ranging from 1.2 to 3. The capability of matching the input impedance at two frequencies with a single-feed-point probe-feed configuration is also analyzed in [22]. By a proper choice of the feed point, it is possible to obtain 16 dB for the input reflection coefficient at both frequencies.

### 4.2 Slotted square-patch antenna

Dual-polarization behavior can be obtained by using two pairs of orthogonal slots on a square patch. This leads to the slotted square-patch antenna depicted in Figure 3. The physical principle of operation is the same as that for linear polarization. Indeed, for both $TM_{100}$ and $TM_{300}$ modes, the cross-polarized slots do not perturb the actual current distribution. This also provides a good decoupling between the two orthogonal modes $TM_{n00}$ and $TM_{0n0}$ ($n = 1, 3$). A decoupling between the two ports of -35 dB and -40 dB was found at the lower and the upper frequency, respectively.

The lower frequency can be predicted according to Equation (1), in which the factor $G$ assumes the different expression [23]

$$G = 1.13 - 0.19\frac{L_s}{W} - 0.73\frac{w}{W}. \qquad (6)$$

Regarding the upper frequency, the same model as described for the slotted rectangular-patch antenna can be applied, thus leading to Equation (5). However, in this case, the accuracy is worse than that for the slotted rectangular-patch antenna, due to the fact that the model does not account for coupling between the parallel slots. This deficiency, which is not significant for the slotted rectangular-patch antenna, assumes here more importance, due to the square geometry. Furthermore, the slotted square-patch antenna exhibits a smaller range of frequency ratio than the slotted rectangular-patch antenna.

### 4.3 Slotted cross-patch antenna

Removing the portion of the metallization close to the slot ends gives rise to the slotted cross-patch antenna, shown in Figure 4. In this configuration, both the slot length, $L_s$, and the dimension $L$ can be reduced simultaneously, thus eliminating the currents that are the main cause of pattern distortion. This also allows a more extended range of the frequency ratio, owing to the additional geometrical parameter. The distance between the slots and the patch edges plays an important role in the antenna properties. This parameter should be kept small to avoid perturbations of the $TM_{100}$ mode.

The lower frequency can be predicted from Equation (1), in which the factor $G$ now assumes the expression [23]

$$G = 1.31 - 0.42\frac{L_s}{W} - 1.3\frac{w}{W}. \tag{7}$$

The upper frequency, $f_{300}$, can be predicted by employing the same model as for the slotted rectangular-patch antenna and the slotted cross-patch antenna. Using the pertinent parameters yields

$$f_{300} = \frac{c}{2(L+d)\sqrt{\varepsilon_e(w/t, \varepsilon_r)}}. \tag{8}$$

In contrast to the slotted square-patch antenna, this expression has been found to be very accurate, due to the coupling reduction between perpendicular slots. In Equation (8), $L - L_s$ is assumed to be very small [23].

Another significant advantage of the slotted cross-patch antenna, with respect to the slotted square-patch antenna, is the improved decoupling of the two orthogonal ports at the lower frequency: it is 5-6 dB better. The slotted cross-patch antenna is also preferable in terms of simultaneous matching by using a single feed-point in the probe-fed configuration [23]. On the other hand, the slotted square-patch antenna occupies less space compared to the slotted cross-patch antenna, the resonant frequencies being equal. Furthermore, it has a slightly better behavior in terms of cross-polar components, since the cross geometry provides further transverse current lines on the patch.

### 4.4 Cross-subarray dual-frequency patch antenna

The cross-subarray patch antenna, shown in Figure 5, allows large values of the frequency ratio [10]. It consists of a cross patch working at the lower frequency, and four square patches operating as a subarray at the upper frequency. The layout is designed to obtain a compact and modular structure, and is well suited for use in dual-frequency arrays. The geometry has two symmetry planes, to provide radiation in dual-linear polarization, as required in SAR antennas. The cross-patch resonance is only slightly perturbed by the four square patches, providing that the radiating edges of the cross patch are sufficiently spaced from the four square patches. Thus, the design of the relevant frequency can be performed by using well-known CAD formulas. On the contrary, the resonance of the square patches may be affected by the presence of the cross patch, which causes a reactive loading of the square patch. Therefore, the upper resonance can be slightly lower than that predicted for four isolated patches. The decrease of this upper frequency is noticeable when the spacing ($g$ in Figure 5) is lower than the substrate thickness. However, even in this case, the use of a simple transmission-line model, with a proper edge-capacitance loading, can provide an accurate prediction of the upper frequency. Eventually, the geometry provides the possibility of a flexible and simple final design. In designing the antenna, one should choose carefully the distances between the square patches, which have to be less than $0.70\lambda$ to avoid scan blindness at the upper frequency. This requirement is significant when the module is placed into an array architecture. In this case, care must also be taken in designing the distances between the cross patches. A possible solution for an array configuration for a frequency ratio equal to 3 (typical for a S-X band SAR) may be obtained by spacing the cross patches at 0.45 wavelength at the lower frequency, and the square patches at 0.7 wavelength at the upper frequency. This choice may provide a compromise between the reduction of the coupling among the cross patches, and the possible scan blindness occurrence at the upper band. However, practical considerations of the coupling and of the possibility to locate the feed network would suggest a larger spacing among cross patches. This would require a basic periodic cell, made of one cross patch and nine square patches.

