Fourth, if we want to compute the electric field $\vec{E}$ only on a certain spatial window of the entire $xy$-domain, whether centered or not, we can easily modify the convolution algorithm by providing as input the function $\vec{E}$ on the desired $xy$-window, as long as a sufficient decay of $\vec{E}$ is ensured. The Green’s function can also be considered on a window smaller than the entire $xy$-domain, provided that such a window is centered at the origin and a decay of at least 60 dB from the peak is ensured. By doing that, the computational time for the convolution drastically diminishes.

Fifth, it is noted that the singularity extraction technique provides the correct aperture field also for nonsingular spectral components and the technique can thus be applied to all components of the PWS without a priori considerations on the absence or presence of the singularity.

4. TEST CASE

To illustrate the effect of the singularity extraction technique presented in Section 3, we investigate here an array of 5 $y$-oriented electric Hertzian dipoles displaced along the $x$-axis at a distance of 2A from each other, see Figure 2. The excitations of the five dipoles are $P, P_{/2}, P_{/5}, P_{/8},$ and $P_{/10},$ respectively, with $P$ being the dipole moment of the dipole at the origin. The exact PWS is first computed from Eq. (2) on the $[-2k_x : 2k_x] \times k_y$-domain with 91 sampling points in both directions, see Figure 3. The PWS clearly shows the singularity in both the $x$- and $y$-components. In the computation of the aperture field from this PWS, only the visible region is taken into account since this is most often the case in practice, while the invisible region is zero-padded.

Figure 4a shows the analytical $x$- and $y$-components of the electric field on the $z = 0.1A$ plane while Figure 4b shows the result of a straightforward IFFT of the singular PWS without the use of the singularity extraction technique. It is evident that while all five dipoles are seen in Figure 4(a), only the first two are clearly distinguished in Figure 4(b). The last three dipoles, having a weaker excitation, cannot be correctly detected since the singularity is not properly taken into account. Figure 4c then shows the result of applying the singularity extraction technique. In this case all five dipoles are clearly detected and the difference in their excitations can also be seen. The slightly wider extensions of these compared to the analytical result is due to the truncation of the PWS to the visible region and the truncation of the two functions involved in the convolution on the finite $xy$-plane.

5. CONCLUSIONS

An effective technique to extract the singularity of plane wave spectra in the computation of antenna aperture fields has been presented. The algorithm is based on the Inverse Fast Fourier Transform and Weyl-identity and allows the accurate computation of the aperture field when a dense sampling in the spectral domain is not possible. The detection of sources of very weak amplitude has been verified by a numerical example and the evident advantages compared to the Inverse Fast Fourier Transform of the PWS without the singularity extraction have been underlined.

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In this letter, a compact and wideband double-sided printed antenna with spiral configuration is designed to operate in both C and X-bands. The antenna was fabricated and characterized by determining the impedance matching condition, the impedance bandwidth, and the radiation patterns. The compact size of the antenna was obtained by optimizing the spiral length, by using a circular capacitive loading (central disk), and by using the double-sided structure, whereas the broadband behavior was achieved by a capacitive coupling (a rectangular patch is connected to the ground plane), and by using a modified-CPW feeding technique. It is shown that, for a fixed spiral length, the impedance bandwidth can be tuned especially by the width of the rectangular coupling-patch.

2. ANTENNA DESIGN

The proposed antenna was etched on F4 substrate with thickness $h = 0.8$ mm and permittivity $\varepsilon_r = 4.4$. The design parameters of the optimized structure are summarized in the caption of Figure 1. The dimension of the antenna is 50 by 40 mm$^2$. The antenna is fed by a balun.

