OPTIMIZATION OF CONFORMAL
LOW PROFILE DIPOLE ANTENNAS

A Thesis in
Electrical Engineering

by
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Abstract

A study of horizontal low profile (close to the ground) open sleeve dipole antennas was carried out. The open sleeve dipole is a well established technique of increasing the bandwidth of a conventional dipole. By introducing parasitic elements in the vicinity of the driven element the impedance is slightly desensitized to changes in frequency. The objective was to see if such open sleeve dipoles at distances of less than $0.1\lambda$ from a PEC ground plane would be a feasible alternative to conventional patch antennas for use in conformal applications.

Typically horizontal dipoles close to the ground have low radiation resistance and act as poor radiators. Also they have a high impedance mismatch with typical feed systems ($50\Omega$ or $75\Omega$) making them impractical for use in certain applications. Simulations using FEKO showed that such low profile antennas based on the open sleeve idea could be constructed close to a PEC ground using thin wires and still have acceptable VSWR characteristics. Conformal variants of such antennas were then built - using thin conducting strips (instead of thin wires) on suitable substrates - and tested to verify the simulation results. These substrates in addition to providing mechanical backing to the antenna also reduce its effective length thus permitting miniaturization of the antenna. The design of these antennas and their results of this study are presented in the thesis. A qualitative comparison between the VSWR obtained using simulations and measurements seems to suggest the possibility of using these open sleeve dipole designs for conformal low-profile antenna applications.
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I wish to thank my advisor, Dr. Breakall - whose timely help and advice enabled me to complete this project successfully.

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Thank you, one and all.
1. Introduction

1.1 Dipole Antenna

Broadly speaking dipole antennas can be characterized by two radiating arms fed by a source. This definition incorporates a larger class of antennas that have two radiating arms; namely bow-ties, bi-conical antenna, wire or cylindrical dipole, folded dipole etc. Conventionally the term dipole is frequently used synonymously with thin wire dipoles. The study of thin wire dipole in free space is dealt in sufficient detail in [1], from which we know that the input impedance, $Z_{in}$, of the infinitesimally thin half-wave dipole is approximately $73 + 42.5j\Omega$. The dipole resonates (has $Im[Z_{in}] = 0$) at a length of about $0.475\lambda$. This dipole has a very narrow bandwidth, measured in terms of input impedance as a function of frequency, as shown in Figure 1.1.

An antenna similar to the dipole antenna is the monopole antenna - with only one radiating arm fed against a ground plane. Analysis using image theory (assuming infinite ground plane) yields similar results for such monopole antennas and their equivalent dipole antennas - except that their impedance is half that of the dipoles.

In practical applications, an antenna is seldom in free space and the effect of ground and other objects in its vicinity play a significant role in its performance. The effect of ideal and lossy ground is also treated to some detail in [1]. The effect of ground on a half-wave dipole antenna depends on the orientation of the dipole - ie. horizontal or vertical. The analysis can be made using image theory (for ideal PEC ground). The rules for image
theory are summarized in Figure 1.2. The effect of PEC ground on horizontal and vertical dipoles are shown in Figure 1.3b and Figure 1.3a as functions of their height above ground. This suggests that while vertical dipoles close to the ground are potentially good radiators, horizontal dipoles very close to the ground act as poor radiators. This is in agreement with the concept of image theory - which suggests a canceling effect of electric current on the horizontal dipole-image pair as shown in Figure 1.2.

The degree of impedance mismatch between the line feeding the antenna and the impedance at the input of the antenna is measure by the Voltage Standing Wave Ratio (VSWR). The
The magnitude of reflection coefficient $\Gamma$ is given by

$$|\Gamma| = \frac{VSWR - 1}{VSWR + 1} \quad (1.1)$$

The return loss (RL) is defined as the amount of power reflected back to the source measured in dB relative to the incident power.

$$RL(dB) = -20 \log_{10}|\Gamma| \quad (1.2)$$

Since VSWR and $\Gamma$ are related, it is important to specify the characteristic impedance with respect to which the VSWR is calculated while stating it. Throughout this work, the VSWR is specified at a characteristic impedance of $50\Omega$ unless specified otherwise. A VSWR of 2 corresponds to a RL of 9.55 dB and is taken as the minimum acceptable return loss throughout this work. A high impedance mismatch results in a high VSWR. As a result of the $\lambda/2$ dipole close to the ground having a low radiation resistance, it cannot be matched to any typical transmission line having $50\Omega, 75\Omega, 100\Omega$ impedance and yields high VSWR and return loss i.e. it reflects most of the power rather than radiating it.
Figure 1.3: Variation of impedance of horizontal and vertical $\lambda/2$ Dipoles above PEC Ground as obtained using FEKO simulations, where the height $h$ is the height of the dipole's feed point from the ground.
1.2 Broadband Dipoles and Monopoles

All practical wire dipoles have a finite radius and are not infinitesimally thin. For antennas to be practically useful, they must have a suitable and significant bandwidth of frequencies over with they can be used effectively.

