Nature has some astonishing camouflaging for survival. With space constraint, electronic circuitry is also struggling to survive within the small devices. The situation demands a camouflaging in electronics also.
The requirements for increased functionality (e.g., direction finding, beam steering, radar, control and command) within a confined volume place a greater burden on today’s transmitting and receiving systems. Reconfigurable antennas (RAs) are a solution to this problem. Reconfiguring an antenna is achieved through deliberately changing its frequency, polarization, or radiation characteristics. This change is achieved by many techniques that redistribute the antenna currents and thus alter the electromagnetic fields of the antenna’s effective aperture. Reconfigurable antennas can address complex system requirements by modifying their geometry and electrical behaviour, thereby adapting to changes in environmental conditions or system requirements (i.e., enhanced bandwidth, changes in operating frequency, polarisation, and radiation pattern). This concept can significantly reduce the number of components and thus hardware complexity, and cost of communication systems. In the early 1930s, the nulls of a two-element array were steered by using a calibrated variable phase changer in order to determine the direction of arrival of a signal [1]. “Reconfigurability” was defined as “the ability to adjust beam shapes upon command [2]”. However the first patent on reconfigurable antennas appeared in 1983 by Schaubert [3]. Excellent overview of reconfigurable antennas, with many examples is given in [4-5].

6.1 Classification and Techniques for Reconfiguration

Six major types of reconfiguration techniques are used to implement reconfigurable antennas, as indicated in Fig.6.1. Electrically reconfigurable Antennas rely on radio-frequency microelectromechanical systems (RF- MEMS), PIN diodes, and varactors to redirect their surface currents.
Antennas that use photoconductive switching elements are called optically reconfigurable antennas and which make use of structure variation are called physically reconfigurable antennas. Smart materials such as ferrites and liquid crystals can also be used to achieve reconfigurability.

Reconfigurable antennas can be classified into four different categories [6].

- **Category 1**: A radiating structure that is able to change its operating or notch frequency by hopping between different frequency bands is called frequency reconfigurable antenna.

- **Category 2**: A radiating structure that is able to tune its radiation pattern is called radiation pattern reconfigurable antenna. For this category, the antenna radiation pattern changes in terms of shape, direction, or gain.

- **Category 3**: A radiating structure that can change its polarisation is called polarisation reconfigurable antenna. In this case, the antenna can change, for example, from vertical to left-hand circular polarisation.
Category 4: This category is a combination of the previous three categories. For example, one can achieve a frequency reconfigurable antenna with polarisation diversity at the same time.

6.2 Electrically Reconfigurable Antennas

An electrically reconfigurable antenna relies on electronic switching components (RF-MEMs, PIN diodes, or varactors) to redistribute the surface currents, and alter the antenna radiating structure topology and/or radiating edges. The integration of switches into the antenna structure makes it easier for designers to reach the desired reconfigurable functionality. The ease of integration of such switching elements into the antenna structure has attracted antenna researchers to this type of reconfigurable antennas despite the numerous issues surrounding such reconfiguration techniques. These issues include the nonlinearity effects of switches, and the interference, losses, and negative effect of the biasing lines used to control the state of the switching components on the antenna radiation pattern. Some important characteristics of a switch are characteristic impedance, bandwidth, insertion loss, isolation, switching speed, power handling and topology [7]. A rule of thumb for the highest operating frequency for the switch is approximately one tenth of resonant frequency [8]. Widely used microwave switch is the PIN diode [9]. Forward biasing a PIN diode creates a very low resistance at high frequencies, while reverse biasing results in an open circuit.

Frequency RAs can change the antenna operating frequency continuously or discretely while maintaining the polarisation and radiation pattern stable across the entire frequency tuning range. A high-gain quasi-Yagi dipole RA operating over the 478-741 MHz UHF TV band is proposed.
for cognitive radio applications [10]. A method for designing affordable, compact, RA by changing the effective length of a resonant slot antenna through control of combinations of electronic RF switches is reported in [11].

