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All-Passive Composite Right/Left-Handed (CRLH) Antenna Array for Sum and Difference Patterns

By

Vu James Le

Thesis Submitted

in

Partial Fulfillment of the

Requirements for the Degree of

MASTER OF SCIENCE IN ELECTRICAL ENGINEERING

Approved by:

Professor _____

Dr. Jayanti Venkataraman, Advisor

Professor _____

Dr. Gill Tsouri

Professor _____

Dr. Panos Markopoulos

Professor _____

Dr. Sohail Dianat

DEPARTMENT OF ELECTRICAL AND MICROELECTRONIC ENGINEERING
KATE GLEASON COLLEGE OF ENGINEERING
ROCHESTER INSTITUTE OF TECHNOLOGY
ROCHESTER, NEW YORK
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Abstract:

Monopulse tracker is a system that uses the sum and difference radiation patterns to track. If there is a target in the beam, the tracker receives an echo from it following each transmitted pulse. The target range or distance is readily determined by multiplying half the time delay between the transmitted and received pulses by the speed of light. Typically, the sum and difference patterns are obtained by adding and subtracting individual beam patterns electronically or using RF components, such as: hybrid junctions, directional couplers. However, RF components are frequency dependent. Once they are used to realize the sum pattern at one frequency, they cannot be used to realize the difference pattern at another frequency. Hence, the objective of this work is to create a dual-band phased-array antenna with Composite Right/Left-Handed (CRLH) transmission lines to obtain the sum and difference patterns directly and passively.

In order to realize the sum and difference patterns directly on a single physical board, the amplitude distribution of an 8 isotropic linear antenna array is designed using the Dolph-Tchebycheff antenna synthesis technique. By exciting all the antenna elements in-phase, the sum pattern is generated; exciting one half of the antenna array out-of-phase (180° phase difference) with respect to the other half, the difference pattern is released. For hardware implementation, a dual-band Wilkinson power divider has been designed for in-phase output at 2.5 GHz and a phase difference of 180° at 1.8 GHz by injecting a CRLH unit cell in one of the output arms. In addition, microstrip power dividers are designed to implement the amplitude distribution. These are incorporated with dual-band microstrip antennas to obtain the sum and difference patterns.

While the amplitude distribution of an 8 isotropic linear array are computed using MATLAB, hardware design optimization and simulation have been done using ANSYS HFSS. A CRLH unit cell, inspired by available research in the literature, has been designed and implemented in one of the output arms of a conventional Wilkinson power divider resulting an output of 180° phase difference at 1.8 GHz and 0° phase difference at 2.6 GHz. This Dual-Band CRLH Wilkinson power divider has been fabricated and validated using Network Analyzer. The difference pattern has been designed with a side lobe level (SLL) of -30 dB. This results a SLL of -12 dB for the sum pattern. In addition, the sum and difference patterns are theoretically generated with MATLAB to verify the designed SLL. Furthermore, four different microstrip power dividers have been designed to split the power properly between antenna elements based on the computed amplitude distribution. Dual-band microstrip antennas operating at 1.8 GHz and 2.6 GHz are designed and equally spaced out at a distance $d = \lambda/2$ to avoid the mutual coupling. The design is modeled and simulated in HFSS and for the Rogers RT/duroid substrate with a dielectric constant of 2.2 and the thickness of 31 mil. All the necessary measurements to validate the design can be done using the Network Analyzer and anechoic chamber.

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1. Introduction:

1.1. Monopulse tracker and metamaterials:

Monopulse tracker is a radar system that uses sum and difference patterns to track. Sum pattern is a broadside beam and difference pattern is a two sharp equal main lobes with a deep null in the broadside direction. By switching between the sum and difference patterns (Figure 1.1), detection and tracking of airborne vehicles can be achieved.

Figure 1.1. Operation of a monopulse tracker [15]

It operates by radiating electromagnetic energy as transmitted pulses and detecting the echo returned from reflecting objects (targets). The target range or distance is calculated by multiplying half the time delay between the transmitted and received pulses by the speed of light. For tracking, the beams are locked on the selected target. This can be done by keeping the antenna axis aligned with a selected target mechanically. In addition, the operating frequencies of the sum and difference patterns cannot be too far apart because the wavelengths would be significantly

different. As a consequence, the target may be visible at one operating frequency but not the other. In general, radar is a complex system. For the scope of this research, only radiating electromagnetic energy to generate the sum and difference patterns for tracking is investigated.

Figure 1.2. Simple block diagram of a radar system [10]

The basic parts of a radar system are illustrated in Figure 1.2. The radar signal (pulse) is generated by the transmitter and radiated into space by antenna (the transmitter is shown as a power amplifier). For detection and tracking, sum and difference pulses are rapidly generated and transmitted. Typically, sum and difference pulses are generated by adding and subtracting individual beam patterns electronically or using RF components, such as: hybrid junctions, directional couplers. There are some disadvantages of obtaining sum and difference patterns this way. The patterns that were added or subtracted would not be the same if one of the antenna elements were missing. In addition, the patterns are affected by mutual coupling. Hence, the ultimate goal of this research is to suggest an alternative design to generate the sum and difference pulses simultaneously, directly, and passively using metamaterials. The sum pattern can be obtained with conventional power divider feed system at one frequency. The difference pattern

can be obtained by incorporating metamaterials in the same feed system and functioning at a different frequency.

Primarily, all materials can be characterized by two main parameters: dielectric permittivity (ϵ) and magnetic permeability (μ). The permittivity measures the resistance of materials in the formation of electric field and the permeability measures the ability of materials to allow the magnetic lines of force to pass through it. As shown in Figure 1.3, all the conventional materials have both positive μ and ϵ (1st quadrant); if either μ or ϵ is negative or positive (2nd and 4th quadrants), there are some ferrites and semiconductors; However, if both μ and ϵ are negative (3rd quadrant), no material can be found in the nature. This is preferred to metamaterials. Metamaterials are artificial structures that can be designed to exhibit specific electromagnetic properties not commonly found in nature. The concept of creating a material that has both negative permittivity and permeability was theoretically predicted by the Russian physicist Veselago back in the 1960s [12]. However, the experimental verification of such a material did not occur until late 1990s, early 2000s by a group of researchers from University of California, San Diego. Since then, metamaterial has led to several novel microwave application and devices.

Figure 1.3. Material parameter space characterized by dielectric permittivity (ϵ) and magnetic permeability (μ) [13]

Furthermore, metamaterials with simultaneously negative permittivity and permeability are commonly preferred to as left-handed materials. Left-handed materials are generally modelled as composite right/left-handed (CRLH) structures, which include right-handed (RH) effects that occur naturally in practical LHMs. Hence, CRLH transmission lines possess the property of both right and left hand materials. Using the transmission line theory, [12] theoretically derives the equivalent circuit model of a CRLH transmission line (Figure 1.4). The CRLH transmission line model consists of a series inductor (L_R) in parallel with a series capacitor (C_L) and a shunt capacitor (C_R) in parallel with a shunt inductor (L_L). While C_R and L_R naturally present in the CRLH transmission line as RH material, C_L and L_L can be physically implemented using surface-mount technology chip components.

Figure 1.4. Theoretical model of a CRLH transmission line [12]

CRLH transmission lines have been used in many novel microwave applications, such as: designing a full-hemisphere frequency scanning antenna, advancing wave guided structures... and specially developing dual-band components/devices. This will be a critical in developing an alternative design for releasing sum and difference patterns directly, and passively.

