Chapter 6

Shared Aperture Design Using Two Element Dual Frequency Microstrip Radiators
6.1 INTRODUCTION

Microstrip antennas are popularly used in communication systems for their low profile and ease of conformability on a curved structure. But its disadvantage is the inherent narrow bandwidth. Frequently, the communication links use both transmission and reception frequencies. Apparently, this demands for a broadband antenna or two different antennas. Broadband antenna is not a preferred choice in order to avoid interference at the intermediate frequencies. Use of two different antennas increases cost and space. A suitable alternative is the use of dual frequency operation of an antenna.

Dual frequency antennas are known for a long time [1]–[16]. A good overview is presented in [1]. In an orthogonal mode dual-frequency patch [2], [3], the orthogonal resonances (TM$_{10}$ and TM$_{01}$) of a single patch are employed. The works in [4], [5], used dual feed and orthogonal slot aperture coupling.

Electromagnetically coupled stacked patches were employed in [6]–[8], where multilayer patches were stacked to produce dual frequency or broadband response. Dual band was achieved using aperture coupling with stacked patches in [9]. About 20% bandwidth for the lower frequency band and 3% bandwidth at the upper band was demonstrated. Multi-frequency antenna comprised of unequal parallel dipoles which were aperture coupled to a microstrip line was reported in [10]. The dimensions of the resonator can be adjusted to produce staggered tuned resonances or an overlapping broadband response. Dual frequency antenna was reported (using single microstrip layer) with resonators of different dimensions [11].

Another popular technique for obtaining dual-frequency behavior is to introduce reactive loading to a single patch [12]–[16]. The reactive-loading was investigated in [12], where a microstrip stub was used. Dual frequency antenna with a high value of frequency ratio (4-5) was reported using two symmetrically positioned varactor diodes between the patch and the ground plane [13]. Slot-loaded patches were investigated in [14]–[16], where two narrow slots were etched close to and parallel to the non-radiating edge of a rectangular patch.
The designs in [1]–[16], used conventional patch geometry. Here, we have explored a new technique. Two identical polarizations at two far apart frequencies have been generated in a composite configuration which uses two elements fed by a single feed line. This design technique provides an independent control to the desired frequencies and reduces element width by 50% at the cost of marginal decrease in gain (0.9dB).

6.2 The Design

The objective of the study is to realize a dual frequency microstrip antenna at C-band. A new shared aperture design has been conceived to produce two frequency bands around 3.93 GHz and 4.85 GHz. Proximity coupled feeding technique has been employed.

The development starts with a conventional patch design at the lower resonance frequency, say \( f \). It consists of two substrate layers as shown in Fig. 6.1. The top layer contains the rectangular patch. No metal is in the back layer. The resonator is proximity coupled by a 50\( \Omega \) microstrip line. 31mil thick RT Duroid 5880 (\( \varepsilon_r = 2.2 \), \( \tan \delta = 0.001 \)) has been considered for this design.

The initial dimensions of a rectangular patch are estimated from the design equations based on transmission-line model analysis. For dominant TM\(_{10}\) mode, the resonance frequency is a function of length [17]

\[
L = \frac{c}{2f \sqrt{\varepsilon_{\text{eff}}}} - 2 \cdot \Delta L
\]  

(6.1)

where \( c \) represents the velocity of light, \( \varepsilon_{\text{eff}} \) is the effective permittivity of the substrate, and \( f \) is the operating frequency. \( \Delta L \) is given by an approximate relation [17]

\[
\Delta L = 0.412 \cdot h \frac{(\varepsilon_{\text{eff}} + 0.3) \cdot \left( \frac{W}{h} + 0.264 \right)}{(\varepsilon_{\text{eff}} - 0.258) \cdot \left( \frac{W}{h} + 0.8 \right)}
\]

(6.2)

where \( W \) represents the patch width and \( h \) is the substrate height. In this design we have
considered \( h_1 = h_2 = h \). The effective dielectric constant \((\varepsilon_{\text{eff}})\) is enumerated from the relation [17]

\[
\varepsilon_{\text{eff}} = \left( \frac{\varepsilon_r + 1}{2} \right) + \left( \frac{\varepsilon_r - 1}{2} \right) \cdot \left[ 1 + 12 \frac{h}{W} \right]^{-1/2}
\]

(6.3)

where \( \varepsilon_r \) is the dielectric permittivity of the substrate.