The printed patch on the top side of the substrate ($W_{\text{sub}} \times L_{\text{sub}}$), is obtained from a circular-spiral, with an inner radius $L_2$ and a width $W_3$, by adding a central disk of diameter $D$ as shown in Figure 1. This modified-spiral patch is fed by a 50-Ω strip line of width $W_2$. On the bottom side of the substrate, a coupling rectangular-patch of dimensions $W \times L$, a rectangular ground plane of dimensions $W_3 \times L_{\text{sub}}$, a rectangular slot of width $W_4$, and a connecting strip line of dimensions $(L_1 \times W_1)$ were printed.

The trace equation of the spiral is $r = r_0 \Theta + r_1$, where $r_1$ is the inner radius, $\Phi$ the associated angle and $r_0$ the growth rate determined from the spiral width $W_2$ and the spacing $s$ between each turn, $r_0 = (s + W_2)/\pi$.

In our case, with the design parameters $W_2 = 1$ mm, $s = 1$ mm, $r_1 = D/2 = 2.5$ mm and $n = 6.8$ (n is the number of turns), the trace equation is $r = 2/\pi \Phi + 2.5$ (mm) with $\varepsilon(0, 13.6\pi)$. Consequently, the spiral length is 379.3 mm, and by adding the microstrip feed line length 20.5 mm, we get the total length of the printed dipole on the top side of the substrate, which is about 400 mm.

![Figure 1](image1.png)

**Figure 1** Printed dipole antenna (Dimensions in mm: \(L_{\text{sub}} = 50, W_{\text{sub}} = 40, L = 35, W = 12, L_1 = 15.5, W_1 = 6, L_2 = 15, W_2 = 1, W_3 = 5, W_4 = 3, D = 5\))

![Figure 2](image2.png)

**Figure 2** Effect of the capacitive coupling element on the measured return loss
During the design procedure, the shape (rectangle or disk) and the position of the coupling patch (when it covers the inferior half, the superior half or the total area of the spiral) were studied numerically and experimentally. It is found that the $-10$-dB impedance bandwidth depends slightly with the shape of the coupling patch and highly with its position. In fact, when the rectangular patch covers the superior half of the spiral as proposed in this article, the excited surfacic currents in this region are clearly higher than the two other cases. From the surfacic current analysis, it seems that the resonating wideband results on the overlap of several resonant bands associated each to the arc-dipole situated in the covered region by the coupling patch.

The structure was optimized in order to design a compact antenna with a small substrate operating in both C- and X-bands. First, Ansoft-HFSS simulations were performed on the structure by optimizing, especially, the spiral-length, and the radius of the central disk. Next, a coupling element (a rectangular patch ($W \times L$) connected to the ground plane with a strip line ($L_1 \times W_1$)) is added. Finally, the effect of the parameters design on the antenna return loss was studied.

Figure 2 shows the effect of the capacitive coupling on the return loss. As it is shown, the $-10$-dB impedance bandwidth is clearly improved by adding the coupling element.

Figures 3 and 4 show the measured return loss for the proposed antenna with various widths $W$ of the rectangular coupling-patch, and various width $W_1$ of the connecting strip line, respectively. The return loss curve corresponding to the optimized structure is plotted with solid thick-line in Figures 3–4. As it is shown, the

![Figure 3](image1)

**Figure 3** Measured return loss for the proposed antenna with different rectangular widths $W$

![Figure 4](image2)

**Figure 4** Measured return loss for the proposed antenna with different stripline widths $W_1$

![Figure 5](image3)

**Figure 5** Prototype of the proposed antenna. [Color figure can be viewed in the online issue, which is available at www.interscience.wiley.com]

![Figure 6](image4)

**Figure 6** Measurement and simulation results of the antenna return loss
impedance matching of the printed dipole antenna dependent critically on $W$ and slightly on $W_1$.

Figure 3 shows several typical results for the antennas with various values of $W$ when $W_1$ is fixed to be 6 mm. It is noted that, the performance of the wideband 6.27–12.47 GHz is mainly influenced by the width $W$. The -10-dB impedance bandwidth can be improved by tuning the value of the parameter $W$. When the parameter $W$ is shifted from the optimum value ($W = 12$ mm), the bandwidth is remarkably narrowed, or divided into smaller bands.