1.2. Contributions to Present Work:

The main contributions of the present work are as follows:

- Designed a linear array of isotropic sources using Dolph-Tschebyscheff antenna synthesis technique to obtain sum and difference patterns passively:
 - Developed the amplitude distributions for sum and difference patterns using MATLAB for a required difference pattern side lobe level, R_o .
 - Calculated far field array factor from general expressions for N isotropic elements to theoretically plot the designed sum and difference patterns and implemented, using MATLAB.
- Recreated a Dual-Band CRLH Phased-Array antenna to direct the main at $\pm 10^\circ$ operating at 1.8 GHz and 2.5 GHz respectively:
 - Followed the procedure described in [4] to calculate values of lumped components, C_L and L_L .
 - Modeled, simulated the Dual-Band CRLH Phased-Array feed system in HFSS.
 - Fabricated the antenna array.
 - Experimentally validated the fabricated board with Network Analyzer and obtained the radiation patterns using anechoic chamber.
- Designed a Dual-Frequency CRLH Wilkinson power divider for in-phase output at 1.8 GHz and a phase difference of 180° at 2.6 GHz as well as an equal split conventional Wilkinson power divider.
 - Studied different L-C unit cell to improve the design.
 - Fabricated and validated the power dividers with return loss measurements.

- Designed microstrip power dividers to implement the Dolph-Tschebyscheff excitation coefficients for sum and difference pattern.
- Designed wideband microstrip-fed monopole and dual-band microstrip patch antennas.
- Modeled and simulated an all-passive CRLH Antenna Array for sum and difference patterns in HFSS.

1.3. Organization:

This thesis is divided into 7 chapters and is organized as follows:

- Chapter 1: An introduction outlining the motivation and background for the thesis.
- Chapter 2: Presenting an overview of sum and difference patterns and how they can be released physically. Describing a procedure of computing Dolph-Tschebyscheff excitation coefficients for a linear array isotropic sources to obtain sum and difference patterns.
- Chapter 3: Recreating a Dual-Band CRLH Phased-Array antenna. Exclusively done to study the CRLH transmission line concept and hardware implementation.
- Chapter 4: Presenting the design process of a Dual-Frequency CRLH Wilkinson power divider as well as studying the effect of L-C unit cell for potentially improving the design. Ultimately, designing a feed system for generating sum and difference patterns.
- Chapter 5: Presenting the design of a dual-band microstrip patch antenna.
- Chapter 6: Presenting the design of an all-passive CRLH Antenna Array for sum and difference patterns.
- Chapter 7: Conclusions and future research.

2. Linear array of isotropic sources with Tschebyscheff excitation coefficients for sum and difference patterns:

2.1. Sum and difference patterns design concept:

In general, the sum pattern is simply a broadside array, and the difference pattern is a broadside array splitting into two equal main lobes with a sharp null in between. An example of the sum and difference patterns is shown in Figure 2.1 [1]. Sum pattern can be generated from isotropic antenna elements by exciting all the elements in-phase (0° phase difference). In addition, exciting one half of the antenna elements out-of-phase (180° phase difference) with respect to the other half will release the difference pattern. However, depending on the designed side lobe level (SLL), the amplitude distribution of the excitation coefficients need to be properly and specifically designed for the sum and difference patterns. As a consequence, two set of excitation coefficients are normally required to generate the sum and difference patterns correspondingly. Due the fact that, all RF components are frequency dependent, it would require two separate physical boards to implement those two excitation coefficient sets. Since the main goal of this study is to generate the sum and difference patterns on a single physical board directly and passively, only one set of the excitation coefficient can be implemented on hardware.

Figure 2.1. Example of sum and difference patterns [1]

In the literature, [2] demonstrates the calculations of the difference pattern amplitude distribution from the corresponding sum pattern amplitude distribution for a specific required SLL. In other words, to find the difference pattern amplitude distribution, a corresponding sum pattern needs to be defined. However, this sum pattern is not what to be generated. It is just an intermediate step to calculate the difference pattern amplitude distribution. Additionally, the Dolph-Tschebyscheff antenna synthesis technique [3] can be used to find the sum pattern amplitude distribution for a required SLL. To release the sum and difference pattern simultaneously, the difference pattern amplitude distribution is implemented out-of-phase at one frequency. In addition to that, the same amplitude distribution is implemented in-phase at a difference frequency. Figure 3.2 demonstrates an amplitude distribution for the sum and difference patterns.

Figure 2.2. Distribution of excitation coefficients. (a) Sum pattern. (b) Difference pattern [2]

Figure 2.2 shows an example of an 8 isotropic elements array. The amplitude distribution is symmetrical and the antenna elements are equally spaced out at a distance d . For the difference pattern, the left half array elements are out-of-phase with respect to the right half array elements. For the sum pattern, they are all in-phase. In addition, since the amplitude distribution for the sum and difference patterns are identical, both sum and difference patterns can be released on a single physical board. However, due to the fact that the sum pattern amplitude distribution is not calculated directly for a specific SLL, the actual SLL when it is generated with the in-phase difference pattern amplitude distribution will be varied. This factor is considered in the design process.

2.2. Designed excitation coefficients to obtain the sum and difference radiation patterns:

In general, there are four main steps that can be followed to calculate the difference and sum patterns amplitude distribution.

- **Step 1:** For a required difference pattern SLL, sum pattern SLL (R_0) is found by using the graph in Figure 2.3. [2]

Figure 2.3. Sidelobe relationship between sum and difference patterns [2], $n = 2r$, r is the number of elements

- **Step 2:** Using the Dolph-Tschebycheff antenna synthesis technique [3], calculate the corresponding sum pattern amplitude distribution, A'_n .

$$A'_n = \sum_{q=n}^r (-1)^{r-q} (z_o)^{2q-1} \frac{(q+r-2)!(2r-1)}{(q-n)!(q+n-1)!(r-q)!} \quad (1)$$

$$\text{Where: } z_o = \cosh\left(\frac{1}{r-1} \cosh^{-1}(R_o)\right),$$

R_o : magnitude of SLL

- **Step 3:** Obtain difference pattern amplitude distribution as defined in [2]:

$$A_n = (2r - 1)A'_n \quad (2)$$

- **Step 4:** Sum pattern amplitude coefficients are re-defined as A_n

Since the designed difference pattern SLL affects the actual sum pattern SLL, various difference pattern amplitude distribution are computed in order to determine the best design. For conveniences, the three steps of computing difference pattern amplitude distribution are programed in MATLAB.

Furthermore, for an even number of elements (r), the array factor (AF) for the sum and difference patterns are found and used to theoretically plot the design sum and difference patterns.

$$AF_{sum} = 2 \sum_{n=1}^{r/2} A_n \cos\left[\frac{(2n-1)kd}{2} \sin(\theta)\right] \quad (3)$$

$$AF_{dif} = 2 \sum_{n=1}^{r/2} A_n \sin\left[\frac{(2n-1)kd}{2} \sin(\theta)\right] \quad (4)$$

(a)

(b)

Figure 2.4. Sum and difference patterns with SLL (Difference) = -20 dB (a) 4-Elements, Required $R_0 = -32$ dB. (b) 8-Elements, Required $R_0 = -30$ dB.

(a)

(b)

Figure 2.5. Sum and difference patterns with SLL (Difference) = -30 dB (a) 4-Elements, Required $R_0 = -45$ dB. (b) 8-Elements, Required $R_0 = -40$ dB,

(a)

(b)

Figure 2.6. (a) 4-Elements, Difference SLL = -40 dB, Required Sum SLL = 58 dB, (b) 8-Elements, Difference SLL = -40 dB, Required Sum SLL = -51 dB,

(a)

(b)

Figure 2.7. (a) 4-Elements, Difference SLL = -50 dB, Required Sum SLL = 70 dB,
(b) 8-Elements, Difference SLL = -50 dB, Required Sum SLL = -63 dB

For the fabrication and demonstration purposes, only 4-element and 8-element antenna array are considered. Amplitude distribution for difference pattern with SLL = -20 dB, -30 dB, -40 dB, and -50 dB are computed. The theoretical sum and difference patterns are plotted above. Additionally, the results are summarized in table 2.1.