The disadvantage of the above configuration for the SAR application is its intrinsic limitation of bandwidth (especially at the lower frequency), which derives from the use of a single substrate. A more convenient structure, from this point of view, consists of printing the cross patches on a superstrate with a very low dielectric constant, located on the square-patches' substrate. This structure, and the relevant feed network inside a basic periodic cell, with an element of one cross patch by nine square patches, is presently under investigation.

## 5. Concluding remarks

An overview of dual-frequency patch antennas has been carried out, with special emphasis on configurations that are particularly attractive for their simplicity and design flexibility. Attention has been focused on the geometry of the radiators, avoiding the important problem of the dual-frequency feed network. The work that is discussed in the literature on this latter subject is very insufficient, and further research is needed. The use of a single feed network for both frequencies may be practical only when the two frequencies are very close to each other. For large separations between the frequencies (a frequency ratio of 2.5 to 3), two different microstrip networks have to be designed. This must be done taking into account the coupling between them; they are forced to coexist in a restricted space, especially when dual polarization is required.

The current trend toward high values of frequency ratio involves separating the structures relevant to the two frequencies on different dielectric layers, in such a way as to obtain two antennas that are, as much as possible, independent. In this case, two options should be investigated. The first is to print the smaller (upper-frequency) patches on a dielectric layer above the larger patches. This could have the advantage of limiting the blockage of the low-frequency radiation. On the other hand, in this condition the larger patches work on a substrate presumably thin in terms of a wavelength, so that their bandwidth could be too narrow. The second possibility is to print the larger patches on the upper layer, to broaden their bandwidth. Although this latter solution has the impairment of causing blockage of the upper-frequency radiation, and possible excitation of cross-polar components, in our opinion it seems to be more interesting, provided the lower-frequency patches can be shaped properly.

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## 7. References

1. J-S. Chen, K-L. Wong, "A Single-Layer Dual-Frequency Rectangular Microstrip Patch Antenna Using a Single Probe Feed," *Microwave and Optical Technology Letters*, **11**, 2, 1996, pp. 38-84.

2. Y. M. M. Antar, A. I., Ittipiboon, A. K. Bhattachatyya, "A Dual-Frequency Antenna Using a Single Patch and An Inclined Slot," *Microwave and Optical Technology Letters*, **8**, 6, 1995, pp. 309-310.

3. Y. Murakami, W. Chujo, I. Chiba, M. Frujise, "Dual Slot Coupled Microstrip Antenna for Dual Frequency Operation," *Electronics Letters*, **29**, 22, 28 October 1993, pp. 1906-1907.

4. M. Deepukumar, J. George, C. K. Aanandan, P. Mohanan, K. G. Nair, "Broadband Dual Frequency Microstrip Antenna," *Electronics Letters*, **32**, 17, 15 August 1996, pp. 1531-1532.

5. S. A. Long, M. D. Walton, "A Dual-Frequency Stacked Circular-Disc Antenna," *IEEE Transactions on Antennas and Propagation*, **AP-27**, 3, March 1979, pp. 1281-1285.

6. J. S. Dahele, K. F. Lee and D. P. Wong "Dual Frequency Stacked Annular-Ring Microstrip Antenna," *IEEE Transactions on Antennas and Propagation*, **AP-35**, 11, November 1987, pp.1281-1285.

7. J. Wang, R. Fralich, C. Wu and J. Litva, "Multifunctional Aperture Coupled Stack Antenna." *Electronics Letters*, **26**, 25, December 1990, pp. 2067-2068.

8. D. Mirshekar-Syankal, H. R. Hassani, "Characteristics of Stacked Rectangular and Triangular Patch Antennas for Dual Band Application," IEE 8th International Conference on Antennas and Propagation, Edinburgh, March 1993.

