PCS Antenna Design: The Challenge of Miniaturization

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Abstract

PCS (personal communication system) devices have become an important part of everyday life. The pressure to design small, lightweight, and user-friendly mobile-communication devices has increased accordingly, creating the need for optimal antennas for mobile applications. In this paper, we present some basic rules about electrically small antennas, give clues and guidelines about efficient antenna miniaturization, and, finally, show some examples of miniature antennas developed in our laboratory for practical applications.

Keywords: Antennas; mobile antennas, antenna theory; loaded antennas; Global Positioning System; slot antennas; conformal antennas; SMILA antennas; microstrip antennas; PIFA; land mobile radio cellular systems; land mobile radio; Land mobile radio equipment; electrically small antennas; antenna limitations; antenna measurements; multi-frequency antennas; PCS antennas

1. Introduction

Mobile communications have become an important part of telecommunications. Original applications – such as paging, mobile phones, or GPS – have shown a tremendous growth, and new applications are emerging every day: tagging, wireless computer links, wireless microphones, remote control, wireless multimedia links, satellite mobile phones, wireless Internet; just about everything "goes mobile."

Mobile means practical for the user and easily transportable. The mobile terminals of wireless applications must be light, small, have low energy consumption, and have an appealing design. Technology has evolved very quickly to satisfy these needs in a rapidly growing market: chips are becoming smaller, they consume less current, they are more efficient, and they perform more-complex operations. Thus, the sizes of the electronics needed for a mobile application have decreased drastically during the past few years, whereas their functionality has increased. The antennas, however, have not experienced the same evolution, and marketing staff and designers are pressing engineers to develop smaller antennas.

In opposition to electronic chips, the size of the antenna for a given application is not related mainly to the technology used, but is determined by the laws of physics: the antenna size with respect to the wavelength is the parameter that will have the preponderant influence on the radiation characteristics. This follows from the fact that an antenna is used to transform a guided wave into a radiated wave, and vice-versa, and one understands that to perform this transformation efficiently, the size should be of the order of half a wavelength or larger. Antennas can, of course, be made smaller, but at the expense of bandwidth, gain, and efficiency.

The art of antenna miniaturization is an art of compromise: one has to design the smallest possible antenna that is still suitable for a given application with regard to its radiation characteristics. Or, in other words, one looks for the best compromise among volume, bandwidth, and efficiency. And this best compromise is usually obtained when most of the allotted volume participates in the radiation.

In this paper, we will first state the limitations of antenna miniaturization, by reminding the reader of the well-known laws linking gain, bandwidth, and antenna size. We will then review different ways to make antennas smaller and, when possible, explain the effect of a given miniaturization technique on the antenna-radiation properties. The next section will be devoted to examples of practical, small antennas, designed and realized in our laboratory. These examples include small PCS antennas, dual-frequency antennas, and dual-polarized antennas. The fifth section deals with the measurement of small antennas, and will point to most of the specific problems encountered when measuring electrically small antennas, and will give some clues as to how to proceed correctly. Finally, in the conclusion, we will sum up the important steps of antenna miniaturization, and give some design hints.

2. Physical Limitations on Electrically Small Antennas

As is well known (see, for instance, [1-9]), miniaturizing an antenna will affect its radiation characteristics. Reducing an antenna's size will influence its bandwidth, gain, efficiency, and polarization purity. Moreover, we will see that it is not always easy to feed a small antenna efficiently. This follows directly from the fact that an antenna is a device used to transform a guided wave into a radiated wave, or vice-versa. This transformation process is related to the wavelength, and the antenna size is much more important than the antenna technology in determining how well and for which frequencies this transformation will be satisfactory. Many authors have studied the relationship between the electrical dimension of an antenna and its radiation performances for many years. The main results of these studies are summarized below.

The analysis of electrically small antennas and the study of the effect of size reduction on radiation properties was initiated in the mid-forties. Wheeler [1] used the radiation power factor, defined as the quotient between the antenna resistance and the antenna reactance, to quantify the radiation of an antenna. Using a simple equivalent lumped circuit, he deduced that this quotient was equivalent to the bandwidth multiplied by the efficiency, in those cases where the antenna was matched by a simple circuit. This early paper was a first attempt to confirm mathematically the intuition we have that the product of (efficiency × bandwidth) is directly related to the volume occupied by an antenna. Indeed, as the size of an antenna decreases, its reactance increases, but its radiation resistance decreases.

The work of Wheeler was generalized by Chu [2], who derived the equivalent lumped-circuit element from an expansion in partial fractions of the spherical-wave impedance functions around the antenna. Thus, the relationship between the minimum quality factor, Q, of an omnidirectional antenna and its volume was formally established. He obtained the following approximate formula, valid for linearly polarized antennas:

$$\frac{1+2(ka)^2}{(ka)^3+\left[1+(ka)^2\right]}, \qquad (1)$$

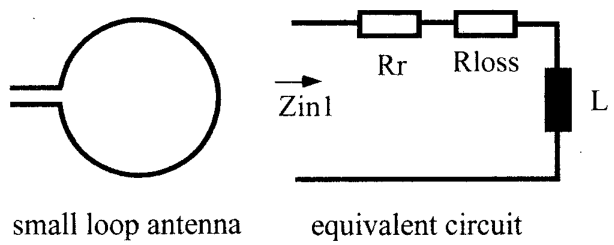


Figure 1a. A loop antenna and its equivalent circuit: $Z_{in1} = R_r + R_{loss} + j\omega L$.

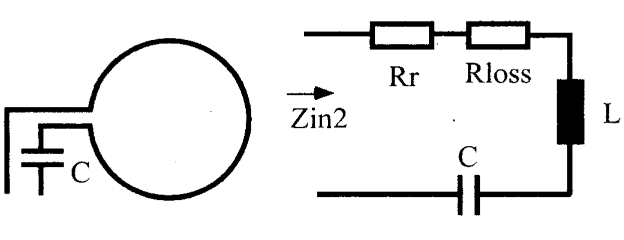


Figure 1b. A loop antenna, matched by loading it with a capacitor, and the equivalent circuit: $Z_{in2} = R_r + R_{loss} + j\frac{\omega^2 LC-1}{\omega C}$.

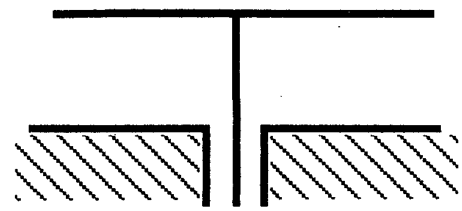


Figure 2a. Examples of loaded antennas: A two-element top-loaded monopole (loaded to reduce its height).

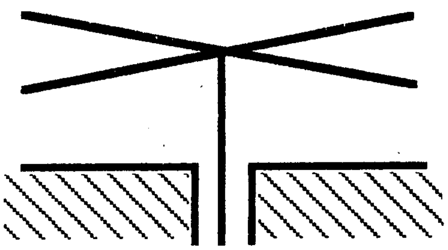


Figure 2b. Examples of loaded antennas: A four-element top-loaded monopole (loaded to reduce its height).

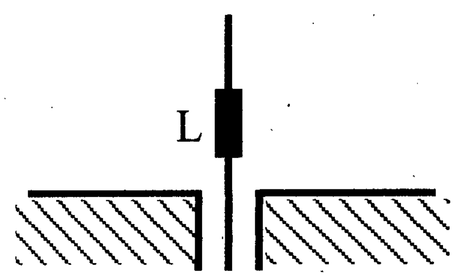


Figure 2c. A coil-loaded antenna (loaded to reduce its height).

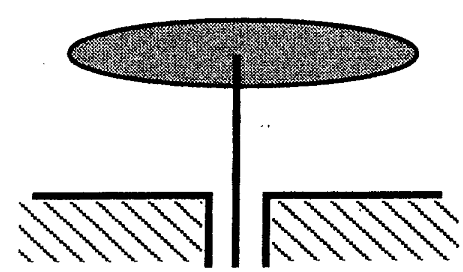


Figure 2d. A capacitor-plate antenna (loaded to reduce its height).

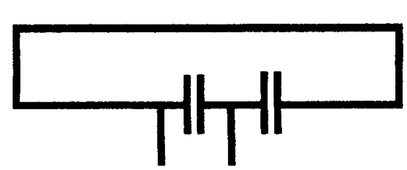


Figure 2e. A loop loaded by capacitances to reduce its size.

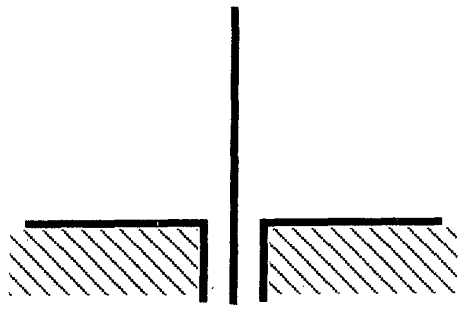


Figure 3a. An example of a material-loaded antenna: $h = \frac{\lambda_0}{4}$.

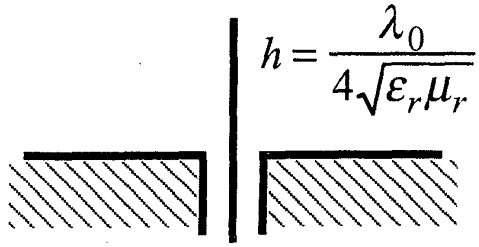


Figure 3b. An example of a material-loaded antenna. $h = \frac{\lambda_0}{4\sqrt{\varepsilon_r \mu_r}}$.