Using these relations, the length of the patch is found to be \( L = 25 \text{mm} \). We have considered the patch width as \( W = 1.2L \), which gives acceptable gain, bandwidth and cross polar discrimination.

To improvise the concept of shared aperture dual frequency operation we have: (i) reduced the patch width by 50\%, and (ii) have introduced a 2nd patch which is easily accommodated in the same space. In this composite configuration, both elements are fed by a single microstrip feed line. The geometry is shown in Fig. 6.2. Their resonance frequencies are:

1. \( f_1 \approx f \) for \( L_1 \approx L \)
2. \( f_2 > f_1 \) for \( L_2 < L_1 \)
Fig. 6.2. Geometry of the proposed dual frequency antenna. The footprint of the basic patch in Fig. 6.1 is shown by the red dashed line in Fig. 6.2(a). An enlarged view of the central portion of the proposed antenna is shown in Fig. 6.2(b).

The patch offsets \(D_1\) and \(D_2\), widths \(W_1\) and \(W_2\), and depth of feed \(D_F\), as shown in Fig. 6.2, have been optimized to achieve desired radiation performances at the target frequencies. The design optimizations have been carried out through successive simulations using [18].
6.2.1 Effect of Patch Offsets ($D_1, D_2$) on Matching

Initially, the individual patch offsets, determined as $D_1$ (larger patch) and $D_2$ (smaller patch), are optimized independently for each element in absence of the other patch. The variation of the reflection coefficient with normalized patch offset is shown in Fig. 6.3(a). As the patch offset is increased, the matching improves monotonically up to certain value at which the best matching is achieved. Beyond that the matching degrades monotonically. These near optimum patch offsets are $D_1 = 0.009 \lambda_1$ and $D_2 = 0.0035 \lambda_2$. The final offsets are optimized in presence of both the patches. The final offsets are $D_1 = 0.00916 \lambda_1$ and $D_2 = 0.00483 \lambda_2$.

6.2.2 Effect of Patch Width ($W_1, W_2$) on Matching

The patch widths ($W_1, W_2$) have a significant role on the impedance matching of the antenna. The simulated reflection coefficients as a function of the patch width are plotted in Fig. 6.3(b). The matching improves monotonically up to a certain patch width and then degrades. The best matching is obtained for $W_1 = 0.17 \lambda_1$ and $W_2 = 0.217 \lambda_2$. While varying the widths, the patch offsets are kept unchanged at their pre-optimized values ($D_1 = 0.00916 \lambda_1$ and $D_2 = 0.00483 \lambda_2$).

6.2.3 Effect of Feed Depth ($D_F$) on Matching

Having optimized the offsets and widths, the depth of the feed line beneath the patches has been varied. The impedance matching has a significant dependency on the feed depth ($D_F$) as shown in Fig. 6.4. It is observed that $D_F > 0$ is not a suitable choice; rather $D_F < 0$ is preferred from matching point of view. We have used $D_F = -0.005 \lambda_1 = -0.005 \lambda_2$ for our purpose.

6.2.4 Effect of Patch Width ($W_1, W_2$) on Resonance

The patch width also affects the resonance frequency. The variation with patch widths is shown Fig. 6.5. The resonance frequency monotonically decreases as $W_1$ and $W_2$ increases.
Fig. 6.3. Simulated reflection coefficients as a function of (a) patch offset, and (b) patch width. $L_1 = 25\text{mm}$, and $L_2 = 20\text{mm}$.
Fig. 6.4. Variation of reflection coefficient as a function of depth of the feed line. $L_1 = 25$mm, $L_2 = 20$mm, $D_1 = 0.00916\lambda_1$, $D_2 = 0.00483\lambda_2$, $W_1 = 0.17\lambda_1$, and $W_2 = 0.217\lambda_2$.

