Passive Reconfigurable Dual Linear and Dual Circular Polarization with CLRH-TL for Microstrip Patch Antennas

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Passive Reconfigurable Dual Linear and Dual Circular Polarization with CLRH-TL for Microstrip Patch Antennas

by

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Thesis
Submitted in Partial Fulfillment of the Requirements for the Degree of
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Passive Reconfigurable Dual Linear and Dual Circular Polarization with CLRH-TL for Microstrip Patch Antennas

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Abstract

Passively controlled reconfigurable antennas are desirable for their relatively inexpensive cost compared to the intensive implementation of RF integrated circuits. Extensive research continues to go into structures utilizing frequency selective metamaterials and their applications in reconfigurable antennas in order to find alternatives to active components. Composite left/right-handed transmission lines (CLRH-TLs), consisting of series capacitors and shunt inductors, are one such structure that is capable of operating as a left-handed transmission line at select frequencies. In this thesis, we propose and investigate a unique frequency selective open circuit utilizing CLRH-TLs that operates as an electrical open and a transmission line at two different frequencies, as well as a quadrature hybrid that is configured with these structures to create the required power and phase distribution for dual linear and dual circular polarizations at four different frequencies. The frequency selective open circuit is verified through simulation and validated through prototyping. The modified quadrature hybrid is integrated in the feed system of an dual linear polarized aperture coupled square microstrip patch antenna that is set in a stacked arrangement to increase the bandwidth to 17% (10.7-12.7GHz). Equivalent circuits to the modified quadrature hybrid demonstrate the four polarizations at four frequencies.
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1. Introduction:

The distribution of electromagnetic waves in accordance to design criteria lies at the heart of microwave engineering; without distribution networks there is no microwave transmission. Antennas still operate entirely in the analog domain, propagating sinusoids that have undergone modulation, and most likely will continue to do so in the near future. It follows, then, that improvements to the techniques of feed network design serve to benefit a broad range of applications across the industry. Research into microwave frequency technologies has been performed globally since the mid-20th century, spanning countless applications; a relatively recent development, however, is the discovery of manufacturable metamaterials.

First theoretically proposed as early as 1904[2][3], metamaterials are a type of artificial structure that exhibits electrical properties not seen in natural materials. Negative-index metamaterials, a type of metamaterial that possesses negative permeability and permittivity, were not conceived until 1967[4], and due to the limitations of manufacturing capabilities it was not until the year 2000 that the first negative-index metamaterial structures were produced and validated through measurement[1]. With their negative index of refraction these structures alter the phase velocity of incoming wave, altering it to be antiparallel to the group velocity; this effectively phases the wave in reverse inside the material, hence they are also referred to as left-handed materials as they follow...
an equivalent “left hand rule” for the relation of E and H-fields rather than the right hand rule of natural materials[5].

\[
\left( \frac{1}{j\omega C'} \right) \, dz \\
\left( \frac{1}{G'} \right) \, dz \\
\left( \frac{1}{j\omega L'} \right) \, dz \\
\left( \frac{1}{R'} \right) \, dz
\]

Figure 1.2: Infinitesimal Circuit Model of First Proposed Left-Handed Transmission Lines[6]

Applications for left-handed materials in microwave engineering were found soon after their experimental validation; included among them is concept of a left-handed transmission line. Opposing a right-handed transmission line that phases propagated waves forwards, a left-handed transmission line phases propagated waves backwards. As left-handed transmission lines, similarly to left-handed materials, do not exist naturally, the artificial induction of left-handed behavior has been investigated through the use of unit cells[5]. Work has been done to examine an equivalent unit length model of left-handed transmission lines, determining it to function conversely to the unit length model of a right-handed transmission line. While a right-handed transmission line functions as a continual series inductance and shunt capacitance, a left-handed transmission line would function as a continual series capacitance and shunt inductance; the introduction of electrically small lumped capacitances and inductances to a standard transmission line has been used to emulate this and induce left-handed behavior at select frequencies[5]. Because these transmission lines propagate waves right-handedly at frequencies outside of left-handed operation they have been termed composite left/right-handed transmission lines (CLRH-TLs).
CLRH-TLs have been extensively researched for a variety of applications due to their relatively inexpensive components and frequency selective behavior. Previous work has utilized the frequency selective phasing to alter an existing circuit to function in multiple operating bands\cite{7}\cite{8}, provide passive control over phased feed networks\cite{9}\cite{10}, or create functionality that would otherwise be impossible for right-handed transmission lines alone\cite{11}. In the field of reconfigurable antennas, a type of antenna that is capable of reversibly altering any of its design parameters, CLRH-TLs are utilized to provide passive control of the reconfigurability as opposed to the active control of semiconductor or MEMs devices\cite{12}\cite{13}. Reconfigurable antennas are desirable for the flexibility they add to systems, and passively controlled reconfigurable antennas are desirable on top of that for their reduced component cost as well as their usage in zero power devices.