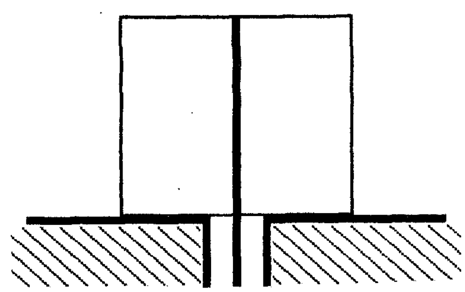


Figure 3c. An example of a material-loaded antenna: $\frac{\lambda_0}{4\sqrt{\varepsilon_r \mu_r}} < h < \frac{\lambda_0}{4}$.

where k is the wave number and a is the radius of the smallest sphere enclosing the antenna. The computation of the quality factor was refined by several other authors [4-6], who derived an exact expression for the smallest possible Q for a linearly polarized antenna:

$$Q = \frac{1}{(ka)^3} + \frac{1}{ka}. \qquad (2)$$

It is interesting to notice that these two expressions give identical results for very small antennas ($ka \ll 1$), such as those that we will mostly discuss in this paper.

The work of Wheeler and Chu, on fundamental limits of small antennas, was extended by Harrington [7] to include the effect of losses. Indeed, a miniaturized antenna will show a higher concentration of surface currents than standard antennas; thus, the ohmic losses will be enhanced. Moreover, Harrington proposed a very useful and simple formula, which gives a practical upper limit for the gain that a small antenna can achieve while still having a reasonable bandwidth:

$$G = (ka)^2 + 2ka. \qquad (3)$$

It is important to remember that Equation (3) gives an upper limit for the gain. The approximation is good for standard antennas, but this limit is more difficult to reach as the antenna becomes very small. Indeed, in those cases the losses may increase drastically, as was already mentioned by Harrington.

Antenna miniaturization affects gain, bandwidth, and efficiency, but can, in many cases, also affect polarization purity [8, 9]. Indeed, we will see in the next section that an efficient way to make antennas smaller is to bend their geometry, in order to force the current to meander, so that the antennas seems electrically larger than it really is. A good example of this concept is the planar inverted-F antenna (PIFA) [8]. The surface current on the strip that

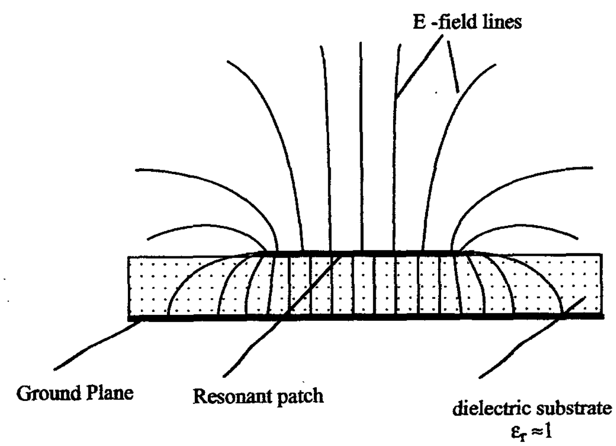


Figure 4a. The effect of the permittivity value on a microstrip-patch antenna: $\varepsilon_r \approx 1$.

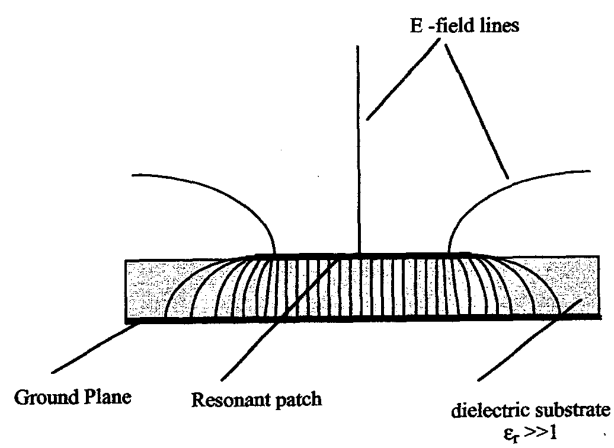


Figure 4b. The effect of the permittivity value on a microstrip-patch antenna: $\varepsilon_r \gg 1$.

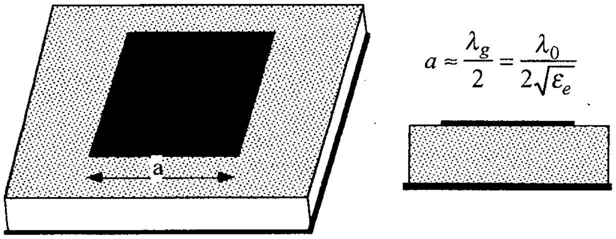


Figure 5a. An un-shorted microstrip antenna: $a \approx \frac{\lambda_g}{2} = \frac{\lambda_0}{2\sqrt{\varepsilon_e}}$.

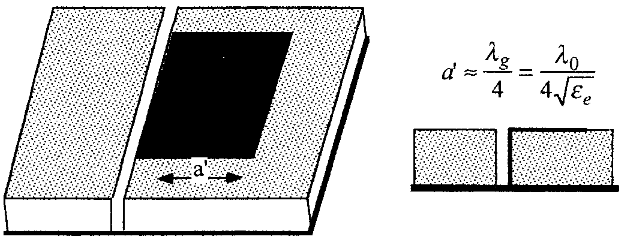


Figure 5b. A shorted microstrip antenna: $d \approx \frac{\lambda_g}{4} = \frac{\lambda_0}{4\sqrt{\varepsilon_e}}$.

is short-circuiting a PIFA to the ground plane is perpendicular to the current on the main radiating plane of the antenna, and will thus lead to cross-polarized radiation. However, high polarization purity is often not required for mobile applications, as the multipath propagation distorts the polarization of the signals, anyway.

A last limitation in antenna miniaturization is the difficulty of correctly feeding small antennas [10]. The correct feeding of a very small antenna is often neither balanced nor unbalanced, but somewhere in between. The microstrip antenna is a simple example that illustrates this effect: A microstrip patch (a typical case of an unbalanced antenna) can be made electrically small by using a substrate with a high permittivity. However, the overall size of the microstrip antenna will also depend on the size of its ground plane. Thus, to make it small, one has to cut the ground plane, making the microstrip antenna closer to a balanced structure than to an unbalanced one, and rendering it difficult to feed it in an efficient way.

3. Miniaturization Techniques and their Effect on the Radiation Characteristics

Techniques for making antennas smaller have been known for a long time, and many of them are described in standard textbooks (see for instance [11, 12] or [13] for more-exotic antennas). The principle behind these techniques will be described below, with an emphasis on its effect on the radiation characteristics of the antenna.

The main miniaturizing tools used are loading the antenna with lumped elements, high-dielectric materials, or with conductors; using ground planes and short circuits; optimizing the geometry; and using the antenna environment (such as the casing) to reinforce the radiation. These techniques will be illustrated and explained with very simple examples in the following paragraphs. The reader should bear in mind that these techniques have been extensively used in the mobile-communication business, where the most interesting results were obtained by combining several of them for the design of one antenna, as is the case with the PIFA. References on small antennas for communications can be found in [8, 9, 14].

A new trend in mobile communications is to combine two antennas into one, instead of making two small antennas. Thus, the radiating element is not the smallest possible, but it covers two or more frequency bands and/or several polarizations [15-22]. Strictly speaking, these dual-band antennas are not miniature antennas, but this technique nevertheless will be described here, as the scope of this paper is how to save space and weight when designing antennas.

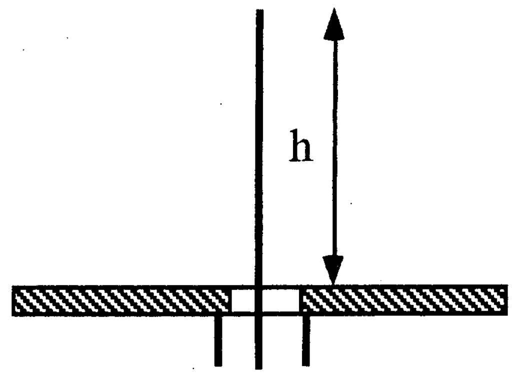


Figure 6a. Geometrical transformations of a monopole: the monopole antenna, $h \approx \frac{\lambda_0}{4}$.

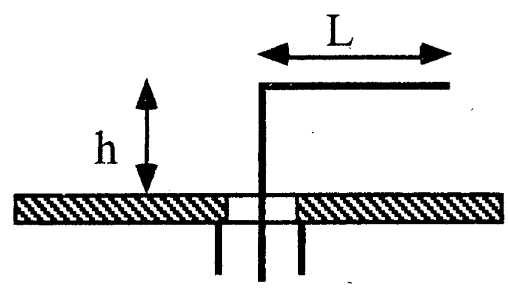


Figure 6b. Geometrical transformations of a monopole: the inverted-L antenna (ILA), $h + L \approx \frac{\lambda_0}{4}$.

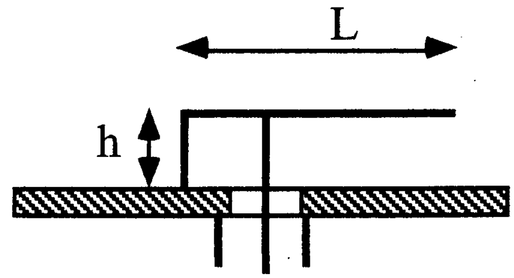


Figure 6c. Geometrical transformations of a monopole: the inverted F antenna (IFA).