Fig. 6.5. Variation of resonance frequency as a function of element width. $L_1 = 25$mm, $L_2 = 20$mm, $D_1 = 0.00916\lambda_1$ and $D_2 = 0.00483\lambda_2$. 
Using these data, and curve fitting technique, an empirical relation has been proposed to predict the resonance frequency as

\[
f = \frac{c}{2 \cdot (L + 2 \cdot \Delta L) \cdot \sqrt{\varepsilon_{\text{eff}}}} \cdot \frac{1}{2} \left( \sqrt{\varepsilon_{\text{eff}}} \cdot 0.454 \cdot \left( \frac{L}{W} \right)^{1/6} \right)
\]

(6.4)

The relation is found to be valid for \(0.1 \lambda \leq W \leq 0.4 \lambda\). The relation predicts resonance frequency with better than \(\pm 0.6\%\) accuracy.

**6.3 Simulation with Optimized Patch Dimensions**

In this section, a comparison between the basic antenna (full width, Fig. 6.1) and the two-element shared aperture design (Fig. 6.2) has been examined. The simulated reflection coefficients are shown in Fig. 6.6. The basic full width antenna resonates at 3.83GHz. The bandwidth \((S_{11} \leq -10 \text{ dB})\) is \(\approx 90\text{MHz}\) (2.35%). The two-element design exhibits dual resonances at 3.93GHz and 4.85GHz. The impedance bandwidths of the proposed antenna are \(\approx 55\text{MHz}\) (3.90-3.955 GHz, \(\approx 1.4\%\)) at the lower resonance and \(\approx 55\text{MHz}\) (4.822-4.877 GHz, \(\approx 1.2\%\)) at the higher resonance frequency.

The gain comparison is shown in Fig. 6.7. The basic antenna shows a gain = 7.1dBi, whereas the proposed design yields 6.8dBi gain at 3.93GHz and 6.2dBi at 4.85GHz. A marginal decrease by 0.3dBi at the lower resonance and 0.9dBi at the higher resonance are documented. Fig. 6.7 shows a positive gain over 3.73-4.04GHz (\(\approx 310\text{MHz}\)) and 4.70-5.03GHz (\(\approx 330\text{MHz}\)).

In Fig. 6.8, the H-plane patterns are compared. The co-polar patterns are almost similar. The peak power transmission is \(\approx 0.3\text{dB}\) lower in the case of composite antenna. However, the cross polar component is bit higher for the proposed one. The E-plane patterns are examined in Fig. 6.9. Similar observation is revealed in this also. A summary is presented in Table 6.1.
Fig. 6.6. Simulated reflection coefficients of the single element (basic) antenna and proposed antenna. $W = 30\text{mm}$, $L = 25\text{mm}$, $L_1 = 25\text{mm}$, $W_1 = 13.5\text{mm}$, $L_2 = 20\text{mm}$, $W_2 = 14\text{mm}$, $D_1 = 0.7\text{mm}$, and $D_2 = 0.3\text{mm}$.

Fig. 6.7. Simulated gain of the single element (basic) antenna and proposed antenna. Parameters are as mentioned in Fig. 6.6.
Fig. 6.8. Simulated H-plane radiation patterns of the basic antenna (3.83GHz) and proposed antenna (a) 3.93 GHz, and (b) 4.86GHz. Parameters are as mentioned in Fig. 6.6. Single element (basic) antenna, and proposed shared aperture antenna.
Fig. 6.9. Simulated E-plane radiation patterns of the basic antenna (3.83GHz) and proposed antenna (a) 3.93 GHz, and (b) 4.86GHz. Parameters are as mentioned in Fig. 6.6. Single element (basic) antenna, and proposed shared aperture antenna.
6.4 Prototype and Experimental Verification

To establish the design, experimental studies have been carried out using the fabricated prototype. The fabrication and measurement results are discussed in this section.

6.4.1 Fabricated Prototype

Prototype of the antenna has been realized using 31 mil RT Duroid 5880 substrates. A photograph is shown in Fig. 6.10. The dimensional details are provided in the figure caption (Fig. 6.10). An SMA connector has been used to feed the microstrip line.

6.4.2 Measurement Results

The reflection coefficient of the antenna has been measured using a vector network analyzer (Fig. 2.12 of section 2.4.2) and are shown in Fig. 6.11. They exhibit similar nature as indicated by the simulated results. But the resonance frequencies are shifted to higher values. The shift is ≈126MHz at the lower resonance and ≈110MHz at the higher resonance. The patches got over-etched to some extent from the designed values.