Table 2.1. Sum and difference patterns SLLs with the computed amplitude distribution

Required Difference Pattern SLL (dB)	Sum SLL, R_o (dB)		Excitation coefficients						Actual Sum Pattern SLL (dB)	
	4-	8-	4-Elements		8-Elements				4-	8-
	Elements	Elements	A_1	A_2	A_1	A_2	A_3	A_4	Elements	Elements
-20	-32	-30	0.8	1	0.39	0.94	1	0.71	-10.5	-9.01
-30	-45	-40	0.92	1	0.44	1	0.92	0.45	-12	-10.2
-40	-58	-51	0.97	1	0.47	1	0.8	0.3	-12.6	-11.1
-50	-70	-63	0.99	1	0.49	1	0.72	0.22	-12.89	-11.6

As seen in Table 2.1, increasing SLL of the difference pattern does result higher SLL in the sum pattern. However, the increasing is not significant after the difference pattern SLL is increased to -30 dB. In fact, the SLL of the sum pattern will not exceed -13 dB [3]. Additionally, the theoretical sum and difference pattern plots show that with higher number of antenna elements, the beam gets narrower, which is required for tracking. Due to these reasons, an 8 isotropic linear antenna array with the amplitude distribution of [0.44, 1, 0.92, 0.45] is selected for hardware implementation.

In conclusion, while the amplitude distribution can be implemented by power dividers, the phase difference would require extra RF component, such as a phase shifter. This would defeat the purpose of this study: Generating the sum and difference patterns directly and passively. To eliminate the phase shifter, the novel Composite Right/Left-Handed (CRLH) metamaterial could be used to manipulate the phase angle of the transmission lines. In addition, a dual-band CRLH Phased-Array Antenna [4] has been designed to exclusively study the novel CRLH metamaterial.

3. Dual-Band Composite Right/Left-Handed (CRLH)

Phased-Array Antenna:

Phased-array can be achieved with the conventional materials by manipulating phase delay (α) between the transmission lines (TL) feeding the antenna elements. However, due to the nature of the conventional materials, phase propagation only happens in one direction. As a result, the radiated beams can only be steered in the first quadrant (Figure 3.1-a). On the other hand, CRLH TLs, phase progresses in both directions, which provides not only phase delay but phase advance. This enhances the ability of steering the radiated beams in both first and second quadrants (Figure 3.1-b).

Figure 3.1. Comparison between (a) Conventional and (b) CRLH all passive dual-band phased-array antenna [4]

CRLH metamaterial can be modelled by an equivalent TL (Figure 3.2-a). Comparing to the conventional (i.e., RH) material, which is modelled by a series inductor (L_R) and a shunt capacitor (C_R), CRLH TL model has an extra series capacitor (C_L) and shunt inductor (L_L). These

capacitance and inductance are forced into the RH TL providing the phase advance property. In practice, C_L and L_L can be inserted into the conventional TL as lumped components. As a consequence, the values of C_L and L_L can be properly designed to steer the radiated beam simultaneously in first and fourth quadrants.

Figure 3.2. CRLH TLs models (a) Theoretical circuitry (b) Hardware implementation

3.1. Dual-Band CRLH feed system:

The objective in this chapter is to recreate the design discussed in [4], using HFSS for simulation to obtain the same results. Further, hardware implementation and experimental validation is also done to show that the design presented in [4] has been successfully obtained.

The objective in [4] is to passively direct the main beam angles (θ_0) at $+10^\circ$ and -10° for the operating frequencies of 1.8 GHz and 2.5 GHz, respectively. To avoid mutual coupling and grating lobe, the interelement spacing (d) [5] is designed to be 7.5 cm. Following the procedures described in [4], the required relative phase differences (α) that would provide the phase lag and phase advance to direct the main beams at $+10^\circ$ and -10° are 28° and -39° correspondingly. These phase lag and phase advance are achieved by using CRLH TLs. With the LH material's characteristic impedance of 50Ω , the C_L and L_L are computed $C_L = 1.73$ pF and $L_L = 4.33$ nH. In addition, the

dual-band CRLH phased-array antenna will be fabricated on the Rogers RT/duroid 5880 substrate with the dielectric constant, $\epsilon_r = 2.2$ and thickness of 31 mil. Table 3.1 summarizes the all the design parameters and constraints.

Table 3.1. Dual-Band CRLH feed system designed parameters and constraints

Operating frequency	$f_1 = 1.8 \text{ GHz}$	$f_2 = 2.5 \text{ GHz}$
Relative phase difference (α)	$+28^\circ$	-39°
Desired main beam angle (θ_0)	$+10^\circ$	-10°
C_L	1.73 pF	
L_L	4.33 nH	
Interelement spacing (d)	7.5 cm	
Substrate dielectric constant, (ϵ_r)	2.2	
Substrate thickness, (h)	31 mil	
LH Characteristic impedance, (Z_0)	50 Ω	

To implement the designed parameter and obtain the main beams directed at $+10^\circ$ and -10° , a CRLH feed system is designed (Figure 3.3). The feed system contains two main parts. Part I is a cascaded conventional Wilkinson Power Divider. The power dividers are designed at 2 GHz to evenly divide the power in-phase to the CRLH dispersive lines (25% of the input power). Part II are the dispersive lines, which are responsible for creating the correct phase delay and phase advance. The hardware implementation structure of C_L and L_L is described early (Figure 3.2). Three lumped capacitors are in series with the LH material; two lumped inductors are in parallel with the LH material and directly connected to the ground plane. As seen, the third lumped capacitor has a value of $2 C_L$. This is done for matching purposes. In addition, this structure is defined as a unit cell. The unit cell is capable of creating a phase delay of $+28^\circ$ at 1.8 GHz and a phase advance of -39° at 2.5 GHz simultaneously and passively. To ensure a consistent progressive phase delay and

phase advance between the dispersion lines, the number of lumped components C and L are properly increased and injected. Theoretically, phase delay and phase advance are directly proportional to the number of unit cell. For example, if a unit cell creates a $+28^\circ$ phase delay and -39° phase advance, then two unit cells will create $+54^\circ$ and -78° in phase delay and phase advance correspondingly. The exact number of lumped C and L used are shown in table 3.2.

Figure 3.3.. Dual-Band CRLH Feed system with lumped components implemented.

(I) 4-way Power Divider. (II) CRLH Dispersive Lines

Table 3.2. Number of L and C lumped components injected in each dispersion lines

Dispersion Lines	Number of cells	# of C	# of L	# of C for matching
#1	1	2	2	1
#2	2	4	4	1
#3	3	6	6	1
#4	4	8	8	1

The Dual-Band CRLH feed system is simulated in HFSS. The simulated Return Loss (RL) is shown in figure 3.4. The RL is below -15 dB at the interesting frequencies confirming the performance of the feed system in terms of power.

Figure 3.4. Simulated Return loss of the Dual-Band CRLH feed system

In addition, the simulated progressive phase delay and phase advance between the dispersion lines are generated from 1.5 GHz to 3 GHz (Figure 3.5). As seen, the CRLH TLs behave as expected.

At the designed frequencies 1.8 GHz and 2.5 GHz, the phase responses are lagging and leading simultaneously, which generates the correct phase delay and phase advance to steer the beams at $+10^\circ$ and -10° . The simulated results are comparing to [4] and presenting in Table 3.3. As seen in Table 3.3, the phase delays are very close to the desired ones: -39° at 2.5 GHz and $+28^\circ$ at 1.8 GHz. However, α_2 is only $+18.34^\circ$ at 1.8 GHz. Although, beside the effect of the L-C unit cell, the relative phase difference can also be manipulated by tuning the physical length of the dispersion

Figure 3.5. Progressive phase delay between the dispersion lines

lines. Changing relative phase difference at one interesting frequency will also change relative phase difference at another interesting frequency. In conclusion, this feed system will be integrated with wideband monopole antennas for fabrication and validated using Network Analyzer and anechoic chamber.

Table 3.3. Simulated relative phase difference comparing to [4]

Frequency	Desired α		α_1	α_2	α_3
1.8 GHz	28°	Simulated	+27.34 deg	+18.34 deg	+29.56 deg
		[4]	+29.9 deg	+28.8 deg	+28.5 deg
2.5 GHz	-39°	Simulated	-38.67 deg	-39.39 deg	-39.87 deg
		[4]	-38.4 deg	-39.6 deg	-38.2 deg

3.2. Small Wideband Microstrip-fed Monopole Antenna:

To validate the Dual-Band CRLH feed system design and for the feasibility of the fabrication, a wideband microstrip-fed monopole antenna [7] is designed to operate cross the interesting frequencies, 1.8 GHz and 2.5 GHz. This antenna has a planar structure and is designed in microstrip figure, which are suitable for printed circuit board.