Zero power devices are a type of device that requires no additional input power to operation. Passively reconfigurable antennas can be seamlessly integrated into zero power devices with proper system design as the reconfigurability is not dependent active devices; through different input frequencies different functionality may be achieved. This has applications in IoT devices as well as battery-free systems, or any type of device that is not guaranteed consistent power; for example, a reader device could transmit multiple polarizations in order to provide polarization diversity to the zero power device through decreased polarization loss factor without any computational overhead on the zero power device.

Figure 1.3: Infinitesimal Lossless Circuit Model of First Designed CLRH-TL\cite{5}
A property that is frequently targeted for reconfigurability is the polarization of the antenna design; by adjusting the polarization adaptively significant improvements in system gain and channel capacity can be made without requiring additional transmission power or moving parts\[15\]. Reconfigurable polarization patch antennas are further desirable as they possess the benefits of a patch antenna (inexpensive manufacturing, low profile, and easy integration) along with the benefits of polarization diversity\[16\]\[14\]\[17\]. The patch antenna platform has many well documented alterations and techniques that can be implemented with it to meet various design criteria, such as aperture coupling, stacked resonators, and symmetrical design.

Figure 1.5: Modern Applications of Aperture-Coupling

(a) Wideband Reflect-Array\[18\]  (b) Bandpass Frequency Selective Surface\[19\]
Aperture-coupling is a patch feed mechanism first proposed in 1985 [20] that utilizes the coupling of a transmission line’s H-field with a slot in its ground to project the field onto a microstrip patch. It has been analyzed in numerous ways, including but not limited to directly through the fields [21] and as a pair of coupled inductors [22]. Aperture-coupled designs offer wider bandwidth [23], improved ability to tune the patch separately from the feed [22], and allow greater flexibility in the layout of the feed network as the patch itself resides on another layer isolated by a ground plane [20][21]. Its broad utility has brought it widespread use, with many modern designs utilizing it for their own purposes [18][19]. When utilized as the feed of an antenna with a wide bandwidth it is frequently paired with stacked element resonators.

Stacked element resonators are what is entailed by their name: several similar elements stacked in sequence in order to produce a desired effect, usually a wider bandwidth [25]. Stacked element resonators are commonly utilized with aperture-coupled microstrip patches [24][26] due to their utility in widening the narrow bandwidth of a single microstrip patch along with their relatively simple implementation. In designs where aperture radiation control is necessary, such as a dual linear symmetric patch design in which the apertures are in close proximity, they can also be used to compensate for the decreased bandwidth from offsetting the resonance of the aperture [27].
The dual linear polarization of a microstrip patch can be accomplished in several ways and aperture feeds aligned with adjacent edges is one such popular technique for doing so [28]. Apertures may either be crossed, with their feedlines present on separate copper layers, or offset, with their feedlines present on the same copper layer. Offsetting the apertures from the center provides measurably worse performance compared to crossed apertures, however the losses are not significant for most applications [28] and decrease the overall cost of the antenna by requiring less layers for the design.

This thesis proposes a novel device consisting of a frequency selective transmission line that is capable of operation as both an electrical open and a transmission line at different frequencies through the utilization of CLRH-TLs. The proposed device allows for new techniques in the design of microwave networks to combine two physically distinct structures into a single layout through this frequency selective behavior; because an electrical open has the physical analog of an air gap, structures implementing this frequency selective open circuit (FSOC) possess two layout equivalents: one in which the FSOCs are operating as an open and one in which the FSOCs are operating as a transmission line. Through proper utilization of FSOCs, then, a design may redirect power in the same manner that RF switches allow for through only passive components. This thesis explores the use of these FSOCs in a modified quadrature hybrid structure used to feed an aperture-coupled stacked element microstrip antenna, therefore creating frequency controlled
reconfigurable behavior.

In Chapter 2 we comprehensively cover the design process of a dual linearly polarized aperture-coupled stacked patch. It begins with an explanation of the theoretical basis for the technologies implemented and concludes with the technique of iterative design phases for combining the technologies into a singular antenna.

In Chapter 3 we introduce the concept of a path-length phased open circuit in the terms of left and right-handed lines. We then demonstrate and derive a utilization of CLRH-TLs to combine right and left-handed operation, producing a frequency selective open circuit. Following this we provide simulation verification and experimental validation of the derived device.

In Chapter 4 we implement the novel device in the base structure of a quadrature hybrid, beginning with an explanation of the theoretical basis of a quadrature hybrid and the desired polarization outputs. We follow this with the presentation of the design process of this modified quadrature hybrid, providing verification through simulation and validation through experimental results across the stages.

In Chapter 5 we conclude this work with an analysis of the modified quadrature hybrid and propose alterations along with a suggested direction for iterative works.
Major Contributions of Present Work

• **Frequency Selective Open Circuit** — We present the derivation, verification, and experimental validation of a novel CLRH-TL device that displays the behavior of both a transmission line and an open circuit at different frequencies.