9. F. Croq, D. Pozar, "Multifrequency Operation of Microstrip Antennas Using Aperture Coupled Parallel Resonators," *IEEE Transactions on Antennas and Propagation*, **AP-40**, 11, November 1992, pp. 1367-1374.

10. C. Salvador, L. Borselli, A. Falciani, S. Maci, "A Dual Frequency Planar Antenna at S and X Bands," *Electronics Letters*, **31**, 20, October 1995, pp. 1706-1707.

11. T. Chan, Y. Hwang, "A Dual-Band Microstrip Array Antenna," IEEE International Symposium on Antennas and Propagation *Digest*, Newport Beach, California, June 18-23, 1995, pp. 2132-2135.

12. W. F. Richards, S. E. Davidson, S. A. Long, "Dual-Band Reactively Loaded Microstrip Antenna," *IEEE Transactions on Antennas and Propagation*, **AP-33**, 5, May 1985, pp. 556-560.

13. S. E. Davidson, S. A. Long, W. F. Richards, "Dual-Band Microstrip Antenna with Monolithic Reactive Loading," *Electronics Letters*, **21** , 21, 1985, pp. 936-937.

14. H. Nakano, K. Vichien "Dual-Frequency Patch Antenna with a Rectangular Notch," *Electronics Letters*, **25**, 16, 1989, pp. 1067-1068.

15. D. Sanchez-Hernandez and I. D. Robertson "Analysis and Design of a Dual-Band Circularly Polarized Microstrip Patch Antenna," *IEEE Transactions on Antennas and Propagation*, **AP-43**, 2, February 1995, pp. 201-205.

16. S. S. Zhong and Y. T. Lo, "Single Element Rectangular Microstrip Antenna for Dual-Frequency Operation," *Electronics Letters*, **19**, 8, 1983, pp. 298-300.

17. D. H. Schaubert, F. G. Ferrar, A. Sindoris, S. T. Hayes, "microstrip Antennas with Frequency Agility and Polarization Diversity," *IEEE Transactions on Antennas and Propagation*, **AP-29**, 1, January 1981, pp. 118-123.

18. R. B. Waterhouse, N. V. Shuley, "Dual Frequency Microstrip Rectangular Patches," *Electronics Letters*, **28**, 7, 1992, pp. 606-607.

19. B. F. Wang and Y. T. Lo, "Microstrip Antenna for Dual-Frequency Operation," *IEEE Transactions on Antennas and Propagation*, **AP-32**, 9, September 1984, pp. 938-943.

20. S. Maci, G. Biffi Gentili, G. Avitabile, "Single-Layer Dual-Frequency Patch Antenna," *Electronics Letters*, **29**, 16, August 1993.

21. M. L. Yazidi, M. Himdi and J. P. Daniel, "Aperture Coupled Microstrip Antenna for Dual Frequency Operation," *Electronics Letters*, **29**, 17, August 1993.

22. S. Maci, G. Biffi Gentili, P. Piazzesi, C. Salvador "A Dual Band Slot-Loaded Patch Antenna," *IEE Proceedings H*, **142**, 3, March 1995, pp. 225-232.

23. P. Piazzesi, S. Maci, G. Biffi Gentili, "Dual-Band Dual-Polarized Patch Antennas," *Int. MiMiCAE Jour.*, **5**, 6, December 1995, pp. 375-384.

24. C. S. Lee, and V. Nalbandian, "Impedance Matching of a Dual-Frequency Microstrip Antenna with an Air Gap," *IEEE Transactions on Antennas and Propagation*, **AP-41**, 5, May 1993, pp. 680-682.

25. A. Boag, Y. Shimony, A. Boag, and R. Mittra "Dual-Band Cavity Backed Quarter Wave Patch Antenna," IEEE International Symposium on Antennas and Propagation *Digest*, Newport Beach, California, June 18-23, 1995, pp. 2124-2127.

## Introducing Feature Article Authors

**Stefano Maci** was born in Rome in 1961. In 1987, he obtained his doctoral degree in Electronic Engineering from the University of Florence. In 1990, he joined the Department of Electronic Engineering of the University of Florence as Assistant Professor. Since 1993, he has also been an Adjunct Professor at the University of Siena. In 1988, he won the national Young Scientists "Francini" award for the Laurea thesis, and in 1996 he was awarded the "Barzilai" prize for the best paper at the National Italian Congress of Electromagnetism (XI RiNEm). In 1997, he was an invited Professor at the ElectroMagnetic Institute of the Technical University of Denmark. His interests are focused on electromagnetic theory, mainly concerned with high- and low-frequency methods for antennas and electromagnetic scattering. He has also developed research activities on specific topics concerned with microwave antennas, particularly focused on the analysis, synthesis, and design of patch antennas.