The design process is very similar to designing a microstrip rectangular patch antenna. The rectangular patch must be designed such that it resonances within the desired bandwidth. This can be easily achieved by tuning the length and width of the rectangular patch. However, one of the

Figure 3.6. Microstrip-fed Monopole Antenna Layout

disadvantages of microstrip patch antenna is the narrow operating bandwidth (usually just a couple of kHz). To open up the bandwidth, instead of having a ground plane underneath of the rectangular patch, the ground plane is partially removed and separated at a certain distance from the patch. This allows the electromagnetic fields coupling between the rectangular patch and the ground plane. In addition, two small notches are cut on the rectangular patch to enhance the coupling and achieve a better wideband. The final design with detailed dimensions are shown in Figure 3.6.

The wideband microstrip-fed monopole antenna will be fabricated on the Rogers RT/duroid 5880 substrate with the dielectric constant, $\epsilon_r = 2.2$ and thickness of 31 mil. The design is

modelled and simulated in HFSS. The simulated RL are obtained (Figure 3.7). The simulated return losses are below -15 dB for operating frequencies, 1.8 GHz and 2.5 GHz.

Figure 3.7. Simulated RL of the microstrip-fed Monopole Antenna Layout

3.3. Simulations and hardware validations of the Dual-Band CRLH

Phased-Array Antenna:

3.3.1. Dual-Band CRLH Phased-Array Antenna simulations:

The designed Dual-Band CRLH feed system is integrated with the designed wideband microstrip-fed monopole antenna (Figure 3.8) to create a four-element linear array. Input power is equally delivered to the dispersive lines in-phase (part I); the dispersive lines (part II) create the phase delay and phase advance necessary to steer the beams; linear antenna array (part III) allows the beam to be formed. The design is modelled and simulated in HFSS.

Figure 3.8. Simulated layout of the Dual-Band CRLH Phased-Array Antenna.

(I) 4-way Dower Divider. (II) CRLH Dispersive Lines. (III) Dual-Band Antenna

Figure 3.9.. Simulated Return Loss of the Dual-Band CRLH Phased-Array Antenna

The return loss of the entire circuit (both feed system and antenna) is generated and shown in Figure 3.9. The return losses are both under -9 dB for both 1.8 GHz and 2.5 GHz. In addition, the radiation pattern in the E-plane are simulated and obtained for both desired frequencies (Figure 3.10). The simulated results show that the main beam is directed at $+10^\circ$ at 1.8 GHz and -9° at 2.5 GHz.

Figure 3.10. Simulated E-plane radiation pattern of the CRLH Phased-Array Antenna

3.3.2. Dual-Band CRLH Phased-Array Antenna experimental validations:

The Dual-Band CRLH Phased-Array Antenna is fabricated on the Rogers RT/duroid 5880 substrate. The final design layout is done in CAM315 before sending out for etching. Figure 3.11 shows the simulated and fabricated layouts side-by-side.

Figure 3.11. Dual-Band CRLH Phased-Array Antenna. (a) Simulated (b) Fabricated

The return losses are measured with PNA Network Analyzer (Agilent E8363B). Measurement data is collected and plotted against the simulated result (Figure 3.12).

Figure 3.12. Simulated and Measured RL of the Dual-Band CRLH Phased-Array Antenna

As seen in Figure 3.12, the resonating regions are shifted resulting better RLs than predictions.

The RLs are both below -14 dB for both desired frequencies, 1.8 and 2.5 GHz.

In addition, to verify the beam angles, radiation patterns of the fabricated hardware are obtained in the anechoic chamber. All measurements are done and necessary data are collected at 1.8 GHz and 2.5 GHz. The measured data is normalized and plotted against the simulated data (Figure 3.13).

Figure 3.13. Simulated and Measured E-plane radiation pattern of the Dual-Band CRLH Phase- Array Antenna (a). $f_1 = 1.8$ GHz. (b). $f_2 = 2.5$ GHz

As seen in Figure 3.13, the simulated and measured radiation patterns show the main beam angles towards $\theta_0 \approx 10^\circ$ as desired with good agreement between them. Table 4

Table 4. Dual-Band CRLH Phased-Array Antenna experimental validations

Frequency		RL	Main beam angle (θ_0)	Desired angle (θ_0)
1.8 GHz	Simulated	-9.4 dB	+ 10 deg	+ 10 deg
	Measured	-25 dB	+ 11 deg	
2.5 GHz	Simulated	-33.8 dB	-9 deg	-10 deg
	Measured	-14.1 dB	-8.5 deg	

In conclusion, the main beams are directed at $+10^\circ$ at 1.8 GHz and -10° at 2.5 GHz passively and simultaneously. This is made possible with CRLH transmission lines, which possess the ability of providing phase delay and phase advance naturally. The novel CRLH transmission lines nature are simulated and validated with hardware fabrication. In this particular case, the C-L unit cell was designed for $\pm 10^\circ$ phase delay and phase advance. However, this phase difference can be any desired angle. In other words, if properly designed, a phase difference of 180° can be implemented in the same structure. Hence, a phase shifter could be potentially replaced by an equivalent CRLH transmission line.

4. Sum and difference patterns feed system:

To generate the sum and difference patterns on a single physical board, the designed 8-element isotropic linear array must be released at two different frequencies simultaneously and passively. While the computed amplitude distribution can be implemented by controlling the power distributed to each antenna element, phase difference between them can be manipulated with CRLH metamaterial. To be more specific, conventional Wilkinson power divider and microstrip power dividers are designed to control the power distribution and a dual-frequency CRLH Wilkinson power divider is designed for phase difference. In addition, the sum and difference patterns will be released at 2.6 GHz and 1.8 GHz correspondingly.

4.1. Wilkinson Power Divider with left-handed material:

4.1.1. Conventional Wilkinson Power Divider:

To equally split the input power, different kind of power dividers may be considered, such as a T-junction divider, a resistive divider... However, each kind has its own disadvantages. A T-junction divider cannot be match at all ports, and while a resistive divider can be matched at all ports, it is lossy. Furthermore, both devices don't have any isolation between output ports. As a consequence, the equal-split conventional Wilkinson power divider is selected.

Figure 4.1 shows the diagram of an equal-split Wilkinson power divider in microstrip form and the equivalent transmission line circuit. As seen, the power divider contains two branches with the physical length of $\lambda/4$ (calculated at the designed frequency) with the characteristic impedance of $\sqrt{2}Z_o$ (Z_o is the characteristic of the feeding lines). To make it possible for all three ports matched,

a resistor of $2Z_0$ is added across the two output ports. Using the Even-Odd Mode Analysis, [6] proves that when the divider is driven at port #1 and the outputs are matched, no power is dissipated in the resistor, only the reflected power from ports 2 and 3 is dissipated in the resistor. Thus the divider is lossless. In addition, [6] proves that there is no power associated between port #2 and #3 (the reflection coefficients $S_{23} = S_{32} = 0$), ports 2 and 3 are isolated.

Figure 4.1. The Wilkinson Power Divider. (a) An equal-split Wilkinson power divider in microstrip form. (b) Equivalent transmission line circuit. [6]

To implement the equal-split conventional Wilkinson power divider, a model is developed using the High-Frequency Structure Simulator (HFSS) and presented in Figure 4.2. The microstrip power divider is designed to operate at 2.5 GHz and will be fabricated on a substrate, Rogers RT/duroid 5880 material with a dielectric constant of 2.2. All the physical dimensions are synthesized using the TLR tool and shown in table 4.1:

Figure 4.2. A Model of the conventional Wilkinson Power Divider in HFSS

Table 4.1. Designed dimensions of the equal-split conventional Wilkinson power divider

	Impedance (Ω)	W (mm)	Theoretical length	Physical length (mm)
Feed line (#1)	$Z_o = 50$	2.34	N/A	338
Output lines (#2, #3)	$Z_o = 50$	2.34	$\lambda/2$	1719
Half of the ring	$\sqrt{2}Z_o = 70.7$	1.35	$\lambda/4$	873
Resistor (603)	$2Z_o = 100$	1.52	N/A	0.75

Figure 4.3. Return Loss of the Conventional Wilkinson Power Divider.