• **Dual Linear and Dual Circular Polarizations passively through circuitry** — Incorporating the FSOC within a quadrature hybrid power divider, we demonstrate a method for achieving dual linear polarization and dual circular polarizations passively through simulation and experimental validation.

• **Modified Quadrature Hybrid Power Divider** — We present the design of a modified quadrature hybrid power a combination of using the CLRH cells and FSOC placed strategically in the branches of a quadrature hybrid power divider to obtain dual linear and dual circular polarizations.
2. Dual Linearly Polarized Aperture-Coupled Stacked Patch

To demonstrate the novel device and power distribution techniques proposed in this work an antenna possessing specific properties is designed. It must provide linear polarization on both the horizontal and vertical axis, with approximately equivalent radiation patterns and gain between them, as well as a sufficiently wide bandwidth to encompass the frequency controlled behavior. To this end, a symmetrical aperture coupled stacked square patch is chosen. A -10dB bandwidth of 17%, 10.7GHz to 12.7GHz, is selected as the design goal.

2.1. Theoretical Basis

![Figure 2.1: Linearly Polarized Stacked Element Square Patch](image)

The function of a dual linearly polarized aperture-coupled stacked patch can be divided into three key components; an aperture-coupled patch, a stacked element, and a symmetrical second feed. Aperture coupling a feed to a patch consists of cutting a slot in a shared ground plane in order to utilize aperture resonance to couple the fields of the feed line to the patch itself.
Utilizing the right hand rule, it can be found that the H-field of a TEM wave propagating along the Y-axis with a Z-axis polarized E-field is X-polarized. Aligning the rectangular aperture in the ground plane with the X-axis will then couple the H-field to it as a slot aperture. To satisfy boundary conditions this slot will radiate an opposite direction H-field on the other side of the ground plane where the microstrip patch is placed; this coupling can be modeled as a pair of coupled inductors between the feed line and patch [22].

As current is induced in the patch through the coupled inductors in the equivalent circuit model a potential difference builds across the patch, with opposite edges gaining equal magnitude but inverted phase voltages. With the appropriate patch length, then, the voltage at a specific
wavelength (and therefore frequency) will reach magnitude maxima with opposite signs at both ends of the patch at the same time, an equivalent state to traditional edge fed patch operation. The coupling coefficient of the coupled inductors is mainly determined by the length of the aperture and length of the stub. A longer aperture generally provides stronger coupling between the patch and feed line due to increased coupling surface area, while the open stub controls the location of current maxima beneath the aperture; the H-field of a transmission line at a given location is proportional to the current, therefore adjusting the line length of the open stub to produce current maxima beneath the aperture ($\lambda/4$) maximizes coupling. With increased coupling comes a larger load on the original transmission line, and therefore a higher impedance; adjusting this coupling such that the impedance of the transmission line is matched is the primary purpose of tuning.

Stacking resonators to improve the bandwidth of an otherwise narrowband geometry is a highly proven method, especially for patch designs\[27\][25]. In the case of patch antennas, stacked resonators can be simply conceptualized by analyzing the field lines and charge distributions of a normal patch.

Figure 2.4: Field Lines and Charge Distributions of Idealized Patch\[29\]

Figure 2.4(a) depicts the charge distribution of a singular ideal patch antenna and ground plane. The ground plane consists of an infinite copper plane that is divided in charge polarity by the center of the patch, while the patch consists of a finite copper plane that also displays a charge distribution centered on the patch; Figure 2.4 (b) depicts the equivalent field lines for this charge distribution. Assume now that a second (parasitic) patch of similar dimensions is placed upon the first patch, separated by a dielectric layer.

11
If placed in close enough proximity to the first (driven) patch, it would begin to terminate some of the field lines from the driven patch. If the parasitic patch is aligned with the driven patch it is terminating field lines on both ends of the driven patch simultaneously, producing a similar but inverted charge distribution on the parasitic patch. This distribution produces surface currents in the same manner that feeding the parasitic patch directly would, therefore producing radiation from the parasitic patch. If the parasitic patch is of a slightly different size than the driven patch then the resonant frequency of this radiating behavior of the parasitic patch differs slightly. This offset resonance of the parasitic patch then expands the bandwidth of the composite antenna by combining the parasitic patch’s bandwidth with the bandwidth of the driven patch.
The second feed of the designed patch operates on diagonal symmetry; as long as the first feed is adjusted such that it has no intersection with the diagonal of the square patch, it can be assumed that with some additional tuning the two feeds will both function identically. This concept of two offset slots in close proximity to provide two oppositely polarized feeds to a patch antenna has been proven to not introduce infeasibly negative effects in prior work[28].
2.2. Verification and Validation

The design of the dual linearly polarized aperture-coupled stacked patch is divided into phases as described in the theory section: aperture-coupled patch, aperture-coupled stacked patch, and dual linearly polarized aperture-coupled stacked patch. The design flow (Figure 2.7) describes the process used for each design phase. Upon modeling the current design phase using variable lengths and applying initial values based on theoretical calculations, the inbuilt optimizers in ANSYS HFSS are used to iteratively alter the select design variables of the current phase.