Figure 4 gives the effect of the width $W_1$ on the impedance bandwidth when $W$ is fixed to be 12 mm. As it can be remarked, the width $W_1$ of the connecting strip can be adjusted to slightly improve the antenna in-band impedance matching. The optimum value was found to be $W_1 = 6$ mm.

Figure 7 Measured x-z plane radiation patterns at resonant frequencies 6.68, 8.88, 9.56, and 10.61 GHz, respectively.
In fact, the variation of the impedance matching, with \( W \) and \( W_1 \), is due to the modification of the coupling region condition between the capacitive coupling rectangular-patch \( W \times L \) and the spiral dipole.

3. EXPERIMENTAL RESULTS

Prototype of the optimized printed dipole antenna (see Fig. 5) was fabricated based on the dimensions specified in Figure 1.

The Return loss of the structure was measured over the frequency range 5–13 GHz by using a Wiltron 360B vector analyzer. Measurements on the radiation patterns were performed in anechoic room.

Figure 5 illustrates both measured and simulated results on the return loss. As it is shown, a relatively good agreement between the simulated and measured return loss was observed. The small discrepancies between simulated and measured can be attributed to

Figure 8  Measured y-z plane radiation patterns at resonant frequencies 6.68, 8.88, 9.56, and 10.61 GHz, respectively
the effect of the SMA connector and fabrication imperfections. The realized antenna presents a wide resonant band that extends from 6.27 GHz to 12.47 GHz, with an impedance matching of ~66% (for $S_{11} < -10$ dB). As a result, it can be seen that the impedance bandwidth of this resonant band is suitable for communication systems that operate in the C and X-bands.

Otherwise, a relatively good agreement between simulated and measured return loss results can be observed from Figure 6. The discrepancy between measured and simulated results might be attributed to the effect of the SMA connector and fabrication imperfections.

The measured far-field radiation patterns of both $E_{\theta}$ and $E_{\phi}$ components for the (x-z) plane and (y-z) plane, at the resonance frequencies 6.68, 8.88, 9.56, and 10.61 GHz, are shown in Figures 7 and 8, respectively.

As it is shown, the radiation patterns for the $E_{\theta}$ and $E_{\phi}$ components present a certain degree of similarity in both x-z and y-z planes with several dips. These dips might be attributed to the possible destructive interaction in radiation between the modified spiral dipole printed on the top of the substrate and the coupling element printed on the bottom side. However, the radiation is directed toward both negative and positive y-axis. This effect is explained by the radiated fields, at the same time, of the spiral dipole and the capacitive coupling element printed on both sides of the substrate.

The power intensity of the far-field radiation of $E_{\theta}$ and $E_{\phi}$ components are approximately similar in both x-z and y-z planes, except for the frequency 6.68 GHz in the y-z plane. This effect can be explained by the spiral topology of the propagation path and the feeding technique.

The maximum measured gains observed in the range 6.27–12.47 GHz at frequencies 6.68, 8.88, 9.56, and 10.61 GHz are 3, 4.1, 2.9, and 3.5 dBi, respectively.

4. CONCLUSION

A novel compact and wideband printed dipole antenna for multiple wireless services has been realized. The proposed double-sided printed dipole antenna with a spiral dipole exhibits a wideband impedance behavior. The compact size of the antenna was obtained by optimizing the antenna geometry, whereas the broadband behavior was achieved especially by a capacitive coupling. The resonant band is about 6.2 GHz (6.27–12.47 GHz) with a fractional bandwidth of 66%. The measured peak gains over this band at frequencies 6.68, 8.88, 9.56, and 10.61 GHz are 3, 4.1, 2.9, and 3.5 dBi, respectively.

The wide impedance bandwidth, the relatively good gain and radiation characteristics, and the simple feeding of the proposed antenna make it a good choice for communication systems that operate in the C and X-bands applications.

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