The power divider is simulated and the return loss (RL) is plotted and shown in figure 4.3. As seen the RL is -20.4 dB at the designed frequency of 2.5 GHz. In addition, the S parameter magnitudes are plotted in figure 4.4. S_{21} and S_{31} are almost identical, which demonstrates the isolation between the output ports 2 and 3.

Figure 4.4. Frequency response of the equal-split Wilkinson power divider.

To confirm the power division ratio, the Output Power Ratio (OPR) is calculated and plotted in figure 4.5. The output power at one port with respected to the input port can be calculated as a product of the reflected wave and its conjugate, which is the magnitude of the reflection coefficient at that port squared. Thus $OPR = \left| \frac{mag(S_{21})}{mag(S_{31})} \right|^2$. As seen, OPR is a flat line equal to 1. This demonstrates that the input power is equally split between the two output ports 2 and 3.

Figure 4.5. Output Power Ratio between Port #2 and Port #3 of the Conventional Wilkinson Power Divider

Finally, the phase responses of the two output ports with respected to the input port are plotted in figure 4.6. Since, it's a conventional power divider, the phase response is the same at the two output ports. This is demonstrated in figure 4.6. The two output ports are always in-phase. However, to release the sum and difference radiation patterns simultaneously on a single physical board, all the antenna elements need to be fed in-phase at one frequency (for the sum pattern) and one half out-of-phase with respected to the other half at a different frequency (for the difference

pattern). In other words, a dual-frequency power divider that is in-phase at one frequency and out-of-phase at a different frequency needs to be designed.

Figure 4.6. Phase Angles of the output Ports respected to Input Port

4.1.2. Dual-Frequency CRLH Wilkinson Power Divider:

To design such a device described in the previous section, the CRLH metamaterial was intentionally studied in the Dual-Band CRLH Antenna Phased Array project (section 3). Due to the property of the left-handed material, the phase response can be manipulated, so that instead of being identical, the phase response of the two output ports are lagging with respect to each other. This will guarantee some phase difference (out-of-phase) before they're crossing (in-phase).

Figure 4.7. The Dual-Frequency CRLH Wilkinson power divider

Using the concept of implementing L and C lumped component in the dispersion line inspired by [4], a dual-frequency Wilkinson power divider is designed (figure 4.7), and will be fabricated on a substrate, Rogers RT/duroid 5880 material with a dielectric constant of 2.2. The objective is that the output power at ports 2 and 3 will be out-of-phase at 1.8 GHz and in-phase at 2.6 GHz. This is

achieved by injecting a L-C unit cell, which contains 2 capacitor, $C_L = 1.7$ pF, 2 inductors $L_L = 4.3$ nH, and a matching capacitor with the capacitance of $2C_L = 3.4$ pF. In addition, the unit cell is specifically designed and tuned to work at 1.8 GHz. All the physical dimensions are synthesized by the TRL tool and presented in table 4.2. Furthermore, to control the in-phase frequency, beside the effect of the L-C unit cell, extra physical length is added to the port #3 output transmission line.

	Impedance (Ω)	Theoretical length	Physical width (mil)	Physical length (mil)
Feed line (port 1)	$Z_o = 50$	N/A	92	338
Output line (port 2)	$Z_o = 50$	$\lambda/2$	92	1719
Output line (port 3)	$Z_o = 50$	N/A	92	1719 + 3739
Half of the ring	$\sqrt{2}Z_o = 70.7$	$\lambda/4$	53	873
Resistor (603)	$2Z_o = 100$	N/A	60	30
Capacitor, C_L (603)	1.7 pF	N/A	60	30
Inductor, L_L (603)	4.3 nH	N/A	60	30

Table 4.2. Designed dimensions of the dual-frequency CRLH Wilkinson power divider

Figure 4.8. Output power out of phase at 1.8 GHz and out of phase at 2.61 GHz

The dual-frequency CRLH Wilkinson power divider is simulated in HFSS. Figure 4.8 shows the simulated phase response of the power divider. As seen, since the unit cell is only implemented on the port #3 output line, S_{21} is identical in both cases, with and without the L-C unit cell (black and purple plots). However, under the effect of the unit cell, S_{31} is shifted from red to blue plot, resulting a 179.51° phase difference at 1.8 GHz and 0.38° phase difference at 2.61 GHz. In other words, the output power is out-of-phase at 1.8 GHz and in-phase at 2.61 GHz. Table 4.3 summaries the phase response changed in both cases. In addition, the RL is plotted in figure 4.9. At the operating frequencies of 1.8 GHz and 2.61 GHz, the RL is -21.2 dB and -15.1 dB respectively.

Table 4.3. Simulated phase response of the Dual-Frequency CRLH power divider

Frequency	1.8 GHz		2.61 GHz	
	Without CRLH cell	With CRLH cell	Without CRLH cell	With CRLH cell
S_{21}	143°	140.5°	41.1°	41.2°
S_{31}	-159.3°	-39.01°	-38.5°	40.82°
Phase difference	302.2° or 122.3°	179.51°	79.6°	0.38°

Figure 4.9. Return Loss of a dual-frequency power divider

4.1.3. Validation of a fabricated dual-frequency CRLH Wilkinson Power

Divider:

4.1.3.1. Cascaded Dual-Frequency CRLH Wilkinson Power Divider:

To test the designed conventional and the Dual-Frequency CRLH Wilkinson power dividers, two conventional Wilkinson power dividers are cascaded with a Dual-Frequency CRLH Wilkinson power divider (Figure 4.10). Theoretically, the Dual-Frequency CRLH Power divider (part I) will not only behave as a power divider (splitting the input power equally), but also a 180 degrees phase shifter at 1.8 GHz. Consequently, at this frequency, the left half of the circuitry (P2 and P3) are out-of-phase with respect to the right half of the circuitry (P4 and P5); while P2 and P3 are in phase as well as P4 and P5. In addition, they will all be in-phase at 2.5 GHz.

Figure 4.10. Cascaded Dual-Frequency CRLH Wilkinson Power Divider. (I) Dual-Frequency CRLH Wilkinson Power Divider (II) Conventional Wilkinson Power Dividers

Figure 4.11. Simulated phase responses of the Cascaded Dual-Frequency CRLH Power Divider

The Cascaded Dual-Frequency CRLH Wilkinson Power divider is simulated in HFSS. The phase responses between the dispersive lines are generated and plotted in Figure 4.11. As shown, P2 and P3 are in-phase cross the bandwidth as well as P4 and P5. However, at 1.8 GHz, P2 and P3 are out-of-phase with respect to P4 and P5 (180° phase difference); at 2.5 GHz, they are all in-phase (0° phase difference). Furthermore, the return loss is obtained and plotted in Figure 4.12. The RLs are both below -12 dB for both interesting frequencies of 1.8 GHz and 2.5 GHz.

Figure 4.12. Simulated RL of the Cascaded Dual-Frequency CRLH Wilkinson Power Divider

4.1.3.2. Fabrication and experimental validations:

The Cascaded Dual-Band CRLH Wilkinson power divider is fabricated on the Rogers RT/duroid 5880 substrate. The final design layout is done in CAM315 before sending out for etching. Figure 4.13 shows the simulated and fabricated layouts side-by-side.

Figure 4.13. Cascaded Dual-Frequency CRLH Power Divider. (a) Simulated. (b) Fabricated.

Figure 4.14.. Measured dispersive line phase responses

The phase responses of the dispersive lines are measured using the PNA Network Analyzer (Agilent E8363B). Measurement data is collected and plotted in Figure 4.14. As seen, S_{21} and S_{31} are always in-phase as well as S_{41} and S_{51} . In addition, S_{21} and S_{31} are out-of-phase with respect to S_{41} and S_{51} at 1.8 GHz; they are all in-phase at 2.5 GHz. Table 4.4 summarizes the experimental validation results comparing to the simulated results.