Optimization in HFSS utilizes cost functions in which the user defines certain conditions that have a cost to them; for example, if the cost function is “\( dB(S_{11}) \leq -20 \) at 11.7GHz,” then an iteration with an \( S_{11} \) magnitude of -13dB at 11.7GHz would have a cost of 7. In the case of the design of this antenna the sequential nonlinear programming optimizer is used; this optimizer attempts to meet the cost function by solving it as a nonlinear programming problem, with the problem’s variables set as the chosen design variables from the 3D model. By converging on solutions to this problem through iteratively altering the design variables, the optimizer ideally eventually reaches a combination of the variables selected that meets the requirements of the cost function. To ensure the optimizer provides good results, reasonable bounds must be placed on its
ability to alter each variable; if this is not done there is the potential for a “functional” design that exists in simulation only.

Figure 2.8: Schematic of Stacked Patch with One Feed

The first phase of the design is the aperture-coupled square patch, simulated on 31mil Rogers 5880 Duroid. It is defined in terms of patch dimensions, aperture dimensions, feed length, stub length, and aperture offset (Figure 2.8 (a)). The stackup is identical to Figure 2.8 (b) with the exception of the parasitic patch; it is not included in this design phase. For initial values the design frequency is set to the bandwidth center of 11.7GHz: the patch dimensions are set to $0.33\lambda$ (semi-perimeter of $\lambda/2$), the aperture length is set to $0.25\lambda$, the aperture width to a variable that is always a tenth of the aperture length, the feed length to $\lambda/2$ to allow for a full standing wave wavelength, the stub length to $\lambda/4$ (current maximum beneath aperture), no aperture offset on the X-axis, and an aperture offset set to a third of the length of the patch edge on the Y-axis. The aperture Y-axis offset is set in preparation of the implementation of symmetry, as the farther the slot is from the center the longer it can be afforded to be; if accounted for properly its effect on the design is relatively minimal[28]. The variables of patch dimensions, aperture length, stub length, and Y axis aperture offset are then swept with tolerances of $\pm 25\%$ and a cost function that indicates
$S_{11}$ at 11.7GHz should be minimized.

![Graph](image1.png)

(a) $S_{11} (dB)$

(b) Radiation Pattern at 11.7GHz (dB)

Figure 2.9: Aperture-Coupled Patch Parameters (Simulation)

Figure 2.9 showcases the results of the optimization of the first design phase. The sole patch displays the expected narrowband frequency response of a single microstrip patch as well as a normal radiation pattern. As these results are acceptable the design flow moves on to the next phase, the stacked aperture-coupled patch (Figure 2.8). In this design phase the parasitic patch is added, defined in terms of the driven patch length plus an additional new variable that determines how much larger the parasitic patch is; its initial value is set to 1mm, the difference in side lengths between a square patch designed at 10.7GHz and a square patch designed at 12.7GHz. This variable is added to the optimizer in addition to the previously swept variables and the new design is run with a cost function that indicates $S_{11} \leq -15\text{db}$ across the bandwidth of 10.7GHz to 12.7GHz.
As anticipated, this design possesses a significantly enhanced bandwidth compared to the single patch with a -10dB bandwidth of 20%. The predicted dual resonances of the stacked patch are clearly visible, and the radiation pattern appears normal for a patch antenna. While the optimization goal of -15dB across the entire bandwidth was not fully met, the -10dB bandwidth is more than sufficient to advance to the next design phase, the dual linearly polarized aperture-coupled stacked patch.
To create the model for this design phase, the feed and aperture of the previous design phase are mirrored across the diagonal axis. It is at this point that the X-axis aperture offset variable is utilized; after mirroring the feed system the same cost function from the last design phase is run, this time only with the variables of X-axis aperture offset and aperture length. This is performed in order to recoup any losses created by the introduction of another aperture in close proximity to the original. After running this optimization, one final optimization with all of the previously utilized variables is run with tolerances of 10%. The final results of this design are viewed in Figure 2.12 and Figure 2.13.

Figure 2.12: Final Design Phase S-Parameters (Simulation)
The dimensions produced by the final optimization pass are as follows: driven patch length of 6.17mm, parasitic patch length of 6.93mm, aperture dimensions of 6.22mm x 0.62mm, aperture Y offset of 3.23mm, and aperture X offset of 0.98mm. From the simulation data it is apparent that the addition of a second feed has a noticeable effect on the resonance of the patch configuration, however the target bandwidth is still attained. The radiation patterns suggest noticeable cross
polarization but also good symmetry; the equivalent patterns between the two ports are nearly identical. With properly phased inputs the capability of circular polarization is also demonstrated; this is pertinent to the demonstration of the modified quadrature hybrid.