**Guido Biffi Gentili** was born in Lucca in 1943. He received his doctoral degree in electronic engineering from the University of Pisa in 1970. From 1970 to 1976 he was with the Department of Electrical Engineering, University of Pisa. In 1977, he joined the Department of Electronic Engineering of the University of Florence, where he taught courses on radar theory and techniques. He is an Associate Professor of Electromagnetic Theory and Techniques. His research interests are in the areas of microwaves, remote sensing, CAD modeling, and biological applications of EM waves.

**Editor's Comments** *Continued from page 6*

was going to be used for direction finding. Neural-network techniques are now more common. Although there are some fundamental differences, there is also a very interesting analogy between the two approaches. Bob Mailloux and Hugh Southall explain that analogy in their article. The significance is that they use this analogy as the basis for a tutorial on how a neural network can be used to control an array, particularly for direction finding. The article is of interest for at least two reasons. First, it's a good tutorial and introduction to neural networks for phased arrays. Second, you will find the analogy itself interesting, even if you are familiar with phased arrays. I think you'll like this article.

The finite-difference time-domain technique is one of the most commonly used numerical techniques in our area of engineering. One of the major issues affecting computational speed and accuracy is how to terminate the boundaries of the computational domain for exterior, or open-region problems. Some type of absorbing boundary condition has been the usual answer. However, Omar Ramahi has developed an alternative, called the Complementary Operators Method. This synthesizes a set of operators for the FDTD problem that have the property that first-order reflections are canceled at the boundaries of the computational domain, independently of the wave number. This has some important potential advantages, and these are demonstrated in several examples in the article. In reading this article, be aware of one point. As developed in this article, the Complementary Operators Method requires running the FDTD problem twice. That does not imply a doubling of the computational time, because the size of the computational domain is made smaller, and it also generally results in greater accuracy. However, an extension of the method, called the Concurrent Complementary Operators Method, has just been developed. It was presented by the author at the Montreal Symposium, and can be found in the references. This extension removes the need to run the computation twice. In an appendix, the author gives the detailed equations necessary to actually implement the basic method. If you have an interest in computational electromagnetics, I think you will enjoy this article. Even if your interest is primarily as a user of such results, you will find the discussion of absorbing boundary conditions and the examples illuminating.

**Other "presents" in this issue.** High efficiency and small reflectors are two sets of parameter constraints that are not usually satisfied together in dual-reflector antenna designs. However, Alex Popov has come up with a design approach that does accomplish this, and he presents it in Tom Milligan's Antenna Designer's Notebook. Tom has also run a number of examples of design tradeoffs using the approach, and these give a good feeling for what is possible with it.

Predicting the radiation characteristics of an antenna in the presence of a dielectric radome is a difficult problem. In John Volakis' EM Programmer's Notebook, James Shifflett provides a demonstration of how Physical Optics methods can be successfully used to make such predictions. The agreement with measurements for both the amplitude and the phase, both on bore-sight and off-axis, is quite impressive.

The results of the AP-S elections for officers and new AdCom members appear in this issue. Also announced are those newly elected to the grade of Fellow of the IEEE. Ahmed Kishk, one of our two Editors, was elected this time, and it is a very well deserved honor. Congratulations!

In addition to the many important calls for papers in this issue, there are several announcements of competitions for support for attending meetings. The Raj Mittra Travel Grants are tied to having a paper accepted at the AP-S/URSI Symposium, and thus the deadline for applying for the grants is the same as for submitting a paper to the Symposium. The AP-S AdCom approved waiving the registration fees for the Symposium for a limited number of people who have received grants for which such a waiver is a requirement. An announcement of this, and the method of applying for such a waiver, appear in this issue. The annual Conference Travel Grant for ex-USSR scientists and engineers is also now available, and announced in this issue.

If you are a US electrical engineer, how would you like to be in a position to advise a Senator or Representative regarding technology issues? You can be. The IEEE-USA's Congressional Fellow program is described in Kathie Virga's PACE column. The deadline for application is near the end of February. There is also an Executive Fellowship program, supplying advisers to the Executive branch of the government.

**It seems like it should only be August.** This year has gone by very quickly. Shortly after (or maybe even, before) you get this issue, 1997 will end, and the *Magazine* (and its predecessor, the *Newsletter*) will enter its fortieth year of publication. For most, the end of the year brings celebrations of joy and faith, and a renewal of hope that is a most fitting prelude to the new year. I hope that you, your family, and your colleagues have a most joyous holiday, and that the New Year is happy and prosperous for you.

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