Table 4.4.. Simulated and measured phase angle of the dispersion lines

Frequency		S ₂₁	S ₃₁	S ₄₁	S ₅₁
1.8 GHz	Simulated	-70.6°	-70.9°	107.8°	107.9°
	Measured	-108°	-107.2°	72.5°	74°
		In- Phase (Left half)		In-phase (Right Half)	
		180 ⁰ phase difference between the left and right half			
2.5 GHz	Simulated	93.2°	93°	93.9°	94°
	Measured	59.3°	61°	57.2°	59.1°
		all elements in-phase			

In conclusion, the designed Dual-Frequency CRLH behaves as desired. Under the effect of the CRLH transmission line (on one of the output arms), half of the circuitry is out-of-phase with respect to the other half at one frequency and all is in-phase at a different frequency. This is a key component to directly generate the sum and difference patterns on a single physical board simultaneously and passively.

4.1.4. Improving Dual-Frequency CRLH Wilkinson Power Divider:

As mentioning in the previous section, to control the in-phase frequency, extra physical length was added to the port #3 output transmission line. In the attempt of controlling the in-phase frequency completely with left-handed material, different unit cells were studied by increasing the number of L and C lumped components as well as changing the capacitance and inductance of the unit cell. Figure 4.15 shows the equivalent of a CRLH metamaterial and a conventional material TLs at 1.8 GHz. In this particular case, the unit cell contains 4 lumped capacitors, $C_L = 1.7$ pF, 4 lumped inductors, $L_L = 4.7$ nH, and 1 matching lumped capacitor, $2C_L$. As shown, a 60.2 mm conventional material TL can be replaced by a L-C unit cell and still have the same effect. Figure 4.16 presents the simulated phase response of the two cases. At the designed frequency of 1.8 GHz, the phase delay is 175.7 degrees resulting from the conventional material TL and 177.3 degrees from the CRLH metamaterial TL. This demonstrates that for a certain degrees of phase delay created by conventional TL, it can also be done by a proper set of L and C lumped components. Furthermore, in general, the phase delay can be increased by adding more lumped capacitors and inductors to the unit cell.

Figure 4.15. Equivalent physical length and L-C unit cell, $C_L = 1.7$ pF, $L_L = 4.7$ nH

Figure 4.16. Phase response of the actual physical length and equivalent unit cell

The equivalent L-C unit cell is then implemented into one output line of the conventional Wilkinson power divider (Figure 4.17). Figure 4.18 shows the phase difference of 179.65 degrees (out-of-phase) at the designed frequency of 1.8 GHz. However, the in-phase frequency is pushed up to 3.85 GHz. In a monopulse tracker, the sum and difference pattern can be released at different frequencies, but ideally these two frequencies need to be closer or within in a same frequency band. Originally, the difference and sum patterns are designed to be realized at 1.8 GHz and 2.6 GHz correspondingly. With this alternative design of the power divider, the two operating frequencies are 1.8 GHz and 3.85 GHz. This large gap would potentially affect the detection ability. Since the wavelengths are significantly different at 1.8 GHz and 3.85 GHz, the tracking object may be visible at one operating frequency but not at the other. In addition, larger gap between the operating frequencies would require antennas with larger bandwidth.

Figure 4.17. An equal-split Wilkinson power divider with increased number of L and C lumped components.

Figure 4.18. Phase response at the output ports

In conclusion, even though it is possible to reduce the physical length by designing a proper L-C unit cell, it is more challenging in terms of controlling the in-phase frequency. Because of the large gap between the in-phase and out-of-phase frequencies of this design, it will not be used for the final design due to the fabrication and feasibility purposes.

4.2. Implementation of excitation coefficients:

In order to implement the magnitude of the excitation coefficients, the power feeding into each antenna element needs to be properly controlled. In other words, the excitation coefficient magnitudes are directly proportional to the percentage of the power distributed into each antenna element.

To implement $[0.44:1:0.92:0.45]$, two power dividers are needed. One is to split the power at a division ratio of $0.44:1 = 0.44$ and another one at $0.92:0.45 \sim 2.04$ ratio. This is done by two microstrip power dividers.

4.2.1. Microstrip power divider:

Figure 4.19. Equivalent transmission line circuit of a microstrip power divider.

A microstrip power divider is simply a T-junction power divider, which is a 3-port network. Since, any 3-port network cannot be lossless, reciprocal, and all ports matched simultaneously, the lossless and reciprocal microstrip power divider cannot be perfectly

matched at all three ports. In other words, depending on the designed power division ratio, the characteristic impedance of the transmission lines, Z_{01} and Z_{02} , need to be calculated accordingly.

Based on the transmission line theory, the input impedance of a transmission line can be calculated as: $Z_{in} = Z_0 \left[\frac{Z_L + jZ_0 \tan(\beta l)}{Z_0 + jZ_L \tan(\beta l)} \right]$

For $l = \frac{3\lambda}{4}$, Z_{in} is simplified as $\frac{Z_0^2}{Z_L}$. Hence, $Z_{1,AA}$ and $Z_{2,AA}$ are obtained below:

$$Z_{1,AA} = \frac{Z_{01}^2}{Z_L} \text{ and } Z_{2,AA} = \frac{Z_{02}^2}{Z_L} \rightarrow Z_{TAA} = \frac{Z_{1,AA} Z_{2,AA}}{Z_{1,AA} + Z_{2,AA}} = \frac{Z_{01}^2 Z_{02}^2}{(Z_{01}^2 + Z_{02}^2) Z_L}$$

$$\text{For impedance matching, } Z_{TAA} = Z_0 \rightarrow \frac{Z_{01}^2 Z_{02}^2}{(Z_{01}^2 + Z_{02}^2) Z_L} = Z_0 \quad (4)$$

In addition, $P_1 = \frac{V_0^2}{2Z_{1,AA}} = \frac{V_0^2 Z_L}{2Z_{01}^2}$. Similarly, $P_2 = \frac{V_0^2 Z_L}{2Z_{02}^2}$.

For a power division ratio, $\alpha = \frac{P_1}{P_2}$, we have:

$$\frac{P_1}{P_2} = \frac{Z_{02}^2}{Z_{01}^2} = \alpha \rightarrow Z_{02} = \sqrt{\alpha} Z_{01} \quad (5)$$

From (4) and (5), the expressions for Z_{01} and Z_{02} are derived as:

$$Z_{01} = \sqrt{Z_0 Z_L} \sqrt{1 + \frac{1}{\alpha}} \quad (6)$$

$$Z_{02} = \sqrt{Z_0 Z_L} \sqrt{1 + \alpha} \quad (7)$$

One might notice, for this to work, the transmission line lengths must be $\frac{(2n+1)\lambda}{4}$, with $n = 0, 1, 2, \dots$ so that for $\tan(\beta l)$ is undefined, $Z_{in} = \frac{Z_0^2}{Z_L}$. Since the spacing between the elements is $d = \lambda/2 \sim 83.33\text{mm}$, the length of the TLs must $\frac{3\lambda}{4}$ ($n = 1$) at minimum to ensure a proper physical layout.

4.2.2. Microstrip power divider with a power division ratio of 0.44 and 2.04:

With $Z_0 = 50 \Omega$, operating frequency of 1.8 GHz, using the derived expression (6) and (7), table 4.5 summarizes of physical dimensions of the power dividers. In addition, using the TLR tool, the physical length of $\frac{3\lambda}{4}$ at 1.8 GHz on the Rogers RT/duroid substrate is synthesized to be roughly 91.2 mm.