6.3 Frequency Reconfigurable Antenna derived from Asymmetric Coplanar Stripline (ACS)

In this work, metamaterial spiral is loaded on an ACS to achieve frequency RA. The antenna exhibits zeroth order characteristics, making it flexible to change the dimensions without affecting the resonant frequency. As explained in Chapter 3, spiral in combination with microstrip transmission line exhibits CRLH property. A CRLH Transmission Line is composed of periodic repetition of CRLH unit cell of size ‘p’, (shown in Fig.3.11 is reproduced for reference) is shown in Fig.6.2. The structure behaves as a uniform transmission line and may therefore be transformed into a resonator by using discontinuous (short/open) terminations.

![Fig.6.2 Equivalent circuit of CRLH Transmission Line unit cell](image)
As seen in Fig.6.2, the series resonance frequency is given by \( \omega_{se} = \frac{1}{\sqrt{L_R C_L}} \) and shunt resonance is given by \( \omega_{sh} = \frac{1}{\sqrt{L_L C_R}} \). The case of equal series and shunt resonances is referred to as the balanced resonance condition. This is also the condition for optimal matching over a broad frequency range, hence large bandwidth. In the unbalanced resonance condition, \( \omega_{se} \) and \( \omega_{sh} \) are unequal, leading to a narrow bandwidth [12-13]. The uniplanar equivalent of a CRLH Microstrip Transmission Line may be obtained using a Coplanar Waveguide. For further size reduction, an Asymmetric Coplanar Stripline (ACS) may be used [14].

Type 1 and Type 2 spirals explained in Chapter 3 are the basic blocks of this application. It is recalled that the difference between Type 1 and Type 2 is the gap (\( w_3 \)) between the spiral and transmission line. The evolution from spiral embedded transmission line to ACS antenna is explained below. Fig.6.3(a) shows an Asymmetric Coplanar Stripline of length ‘L’ and width ‘W’ corresponding to 50 \( \Omega \) characteristic impedance. Width and length of ground are ‘Wg’ and ‘Lg’. Fig.6.3(b) shows type 2 spiral of length ‘L’ inserted between transmission line sections of length L_1 and L_2. Fig.6.3(c) shows how transmission line L_2 is terminated in open circuit to transform the transmission line structure to ACS open ended resonator. The resonating frequency of this open ended resonator is 4.35 GHz, which is near to resonating frequency (4.6 GHz) of Type 2 spiral embedded in microstrip transmission line. When a gap (\( gt_2 \)) of 0.2 mm is introduced between spiral and transmission line, ‘t_2’ as shown in Fig.6.3(d) the resonating frequency is lowered to 2.86 GHz, which is similar to resonating frequency (2.65 GHz) of Type 1 resonator explained in Chapter 3.
Chapter 6

(d) Frequency (GHz) 2.0 2.5 3.0 3.5 4.0 4.5 5.0
Reflection characteristics (dB) -8 -6 -4 -2 0

(e) Frequency (GHz) 2.0 2.5 3.0 3.5 4.0 4.5 5.0
Reflection characteristics (dB) -14 -12 -10 -8 -6 -4 -2 0

(f) Frequency (GHz) 2.0 2.5 3.0 3.5 4.0 4.5 5.0
Reflection characteristics (dB) -14 -12 -10 -8 -6 -4 -2 0
Fig.6.3 (a) ACS transmission line (b) Type 2 spiral embedded into ACS transmission line (c) Type 2 spiral embedded ACS transmission line transformed to open ended resonator (d) gap ‘gt2’ introduced between t2 and Type 2 spiral (e) t2 completely removed to yield Type A ACS (f) gap ‘gt2’ shorted to get Type B ACS antenna (g) PIN diode at gt2 to achieve frequency reconfiguration \( L_1=2.6\text{mm}, L=4.7\text{mm}, W=3\text{mm}, w_1=0.3\text{mm}, w_2=0.3\text{mm}, W_g=4\text{mm}, L_g=3.5\text{mm}, g=0.3\text{mm} \) (h) Experimental and simulation results