Antenna bandwidth can be defined as the range of frequencies over which the antenna can be utilized for a particular application. It can be defined in terms of impedance - called impedance bandwidth or with respect to the maximum gain/directivity - called pattern bandwidth. The impedance bandwidth is defined as the range of frequencies over which the antenna operates within an acceptable VSWR (2:1). Input impedance and radiation efficiency of the antenna are related to the impedance bandwidth. Pattern bandwidth is related to the maximum directivity, side-lobe levels, polarization, beam-squint, beamwidth etc. Typically it can be defined as the range of frequencies over which the gain does not drop below 3dB of the maximum gain.

A typical design process involves determination of antenna length (approximately 0.475λ if dipole and 0.238λ for monopoles) for resonance and then adjustment of the radius to meet bandwidth requirements. In general, thicker antennas have wider bandwidth than their infinitesimally thin equivalents as this reduces the sensitivity of the reactance of the dipole with respect to frequency. This follows directly from the size, gain and bandwidth relation as described in [2] - which suggests that the bandwidth can be increased by more effectively utilizing the space inside the smallest sphere enclosing the antenna. The thin wire dipole is one dimensional and hence has the worst utilization of the space and therefore poor bandwidth. On the other hand [2] and [3] suggest efficiency for a lossy antenna decreases as the size (or utilization of the sphere) increases.

A conformal alternative to the wire or cylindrical dipole is the printed dipole or microstrip dipole (backed by a ground plane). As suggested in [4] the parallel between cylindrical dipole and printed/microstrip dipoles can be built using the following assumptions :-
1. The cylindrical dipole of radius $r_e$ and a microstrip dipole of width $w$ are equivalent if

$$r_e = w/4$$  \hspace{1cm} (1.3)$$

$r_e$ is called the equivalent radius \(^\ast\) of the microstrip dipole.

2. The length of the printed dipole $L$ must be adjusted for dielectric loading and open end effects such that

$$L = L_e - 2l_{oc}$$ \hspace{1cm} (1.4)$$

where $L_e$ is the effective length of the dipole adjusted for the dielectric loading by a dielectric of effective relative permittivity $\epsilon_{re}$, and $l_{oc}$ is the open end effect correction.

An approximate expression for percentage bandwidth for these dipoles is given by [1]

$$BW = \frac{1}{\sqrt{2}} \frac{w h}{\lambda \lambda 5 \sqrt{\epsilon_{re}}}$$ \hspace{1cm} (1.5)$$

where $w$ is the width of the strip, $h$ is the height of the strip above the ground plane and $\epsilon_{re}$ is the effective relative permittivity. The bandwidth increase with an increase in height above the ground and with an increase in the width of the strips.

A variety of geometries yield better bandwidth than the thin wire dipole or monopole. These include biconical antennas, conical skirt monopoles, discones etc. - as extensions of the cylindrical geometry and bow-tie antennas for the planar geometry. The following are some closely related examples of antenna having wider bandwidth of relevance to this work:-

**1.2.1 Sleeve Antenna**

The more common form of the sleeve antenna is the sleeve monopole, which consists of a monopole antenna fed by a coaxial cable. Just like the feed for a regular wire monopole,

\[^\ast\]A table of some geometries and their equivalent cylindrical conductor radii is given in [1], pp. 456-457.
the inner conductor of the coaxial cable feeds the antenna. However, unlike the feed of a regular monopole the sleeve (outer conductor) connects to the ground and is extended along the axis making the sleeve a part of the radiating element of the antenna to provide the broadband characteristics of the antenna. The length of this extension and position of the feed point allows for the variation of feed point characteristics like impedance. The inner surface of the sleeve acts as the return path for the current while the outer surface acts like a radiating element.

A detailed analysis of this antenna using image theory and approximating the sleeve dipole as a superposition of two asymmetrically driven antennas is provided by R. King in [5]. In the analysis, the sleeve monopole and its image form a doubly-fed sleeve dipole as shown in Figure 1.4. Using the linearity of Maxwell’s equations, King obtains expressions for the impedance by superimposing the impedance obtained by exciting the antenna with one source at a time. The analysis and results correspond to an antenna of total length $l + s$ such that $l = \lambda/4$ and $s = \lambda/8$.

![Figure 1.4: Sleeve Dipole and its equivalents - First as a doubly excited antenna and then as a superposition of two asymmetrically excited antennas](image)

Typically, the overall length $l + s$ is chosen to be $\lambda/4$ corresponding to the first resonance and a ratio of $l/s$ is adjusted to approximately 2.25 as described in [6] and [7] to provide near constant radiation pattern over a 4:1 bandwidth.
1.2.2 Open Sleeve Dipole

A modification of the sleeve monopole/dipole is the open sleeve dipole. This consists of a dipole antenna fed by coaxial feed and a coaxial balun arrangement. Two or more (typically two) parasitic elements on either side of the dipole act as simulated sleeves as shown in Figure 1.5. This results in a broadband/multiband performance as compared to a simple wire dipole. The parasitic elements add an additional resonance which in combination to the resonance of the feed gives rise to the broadband performance.