3.1 Lumped-Element Loaded Antennas

This may be the simplest and most immediate way to make an antenna smaller than resonant size, and yet keeping resonant features. The idea is that an antenna smaller than a half-wavelength will have a strong reactive input impedance, which has to be compensated for by loading. Figure 1 shows an example of a small loop antenna, which is matched by simply loading it with a capacitor. The effect of this loading will be to either reduce the efficiency of the antenna, if the added element has losses, or to increase the quality factor of the antenna [23, 24] if the latter are small, and to thus reduce its bandwidth. The loading can be performed by adding lumped elements, as in the example of Figure 1, or by adding conductive parts. Some well-known examples are depicted in Figure 2.

3.2 Antennas Loaded with Materials

The loading of the antenna can also be done by modifying the dielectric or magnetic characteristics of the material surrounding it. The principle is illustrated, for a monopole, in Figure 3: The monopole is resonant when its length is about a quarter of a wavelength. Since the wavelength is shorter in a high-permittivity material, the antenna becomes smaller when embedded in such a material. The size reduction will depend on the permittivity and the shape of the dielectric. Again, assuming that no losses are added, this loading will reduce the bandwidth of the antenna by enhancing its quality factor. This is due to the concentration of the electric field in high-permittivity regions, which makes the adaptive launching of a guided wave into free space more difficult (see Figure 4 for the case of a microstrip-patch antenna). Moreover, a higher permittivity is, unfortunately, often equivalent to higher dielectric losses.

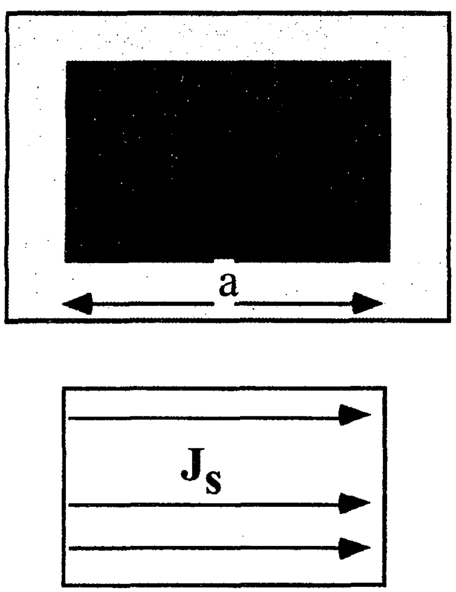


Figure 7a. The effect of slots and notches in a microstrip-patch antenna: the microstrip antenna, $a \cong \frac{\lambda_g}{2}$.

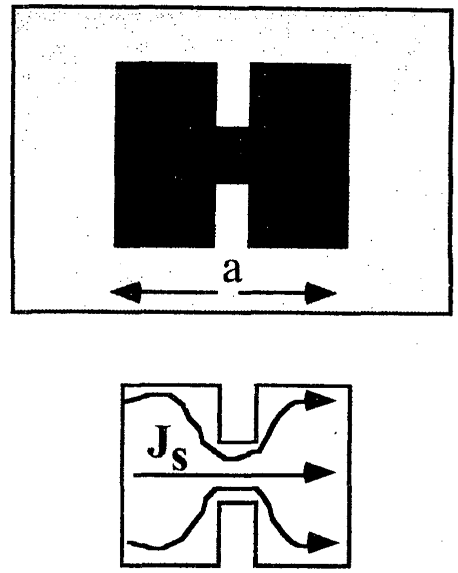


Figure 7b. The effect of slots and notches in a microstrip-patch antenna: the microstrip antenna with notches, $a < \frac{\lambda_g}{2}$.

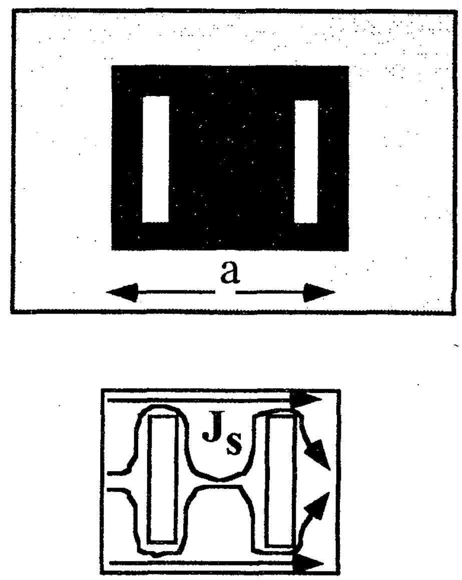


Figure 7c. The effect of slots and notches in a microstrip-patch antenna: the microstrip antenna with slots, $a < \frac{\lambda_g}{2}$.

3.3 Using Ground Planes and Short Circuits

Another popular strategy for making antennas smaller and lowering their profile is to make use of ground planes and short circuits. The principle is easily explained by the well-known example of the monopole compared to a dipole. To be resonant, a dipole must have a length of roughly half a wavelength. This dimension can be halved by replacing one dipole arm by a ground plane, which will, in turn, create a virtual dipole arm, according to image theory.

The principle can be easily extended to planar antennas by adding short circuits to the ground planes, as is the case in the example illustrated in Figure 5. Here, the size of a microstrip patch has been reduced by half by inserting a short-circuiting strip, which acts as a mirror, in the middle of the antenna.

3.4 Modifying and Optimizing the Geometry

Another way to make an antenna smaller and less bulky is to modify its geometry and shape. A good example is the well-known inverted-L antenna, which was derived from the monopole antenna by just bending it (Figure 6). Adding a short circuit to the inverted-L antenna, leads to another popular design, the inverted-F antenna (also shown in Figure 6), which is a textbook example of an antenna design that combines several miniaturizing techniques.

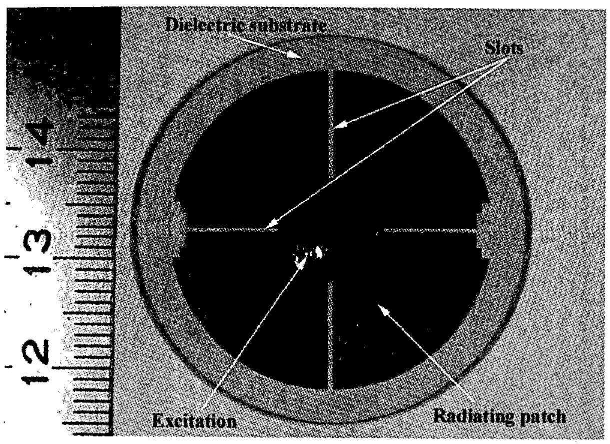


Figure 8. An example of a printed patch antenna for GPS, optimized for small dimensions.

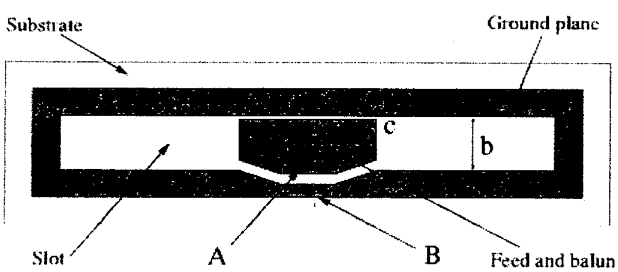


Figure 9. The conformable slot antenna: The slot antenna, printed on a flexible thin substrate, will be conformed around a cylinder. The regions in light gray are dielectric, and the dark regions are copper printed on the dielectric substrate. The antenna is fed by a patch printed in the middle of the slot, the shape of which insures a 50 ohm unbalanced-to-balanced transition.

Table 1. The bandwidth as a function ot the proximity of the metallic plane for a printed slot antenna.

Distance from Ground Plane (mm)	S_{11} (dB)	Bandwidth (MHz)
No ground plane	–20.7	313 (16%)
0	–22.7	196 (8%)
1	–26.8	205 (10%)
2	–16.9	175 (9%)
3	–17.3	203 (10%)

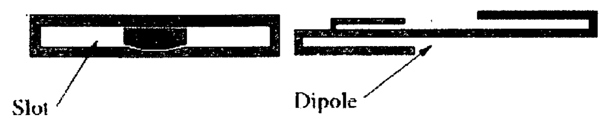


Figure 10a. The slot-dipole antenna, with ground and excitation. This shows the slot and the dipole, which are printed on one side of a flexible thin substrate.

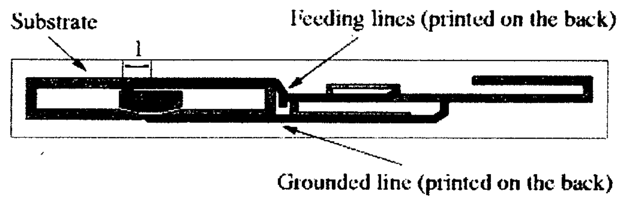


Figure 10b. The slot-dipole antenna, with ground and excitation. This shows the antennas on one side (gray) and the feeding lines (black), which are printed on the other side of the substrate. The antenna will be conformed around a cylinder.

Another example of modified antenna geometry leading to a smaller size is depicted in Figure 7. It consists of a microstrip-patch antenna, the size of which was reduced by inserting slots into it. These slots force the surface currents to meander, thus artificially increasing the antenna's electrical length without modifying its global dimensions.

The effects on the antenna performance of the techniques described in Sections 3.3 and 3.4 are of two kinds. On the one hand, they produce greater current concentrations on the antennas (see, for instance, Figure 7), and therefore increase the ohmic losses and decrease the gain of the antennas. On the other hand, the techniques used (the image effect, the position of short circuits, the position of slots) are often very frequency sensitive. The antenna bandwidth is thus reduced, compared to standard antennas.