This similarity of open ended structures to embedded counter parts is important as it reveals that the resonating frequency is almost independent of mode of termination. Hence, to simplify the structure of ACS, the transmission line \( t_2 \) is completely removed to expose a gap ‘gt2’ to introduce capacitance. This is shown in Fig.6.3(e) and is named Type A resonator. To get the effect of Type 2 spiral, a small short is introduced at the gap ‘gt2’ and is named Type B resonator. Type B resonator is shown in Fig.6.3(f). From the theory of RF switches, it is well known that a PIN diode can act as open circuit in OFF condition and act as short circuit in ON condition. Hence, the short at ‘gt2’ is replaced by a PIN diode to switch between Type A and Type B resonators as shown in Fig.6.3(g). This results in a frequency RA ACS antenna. Experimental and simulation results of this antenna are shown in Fig.6.3(h).
In Fig. 6.3 (g) the spiral resonator is embedded into the signal strip of a 50 \( \Omega \) ACS transmission line on a substrate with \( \varepsilon_r \) 4.4 and thickness 1.6 mm resulting in a CRLH type antenna. The length \((L_1)\) of transmission line \((t_1)\) is optimized as 2.6 mm to achieve good matching. Width of the spiral resonator \((W)\) is 3 mm, which is the width of the 50 \( \Omega \) transmission line. The width of arms ‘\( w_1 \)’ is 0.3 mm and gap between arms of spiral ‘\( w_2 \)’ is 0.4 mm. Reduction of gap can deteriorate the antenna performance as the structure becomes more and more capacitive. The substrate dimension is optimized as 10 mm x 10 mm for better impedance matching. The gap \((g)\) between \( t_1 \) and ground is optimized at 0.3 mm. The width of ground \((W_g)\) is 4 mm and length of ground \((L_g)\) is 3.5 mm as a compromise between matching and bandwidth. When PIN diode is ON the antenna resonates at 4.35 GHz with a narrow bandwidth of 54 MHz as seen in Fig. 6.3(h). This narrow bandwidth is due to the unbalanced nature of CRLH resonator. The input impedance of the structure at resonance is \( 48+j2.64 \Omega \). When PIN diode is OFF it introduces an additional series capacitance, reducing the resonant frequency from 4.35 GHz to 2.9 GHz as observed in Fig. 6.3(h). The impedance bandwidth is reduced to 40 MHz. The input impedance is \( 56.6-j8.43 \Omega \) at resonant frequency, which is more capacitive. This capacitive nature of the input impedance leads to lower bandwidth [15].

A varactor diode can be used at the gap ‘\( gt_2 \)’ instead of PIN diode with varying bias voltages [16] to fine tune the resonant frequencies. Chip inductance also can be inserted at the gap \((gt_2)\) in order to fine tune resonant frequency. Variation in the simulated resonant frequency due to varying capacitor/inductor at the gap is listed in Table 6.1. The variation in
frequency becomes minimal for inductance above 30 nH and capacitance 10 pF.

<table>
<thead>
<tr>
<th>Sl.no</th>
<th>C(pF)</th>
<th>Res.freq (simulation) (GHz)</th>
<th>L(nH)</th>
<th>Res.freq (simulation) (GHz)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>short</td>
<td>4.35</td>
<td>4</td>
<td>4.14</td>
</tr>
<tr>
<td>2</td>
<td>2</td>
<td>4.66</td>
<td>8</td>
<td>3.952</td>
</tr>
<tr>
<td>3</td>
<td>5</td>
<td>4.52</td>
<td>12</td>
<td>3.816</td>
</tr>
<tr>
<td>4</td>
<td>7</td>
<td>4.488</td>
<td>16</td>
<td>3.71</td>
</tr>
<tr>
<td>5</td>
<td>10</td>
<td>4.472</td>
<td>20</td>
<td>3.64</td>
</tr>
</tbody>
</table>

Frequency tuning of antenna is same as that of spiral resonators (SR) Type 1 and Type 2 explained in Chapter 3. When relative permittivity or height of substrate is varied, corresponding change occurs in width (W) of 50 Ω transmission line. Hence, width (W) and length (L) of spiral also change.

6.4 Zeroth Order Resonant Antenna

CRLH can exhibit zeroth order resonant (ZOR) mode in which the resonant frequency is independent of the number of unit cells. Hence, ZOR antenna is also known as infinite wavelength resonant antenna. The phase constant $\beta = 0$ at resonant frequency implies the infinite guided wavelength $\lambda_g = 2\pi/|\beta| = \infty$, and zero phase shift $\theta_m = -\beta l = 0$. This phenomenon enables the realization of zeroth-order resonance [16] in which the length of the resonator is independent of the resonance condition (i.e. the multiple of the half wavelength in case of the open circuited TL). The principal voltage
distribution along the resonant length for the negative and zero resonances is shown in Fig. 6.4.