Figure 2.14: Manufactured Patch Design

To validate the simulation results, the design featured in Figure 2.11 is manufactured and measured with the E8363B PNA Series Network Analyzer. The measurements of the S-parameters reflect the shape of the simulation, however there is a linear shift in the frequency of resonance, with the measured patch resonating from 11.4GHz to 13.8GHz on its port 1 equivalent (20% BW). The
$S_{22}$ magnitude has a similar shape to the $S_{11}$ magnitude, however it displays a worse bandwidth of 1.7GHz (15%) as well as a slight offset in resonance. Connector loading is likely responsible for the shift in resonance, while the tool-assisted hand alignment of the antenna layers is likely the source of the asymmetry between ports. Within the examined implementations of this antenna further in this work the connector loading is separated by a distribution network, therefore the resonance is likely to shift back to its expected position; this is confirmed in Chapter 4.3.

From the radiation pattern measurement it is apparent that the patch demonstrates linear polarization from each port. While there is some pattern distortion the cross-polarization ratio is sufficiently low enough to say that it exhibits linear polarization. This antenna is therefore sufficient for demonstrating the power distribution networks proposed in this work.
3. Frequency Selective Open Circuits

Electrical opens in microwave circuitry describe a special condition of full reflection in which the returned wave is fully in phase with the input wave. If a circuit possesses an electrical open between two arbitrary ports it is equivalent to those ports being fully isolated by an air gap. If an electrical open could be selectively induced through some sort of controlling mechanism, then, the equivalent layout of a circuit could be altered in the same way that physically removing sections would without the likely irreversible damage involved in doing so.

3.1. Combined Left-Right Handed Transmission Lines

CLRH-TLs are a form of metamaterial-altered transmission lines that at specific frequencies provide a negative index of refraction, switching the transmission line from right-handed (RH) to left-handed (LH) operation. In homogeneous left-handed propagation the transmission line displays negative permeability and permittivity, effectively phasing any propagated wave "backwards" rather than forwards. If the transmission line is “balanced” it also provides a smooth transition in index of refraction at the zero crossing frequency as either the left-handed or right-handed component begins to dominate the propagation constant. Previous works have utilized this property to produce a phase shift at specific frequencies. These works have been centered around increasing bandwidth or providing two operating frequencies, however this work instead seeks to use this phase shift to produce frequency selective wave cancellation, altering the overall functionality of the circuit though the production of electrical opens.
At the core of a CLRH-TL is the inversion of the standard transmission line model of operation; while a traditional homogeneous right-handed transmission line consists of a series inductance and a shunt capacitance, a homogeneous left-handed transmission line is instead composed of a series capacitance and a shunt inductance as seen in Figure 3.1 (a) and (b). While a homogeneous LH-TL does not exist in nature due to the parasitics involved in physical transmission line structures [5], LH-TL behavior can be induced at specific frequencies by artificially introducing the requisite inductance and capacitance; this thesis utilizes lumped components to do so. By including electrically small series capacitors and shunt inductors, a unit cell of the CLRH-TL is produced (Figure 3.2).
A CLRH-TL is considered balanced under the condition of (1), that the characteristic impedance of the transmission line is equivalent to the root of the inductance over the root of the capacitance for both RH and LH capacitance/inductance pairs [5].

\[ Z_0 = \sqrt{\frac{L'_L}{C'_L}} = \sqrt{\frac{L'_R}{C'_R}} \]  

(1)

In the balanced case there is no discontinuity in characteristic impedance regardless of the mode of operation, allowing for a continuous transition of the propagation constant between positive and negative at the resonance of the induced left-hand components. The propagation constant itself can be broken down into two components, one for each handedness; in the balanced case this expression simplifies to (2) [5].

\[ \beta = \beta_R + \beta_L = \omega \sqrt{L'_R C'_R} - \frac{1}{\omega \sqrt{L'_L C'_L}} \]  

(2)

From this equation it is apparent that at frequencies higher than the balanced condition’s crossover frequency the line is dominated by its positive right-handed propagation, while at lower frequencies it is dominated by the negative left-handed propagation, therefore introducing an exploitable frequency selective behavior.
3.2. Theoretical Basis of Frequency Selective Open Circuits

Consider the first case of two equal length transmission lines of equal impedance $2Z_0$; should they be connected in parallel their equivalent impedance would then be $Z_0$, with full transmission across the two lines if both ends are matched to $Z_0$. Likewise, consider the second case of two parallel lines each with impedance $2Z_0$, but with one line having a length that is $\lambda/2$ longer than the other; as one line is electrically $180^\circ$ shifted from the other, it will always have a voltage of equal magnitude but opposite phase at the end compared to the shorter line. This leads to full wave cancellation at the output, producing open equivalent circuit across its ports. The only difference between these two circuits is the phase shift in one of the parallel lines; if that phase shift could be manipulated then a design could effectively swap between full transmission and full reflection at a port.
Figure 3.4: Altered Case In Phase and Out of Phase Parallel Transmission Lines