Table 4.5. Physical dimension of the microstrip power dividers

Amplitude ratio	Characteristic impedance (Ω)		Physical width (mm)
0.44	Z_{01}	90.45	0.85
	Z_{02}	60	1.77
2.04	Z_{01}	61	1.76
	Z_{02}	87.24	0.9

Figure 4.20. Modeled microstrip power divider with a power division ratio of 0.44

Figure. 4.21. Modeled microstrip power divider with a power division ratio of 2.04

To verify the power division ratio, the output power at one port with respected to the input port can be calculated as a product of the reflected wave and its conjugate, which is the magnitude of the reflection coefficient at the port squared. Thus the output power ratio between ports #2 and #3

is $\alpha = \left| \frac{mag(S_{21})}{mag(S_{31})} \right|^2$. The results are summarized in table 4.6.

Table 4.6. The microstrip power divider power division verification

Amplitude ratio	Mag(S_{21})	Mag(S_{31})	$\alpha = \left \frac{\text{mag}(S_{21})}{\text{mag}(S_{31})} \right ^2$
0.44	0.56	0.82	0.47
2.04	0.8	0.57	1.97

4.2.3. Final design of the feed system:

For the designed 8-element antenna array, all the excitation coefficient magnitudes are: 0.45:0.92:1:0.44:0.44:1:0.92:0.45. It is symmetric. The left half, [0.45:0.92:1:0.44], is mirrored from the right half, [0.44:1:0.92:0.45]. Considering the left half of the feed system and assuming 100% of the input power is distributed, according to the excitation coefficient magnitudes, the power distributed in to each element are approximately 16, 36, 33, and 16% or [0.16, 0.36, 0.33, 0.16]. Figure 4.22 shows a model of the left half of the feed system. The model contains two microstrip power dividers and a conventional Wilkinson power divider. The two microstrip are designed with a specific power division ratio previously, which will ensure the correct amount of power distributed in to the antenna. In addition, a conventional Wilkinson power divider is used to equally split the power in to each microstrip power divider. Furthermore, all the elements are spaced with the same distance, d .

Figure 4.22. Modeled the left half of the feed system

Figure 4.23. Power ratio between the output ports, left half of the feed system

The Output Power Ratio (OPR) is calculated and plotted in Figure 4.23. Since all the RF components are frequency dependent, the microstrip power dividers were designed at 1.8 GHz with specific power division ratios, these ratios are expected to be different at 2.6 GHz. The output power ratios between ports 2, 3, 4, and 5 and the input port 1 at the operating frequencies are obtained in table 4.7.

Table 4.7. Output power ratios at the operating frequencies, left half of the feed system

	1.8 GHz		2.6 GHz
	Expected	Simulated	
P ₁	0.16	0.14	0.19
P ₂	0.36	0.31	0.26
P ₃	0.33	0.3	0.27
P ₄	0.16	0.15	0.18

As seen in Table 4.7, due to the power loss in the transmission lines, the output power ratios are slightly lower than the expected values at 1.8 GHz. In fact, the simulated RL at 1.8 GHz is -12.9 dB.

Similarly, considering the right half of the system, [0.44:1:0.92:0.45] and assuming that 100% of the input power is distributed, the power distributed in to each element are approximately 16, 33, 36, and 16% or [0.16, 0.33, 0.36 0.16]. A mirrored version of the left half of the feed system (Figure 4.24) is simulated as the right half.

Figure 4.24. Modeled the right half of the feed system

Figure 4.25. Power ratio between the output ports, right half of the feed system

Table 4.8. Output power ratios at the operating frequencies, right half of the feed system

	1.8 GHz		2.6 GHz
	Expected	Simulated	
P ₁	0.16	0.14	0.19
P ₂	0.33	0.31	0.25
P ₃	0.36	0.31	0.27
P ₄	0.16	0.15	0.18

The OPR of the right half is calculated and plotted in Figure 4.25. The power ratios are obtained and presented in Table 4.8. In addition, the simulated RL is approximately -13.3 dB in this case.

To ensure the spacing, d , between the elements, the output TLs of the Dual-Frequency CRLH Wilkinson Power Divider are stretched out (Figure 4.26). The physical length of each TL is tuned so that it is in-phase at 1.8 GHz and out-of-phase at 2.6 GHz.

Figure 4.26. Modified the Dual-Frequency CRLH Wilkinson Power Divider

Figure 4.27 shows the final design of the feed system. It contains the left and right half and the modified Dual-Frequency CRLH Wilkinson Power Divider.

Figure 4.27. Final design of the feed system layout

The feed system design is simulated in ANSYS HFSS. Figure 4.28 and 4.29 show the simulated results of RL and phase response at the output ports.

Figure 4.28. Final design of the feed system RL

Figure 4.29. Final design of the feed system phase response of the output ports

In addition, the power distributed at each output port is calculated by squaring the reflection coefficient and plotted in Figure 4.30. Table 4.9 summaries the phase angles and power distribution of the feed system.

Figure 4.30. Final design of the feed system power ratios of the output ports

Table 4.9. Phase angle and power distribution at the output ports

	1.8 GHz			2.5 GHz	
	Phase angle [degrees]	Power distribution		Phase angle [degrees]	Power distribution
		Expected	Simulated		Simulated
P ₂	-6.19	0.08	0.08	-138.75	0.08
P ₃	-6.25	0.164	0.168	-125.55	0.12
P ₄	-9.54	0.178	0.17	-131.163	0.12
P ₅	-3.22	0.078	0.078	-132.79	0.09
P ₆	173.04	0.078	0.0625	-128.45	0.08
P ₇	166.94	0.178	0.14	-127.45	0.11
P ₈	170.41	0.164	0.14	-121.81	0.12
P ₉	170.39	0.08	0.07	-134.33	0.08

As seen in Table 4.9, at 1.8 GHz, the left half of the feed system is out-of-phase with respect to the right half. This implements the sign of the excitation coefficients and it is critical to obtain the difference pattern. In addition, the power distributions on the left half of the feed system have a good agreement with the theoretical expectations. However, the power distributions on the right half of the feed system follow the power ratios [0.44:1:0.92:0.45] but are lower than expected. This could be potentially caused by the power dissipated on the L and C lumped components in the output line of the Dual-Frequency CRLH Wilkinson Power Divider. In other words, the Dual-Frequency CRLH Wilkinson Power Divider did not split the input power equally when implementing with other parts. Furthermore, at 2.5 GHz, all the output ports are in-phase and the power are more evenly distributed. This will give a lower SLL for the sum pattern due to the nature of isotropic linear antenna array.

In conclusion, the sign and magnitude of the designed excitation coefficients are implemented in microstrip transmission line hardware figure. However, in order to obtain the sum and difference patterns, the feed system needs to connect to antennas that allow the electromagnetic waves radiated.

5. Dual Band Microstrip Patch Antennas:

5.1. Design concepts:

Since the sum and difference patterns are realized at different frequencies, a dual-band or a wideband antenna is needed. Previously, by properly cutting the ground plane, a monopole antenna was designed to work cross a wide band. However, due to this reason, the electromagnetic fields radiates in both directions, above and below the microstrip patch. For monopulse tracker, the beam needs to be directed in one direction as strong as possible. To block the waves radiating below the microstrip patch, a solid ground plane is required. In other words, the antennas need to be designed in the microstrip figure.

However, the microstrip patch antenna has a very narrow bandwidth. Fortunately, different techniques have been developed to realize a dual band microstrip patch antenna in literature. The simplest way is to cut several slots (E shape, U shape, V shape, or rectangular shape...) on the microstrip patch to make it resonating at an extra frequency. By doing this, neither the patch size is increased nor the radiation pattern is significantly affected. Figure 5.1 presents the layout of a dual band microstrip patch antenna with two rectangular slots. This layout is used to design a dual band antenna operating at 1.8 GHz and 2.5 GHz. Note that without the two slots, it is just a rectangular microstrip patch antenna. Hence, the resonate frequency can be realized by tuning the dimensions of the patch, L and W. It depends on the dielectric constant of the fabricated material and the designed frequency, the actual physical dimensions can be calculated.