3.5 Using the Antenna's Environment

Designing an electrically small antenna reduces, in most cases, to finding the best possible compromise among antenna dimensions and radiation characteristics. A very efficient way to do this is to make the antenna's environment (e.g., the casing of the device, etc.) participate to the radiation. In extreme cases, the casing radiates most of the power, whereas the actual antenna merely acts as a resonator to set the appropriate working frequency. An example is the SMILA (smart monobloc integrated-L antenna), described in Section 4.4

It should be noted that small antennas are very often affected by their immediate surroundings, even when this effect is not directly wanted. A good example of this can be found in the data sheets of some commercially available small antennas, where the indicated measured gain is due more to the measurement setup's cables than to the antenna itself [10]. Fortunately, awareness of the particular behavior of very small antennas is growing, and, nowadays, many suppliers provide their customers with clear rules about how their antennas should be integrated into a system.

4. Examples of Practical Miniaturized Antennas Developed at LEMA

4.1 GPS Antennas

GPS ($f_{GPS} = 1.56\,\text{GHz}$) antennas are a rather special case, since here, the required bandwidth is small (1% bandwidth is sufficient, and this mainly serves to accommodate the dispersion from sample to sample, and the thermal effects), and the polarization is circular (RHCP). In the early days of GPS, the most frequently used antenna was the so-called "quadrifilar helix," a three-dimensional antenna made of wires. To reduce the price and dimensions, most modern GPS antennas are now of planar type, essentially microstrip antennas. In the first generation, quasi-square patches, with dimensions of about 60 mm × 60 mm, were used, etched on a standard ceramic substrate (alumina), with a substrate thickness of a few millimeters. Nowadays, smaller antennas are offered in the market, some being as small as about 25 mm × 25 mm, with a thickness of 3 to 4 mm. In this case, it must be noticed that the claimed gain is most frequently specified for the antenna mounted on a large ground plane!

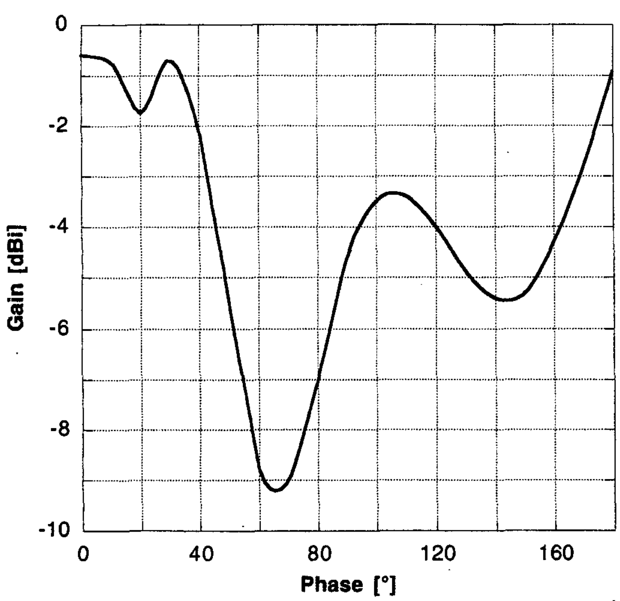


Figure 11. The slot-dipole antenna gain versus the relative phase excitation.

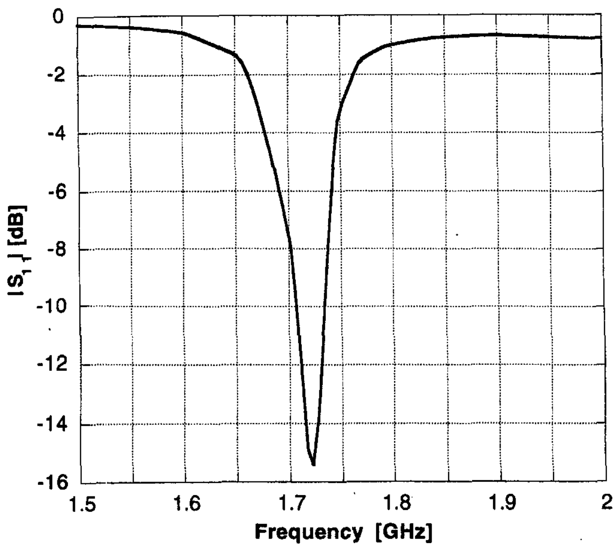


Figure 12. The S_{11} bandwidth for a center frequency of 1.715 GHz. The −10 dB bandwidth is 20 MHz.

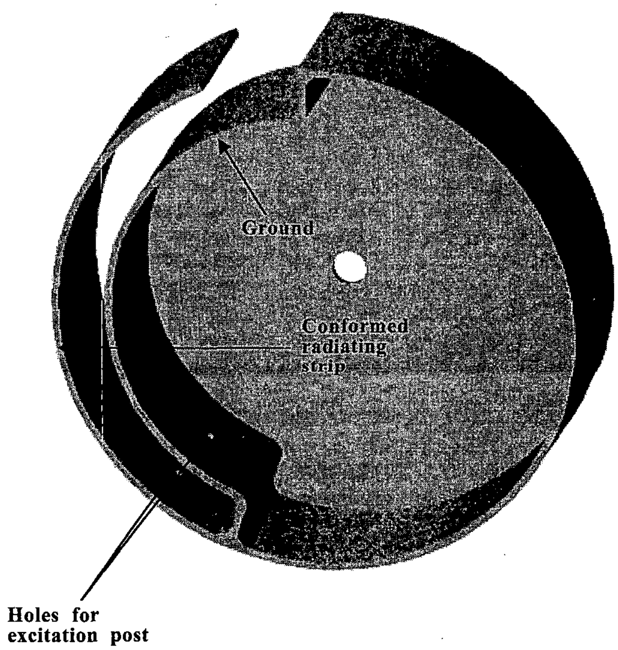


Figure 13. The smart integrated-L antenna (SMILA).

A very small GPS antenna (patented) has been realized in our laboratory [25]. The substrate used was a temperature-stable TMM-10i substrate (made by Rogers), with a diameter of 35 mm, and a thickness of 1.27 mm. The patch-size reduction was achieved by the use of etched slots (see Figure 8). The circular polarization resulted from small notches that introduced an asymmetry in the structure, allowing two orthogonal modes to be excited at two slightly different frequencies. The antenna has been tuned exactly to the GPS frequency, where the measured gain was −3 dBi. The maximum theoretical gain that could have been achieved in the available volume was 1.73 dBi. Our gain was sensibly smaller, but this should be expected, since only a thin slice of the volume of the sphere of Equation (3) was actually used to realize this antenna.

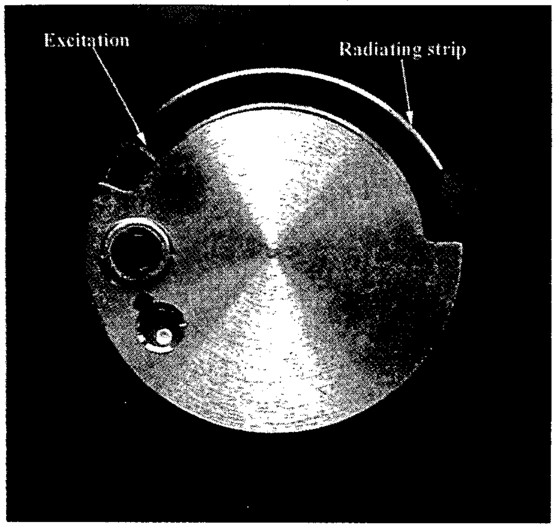


Figure 14. A photograph of the realized SMILA.

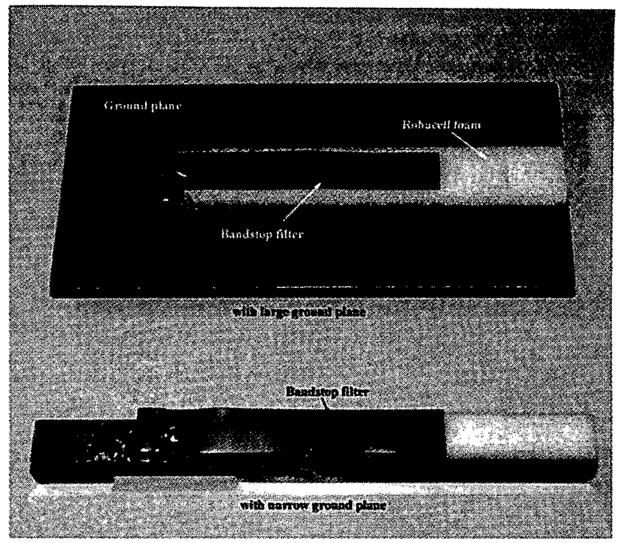


Figure 18. A photograph of realized dual-frequency PIFA.

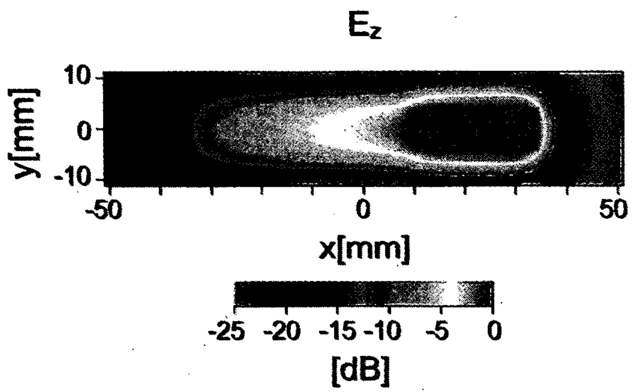


Figure 19a. The near-field measurements of the dual-frequency PIFA: low frequency (925 MHz).