![Voltage distribution of open-circuited TL](image)

**Fig.6.4.** Voltage distribution of open-circuited TL of length $l$. Mode $m = 0$ represents zeroth-order resonator with infinite guided wavelength.

For a CRLH resonator with $N$ unit cells, its resonance occurs when

$$\beta_n = \frac{m \pi}{l} \left( m = 0, \pm 1, \pm 2 \ldots \pm (N - 1) \right)$$  \hspace{1cm} (6.1)

where $l = N^* p$ is the size of resonator and $m$ is the mode number. As a result, a CRLH resonator with $N$ unit cells exhibits finite ‘$2N-1$’ resonance frequencies. The zeroth-order resonance mode corresponds to an infinite wave-length and thereby to a flat field distribution. In the unbalanced CRLH case, the zeroth order resonance corresponds to either $\omega_{se}$ due to the short ended structure or $\omega_{sh}$ due to the open ended structure. The active immittance elements for $\omega_{se}$ are $L_R-C_L$ and for $\omega_{sh}$ are $L_L-C_R$ (Fig.6.2). In the case of open ended structure, only shunt elements are active and therefore
one of the series elements ($L_R$) is sufficient to couple the shunt resonators of additional unit cells. Under this condition all CRLH equations are simplified by the substitution $C_L$ tends to infinity. No more LH branch exists and only the resonance $\omega_{sh}$ appears in the spectrum. If the size of the structure is enlarged by adding more unit cells, the resonant frequency remains the same at $\omega_{zor} = \omega_{sh}$.

Since the resonant frequency of ZOR is independent of size, it can be enlarged by adding more unit cells to attain a very large electrical size and high directivity. The zeroth order status of type B antenna is verified and shown in Fig.6.5. In order to facilitate the switching action between second order Type A and Type B antennas, the second order antenna structure is modified by placing the spiral and its mirror image laterally as shown in Fig.6.5(b). From Fig.6.5(c), it is clear that the experimental resonance frequency is almost the same for one unit cell (2.93 GHz) and for two unit cells (3.01 GHz). An extremely useful and unique property of CRLH metamaterial structures is that the size and gain can be controlled independent of the resonant frequency of the antenna.
Fig. 6.5 (a) Spiral ACS Antenna Type B with two unit cells arranged one over the other (b) Spiral ACS Antenna Type B with two unit cells arranged side by side (c) reflection characteristics showing zeroth order response

The fabricated first and second order antennas are shown in Fig. 6.6.
Fig. 6.6 Fabricated antennas (a) Type A (b) Type B (c) second order Type A

6.4.1 Dispersion characteristics of developed Zeroth Order Antennas

The dispersion characteristic of the Type B antenna is shown in Fig. 6.7. This curve is plotted based on Floquet theorem explained in Chapter 3, from the scattering parameters of Type B antenna. The antenna is
first converted to a two port network by adding a small strip of transmission line at the top, to which the second port is connected (Fig. 6.3(b1)). Reflection and transmission characteristics are observed and the S-matrix is transformed to ABCD matrix whose eigen values give the dispersion characteristics as explained in section 3.5.1.

\[
\begin{bmatrix}
{s_{11}} & {s_{12}} \\
{s_{21}} & {s_{22}}
\end{bmatrix} \rightarrow \begin{bmatrix}
{A} & {B} \\
{C} & {D}
\end{bmatrix} \rightarrow \text{Eigen vector} \rightarrow [y] \rightarrow [\beta]
\]

The dispersion characteristics shows that phase is zero at the resonant frequency of Type B antenna. This indicates the zeroth order nature of the antenna.

![Fig. 6.7 Dispersion characteristic of Type B antenna proving Zeroth Order nature](image)