Consider now a case identical to the first, with the exception that the lines are both of length \( \lambda/4 \) and one branch is instead homogenously left-handed rather than right-handed. Along the right-handed branch the voltage is phased forward 90°, while along the left-handed branch the wave is instead phased backwards by 90°. Assuming a 0° input, this ultimately produces voltages with phase 90° and 270° respectively at the output, an equivalent result to the aforementioned second case but with the structure of the first. Conversely, if a similar situation to the second case was implemented with a right-handed line of length \( \lambda/4 \) and a homogeneously left-handed line of length \( 3\lambda/4 \), then the output would have both lines phased to 90°, equivalent to the first case but with the structure of the second. As the structures for opposite results are identical between right-handed and left-handed operation, the dual-handed nature of CLRH-TLs can combine the four designs in Figure 3.3 and 3.4 into two designs that operate at two frequencies, with either full reflection or full transmission below the threshold frequency and full transmission or full reflection.
above the threshold frequency depending on the structure; these circuits are referred to as frequency selective open circuits (FSOCs) in this thesis. FSOCs can be further differentiated by their right handed (nominal) behavior; nominally closed frequency selective circuits (NC-FSOCs) act as a transmission line in right-handed operation, while nominally open frequency selective circuits (NO-FSOCs) act as an open in right-handed operation.

![Diagram of Circuit Partitions for Derivation of FSOCs](image)

To rigorously prove these structures, the phase shift induced by the CLRH cell is first modeled as a length of \((\theta/2\pi)\lambda\) for numerical convenience in the derivation and because it is proportional to wavelength; it represents the phase shift induced by the proportion of right-handed to left-handed propagation in the propagation constant. Following this, the symmetry across the structure’s vertical axis is invoked as is featured in Figure 3.5 (b) and (d). Because the two halves are identical, their S-matrices are therefore also identical. Additionally, ports 2 and 3 are intrinsically connected to ports 5 and 6, therefore the power exchanged between the halves is circulated. This circulation can be modeled as an infinite series of cascaded transmissions between the two half-networks, with
the output of each half being used as the input of the next transmission of the other. As power exits the system with each circulation at ports 1 and 4 this circulation is lossy, therefore this series will always converge. For the \( n \)\( ^{th} \) transmission of this circulation, 
\[
\begin{align*}
    a_{2}^{(n)-} &= a_{5}^{(n)+}, \quad a_{3}^{(n)-} = a_{6}^{(n)+}, \\
    a_{5}^{(n)-} &= a_{2}^{(n+1)+}, \quad \text{and} \quad a_{6}^{(n)-} = a_{3}^{(n+1)+}.
\end{align*}
\]

Figure 3.6: Visualization of Cascaded Circulation in FSOC Derivation

To begin this proof, assume that the half-networks can each be represented by a lossless, reciprocal 3-port network that is matched to the exit port (ports 1 and 4). Assuming an input solely into port 1, the first transmitted voltage into the second half-network through ports 5 and 6 may be found.
\[ S_{\text{Half}} = \begin{bmatrix} 0 & s_{21} & s_{31} \\ s_{21} & s_{22} & s_{23} \\ s_{31} & s_{23} & s_{33} \end{bmatrix} \]

\[
\begin{bmatrix}
  a_{1}^{(1)-} \\
  a_{2}^{(1)-} \\
  a_{3}^{(1)-}
\end{bmatrix} =
\begin{bmatrix}
  0 & s_{21} & s_{31} \\
  s_{21} & s_{22} & s_{23} \\
  s_{31} & s_{23} & s_{33}
\end{bmatrix}
\begin{bmatrix}
  a_{1}^{(1)+} \\
  a_{2}^{(1)+} \\
  a_{3}^{(1)+}
\end{bmatrix} =
\begin{bmatrix}
  0 & s_{21} & s_{31} \\
  s_{21} & s_{22} & s_{23} \\
  s_{31} & s_{23} & s_{33}
\end{bmatrix}
\begin{bmatrix}
  a_{1}^{(1)+} \\
  a_{2}^{(1)+} \\
  a_{3}^{(1)+}
\end{bmatrix}
\]

\[
\begin{bmatrix}
  a_{1}^{(1)-} \\
  a_{2}^{(1)-} \\
  a_{3}^{(1)-}
\end{bmatrix} =
\begin{bmatrix}
  0 \\
  s_{21}a_{1}^{(1)+} \\
  s_{31}a_{1}^{(1)+}
\end{bmatrix} =
\begin{bmatrix}
  0 \\
  a_{5}^{(1)+} \\
  a_{6}^{(1)+}
\end{bmatrix}
\]

\[ a_{5}^{(1)+} = (s_{21}^{2} + s_{31}^{2})a_{1}^{(1)+} \]