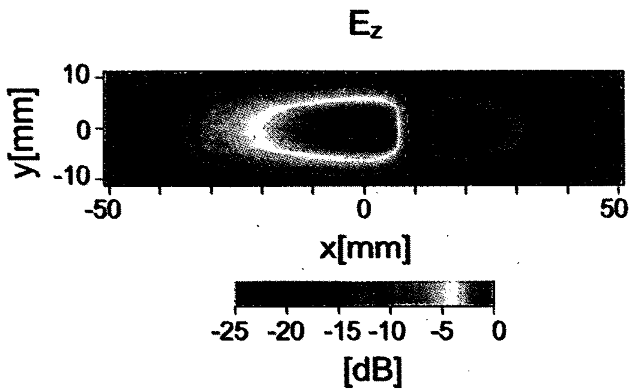


Figure 19b. The near-field measurements of the dual-frequency PIFA: high frequency (1.78 GHz).

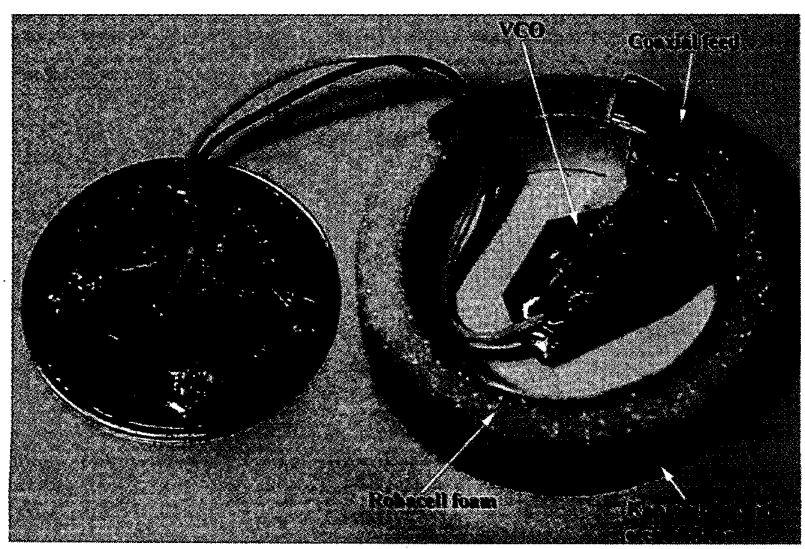


Figure 21. The dual-frequency SMILA (smart monobloc integrated-L antenna).

The bandwidth for a circular polarization with an axial ratio better than 3 dB was very small (0.2%). The maximum gain was lower than for other (larger) commercial antennas, but it was sufficient to allow normal GPS operation, and it represented the best affordable performance within the very limited volume allowed for this prototype.

4.2 Conformal Slot and Slot-Dipole Antennas

A slot antenna, printed on a thin conformable substrate (Epoxy Perstorp FR4, $\varepsilon_r = 4.4$, height = 0.1 mm), is presented in Figure 9. The antenna was first designed to fulfil the DECT (Digital Enhanced Cordless Telecommunications, operating at $f = 1900$ MHz) requirements, when mounted on a small metallic cylinder, which acted as the ground (diameter = 40 mm, height = 10 mm). The main difficulty was to achieve the 6% bandwidth, as the largest dimension allotted was smaller than a quarter of a wavelength. Due to the presence of the metallic cylinder very close to the radiating structure, a slot antenna was chosen. The metal surrounding the slot was reduced to the minimum, and both the balun and the matching network were included in the design [26, 27]. The antenna was etched on a very thin dielectric, so that it could easily be conformed around a cylinder. The flattened-out antenna is presented in Figure 9: It consists of a slot in a small ground plane, fed by a structure that can be assimilated into a very short coplanar waveguide (the feed point is designated by A in Figure 9, whereas point B is grounded). This coplanar-like line was designed to have an input impedance of 50 ohms, and the matching was done by adjusting the geometrical parameters *b*, *c*, and *d* in Figure 9. To test the effect of the metallic cylinder on the slot antenna, several flat slot antennas were measured in front of a ground plane. The effect of this ground plane on the bandwidth was very small, as shown in Table 1. The measured gain of the conformal antenna was quite low, −11 dBi. To improve this figure, a dipole antenna was coupled to the slot antenna [28]. The dipole was folded to fit in the available space, and printed lines were used to ground and feed the antenna, as shown in Figure 10. The antennas were printed on one side of the substrate, and the feeding lines were on the other side. The relative phase between the excitation of both antennas can be adjusted using the parameter *l*, which allows shortening or lengthening the feed line of the slot (Figure 10.). In Figure 11, one can see that the gain is dependent on the relative phase excitation. It has been shown that the maximum gain is achieved when the antennas are excited in equi-phase and equi-amplitude. Compared to the slot antenna, the gain was increased by more than 10 dB, to −0.5 dBi. The theoretical upper limit for the gain is 3 dBi. The S_{11} bandwidth is depicted in Figure 12; the gain bandwidth has the same behavior.

4.3 SMILA

The SMILA (smart monobloc-integrated L-antenna) [29] is derived from the PIFA concept [30]. It is intended to be mounted on the narrow side of a box (in the case described here, the box was a round "pillbox"). To fit onto this case, the PIFA antenna was made narrow and long, with the short-circuited edge being of the same width as the antenna itself. The thickness of the antenna is small when compared to a standard PIFA. This fact would tend to reduce the bandwidth in the case of a large ground plane. The antenna was bent as a round sector to fit the box's shape, and was placed in a recess of the box, so that it was flush with it. In practice, the antenna is manufactured directly as a part of the box. Figures 13 and 14 show this concept clearly.

Even with an antenna close to the ground plane (in our example, 3 mm for a frequency of 1.85 GHz, which is about l/50), the bandwidth is reasonable (about 4%), due to the fact that the dimensions of the ground plane are small. The antenna resonant frequency can be very easily tuned by adjusting the length of the resonant sector, and the impedance can also be matched by properly positioning the feed point. As this antenna was made of a relatively thick metal part, and as no dielectric was present, the losses were very low, which means that the performance was close to the maximum predicted by the theory. Using the measurement technique described in Section 5, the maximum gain at 1.85 GHz for a prototype measuring 35 mm in diameter and 10 mm thick was found to be +0.5 dBi (about 1 dB higher than the best-realized printed antenna of similar dimensions). The highest theoretical gain achievable, according to Equation 3 is 2.8 dBi.

4.4 Dual-Frequency, Dual-Polarization-per-Frequency Four-Port Printed Planar Antenna

Strictly speaking, this antenna is not a miniaturized antenna, but the miniaturization comes from the fact that four different antennas, with individual ports, occupy the place normally required by a single antenna. It uses the principle of combining an SSFIP

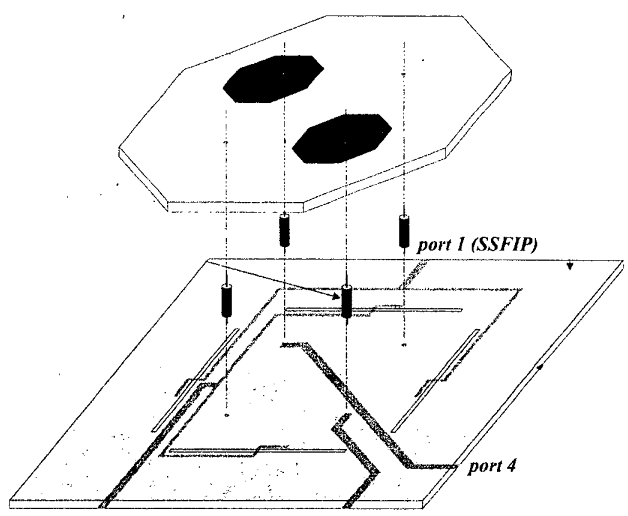
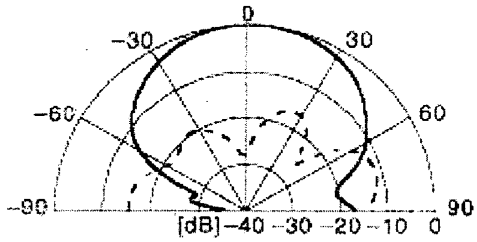
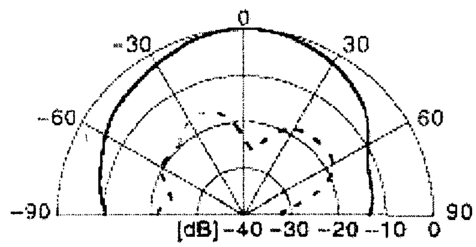


Figure 15. An exploded view of the dual-frequency, dual-polarization-per-frequency four-port printed planar antenna. The two octagonal patches on the top are orthogonally polarized. The ground plane of the octagonal substrate forms the SSFIP patch. This TM10i substrate has $h = 1.91$ mm, $\varepsilon_r = 9.8$. The four supporting posts are made of semi-rigid coaxial cable, with two being used to feed the upper octagonal patches. The ground plane has four slots, and is on an ULTRALAM2000 substrate with $h = 0.76$ mm, $\varepsilon_r = 2.485$.



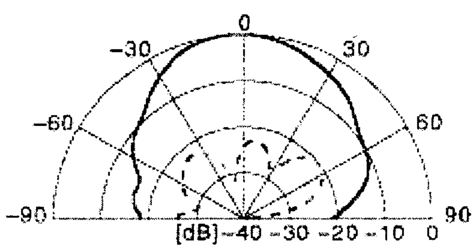
port 1/E-plane

Figure 16a. The radiation pattern of the dual-frequency, dual-polarization-per-frequency, four-port printed planar antenna: port 1, E-plane.