It can be observed that for second order, only the spiral is repeated and the transmission line near the feed end is not considered a part of unit cell. To ensure whether this decision is appropriate, dispersion characteristics of Type 1 and Type 2 spirals are plotted using simulation of eigen solver method. The method and results are briefly summarised as follows.
In order to study the metamaterial property of spiral, an infinite repetition of the unit cell is desired. This requirement is achieved by imposing boundary conditions on the unit cell. Two types of boundary conditions can be used to achieve periodicity. The combination of perfect electric (PE) parallel to electric field in the structure and perfect magnetic (PM) boundary conditions perpendicular to PE simulates the periodic boundary conditions by utilising the symmetry inherited by the metamaterial due to periodic repetition of the unit cell [17]. Periodicity can also be achieved by providing master-slave boundary conditions. The boundary conditions at the master are enforced at the slave's surface, hence realising an infinite periodic repetition. Type 2 spiral is shown in Fig. 6.8(a) enclosed in proper boundary conditions. To get the effect of periodicity of an array of spirals, master slave boundary is applied along x-direction as shown in Fig. 6.8(b). Since, the spiral is rectangular, the electric field is directed along its length and hence, perfect electric boundary is set parallel to length of spiral as shown in Fig.6.8(c) and finally perfect magnetic boundary is applied along y-direction as shown in Fig.6.8(d). This setup provides planar periodicity of spirals. Simulation of the array is done in eigen solver mode. The phase advancement of wave along z-direction is measured for angles '0' to '180'. Calculation is done using linear step option of 48 steps. This ensures sufficient accuracy for the dispersion characteristics.

Dispersion is plotted for first two modes and is shown in Fig. 6.9(a). The same procedure is repeated for Type 1 spiral and its dispersion curve is shown in Fig. 6.9(b). From Fig.6.8(a) it can be observed that for first mode, phase advancement is zero at 4.4 GHz, the resonant frequency of Type B antenna. This confirms the zeroth order characteristics of Type B antenna.
Similarly, Fig. 6.9(b) shows that first mode has zero phase advancement at 2.84 GHz, the resonant frequency of Type A antenna. This confirms the zeroth order nature of Type A antenna also. The curves plotted confirm that spirals are indeed the unit cells of the developed antennas.

Fig. 6.8 (a) Type 2 spiral enclosed in air box (b) master slave boundary simulating periodicity along x-direction (c) perfect electric (PE) boundary parallel to electric field of spiral (d) perfect magnetic (PM) boundary perpendicular to PE.
Fig. 6.9 (a) Dispersion characteristics of Type 2 Spiral Resonator showing zero phase at 4.4 GHz. (b) Dispersion characteristics of Type 1 Spiral Resonator showing zero phase at 2.84 GHz.
6.5 Parameter extraction of ACS antenna

It is observed that Type A and Type B ACS antennas resonate with slightly different frequencies compared to Type 1 and Type 2 spirals explained in Chapter 3 even with same spiral dimensions. This variation is attributed to change in feeding mechanism of spiral, from microstrip structure to ACS structure and hence change in parameters. Therefore parameter extraction of Type A and Type B ACS resonators are done using the same methodology followed in section 3.5.

The inductor and capacitor values of filters discussed in 3.1-3.3 are extracted using equations given in [3.12]. The equations are summarised as follows as given by (6.2-6.4)

\[ L_R = 0.5 \text{ Im} (Z') \text{ at } \omega_{se}; \text{ } Z' \text{ is the first derivative with respective to } \omega \]

\[ C_R = 0.5 \text{ Im} (Y') \text{ at } \omega_{sh} \]

\[ C_L = 1/(\omega_{se}^2 L_R) \text{ and } L_L = 1/(\omega_{sh}^2 C_R) \]

Since, the CRLH TL considered seems to be lossy from experimental results, series resistance and shunt conductance are also extracted as \( R = \text{Re}(Z) \text{ at } \omega_{se} \) and \( G = \text{Re}(Y) \text{ at } \omega_{sh} \). The values are tabulated in Table 6.2.

<table>
<thead>
<tr>
<th>Type of ACS</th>
<th>( L_R \text{ (nH)} )</th>
<th>( C_R \text{ (pF)} )</th>
<th>( L_L \text{ (nH)} )</th>
<th>( C_L \text{ (pF)} )</th>
<th>( R \text{ (Ω)} )</th>
<th>( G \text{ (Ω}^{-1}) )</th>
</tr>
</thead>
<tbody>
<tr>
<td>Type A</td>
<td>2.0253</td>
<td>1.919</td>
<td>1.7061</td>
<td>2.0737</td>
<td>40.17</td>
<td>0.07168</td>
</tr>
<tr>
<td>Type B</td>
<td>1.086</td>
<td>0.1288</td>
<td>8.563</td>
<td>1.2032</td>
<td>23.847</td>
<td>0.0044</td>
</tr>
</tbody>
</table>