Figure 5.1. Layout of Dual Band Microstrip Patch Antenna with rectangular slots [8]

Once the first resonance frequency is defined, [8] shows that the second resonance frequency can be realized without changing the dimensions of the patch. However, [8] doesn't demonstrate how to determine the location of the slots. In addition, the size of the rectangular slots and the spacing between them need to be tuned for different substrate. This is problematic. Furthermore, since the antenna is attached to the end of the feed system and each output line of the feed system has different characteristic impedance, W_f will be different. As a result, F_i needs to be tuned for impedance matching. Fortunately, only 4 different antenna are required to realize the sum and difference patterns since the 8-element antenna array is symmetrical. Throughout the design and tuning process, these patterns are noticed:

- Increasing the slot width will shift the second resonance frequency down in the frequency band.
- Increasing the slot length will shift the second resonance frequency down in the frequency band, but also shift the first resonance frequency up in the frequency band.

5.2. Modeling and simulations:

The High Frequency Structure Simulator (HFSS) is used to model and simulate the dual band microstrip patch antennas, operating at 1.8 GHz and 2.6 GHz. As discussed in 5.1, a rectangular microstrip patch antenna is designed at 1.8 GHz. For the Rogers RT/duroid material ($\epsilon_r = 2.2$) and a 50Ω feed line, W and L are found to be 60mm and 55.5mm correspondingly. To realize the second resonance frequency, two rectangular slots are cut inside the patch. While this is necessary, it will change the impedance of the microstrip patch. In addition, the required characteristic impedance of the feed lines are different. Hence F_i are specifically tuned for impedance matching. Note that to realize the second resonance frequency at 2.6 GHz, the location and dimensions of the slots are critical. As a result, the width of the slots are kept constant at 1mm to reduce the amount of tuning. The length of the slots (L_s), the spacing between them (Y), and the edge-to-edge distance from the microstrip patch and the slots (D) are varied. These three factors (L_s , Y, and D) simultaneously affect the second resonance frequency and are different for different W_f . In other words, L_s , Y, and D need to be tuned separately 4 times since the feed system results 4 different required W_f . In addition, the two rectangular slots are centered from the center of the patch. Table 5.1 summarizes the final design dimensions for each required W_f

Table 5.1 Dimensions and locations of the rectangular slots for different W_f

W_f (mm)	F_i (mm)	L_s (mm)	Y (mm)	D (mm)
0.85	16	23	40	2.5
1.77	20	23	40	2.5
1.76	18	25	40	5.5
0.9	16	23	40	2.5

(a)

(b)

Figure 5.2. Dual Band Microstrip Antenna with $W_f = 0.85$ mm. (a) Simulation model with physical dimensions. (b) Simulated Return Loss

(a)

(b)

Figure 5.3. Dual Band Microstrip Antenna with $W_f = 1.77$ mm. (a) Simulation model with physical dimensions. (b) Simulated Return Loss

(a)

(b)

Figure 5.4. Dual Band Microstrip Antenna with $W_f = 1.76$ mm. (a) Simulation model with physical dimensions. (b) Simulated Return Loss

(a)

(b)

Figure 5.4. Dual Band Microstrip Antenna with $W_f = 0.9$ mm. (a) Simulation model with physical dimensions. (b) Simulated Return Loss

Each antenna with a specific designed dimensions is simulated and the RL is obtained to verify the performance (Table 5.2). Overall, the antennas perform as expected.

Table 5.2. Verifying the designed Dual Band Microstrip Patch Antenna with R

Characteristic Impedances (Ω)	W_f (mm)	Operating Frequencies (GHz)	Return Loss (dB)
90.45	0.85	1.8	- 10.3
		2.58	- 9.8
60	1.77	1.8	- 17.2
		2.6	- 23.6
61	1.76	1.8	- 23.2
		2.56	- 18.8
87.24	0.9	1.8	- 9.7
		2.58	- 9.9

6. Monopulse tracker with sum and difference patterns:

Putting together the feed system and antennas (Figure 6.1), the designed linear array of 8 isotropic elements is implemented to generate the sum and difference patterns. The hardware implementation contains three main parts. While part I is responsible for a 180 degrees phase difference between the left half (P_2, P_3, P_4, P_5) and the right half (P_6, P_7, P_8, P_9), part II controls the power distribution feeding into antennas. Additionally, part III generates the beam to form the sum and difference patterns. The design is simulated in HFSS. The simulated RL (Figure 6.2) and radiation patterns (Figure 6.3 and 6.4) are obtained and plotted.

Figure 6.1. Hardware implementation of the monopulse tracker with sum and difference patterns.

Figure 6.2. Simulated RL of the monopulse tracker with sum and difference patterns

Figure 6.3. Normalized simulated sum and difference radiation patterns in rectangular form

Figure 6.4. Normalized simulated sum and difference radiation patterns in polar form.

As seen in Figure 6.3 and 6.4, the two main lobes in the difference pattern is unequal. This is mainly due to the amount power distributed to the antenna elements, which is controlled by several microstrip power dividers (part II). As emphasizing in 4.2, the length of the output lines are critical and have to be $\frac{3\lambda}{4}$. However, the output lines also have different characteristic impedance or different width (W_f), the antenna feed lines (F_i) are cut into the microstrip patch at a specific distance for impedance matching. These extra lengths are added to the output line of the microstrip power dividers. As a consequence, the power distributions are changed.

Furthermore, the feed lines (F_i) that are inside the microstrip patch can be accounted as a part of the microstrip power divider output lines. The feed system and antennas can be integrated as shown in Figure 6.5. Although this will ensure the correct amount of power distributed to the antennas,

the antenna elements are not placed the same plane. Consequently, the beam forming will be disturbed. Hence the sum and difference patterns cannot be obtained.

Figure 6.5. Different way to integrate the feed system and antennas

7. Conclusions and future research:

In this research, a dual-band phased-array antenna with Composite Right/Left-Handed (CRLH) transmission lines is designed and simulated for releasing sum and difference patterns directly and passively.

In order to realize the sum and difference patterns directly on a single physical board, the amplitude distribution of an 8 isotropic linear antenna array is designed using the Dolph-Tschebycheff antenna synthesis technique. All the antenna elements are excited in-phase to generate the sum pattern; one half of the antenna array are excited out-of-phase (180° phase difference) with respect to the other half to release the difference pattern. For hardware implementation, a dual-band Wilkinson power divider has been designed for in-phase output at 2.5 GHz and a phase difference of 180° at 1.8 GHz by injecting a CRLH unit cell that contains two capacitors ($C_L = 1.7$ pF), two inductors ($L_L = 4.3$ nH), and one matching capacitor ($2C_L$) in one of the output arms. In addition, microstrip power dividers have designed to implement the amplitude distribution. These are incorporated with dual-band microstrip antennas to obtain the sum and difference patterns.

While the amplitude distribution of an 8 isotropic linear array are computed using MATLAB, hardware design optimization and simulation have been done using ANSYS HFSS. A CRLH unit cell, inspired by available research in the literature, has been designed and implemented in one of the output arms of a conventional Wilkinson power divider resulting an output of 180° phase difference at 1.8 GHz and 0° phase difference at 2.6 GHz. This Dual-Band CRLH Wilkinson power divider has been fabricated and validated using Network Analyzer. The difference pattern has been designed with a side lobe level (SLL) of -30 dB. This results a SLL of

-12 dB for the sum pattern. In addition, the sum and difference patterns are theoretically generated with MATLAB to verify the designed SLL. Furthermore, four different microstrip power dividers have been designed to split the power properly between antenna elements based on the computed amplitude distribution. Dual-band microstrip antennas operating at 1.8 GHz and 2.6 GHz are designed and equally spaced out at a distance $d = \lambda/2$ to avoid the mutual coupling. The design is modeled and simulated in HFSS and for the Rogers RT/duroid substrate with a dielectric constant of 2.2 and the thickness of 31 mil. All the fabricated hardware are validated using the Network Analyzer and anechoic chamber.

Several avenues of research are left for future work:

- Fabrication and experimental validation of the dual-band phased-array antenna with Composite Right/Left-Handed transmission lines.
- Improving the power distribution to obtain equal beams for the difference pattern.
- Improving side lobe level for the sum pattern.
- Reducing the length of the CRLH transmission line that yields 180° phase difference.
- For controlling the second interesting frequency, different values of L and C and numbers of lumped components need to be explored.

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