\[ a_{6}^{(1)+} = (s_{21}s_{22} + s_{31}s_{23})a_{1}^{(1)+} \]

\[ a_{5}^{(1)+} = (s_{21}s_{22} + s_{31}s_{23})a_{1}^{(1)+} = (s_{21}s_{22} + s_{31}s_{23}s_{21} + s_{21}s_{23}s_{31} + s_{31}s_{33}s_{23})a_{1}^{(1)+} \]

\[ a_{6}^{(1)+} = (s_{21}s_{22} + s_{31}s_{23}s_{22} + s_{21}s_{23}^2 + s_{31}s_{33}s_{23})a_{1}^{(1)+} = (s_{21}s_{22}s_{23} + s_{31}s_{23}^2 + s_{21}s_{23}s_{33} + s_{31}s_{33}^2)a_{1}^{(1)+} \]

From \( a_{5}^{(1)+} \) and \( a_{6}^{(1)+} \), the first transmission to port 4 and the first transmissions back to ports 2 and 3 may be found. Assuming that port 4 is matched, no voltage that reaches it is reflected back into the system, therefore that voltage exits circulation and the transmission back to the first half-network consists only of \( a_{5}^{(1)-} \) and \( a_{6}^{(1)-} \) as in (4).

\[
\begin{bmatrix}
  a_{4}^{(1)-} \\
  a_{5}^{(1)-} \\
  a_{6}^{(1)-}
\end{bmatrix} =
\begin{bmatrix}
  0 & s_{21} & s_{31} \\
  s_{21} & s_{22} & s_{23} \\
  s_{31} & s_{23} & s_{33}
\end{bmatrix}
\begin{bmatrix}
  a_{1}^{(1)+} \\
  a_{5}^{(1)+} \\
  a_{6}^{(1)+}
\end{bmatrix} =
\begin{bmatrix}
  0 \\
  (s_{21}s_{22} + s_{31}s_{23})a_{1}^{(1)+} \\
  (s_{21}s_{23} + s_{31}s_{33})a_{1}^{(1)+}
\end{bmatrix}
\]

This process is repeatable infinitely to calculate the \( n_{th} \) transmission to each port. Due to the non-reflection of voltage at the exit ports and the circulation between the first and second half-network, the calculation of the next iteration of inter-half transmission can be simplified as follows in (5)
Using the recurrence relation of (5) it is now possible to define the \( n \)th output to ports 1 and 4 \( \alpha_1^{(n)} \) and \( \alpha_4^{(n)} \) in terms of the initial input. Due to superposition the overall transmission to port 4 and reflection to port 1 is equivalent to the sum of every voltage exiting through those ports in the circulation. Combining these two statements results in (6) and (7).

\[
\begin{align*}
\begin{bmatrix}
\alpha_2^{(n+1)} \\
\alpha_3^{(n+1)}
\end{bmatrix}
&= \begin{bmatrix}
s_{22} & s_{23} \\
s_{23} & s_{33}
\end{bmatrix}
\begin{bmatrix}
\alpha_5^{(n)} \\
\alpha_6^{(n)}
\end{bmatrix}
\begin{bmatrix}
2 \\
2
\end{bmatrix}
\begin{bmatrix}
\alpha_2^{(n)} \\
\alpha_3^{(n)}
\end{bmatrix}

\begin{bmatrix}
\alpha_5^{(n)} \\
\alpha_6^{(n)}
\end{bmatrix}
&= \begin{bmatrix}
s_{22} & s_{23} \\
s_{23} & s_{33}
\end{bmatrix}
\begin{bmatrix}
\alpha_2^{(n+1)} \\
\alpha_3^{(n+1)}
\end{bmatrix}
\begin{bmatrix}
2 \\
2
\end{bmatrix}
\begin{bmatrix}
\alpha_5^{(n)} \\
\alpha_6^{(n)}
\end{bmatrix}
\end{align*}
\]

(5)

\[
\begin{align*}
\alpha_1 &= \sum_{n=1}^{\infty} a_1^{(n+1)} = \sum_{n=1}^{\infty} s_{21} a_5^{(n)} + s_{31} a_6^{(n)} = \sum_{n=1}^{\infty} \begin{bmatrix}
s_{21} & s_{31}
\end{bmatrix}
\begin{bmatrix}
\alpha_5^{(n)} \\
\alpha_6^{(n)}
\end{bmatrix}
\begin{bmatrix}
2 \\
2
\end{bmatrix}
\begin{bmatrix}
\alpha_2^{(n)} \\
\alpha_3^{(n)}
\end{bmatrix}

\begin{bmatrix}
\alpha_5^{(n)} \\
\alpha_6^{(n)}
\end{bmatrix}
= \begin{bmatrix}
s_{21} & s_{31}
\end{bmatrix}
\begin{bmatrix}
\alpha_5^{(1)} \\
\alpha_6^{(1)}
\end{bmatrix}
+ \sum_{n=1}^{\infty} \begin{bmatrix}
s_{21} & s_{31}
\end{bmatrix}
\begin{bmatrix}
\alpha_5^{(1)} \\
\alpha_6^{(1)}
\end{bmatrix}
\begin{bmatrix}
2 \\
2
\end{bmatrix}
\begin{bmatrix}
\alpha_2^{(n)} \\
\alpha_3^{(n)}
\end{bmatrix}
\end{align*}
\]