School of Engineering, Faculty of Engineering, CUSAT
From the theory of open ended CRLH resonators, it is expected that type A and type B antennas would resonate at $\omega_{sh}$. This is verified by calculating the series and shunt resonant frequencies (in GHz) using the following equation 6.5. and results are summarised in Table 6.3.

$$\omega_{se} = \frac{1}{\sqrt{L_RC_L}}$$

$$\omega_{sh} = \frac{1}{\sqrt{L_RC_R}}$$

\[\text{(6.5)}\]

<table>
<thead>
<tr>
<th>Type of ACS</th>
<th>Series resonance $\omega_{se}$(GHz)</th>
<th>Shunt resonance $\omega_{sh}$(GHz)</th>
<th>Simulated resonant frequency (GHz)</th>
<th>Experimental resonant frequency (GHz)</th>
</tr>
</thead>
<tbody>
<tr>
<td>A</td>
<td>2.4571</td>
<td>2.7829</td>
<td>2.86</td>
<td>2.91</td>
</tr>
<tr>
<td>B</td>
<td>4.405</td>
<td>4.428</td>
<td>4.35</td>
<td>4.36</td>
</tr>
</tbody>
</table>

It is clear from Table 6.3 that both antennas are resonating at $\omega_{sh}$ as expected and bandwidth of Type A is lower due to unbalanced nature of its resonance ($\omega_{se} \neq \omega_{sh}$).

6.6 Radiation pattern

The radiation pattern of Type B is similar to a dipole as shown in Fig.6.10 (a). However, the pattern is tilted due to the asymmetric feed. Type A antenna radiates with directional characteristics as shown in Fig.6.10 (b). In second order arrangement, since spirals are connected back to back, directive nature of Type A is reformed to omnidirectional pattern as seen in
Fig. 6.10 (c). Front to back ratio for both Type A and Type B is 1.1 for second order. Both antennas exhibit linear polarization. There is minimal variation in the radiation characteristics of Type A and Type B antennas in second order configuration, which enhances the reconfigurable nature of the proposed antenna. The radiation pattern tilt can be resolved using a CPW feed, provided there is no constraint on the size of the antenna as shown in Fig. 6.10 (d). Simulated gain of Type B antenna is 0.146dBi and 0.186dBi for first and second orders, whereas Type A exhibits a low gain of 0.059dBi in both cases, but there is an improvement in radiation characteristics for second order compared to first order of Type A.

As discussed in section 6.3, frequency tuning may be incorporated by introducing reactive components into the gap, \( g_t \). When gap \( g_t \) is closed with capacitor, gain increases considerably. Presence of inductance at \( g_t \) makes antenna behave similar to Type B antenna with frequency tunability. Extremely low values of radiation efficiency do not enable the ZOR structures to work efficiently as an antenna. Therefore the question is whether the structure behaves rather as a lossy transmission line or as an antenna [18]. The simulated radiation efficiency of second order Type A is 11% and second order Type B is 12.8%, which is reasonable among conventional CRLH antennas. The antenna has an added advantage of size reduction as it satisfies the condition for Electrically Small Antennas (ESA) for the two operating frequencies compared to conventional ACS monopoles [19].
Fig. 6.10 (a) radiation pattern of type B (b) radiation pattern of type A antennas. (c) radiation pattern of second order type A and type B antennas. Both shows near dipole characteristics for second order (d) CPW counterpart of type A and type B antennas and its radiation pattern with minimum tilt.
6.7 Electrically Small Antenna (ESA)

An antenna is considered to be an Electrically Small Antenna (ESA), if \( ka = \frac{2\pi a}{\lambda_0} \leq 0.5 \) where \( k \) is the wave number, 'a' is the radius of smallest sphere that surrounds the antenna system and \( \lambda_0 \) is the wavelength at resonance [20]. Most often, antenna itself is a capacitor or inductor and it is tuned to resonance by a reactor of opposite kind. When the antenna becomes electrically small, the propagating modes are replaced by evanescent modes with high Q, where Q is inversely proportional to cube of radius of sphere enclosing the antenna. In short, the maximum bandwidth of an electrically small antenna is regulated by its maximum dimension enclosed within a sphere of radius 'a'.