Literature deals with thick-wire (radius = $\lambda/50$ or more) antennas and report more significant increases in bandwidth or a multi-band performance. Method of Moments based simulations by Gan and Li [8] suggests that for such antennas the first and third resonant frequencies are determined by the driven element while the second resonance depends on the length and separation of the parasitic elements. They numerically established that the sum of the lengths of the parasitic elements and their separation was approximately the wavelength at the second resonance. Their work also deals with antenna having four and eight parasitic elements showing increasing bandwidth 4:1 bandwidth corresponding to a VSWR of 3.5 (return loss = 5 dB). The antenna was studied as a multi-band antenna in another study by the same authors [9]. Use of a thicker wire leads to smaller driven elements and less sensitivity of the reactance to frequency.

Experimental studies on such a cylindrical open sleeve configuration in front of a flat metal reflector in the 225-400 MHz domain reported a 1.8:1 bandwidth in comparison to a 1.25:1 bandwidth for a conventional dipole with the same diameter [10]. The antenna is fed by a folded balun. The results hold true even in the absence of the ground plane. These measurements were reported at a dipole reflector spacing $0.29\lambda$ at the highest frequency and $0.16\lambda$ at the lowest frequency. The radius in this case varied from approximately $\lambda/160$ to $\lambda/20$ at the highest frequency. The driven element had a taper towards the feed - the angle of which was found to be "not critical". The significance of this taper is reported in [11]. The taper can be used to negate the capacitive impedance at the feed and is an
Figure 1.5: Open Sleeve Dipole and Folded Balun Feed
<table>
<thead>
<tr>
<th>Element</th>
<th>Variable</th>
<th>Dimension (λ)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Length of Driven Element</td>
<td>lf</td>
<td>0.415</td>
</tr>
<tr>
<td>Length of Parasitic Element 1</td>
<td>lp1</td>
<td>0.230</td>
</tr>
<tr>
<td>Length of Parasitic Element 2</td>
<td>lp2</td>
<td>0.230</td>
</tr>
<tr>
<td>Separation between Driven and Parasitic Element 1</td>
<td>ls1</td>
<td>0.0381</td>
</tr>
<tr>
<td>Separation between Driven and Parasitic Element 2</td>
<td>ls2</td>
<td>0.0381</td>
</tr>
<tr>
<td>Radius</td>
<td>a</td>
<td>0.0102</td>
</tr>
<tr>
<td>Taper</td>
<td>tp</td>
<td>0.050</td>
</tr>
</tbody>
</table>

Table 1.1: Geometrical details of cylindrical dipole and open sleeve dipole antenna configurations as simulated in FEKO. The result of the simulation is shown in Figure 1.6.

important parameter while designing the antenna.

Figure 1.6 compares bandwidths of various cylindrical dipole and open sleeve dipole geometries as simulated on FEKO. The geometric details for the antenna configurations is listed in Table 1.1. The simulation shows that open sleeve dipole with taper provides over 300% increase in bandwidth over that of a simple cylindrical (thick) dipole for a return loss of greater than 9.55 dB. The results are summarized in Table 1.2. The introduction of the second resonant frequency and its effect on the bandwidth is apparent. Changing various parameters listed in the table allows the adjustment of the antenna’s performance. The influence of the taper on the feed point capacitance is also apparent. Introducing the taper yields better impedance match to 50Ω line.

<table>
<thead>
<tr>
<th>Antenna Configuration</th>
<th>Bandwidth</th>
</tr>
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<tbody>
<tr>
<td>Cylindrical Dipole</td>
<td>17.0%</td>
</tr>
<tr>
<td>Cylindrical Open Sleeve Dipole (no taper)</td>
<td>22.4%</td>
</tr>
<tr>
<td>Cylindrical Open Sleeve Dipole with taper</td>
<td>60.6%</td>
</tr>
<tr>
<td>Cylindrical Open Sleeve Dipole with taper in front of a 1λ - diameter PEC ground plane</td>
<td>59.3%</td>
</tr>
</tbody>
</table>

Table 1.2: Bandwidth for Cylindrical Dipole and Open Sleeve Dipole for results shown in Figure 1.6.

A open sleeve dipole can also be approximated using thin wires for each individual thick
Figure 1.6: Comparison of various configurations of Cylindrical Dipoles and Open Sleeve Dipoles. The bandwidth of the open sleeve dipole with taper exceeds that of an ordinary dipole by over 300%. The performance of the open sleeve dipole with sleeve is better than the dipole without taper. The performance of the dipole does not significantly change when it is placed at a distance of $\lambda/4$ from a $1\lambda$ diameter PEC reflector - the bandwidth remains the same while the resonant frequency shifts slightly. The dimensions of the antenna are as in Table 1.1

wire by constructing a "wire cage" given by the formula as explained in [10]

$$d_{eff} = d\left(\frac{nd_0}{d}\right)^{\frac{1}{n}}$$  \hspace{1cm} (1.6)

where $d_{eff}$ is the diameter of the "wire cage", $n$ is the number of wires in the cage $d_0$ is the diameter of the individual thin wires.
1.2.3 Planar Open Sleeve Dipole (POSD)

The geometry of a open sleeve dipole can be made conformal as shown in Figure 1.7a. By suitably adjusting the planar conducting surface dimensions, performance of the antenna with respect to impedance and pattern bandwidth can be adjusted. The general technique for bandwidth enhancement using parasitic sleeves for microstrip dipoles is dealt in [12]. The author suggests embedding the parasitic elements between the transmission line feed and the dipole antenna or co-planar to the transmission line feed for electromagnetically coupled dipoles. The overall thickness of the dipole from the ground plane is about 0.1λ and a VSWR \( \leq 2 \) bandwidth of about 11% is obtained around a frequency of 10 GHz.