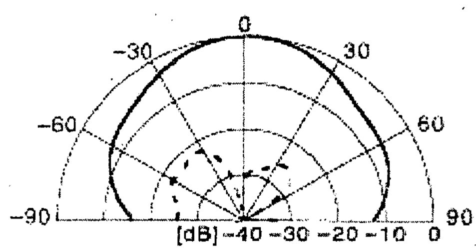
port 1/H-plane

Figure 16b. The radiation pattern of the dual-frequency, dual-polarization-per-frequency, four-port printed planar antenna: port 1, H-plane.



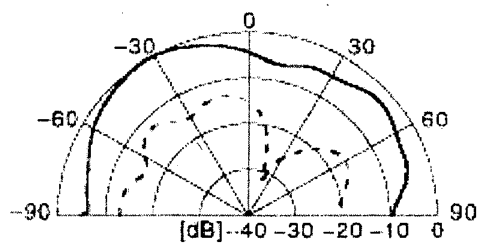
port 2/E-plane

Figure 16c. The radiation pattern of the dual-frequency, dual-polarization-per-frequency, four-port printed planar antenna: port 2, E-plane.



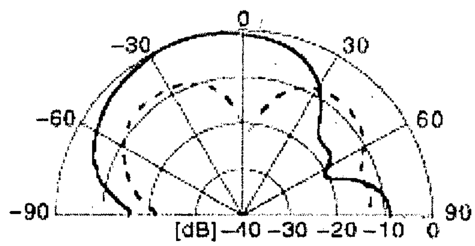
port 2/H-plane

Figure 16d. The radiation pattern of the dual-frequency, dual-polarization-per-frequency, four-port printed planar antenna: port 2, H-plane.



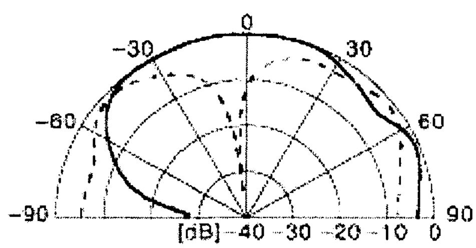
port 3/E-plane

Figure 16e. The radiation pattern of the dual-frequency, dual-polarization-per-frequency, four-port printed planar antenna: port 3, E-plane.



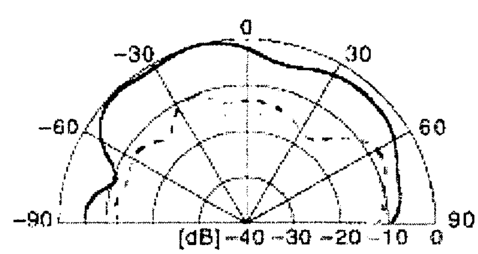
port 3/H-plane

Figure 16f. The radiation pattern of the dual-frequency, dual-polarization-per-frequency, four-port printed planar antenna: port 3, H-plane.



port 4/E-plane

Figure 16g. The radiation pattern of the dual-frequency, dual-polarization-per-frequency, four-port printed planar antenna: port 4, E-plane.



port 4/H-plane

Figure 16h. The radiation pattern of the dual-frequency, dual-polarization-per-frequency, four-port printed planar antenna: port 4, H-plane.

Table 2. The main measured results for the dual-frequency, dual-polarization-per-frequency four-port printed planar antenna. The aperture angle is measured at the –3 dB points.

	f (GHz)	Matching S_{ij} and *Isolation* S_{ij} (dB)				BW (%)	Gain (dBi)	Aperture Angle	
		Port 1	Port 2	Port 3	Port 4			E Plane	H Plane
Port 1	1.83	**–17.1**	*–22.8*	*–17.5 at 2.37 GHz*	*–15.6 at 2.35 GHz*	4.76	5.8	66°	67°
Port 2	1.82	*–22.8*	**–26.8**	*–17.6 at 2.37 GHz*	*–15.8 at 2.35 GHz*	5.04	5.1	51°	57°
Port 3	2.36	*–17.5 at 2.37 GHz*	*–17.6 at 2.37 GHz*	**–21.2**	*–27.0*	0.69	–0.8	58°	71°
Port 4	2.36	*–15.6 at 2.35 GHz*	*–15.8 at 2.35 GHz*	*–27.0*	**–17.3**	0.74	–0.9	95°	61°

(strip-slot-foam inverted-patch) [31] antenna, placed at the bottom, with conventional microstrip-patch antennas on top. The patch of the SSFIP part acts as the ground plane for the upper microstrip patches, and the upper layer is supported by semi-rigid coaxial cables, which act as supporting posts and simultaneously bring the signal to the upper patches. Figure 15 shows an example of an antenna the principle of which has been described in detail in [22]. The lower-frequency antenna is an octagonal patch, fed electromagnetically through four slots in the ground plane. These slots are fed by a microstrip circuit. It is a dual-polarization SSFIP antenna, working at 1.8 GHz, while the higher-frequency (2.4 GHz) comes from the two octagonal upper microstrip patches. The latter exhibit orthogonal linear polarizations, at angles of 45° and 135° with respect to the polarizations of the lower SSFIP antenna. The ground plane for these two upper patches is made by the large octagonal patch, which radiates at the lower frequency. The aim of this design was to show the feasibility of the concept, even for two frequencies that are rather close to each other, as this situation is the most difficult to solve for dual-frequency antennas. The chosen frequencies correspond to DCS (lower) and "Bluetooth" (upper) applications. Despite its compact dimensions, this antenna exhibited good radiation properties and gain, as well as good isolation between the various ports. Table 2 summarizes the antenna's main measured characteristics, and Figure 16 shows the measured radiation patterns.

4.5 Dual-Frequency PIFA and SMILA

The PIFA (planar inverted-F antenna) is one of the most well known and documented [30] compact antennas. It can be viewed as a microstrip antenna where the substrate has been replaced by air, which increases the bandwidth (but also the dimensions!) and limits the losses. A strip short-circuiting one edge of the patch to the ground plane allows the antenna to resonate at approximately a quarter of a wavelength. This very simple and cheap structure, which can be made from just a sheet of metal, is one reason for the success of this antenna. Another is the large choice of parameters (antenna width and length, height above the ground plane, width of the short-circuiting vertical section), which can accommodate a wide range of applications.

Due to the presence of two orthogonal-radiating sections, the polarization purity is bad, as compared, for example, to a microstrip antenna. However, in certain situations, this disadvantage can become an advantage (indoor communications, where the polarization is seldom well known).

It must be noticed that the bandwidth of the PIFA normally increases with a decrease in the ground-plane dimensions, which can be of interest in certain situations.

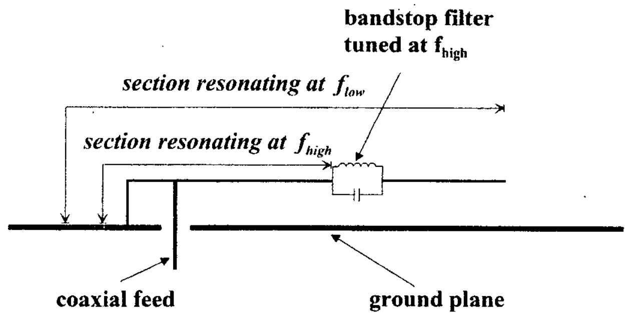


Figure 17a. The principle of a dual-frequency PIFA: general layout.

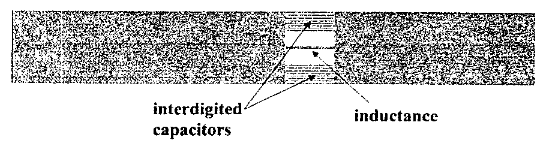


Figure 17b. The principle of a dual-frequency PIFA: A filter with printed inductance and interdigited capacitors.

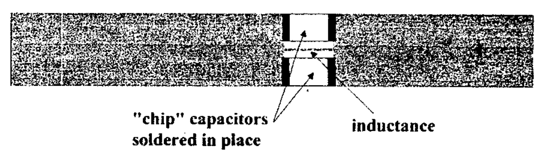


Figure 17c. The principle of a dual-frequency PIFA: A filter with printed inductance and discrete "chip" capacitors.

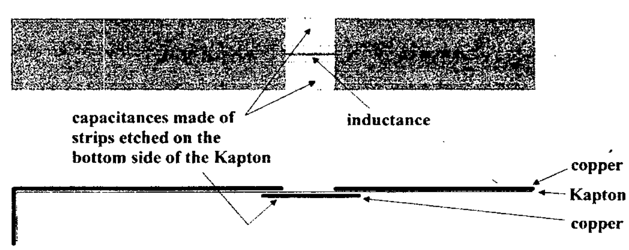


Figure 17d. The principle of a dual-frequency PIFA: A filter with printed inductance and capacitances (double-sided circuit).

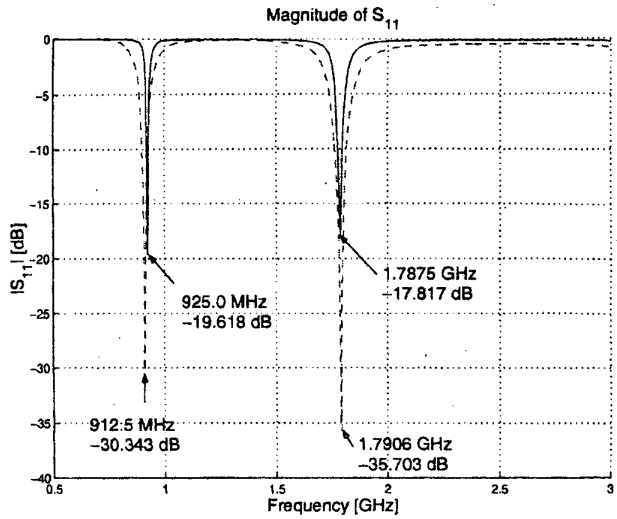


Figure 20. The input reflection coefficient of the dual-frequency PIFA. The solid line is with a large ground plane; the dashed line is with a narrow ground plane.