### Table 6.1

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Basic Antenna</th>
<th>Proposed Shared Aperture Antenna</th>
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<tr>
<td></td>
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<td>Lower Resonance</td>
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<tr>
<td>Resonance Frequency (GHz)</td>
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<td>3.93</td>
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<tr>
<td>Bandwidth (MHz)</td>
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<td>Gain (dBi)</td>
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<td>H-plane 3dB Beamwidth (°)</td>
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<td>E-plane 3dB Beamwidth (°)</td>
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<tr>
<td>E-plane Cross Pol. Peak (dB)</td>
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<td>-25.5</td>
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</table>
causing the up-shift of the resonance frequencies. The realized patch dimensions are $L_1 = 24.8\text{mm}$, $W_1 = 13.4\text{mm}$, $L_2 = 19.9\text{mm}$, $W_2 = 13.8\text{mm}$, and the gap between the patches is $0.98\text{mm}$. This also demonstrates higher impedance bandwidths than predicted. These are $\approx 70\text{MHz} (4.025 - 4.095\text{GHz}, 1.71\%)$ at the lower resonance and $\approx 70\text{MHz} (4.915 - 4.985\text{GHz}, 1.41\%)$ at the higher resonance respectively.

The radiation performances of the antenna are evaluated in a Compact Antenna Test
Fig. 6.11. The measured and simulated reflection coefficient of the proposed antenna. Parameters are as mentioned in Fig. 6.10.

Range, described in section 3.4.2. The transmitting antenna is a linearly polarized horn. The radiation patterns have been measured in $\varphi = 0^\circ$ cut and $\varphi = 90^\circ$ cut. The measured patterns at 4.056 GHz are shown in Fig. 6.12(a). The 3dB beamwidths are 74.8$^\circ$ in $\varphi = 0^\circ$ cut and 74.7$^\circ$ in $\varphi = 90^\circ$ cut. The cross-polar levels are lower than 20 dB. Gain, as much as 7.45dBi is realized. Fig. 6.12(b) shows the measured patterns at 4.96 GHz. The half-power beamwidths are 89.5$^\circ$ in $\varphi = 0^\circ$ cut and 79.22$^\circ$ in $\varphi = 90^\circ$ cut. The cross polar discrimination is better than 19.5dB over the entire scan angle. The gain value is 6.8dBi.

It can be examined from Fig. 6.12 that the H-plane ($\varphi=0^\circ$ cut) patterns are asymmetric with respect to the bore-sight and are tilted by 4$^\circ$-5$^\circ$. This is due to the presence of the other patch in proximity in the same plane.
Fig. 6.12. Measured radiation patterns of the antenna, (a) $f = 4.056$ GHz and (b) $f = 4.96$ GHz. Parameters are as mentioned in Fig. 6.10.
6.5 Conclusion

Figs. 6.11 and 6.12 ensure that, our proposed composite configuration is perfectly generating identical polarizations at two different frequencies. Instead of using a conventional full width patch, two elements of reduced width have been conceived within the original footprint. This design technique provides an independent control to the desired frequencies and reduces element width by 50% at the cost of marginal decrease in gain (0.3-0.9dB). Thus an improvised compact shared aperture antenna has been demonstrated. The antenna finds application where communications at two frequencies with arbitrary separation are desired. The comparison of this development with earlier developments is provided in Table 6.2.

<table>
<thead>
<tr>
<th>Reference</th>
<th>Polarization</th>
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<td>[6]–[8]</td>
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<td>Conventional patch width</td>
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<td>[9]</td>
<td>Single LP</td>
<td>Aperture coupled stacked patch</td>
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<td>[11]</td>
<td>Single LP</td>
<td>Resonators of different dimensions used</td>
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<td>[14]–[16]</td>
<td>Single LP</td>
<td>Slot-loaded patches</td>
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<td>Present work</td>
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<td>Two closely spaced patches of reduced width are proximity coupled to a microstrip feed line.</td>
<td>Element width reduction by 50%. Resonance frequencies can be independently controlled.</td>
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REFERENCES


[18] IE3D, v 13.0. Zeland Software Inc., Fremont, California, USA.