Extending the idea of using parasitic elements to enhance bandwidth to open sleeve type antennas as suggested in [13] parallels the design idea for Cylindrical Open sleeve dipoles. The driven element determines the lower resonant frequency while the length of the parasitics and the separation distance between the driven element and the parasitics determine the upper resonant frequency. Thus the second resonance combined with the first resonance imparts a dual-band or broadband response to the antenna.

A comparison of the bandwidth between the planar and cylindrical geometries is shown in Figure 1.8. The planar elements have a width of 0.0625λ for the dipole and 0.0503λ for the parasitics as suggested in [13] while the cylindrical dipole and parasitic elements have an radius of 0.0143λ which is approximately \( \frac{1}{4} \) the average of the two planar widths. The planar geometry is electrically equivalent to a cylindrical shape of radius approximately one-forth the planar width.

A variation of this geometry is the End-loaded Planar Open Sleeve Dipole (ELPOSD). The electrical length of the antenna can be changed by top loading. Top loading effectively adds capacitance to the antenna, thereby lowering the resonant frequency and increasing the effective electrical length (at that frequency) for a given antenna. Conversely the same technique is used to produce uniform current on an antenna (by top loading) so that it resembles a Hertzian dipole. This permits antenna miniaturization at low frequencies.
Optimization of POSD and its miniaturized variant ELPOSD vertical monopoles mounted over a large ground plane is treated in [13]. They achieved a bandwidth of 49% compared to a planar dipole of bandwidth 14% (VSWR \( \leq 2 \)) and a size reduction between the POSD and ELPOSD of about 19%. A further reduction in length was achieved by dielectric loading - mounting these antennas on a thin dielectric substrate of relative permittivity \( \epsilon_r = 2.33 \) and \( \epsilon_r = 6 \).

A parametric and optimization study of ELPOSD using Ansoft HFSS is undertaken in [14] mounted over a substrate of relative permittivity \( \epsilon_r = 10 \) and height of 0.18\( \lambda \) - optimized value - at a frequency band centered around 2 GHz. A design using parameterized values
versus optimized values reveals the following:

1. Increase in ratio of driven element width to the length improve return loss.

2. The spacing between the driven element and sleeve is claimed to be independent.

3. Long but thin end loading stubs are considered to be better for the design.

4. A sleeve width greater than sleeve length provides better bandwidth response.

The objective of our work is to design and optimize the dimensions of open sleeve dipole configurations for reflector-dipole spacings of less than $0.1\lambda$ to yield good match to a $50\Omega$ thus having improved performance as compared to thin wire dipole antenna this close
<table>
<thead>
<tr>
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</tr>
<tr>
<td>Length of Parasitic Element 2</td>
<td>lp2</td>
<td>0.324</td>
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<td>0.05</td>
</tr>
<tr>
<td>Width of Parasitic Element 1</td>
<td>wp2</td>
<td>0.05</td>
</tr>
<tr>
<td>Length of End load</td>
<td>le</td>
<td>NA</td>
</tr>
<tr>
<td>Width of End load</td>
<td>we</td>
<td>NA</td>
</tr>
</tbody>
</table>

Table 1.3: Geometrical details of cylindrical open sleeve dipole and planar open sleeve dipole antenna configurations. The result of the simulation is shown in Figure 1.8. The POSD dimensions are taken from [13].

to the ground. The qualifiers low-profile meaning "close to the ground" and conformal meaning "lying in plane" are used to describe this geometry.
2. Design, Simulation and Measurements

The design guidelines and dimensions as suggested in [8],[10] serve as an appropriate starting point for thin wire open sleeve dipole antennas. The bandwidth of a $\lambda/2$ dipole and an open sleeve dipole both having thin wire elements of radius $0.005\lambda$ is compared in Figure 2.1. The dimensions of the antennas compared are listed in Table 2.1. With thin wires, the bandwidth of the open sleeve dipole is approximately 17% while that of a simple dipole is around 10% - an increase in bandwidth of 70%. Obviously due to reduced radius, it is unreasonable to expect the bandwidth of the open sleeve dipole to exceed 20-25%, unlike Figures 1.8 and 1.6 where the bandwidth was approximately 40% and 56% respectively.