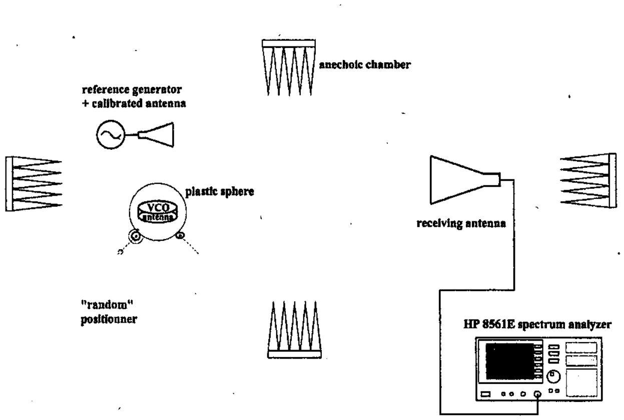


Figure 22. The principle of the gain-measurement setup for electrically small antennas.

The PIFA and the SMILA are good candidates for multifrequency single-feed applications, as shown by the following examples of dual-frequency antennas [32]. The principle is well known from shortwave bands, and is described in Figure 17 for a PIFA antenna. The resonant arm of the PIFA was interrupted at a length corresponding to the upper frequency band by a bandpass filter, tuned at the lower frequency band, with the total length of the antenna corresponding to the lower frequency band. The filter can be realized using lumped or printed elements, the latter being chosen in our case. The inductance was realized by a thin line, and the capacitance, by two interdigital capacitors. Figure 18 shows a photograph of this antenna for two different ground planes. Figure 19 shows the near-field measurements at both frequencies, where the influence of the filter in the upper band can be clearly seen. Figure 20 shows the reflection coefficient of this antenna.

The same concept has been applied to a SMILA (Section 5.3) at the same frequencies for a dual-band GSM application (patent pending, Figure 21). The measured gain for this antenna was −1.9 dBi at 935 MHz (the theoretical upper limit with respect to the antenna size is −0.7 dBi), and −1.5 dBi at 1800 MHz (with a theoretical upper limit of 2.8 dBi). We see that this antenna is very close to optimum at its lower frequency band, which is explained by the fact that, in this case, the radiating surface uses a great percentage of the available volume.

5. Measurement of Electrically Small Antennas

Electrically small antennas are difficult to measure properly because they are neither purely symmetrical nor antisymmetrical, due to the limited size of ground planes or feeding baluns. Therefore, when such an antenna is connected to a measuring device, a current will flow in the outer conductor of the cable connecting the antenna, creating spurious radiation. In fact, this spurious radiation can be so important that it will frequently completely mask the characteristics of the antenna under test, and will cause overestimation of the gain by up to 10 dB or more [10]. Impedance measurements are also affected, as the measurement will yield the impedance of the antenna formed by the small antenna plus the connecting cable. Furthermore, when a small antenna is mounted on a casing, the latter will participate in the radiation, and its effect should be taken into account when characterizing a mobile-communication device. To complicate the problem, the polarization is seldom well defined. Thus, the only way to obtain useful information about the antenna behavior of a mobile-communication device is to test it under operating conditions, and the relevant features will be its gain and efficiency.

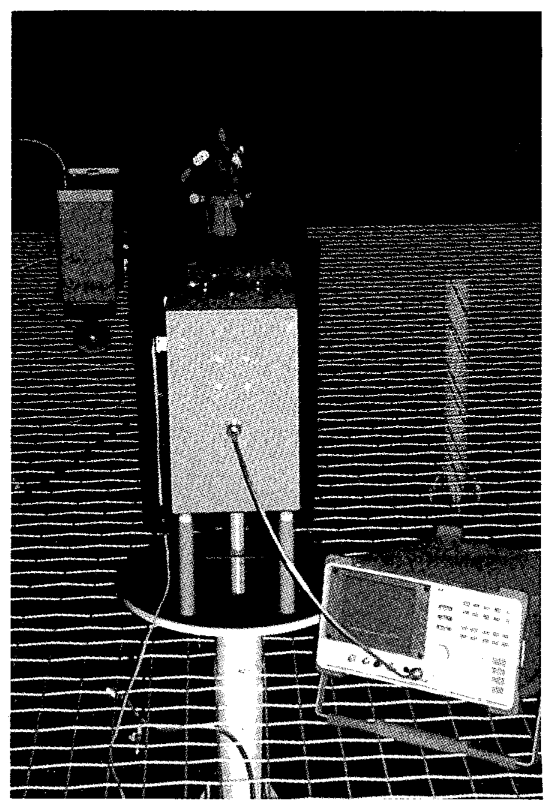


Figure 23. A photograph of the gain-measurement setup.

An original solution to obtain the maximum gain of electrically small antennas matching the above conditions has been developed in our laboratory [33], and will be briefly discussed here. With reference to Figure 22, it consists of the following steps:

- Mounting the electrically small antenna under test in its definitive environment, typically in the shielded case that will later contain the electronic equipment used for the particular application.

- Feeding the antenna with a stable VCO of known (measured) output power, enclosed with the batteries in the shielded case. This forms the Device Under Test (DUT).

- Rotating, in an anechoic chamber, the DUT in all possible orientations and polarizations, using a specially designed "random" positioner, and capturing the maximum level received by the receiving antenna connected to a spectrum analyzer in the "peak hold" mode.

- Replacing the DUT by a reference antenna having an accurately known gain, fed by a calibrated synthesizer at the same frequency as the DUT. The power level is adjusted in order to obtain the same received level as the maximum produced by the DUT.

- Determining the exact maximum gain of the small antenna by a simple calculation.

Figure 23 shows a picture of the actual measurement setup.

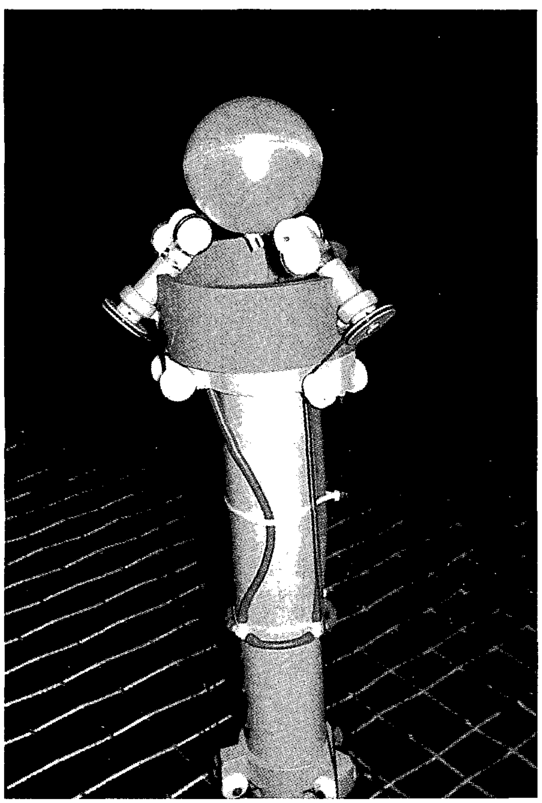


Figure 24. The random positioner used for gain measurements.

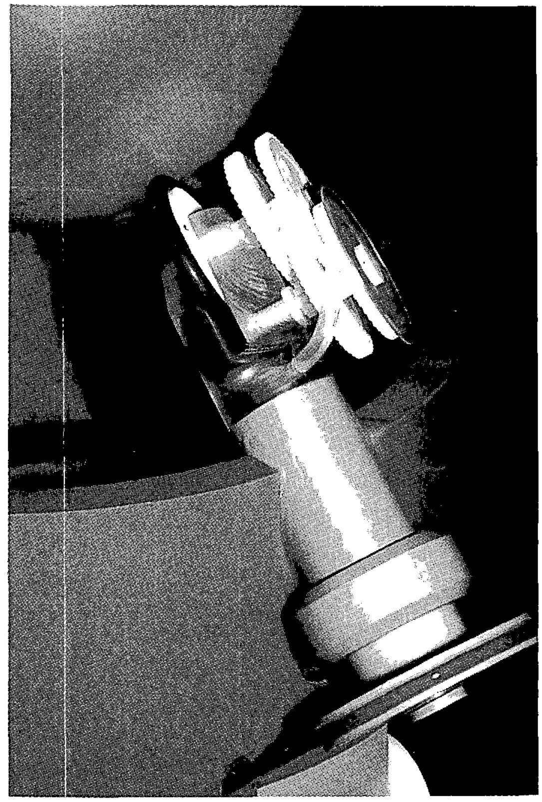


Figure 25. A photograph of the pneumatic motor and the driving belt that induce the rotation for two different axes as part of the gain measurement.

The random positioner used for these measurements was developed specifically for this type of measurement (a zoom view is given in Figure 24). It is made entirely of low-permittivity dielectric materials, in order to perturb the DUT as little as possible. The latter is enclosed in a polypropylene sphere, which is randomly rotated on three wheels. The rotation of each wheel is performed by compressed-air motors, the speed of rotation being adjusted by varying the air pressure. The rotation of the compressed-air motors around their axes is performed by an electrical motor located in the pedestal of the system (far from the DUT), the transmission of movement being made by driving belts with different speed for the three axes. Two axes of rotation (one using compressed air and one using a driving belt) are illustrated in Figure 25. A gain measurement using this type of setup takes just two to three minutes.