The difficulties of designing an electrically small antenna includes:

- impedance matching,
- insertion loss from high current density flowing on a non-perfect conductor, resulting in joule heating, and
- a small radiation aperture with low radiation efficiency.

Though Electrically Small Antennas are very desirable for both military and commercial applications, the primary issues are usually directivity (gain) versus aperture size and bandwidth versus volume (Chu limit). Electrically small antennas are known to be inefficient radiators due to the relative magnitudes of the radiation and ohmic loss resistances [21]. Matching circuits have to be perfect to avoid losses. RFCs may also be employed to avoid return current path to the feed whose size becomes significant compared to ESA. It is concluded that size can only be reduced
at the expense of bandwidth or efficiency. In general, the best performance from ESA will be achieved if the dielectric constant is as low as possible and the aspect ratio, which is the ratio of maximum to minimum dimension of antenna volume, is close to unity. The internal structure of the antenna is also such that the fields fill the minimum size enclosing the sphere with the greatest possible uniformity. Since the reflection characteristics is considered insufficient to describe the performance of an ESA, the product of bandwidth and efficiency, 'Bη' is often chosen as parameter for comparison, where B is the 3 dB bandwidth and 'η' is the efficiency. Since impedance matching is difficult, ESAs tend to be lossy. The parameter 'Bη' is sufficient for characterisation of small antennas because the increase of bandwidth due to loss is met at the expense of loss in radiation efficiency. Thus the product 'Bη' gives the measure of performance for a lossy as well as lossless ESA.

For Type A antenna and Type B antenna, 'ka' is found to be 0.33 and 0.47 satisfying the condition for ESA. A number of ESAs falling under different categories are compared in [22]. Comparison of the antennas discussed in this chapter is also done using this approach.

The drawback of developed antenna is its small gain. However the antenna satisfies Chu limit and has a better performance in comparison with other reported metamaterial antennas of similar size with respect to aspect ratio and substrate used for fabrication [22]. The aspect ratio defined as the ratio of maximum to minimum dimension (the ratio of length to height of substrate) of antenna volume is close to unity. The internal structure of the antenna is also such that the fields fill the minimum size enclosing the
sphere with the greatest possible uniformity. Aspect ratio of developed ACS antennas is 6.25. \( B \eta \) is calculated for first order Type A and Type B antennas and compared with Wheeler's limit defined for a particular substrate and Gustaffson limit defined for a particular aspect ratio. \( B \eta \) (simulated) = 0.057 x 0.076 = 0.00433; where \( B = 0.057 \) is the measured bandwidth and \( \eta = 0.076 \) is the simulated efficiency. \( B \eta \) (theoretical) = 0.054 x 1= 0.054; where \( B = 0.054 \) is the simulated bandwidth and \( \eta = 1 \) is the theoretical efficiency. The ratio of \( B \eta \) (simulated) to \( B \eta \) (theoretical) is 0.08, which is better than other reported metamaterial antennas [22] for a dielectric constant of 4.4. The Wheeler's limit [ratio of \( B \eta \) (measured) to \( B \eta \) (theoretical)] for this substrate is 0.2 [22]. The comparison of both antennas with respect to [22] is shown in Fig.6.11.

\[ \text{Gustaffson's limit for aspect ratio 6.25 is 0.15} \]
\[ \text{Wheeler Chu limit for } \varepsilon_r = 4.4 \text{ is 0.2} \]

Fig.6.11 Comparison of Type A and Type B antennas with respect to [22]

6.8 Inference

The developed antenna is compact and coplanar. It exhibits a frequency reconfiguration of 1.47:1. The radiation characteristics of this antenna in second order remain essentially unaffected by the frequency
tuning. The spiral determines the resonant frequency and hence tuning parameters are length of spiral, number of turns, width of each arm and the gap between arms. The tuning ability of spiral is already discussed in Chapter 3. Due to its small size, it falls under the category of Electrically Small Antennas. The small gain and efficiency may be attributed to the inherent characteristics of ESAs. However, it can be reduced by designing proper matching circuits. The Bloch impedance explained in section 3.5.1 may be utilised for the design of matching circuits. Type A and Type B antennas also exhibit zeroth order characteristics as seen through the calculated dispersion characteristics. The highly reactive and intense near field of these metamaterial ESAs can be utilised for near field applications like sensor networks. This aspect is explained in next chapter.

References


Chapter 6