<table>
<thead>
<tr>
<th>Element</th>
<th>Open Sleeve Dipole</th>
<th>Dimension ($\lambda$)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Length of Driven Element</td>
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<td>Length of Parasitic Element 1</td>
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<td>Length of Parasitic Element 2</td>
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<td>Separation between Driven and Parasitic Element 1</td>
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<tr>
<td>Separation between Driven and Parasitic Element 2</td>
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<tr>
<td>$\lambda/2$ Dipole</td>
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<tr>
<td>Length of Dipole</td>
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<tr>
<td>Radius</td>
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Table 2.1: Dimensions of thin wire open sleeve dipoles and $\lambda/2$ dipoles for antennas compared in Figure 2.1
2.1 Low Profile Open Sleeve Dipole using Thin wires

As a horizontal dipole is moved closer to the ground, its radiation resistance decreases and it is difficult to obtain a good VSWR match to conventional values of characteristic impedance. A resonant horizontal dipole 0.1λ from the ground has a radiation resistance of 20.3Ω and it decreases as we move closer to the ground to 5.6Ω at 0.05λ from the ground [15]. An attempt to increase the radiation resistance of the antenna to 50Ω in a small bandwidth about the design frequency using parasitic sleeves is the goal of the design.

An initial study on thin wire open sleeve dipole near PEC ground was conducted in [15]. OPTFEKO and NECOPT was used to optimize the lengths of thin wire open sleeve dipoles close to the ground. Figure 2.2 shows the geometry of the open sleeve dipole. The lengths of the dipole, lengths of the parasitics and the separation between them were used as
parameters in the optimization process while the antenna’s height above the PEC ground plane was fixed at 0.05λ and 0.1λ above a 1λ × 1λ PEC ground plane. An independent optimization was carried out at an height of 0.075λ. The motivation here was to establish a second resonant frequency (due to the parasitic elements) very close to the resonant frequency of the driven dipole by choosing parasitic elements of approximately the same length of the dipoles - thus giving a broadband response. The resulting dimensions of all three optimization processes are listed in Table 2.2.

Figure 2.2: Geometry of thin wire open sleeve dipole above 1λ × 1λ PEC ground plane.

Figure 2.3 gives us a sense of the impedance bandwidth showing the VSWR as a function of frequency resulting from the optimization. The bandwidth was found to be about 3% , 9.13% and 11.6% for heights of 0.05λ, 0.075λ and 0.1λ respectively.

The radiation pattern for the three heights at the design frequency is shown in figure in 2.4. It is a broadside radiator with a maximum directivity of approximately 9.1 dBi at all three heights - 0.05λ, 0.075λ and 0.1λ - above the ground plane. The beam squint
Table 2.2: Optimized antenna dimensions for thin wire open sleeve dipoles at three heights - 0.05λ, 0.075λ and 0.1λ - above the PEC ground plane. All dimensions are in terms of the freespace wavelength $\lambda$.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Height (h)</th>
<th>0.05λ</th>
<th>0.075λ</th>
<th>0.1λ</th>
</tr>
</thead>
<tbody>
<tr>
<td>lf</td>
<td></td>
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<td>0.4926</td>
<td>0.4670</td>
</tr>
<tr>
<td>lp1</td>
<td></td>
<td>0.4516</td>
<td>0.4400</td>
<td>0.4204</td>
</tr>
<tr>
<td>lp2</td>
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<td>ls1</td>
<td></td>
<td>0.07323</td>
<td>0.09372</td>
<td>0.1602</td>
</tr>
<tr>
<td>ls2</td>
<td></td>
<td>0.04179</td>
<td>0.04024</td>
<td>0.1053</td>
</tr>
<tr>
<td>a</td>
<td></td>
<td>0.005</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

Figure 2.3: Comparison of bandwidth (using VSWR) for thin wire open sleeve dipoles at three heights - 0.05λ, 0.075λ and 0.1λ - above a 1λ × 1λ PEC ground plane using dimensions as specified in Table 2.2.
Figure 2.4

(a) 0.05λ

(b) 0.075λ
Figure 2.4: Radiation Pattern (Figures (a), (b), (c)) in the XZ and YZ planes for the thin wire open sleeve dipoles at a height of $0.05\lambda$, $0.075\lambda$, $0.1\lambda$ above a $1\lambda \times 1\lambda$ PEC ground plane.

Figure 2.5: Variation of directivity with frequency for the thin open sleeve dipole above at three heights above ground plane.
is less than 10° within the frequency band. The total directivity remains fairly constant throughout the impedance bandwidth of each antenna. The pattern is symmetric in the XZ plane but asymmetric in the YZ plane - which is consistent with the geometry. The axial ratio\(^\ast\) is close to 0 - thus indicating a linear polarization.

### 2.2 Low Profile Open Sleeve Dipole Using Planar Elements

Next, in order to develop a conformal design (Figure 2.6), the aforementioned equivalent areas relation was used to establish the widths of the planar open sleeve dipole at the three heights above the 1λ × 1λ PEC ground plane. The equivalent width of the strip was fixed to be 0.2λ (4× radius of the wire). The feed was modeled as using a couple of wire segments connecting to a tapered dipole element. The taper was fixed at 120° for all cases. The lengths from the wire geometry were used as the starting point for another optimization - using OPTFEKO - to establish dimensions for the conformal structure. The optimization yielded lengths that were only slightly different from the corresponding wire structures and are listed in Table 2.3.