6. Conclusions and Perspectives

The demand for new mobile-communication devices and systems will certainly continue for several years, increasing the market pressure for efficient electrically small antennas. These radiating devices should be optimized for each application, in order to get the best compromise among antenna volume, gain, and bandwidth. In most cases, the optimal antenna configuration will be the one using most of the allotted space, having as few as possible dielectric parts, and no high current concentrations in its conducting parts, in order to minimize the losses.

The working frequency of mobile-communication systems is constantly increasing, for two main reasons. First, new frequency bands have to be allotted to take care for the new services proposed. Second, higher working frequencies allow the larger bandwidths requested by multimedia services or high-speed data transfer. The consequence of this shift in frequency is a diminution of the antenna specifications in terms of volume compared to the wavelength, as the latter decreases. The freedom thus gained by the antenna designers should be used to optimize the radiating device in other aspects, namely, the SAR in the human beings using the mobile-communication devices designed. It becomes indeed possible to take this essential aspect into account for antennas having a size of half of a wavelength or more, whereas it is very difficult for antennas having a size of the order of a tenth of a wavelength, as do those presented in this paper.

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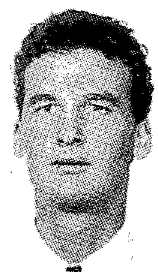
Introducing the Feature Article Authors

Anja Skrivervik received the Electrical Engineer degree in 1986 from the Ecole Polytechnique Fédérale de Lausanne, Switzerland. She obtained her PhD form the Laboratoire D'Electromagnétisme et d'Acoustique of the same institution in 1992. She then worked in industry before joining the Ecole Polytechnique Fédérale de Lausanne as an Assistant Professor at the end of 1995.

Her research interests include small antennas, printed circuits and antennas, and periodic structures in electromagnetics, such as arrays, FFS, and photonic bandgap materials. She represents Switzerland in Commission B of URSI.

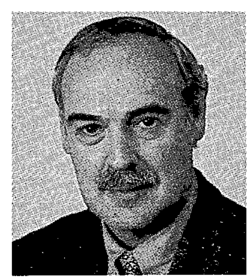
Jean-François Zürcher was born in Vevey, Switzerland, in 1951. He graduated with the degree of Electrical Engineer from Ecole Polytechnique Fédérale de Lausanne (Lausanne Institute of Technology) in 1974. He is presently employed as permanent Scientific Associate with the Laboratoire d'Electromagnétisme et d'Acoustique EPFL, where he is the Manager of the microwave laboratory. His main interests lie in the domain of microstrip circuits and antennas. In 1988, he invented the SSFIP (strip slot foam inverted patch antenna) concept, which became a commercial product. He is presently developing instrumentation and techniques for the measurement of the near fields of planar structures, and for microwave materials measurement and imaging.

Mr. Zürcher is the author or co-author of more than 90 publications, chapters in books, and papers presented at international conferences. He is one of the two authors of the book *Broadband Patch Antennas*, published by Artech in 1995. He holds six patents.



Olivier Staub was born in Lausanne, Switzerland, in 1970. He received the electrical engineering Diploma from the Swiss Federal Institute of Technology at Lausanne (Ecole Polytechnique Fédérale de Lausanne) in 1995.

In 1996, he joined the Electromagnetics and Acoustics Laboratory (LEMA) of the EPFL in order to do a PhD thesis. In this context, he collaborated in several projects of the Swiss Priority Program in Micro- and Nano-System Technology (MINAST), where he collaborated with Swiss universities and industry. His main tasks were the design and analysis of electrically small antennas embedded in small communication devices. Mr. Staub received the IECON Second Best Paper Award at the IEEE IECON'97 conference. He is co-holder of three patents. In 2001, he joined Ericsson AG in Bern, Switzerland, as a radio planning engineer.



Juan R. Mosig was born in Cadiz, Spain. He received the Electrical Engineer degree in 1973 from Universidad Politecnica de Madrid, Spain. In 1976, he joined the Laboratory of Electromagnetics and Acoustics at Ecole Polytechnique Fédérale de Lausanne (EPFL), Switzerland, from which he obtained a PhD degree in 1983. Since 1991, he has been a Professor at EPFL, and since 2000, the Head of the EPFL Laboratory of Electromagnetics and Acoustics (LEMA). In 1984, he was a Visiting Research Associate at Rochester Institute of Technology, Rochester, NY. He has also held scientific appointments at universities of Rennes

(France), Nice (France), the Technical University of Denmark, and the University of Colorado at Boulder, CO.

Dr. Mosig is the author of four chapters in books on microstrip antennas and circuits, and more than one hundred reviewed papers. He is co-organizer and lecturer of yearly short intensive courses in numerical electromagnetics (in Europe and the USA). He is a member of the Swiss Federal Commission for Space Applications, and is responsible of several research projects for the European Space Agency. His research interests include electromagnetic theory, numerical methods, and microstrip antennas. He is a Fellow of the IEEE.

Editor's Comments *Continued from page 8*

Recently, demonstration inflatable array antennas for an L-band SAR, and for X- and Ka-band reflectarrays, have been designed, built, and tested. John Huang discusses the challenges involved in the design and construction of these arrays, and how they were met. He also provides a fascinating look at some of the unique challenges – and at current approaches to addressing these challenges – associated with developing inflatable arrays for space use.

The ability to maximize the number of simultaneous users (and their quality of service) in a cell is fundamental to the economics of almost all cellular communications systems. Smart antennas (antennas that tailor their beams, based on the position of the mobile user and on the directions of arrival and departure of the multipath propagation inherent in many cellular-communication situations) are one very important approach to increasing both the number of simultaneous users and their quality of service. In their feature article, Martin Steinbauer, Andreas Molisch, and Ernst Bonek introduce a new concept for understanding and characterizing the propagation involved with this. They call this the double-directional radio channel. The emphasis is on the need to understand the information about the angular directions of propagation of the multipath at *both* ends of the channel. They detail the method they developed for measuring this information with a high degree of precision. They then develop methods of analyzing the information, and, in particular, of directly relating principal features of the propagation to elements of the physical environment. The results – which are explained in this article in a particularly easy-to-understand manner – lead to some important conclusions regarding popular propagation models. They also demonstrate the ability to provide a full characterization of the amount and nature of the multipath propagation in a particular geometry, which turns out to be the key information necessary to properly exploit such channels.

The title of the feature article by Frank Gronwald and Jürgen Nitsch may strike people as a bit imposing, at first. In fact, the article is simply interesting, necessary reading for anyone who wants to understand the fundamentals of electromagnetics. As the authors explain, some of these fundamentals are rooted in quantum physics. The authors use the electromagnetic potentials and a simple, classic, double-slit experiment that can be explained using both classical and quantum electromagnetics to provide a very understandable connection between the classical and the quantum views. This invited article was based on "Physics Note 10" (from the EMP Note Series). The presentation of that note received the "Best HPE Paper Award – Best Basic Paper" award at the EUROEM 2000 meeting. When you read it, I think you'll understand why.

Also in This Issue

With this issue, we welcome Cynthia Furse as the new Chair of the AP-S Education Committee and as Associate Editor for the Education Column. Her first column, and an "introduction" for her, appear in this issue. Thanks are due to the out-going Chair, Sembiam Rengarajan, for the excellent job he has done as a part of the *Magazine* Staff.

Small mobile-terminal antennas and smart antennas – two topics directly related to some of our feature articles – have also been the topics of ongoing research and development at the Wroclaw University of Technology. In the Wireless Corner, edited by Tuli Herscovici and Christos Christodoulou, Pawel Kabacik describes the results of this work. These are very interesting, and quite useful, antennas.

This issue's EM Programmer's Notebook, edited by John Volakis and David Davidson, is interesting for several reasons. First, the contribution by Prodromos Atlamazoglou, Hristos Anastassiu, and Dimitra Kaklamani provides a nice introduction to the principle of Literate Programming, originally developed by Donald Knuth. It also introduces several public-domain tools that can be used to support Literate Programming. Second, it provides an introduction to the concept and arithmetic of quaternions (used to perform geometrical rotations, among other purposes). Third, it shows how to implement this arithmetic using the object-oriented capabilities of *FORTRAN 90*. Fourth, in doing so, it provides a very nice set of "how to" examples for creating a new arithmetic in an object-oriented language. Since the arithmetic of quaternions is in many ways very similar to the arithmetic of three- and four-vectors, what is in this column could perhaps also be used as a template for developing similar vector operations. Regardless, there is a lot of useful material here.

If you work with radiation problems in cylindrical coordinates – and, in particular, with the directivity of radiators with circular symmetry – you should read the Antenna Designer's Notebook contribution by John Mahony (the Notebook is edited by Tom Milligan). John derives approximations to integrals that regularly occur in such computations in terms of easily evaluated standard functions, and shows that the approximations are very good over a wide range of parameters. As he notes, they should be useful where evaluation by hand or using hand-held calculators is desirable. Because they replace integrations of oscillatory integrands with functions (e.g., trigonometric and Struve functions) for which efficient, well-behaved numerical algorithms already exist, they may also be useful in computational electromagnetics applications.

Jim Lin has some interesting commentary on how cell phone testing, and research on the effects of cell phones on humans, should be carried out. If you enjoy *Dilbert*, or work for a less-than-perfect boss, be sure to read Randy Haupt's Ethically Speaking in this issue.

Your Vote is Critical!

There are expanded statements from the candidates for Division IV Director-Elect in this issue. The participation in IEEE elections is usually very low (with often of the order of only 20% of those eligible to vote actually voting), and the elections are often decided by very few votes. In other words, your vote can really

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