The performance of the planar design is shown in Figures 2.9, 2.7.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Height</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>0.05λ</td>
</tr>
<tr>
<td>If</td>
<td>0.4783</td>
</tr>
<tr>
<td>lp1</td>
<td>0.4559</td>
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<tr>
<td>lp2</td>
<td>0.4575</td>
</tr>
<tr>
<td>ls1</td>
<td>0.07262</td>
</tr>
<tr>
<td>ls2</td>
<td>0.04887</td>
</tr>
<tr>
<td>tp</td>
<td>0.02</td>
</tr>
</tbody>
</table>

Table 2.3: Table showing the values of the parameters for the planar open sleeve dipole at three heights above a 1λ × 1λ PEC ground plane optimized to maximize impedance bandwidth.

\(^\ast\)The axial ratio (AR) here is defined as ratio of the minor axis to the major axis of the ellipse traced by the electric field vector as a wave propagates. Therefore 0 ≤ |AR| ≤ 1. This is the inverse of the definition found in [1].
Figure 2.6: Open sleeve dipole using planar elements above a $1\lambda \times 1\lambda$ PEC ground plane
Figure 2.7

(a) $0.05\lambda$

(b) $0.075\lambda$
Figure 2.7: Radiation Pattern (Figures (a), (b), (c)) in the XZ and YZ planes for the planar open sleeve dipoles at the a height of $0.05\lambda$, $0.075\lambda$, $0.1\lambda$ above a $1\lambda \times 1\lambda$ PEC ground plane.

(c) $0.1\lambda$

Figure 2.8: Variation of directivity with frequency for the planar open sleeve dipole above at three heights above $1\lambda \times 1\lambda$ PEC ground plane.
Figure 2.9: Comparison of bandwidth (using VSWR) for planar open sleeve dipoles at three heights - 0.05λ, 0.075λ and 0.1λ - above a 1λ × 1λ PEC ground plane using dimensions as specified in Table 2.3.
The results of the simulation indicate close resemblance to those obtained with thin wires, providing further validation for the equivalent areas approach. Figure 2.9 shows a bandwidth of 3%, 8.0% and 14.1% for dipole-ground height of 0.05λ, 0.075λ and 0.1λ respectively. The radiation patterns for these cases is shown in Figure 2.7. The radiation patterns indicate a maximum broadside gain of approximately 9.1 dBi in all three cases. The introduction of thin planar elements keeps the polarization the same and does not worsen the axial ratio. Thus this approach suggests a way to develop conformal linearly polarized antennas with very low profile and yet have sufficient bandwidth for the application.

2.3 Dielectric Loading - I

![Figure 2.10: Geometry for a dielectric loaded planar open sleeve dipole.](image)

Next, in order to reduce the size of the antenna, dielectric loading was attempted. The dielectric substrate is backed by a 1λ × 1λ PEC ground plane on one side while the planar
radiating elements are mounted on the other. A CADFEKO model was simulated using a
dielectric of relative permittivity $\epsilon_r = 2$. The thickness of the planar strips was kept the
same while the length of the dipoles was approximately reduced by a factor of $\frac{1}{\sqrt{\epsilon_r}} = 0.707$.
The optimized lengths and separations from Table 2.3 multiplied by 0.707 were used as
a starting point for the optimization. In addition to reducing the size of the antenna
the dielectric backing provides mechanical support to the planar elements. The use of the
dielectric also increases the electrical distance between the radiating elements and the PEC
ground plane.

The results of this optimization are shown in Table 2.4.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Height</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
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<td>$l_f$</td>
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<tr>
<td>$l_{p1}$</td>
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<tr>
<td>$l_{p2}$</td>
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<tr>
<td>$l_{s1}$</td>
<td>0.091807</td>
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<tr>
<td>$l_{s2}$</td>
<td>0.05442</td>
</tr>
<tr>
<td>$t_p$</td>
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</tr>
</tbody>
</table>

Table 2.4: Table showing the values of the parameters for the dielectric loaded ($\epsilon_r = 2$)
planar open sleeve dipole for three different thickness of a substrate backed by a
$1\lambda \times 1\lambda$ PEC ground plane and optimized to maximize impedance bandwidth.

Figures 2.11, 2.12, 2.13 show the performance of the dielectric loaded antenna. The maxi-
mum directivity at the central frequency is 8.1 dBi, 7.9 dBi, 8.1 dBi for substrate thickness
of 0.05\lambda, 0.075\lambda and 0.1\lambda and the beam squint is less than 10° within the impedance
bandwidth. The impedance bandwidth is approximately 4.5%, 11% and 18.1% for die-
electric thickness of 0.05\lambda, 0.075\lambda and 0.1\lambda respectively. The increase in bandwidth is a
result of the increased electrical separation from the PEC ground plane. The decrease in
directivity is expected owing to the smaller size of the antenna.
Figure 2.11: Comparison of bandwidth (using VSWR) for dielectric loaded ($\epsilon_r = 2$) planar open sleeve dipole for three different thickness - $0.05\lambda$, $0.075\lambda$ and $0.1\lambda$ - of a substrate backed by a $1\lambda \times 1\lambda$ PEC ground plane using dimensions as specified in Table 2.4.
(a) Radiation Pattern for dielectric ($\epsilon_r = 2$) thickness of 0.05$\lambda$

(b) Radiation Pattern for dielectric ($\epsilon_r = 2$) thickness of 0.075$\lambda$

Figure 2.12
(c) Radiation Pattern for dielectric ($\epsilon_r = 2$) thickness of $0.1\lambda$

Figure 2.12: Radiation Pattern for dielectric loaded ($\epsilon_r = 2$) planar open sleeve dipole for three different thickness - $0.05\lambda$, $0.075\lambda$ and $0.1\lambda$ - of a substrate backed by a $1\lambda \times 1\lambda$ PEC ground plane.
2.4 Dielectric Loading - II

Next, in order to further reduce the size of the antenna a CADFEKO model was created using a dielectric substrate of relative permittivity $\epsilon_r = 4.3$. Again the substrate was backed on one side by a $1\lambda \times 1\lambda$ PEC ground plane and the radiating elements were mounted on the other side. The thickness of the planar strips was kept the same while the length of the dipoles was approximately reduced by a factor of $\frac{1}{\sqrt{\epsilon_r}} = 0.482$. The optimized lengths and separations from Table 2.3 multiplied by 0.482 were used as a starting point for this optimization. In addition to reducing the size of the antenna the dielectric backing also provides mechanical support to the planar elements.

The results of this optimization are shown in Table 2.5.

Figures 2.14, 2.15, 2.16 show the performance of the FR4 dielectric loaded antenna. The peak directivity at the central frequency is about 6.6 dBi, 5.9 dBi and 5.9 dBi for substrate...
Figure 2.14: Comparison of bandwidth (using VSWR) for dielectric loaded $\varepsilon_r = 4.3$ planar open sleeve dipoles for three different thickness of dielectric substrate - 0.05$\lambda$, 0.075$\lambda$ and 0.1$\lambda$ - backed by a $1\lambda \times 1\lambda$ PEC ground plane using dimensions as specified in Table 2.5.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>0.05$\lambda$</th>
<th>0.075$\lambda$</th>
<th>0.1$\lambda$</th>
</tr>
</thead>
<tbody>
<tr>
<td>$l_f$</td>
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<td>0.1970</td>
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</tr>
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<td>$l_{p1}$</td>
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<td>$l_{s2}$</td>
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<td>$t_p$</td>
<td></td>
<td>0.02</td>
<td></td>
</tr>
</tbody>
</table>

Table 2.5: Table showing the values of the parameters for the planar open sleeve dipole for three thicknesses of the substrate above a $1\lambda \times 1\lambda$ PEC ground plane optimized to maximize impedance bandwidth.
Figure 2.15

(a) Radiation Pattern for dielectric ($\epsilon_r = 4.3$) thickness of $0.05\lambda$

(b) Radiation Pattern for dielectric ($\epsilon_r = 4.3$) thickness of $0.075\lambda$

Figure 2.15
(c) Radiation Pattern for dielectric \( (\epsilon_r = 4.3) \) thickness of 0.1\( \lambda \)

Figure 2.15: Radiation Pattern for dielectric loaded \( (\epsilon_r = 4.3) \) planar open sleeve dipole for three different thickness - 0.05\( \lambda \), 0.075\( \lambda \) and 0.1\( \lambda \) - of a substrate backed by a 1\( \lambda \times 1\lambda \) PEC ground plane.
thickness of 0.05λ, 0.075λ and 0.1λ respectively at the design frequency. The bandwidth is about 6.5%, 9.8% and 24.5% for substrate thickness of 0.05λ, 0.075λ and 0.1λ respectively. The increased dielectric constant provides an additional 2% and 7% bandwidth at 0.075λ and 0.1λ dielectric thickness. Again this is a direct consequence of the increased electrical separation between the PEC ground plane and the radiating elements. The slightly reduced gain can be attributed once again to the reduced size of the antenna.

### 2.5 Modeling, Simulation and Optimization Setup in FEKO

FEKO Suite 5.4 and FEKO Suite 5.5 \[16\] is essentially a 3D-full wave Method of Moments based electromagnetic simulator. Also incorporated in the package are asymptotic methods (GTD/UTD) and the fast multi-pole Method of Moments (FMM) - for analyzing electrically large problems. CADFEKO was used to model all the cases analyzed. The following consideration were made while modeling meshing and optimization :-
1. The optimization process was run with several goals:

   • Minimize VSWR at design/central frequency.
   • Minimize VSWR at several frequencies in a certain bandwidth.
   • Minimize the maximum VSWR in the bandwidth.
   • Simultaneously minimize (with equal weight) the maximum and the minimum VSWR in the specified frequency band.

2. Of these optimizations, the best case - which was not necessarily the same for all - was selected.

3. Occasionally the lengths were manually adjusted to see if a dual band response could be converted to a broadband response by adjusting parasitic element lengths and separations.

4. The feed for the FEKO models was a balanced wire feed. A wire port was attached to a 5 segment piece of wire that formed an inverted-U shaped structure above the substrate as shown in Figure 2.17.

5. Simplex Optimization routine was employed for the optimization. The tendency of this method to converge to a local optimum makes it imperative to judiciously choose a starting point.

6. A more global optimization routine would be more time consuming but it might converge to a better global optimum.

7. Power Balance check was used to verify accuracy of the results.

8. It was observed that using MoM and the surface equivalence principle for the dielectric region yielded poor results on the power balance check. FEM mesh was then adopted for simulations with dielectric.

9. Electrical symmetry in the YZ plane with reference to Figure 2.10 is exploited to reduce problem size and run-time.
Figure 2.17: Modeling of the antenna feed.
2.6 Construction and Measurements

The antennas were constructed to function at 2.4 GHz corresponding to $\lambda = 12.5\,cm$. It was constructed on readily available plastic dielectrics - Teflon $\varepsilon_r = 2$ and G10FR4 $\varepsilon_r = 4.3$. The dielectrics were available with thickness of $0.0508\lambda$, $0.0762\lambda$ and $0.1016\lambda$ - the fractional changes in the thickness did not significantly affect the simulation results. The dielectric was backed by aluminium foil to act as a ground plane. The antennas were modeled in the simulation as lossless antennas. An unbalanced feed using RG-174 coaxial cable was used with BNC connectors to make connections. A string of ferrite beads was used on the cable to impede the unbalanced current on the outer conductor of the cable. The radiating elements were constructed using 63 mils thick copper tape with conductive adhesive and cut to a width of $2.5\,mm\,0.02\lambda$. Figure 2.18 shows the constructed antennas on Teflon and FR4 substrates respectively.

Figure 2.19 shows the VSWR measurements for antennas on a Teflon substrate.
Figure 2.18: Sample constructed antennas with showing RG174 cable feed, BNC connectors and ferrite bead baluns.
(a) VSWR for Teflon dielectric thickness of 0.0508λ

(b) VSWR for Teflon dielectric thickness 0.0762λ

Figure 2.19
(c) VSWR for Teflon dielectric thickness $0.1016\lambda$

Figure 2.19: Comparison of measured and simulated VSWR for $\epsilon_r = 2$ - Teflon

Figure 2.20 shows the VSWR measurements for antennas on the G10-FR4 substrate.
(a) VSWR for G10 FR4 dielectric thickness 0.0508\(\lambda\)

(b) VSWR for G10 FR4 dielectric thickness 0.0762\(\lambda\)

Figure 2.20
(c) VSWR for G10 FR4 dielectric thickness 0.1016λ

Figure 2.20: Comparison of measured and simulated VSWR for $\epsilon_r = 4.3$ - G10FR4
3. Conclusion and Future Work

By having shown that the open sleeve design can be used to create conformal low-profile antennas, several improvements and modifications can be made to the basic design. One approach would be to use strips of greater thickness (0.04λ or more). Existing literature [14], [3] suggests using thicker strips for increasing bandwidth. Another modification that can be considered is miniaturization by end loading. As suggested in [13] end-loaded antennas while being compact maintain their broadband response. Existing optimization techniques can be employed to optimize these designs at the required heights above the ground plane.

A further improvement to the design can be made in developing a circularly polarized antenna by combining ideas found in [12] and [10]. It was suggested in [12] that parasitic elements embedded within the substrate also enhance the bandwidth of the planar dipole. A planar crossed dipole design can be mounted on the surface of a dielectric with the parasitic elements embedded within the substrate. Figure 3.1 shows one such possibility. Once again the lengths, separations and the depth of the embedded parasitic elements must be optimized. The antenna can have an electromagnetically coupled feed as opposed to a coaxial feed. Doing so would facilitate easier feed design for the two arms of the crossed dipole that must be fed in phase quadrature.

More recent developments in the field of artificial High Impedance Electromagnetic Surfaces (HIS), Frequency Selective Surfaces (FSS) etc. have led to the development of novel materials that permit development of low profile antennas [17]. The electric conductor
Figure 3.1: Planar Open Sleeve Crossed Dipole showing surface and embedded elements
introduces currents of opposite phase that inherently cause problems for low profile antennas. The use of artificial magnetic conductors would induce currents in phase in the image, and hence antennas can be constructed still closer to the ground. However, the behavior of these HIS and FSS is restricted to a very narrow frequency bandwidth and hence may not be suitable for very broadband applications.

The choice of optimization also affects design and run-time. The choice for using Nelder Mead Simplex optimization was based on the assumption of being able to obtain good starting conditions and the need for quicker convergence. Global optimization methods like GA or PSO may be used to trade-off time required for convergence of the optimization for improved performance of the antenna.
Bibliography