(6)
therefore, because an normalized input of $a_1$ of the full FSOC. Because the FSOC is a symmetric 2-port network,

$$a_s = a_s$$

Beginning with the half-network S-matrix of the NC-FSOC depicted in Figure 3.5 (a), through

$$s_0 = s_4 = 1$$

can be factored from every term, therefore assuming a normalized input of $a_1^{(1)^+} = 1$ the known values of $a_2^{(1)^-} = s_{21}$ and $a_3^{(1)^-} = s_{31}$ can be used to simplify these equations to (8).

$$a_1^- = \begin{bmatrix} s_{21} & s_{31} \end{bmatrix} \begin{bmatrix} s_{22} & s_{23} \\ s_{23} & s_{33} \end{bmatrix} + \sum_{n=1}^{\infty} \begin{bmatrix} s_{21} & s_{31} \end{bmatrix} \begin{bmatrix} s_{22} & s_{23} \\ s_{23} & s_{33} \end{bmatrix} 2^n \begin{bmatrix} s_{21} \\ s_{31} \end{bmatrix}$$

$$a_4^- = s_{21}^2 + s_{31}^2 + \sum_{n=1}^{\infty} \begin{bmatrix} s_{21} & s_{31} \end{bmatrix} \begin{bmatrix} s_{22} & s_{23} \\ s_{23} & s_{33} \end{bmatrix} 2^n \begin{bmatrix} s_{21} \\ s_{31} \end{bmatrix}$$

With equations for $a_1^-$ and $a_4^-$ from the half-network model it is now possible to derive the S-matrix of the full FSOC. Because the FSOC is a symmetric 2-port network, $s'_{11} = s'_{12}$ and $s'_{12} = s'_{21}$, therefore, because a normalized input of $a_1^{(1)^+} = 1$ is used, the S-matrix is simply $s'_{11} = s'_{22} = a_1^-$ and $s'_{12} = s'_{21} = a_4^-$. 

Beginning with the half-network S-matrix of the NC-FSOC depicted in Figure 3.5 (a), through
utilizing only basic properties (losslessness and reciprocity) it can be found by that it is equivalent to \((9)\).

\[
S_{NC-Half} = \frac{1}{\sqrt{2}} \begin{bmatrix}
0 & e^{-j\pi/4} & e^{-j(\pi/4+\theta/2)} \\
-\frac{e^{-j\pi/4}}{\sqrt{2}} & e^{j\pi/2} & e^{-j(\pi/2+\theta/2)} \\
e^{-j(\pi/4+\theta/2)} & e^{-j\pi} & \frac{e^{-j(\pi/2-\theta)}}{\sqrt{2}} 
\end{bmatrix}
\]  \(9\)

Utilizing this matrix with \((8)\) and combining terms yields \((10)\) and \((11)\).

\[
a_1^- = \frac{e^{-j\theta}(\cos(\theta) - 1)}{2} + \frac{1}{4} \sum_{n=1}^{\infty} \left[ e^{-j\frac{\pi}{4}} e^{-j(n+\frac{\theta}{2})} \right]^T \begin{bmatrix}
-\frac{-e^{-j\frac{\theta}{2}} \cos(\frac{\theta}{2})}{2} & e^{-j\theta} \cos(\frac{\theta}{2}) \\
e^{-j\theta} \cos(\frac{\theta}{2}) & -\frac{e^{-j\frac{3\theta}{2}} \cos(\frac{\theta}{2})}{2}
\end{bmatrix}^n \begin{bmatrix}
e^{-j\frac{\pi}{4}} + e^{-j(\frac{3\pi}{4}+\theta)} \\
e^{j(\frac{3\pi}{4}+\theta)} + e^{j(\frac{\pi}{4}-\frac{3\theta}{2})}
\end{bmatrix}
\]  \(10\)

\[
a_4^- = -je^{-j\frac{\theta}{2}} \cos\left(\frac{\theta}{2}\right) + \frac{1}{4} \sum_{n=1}^{\infty} \left[ e^{-j\frac{\pi}{4}} e^{-j(n+\frac{\theta}{2})} \right]^T \begin{bmatrix}
-\frac{-e^{-j\frac{\theta}{2}} \cos(\frac{\theta}{2})}{2} & e^{-j\theta} \cos(\frac{\theta}{2}) \\
e^{-j\theta} \cos(\frac{\theta}{2}) & -\frac{e^{-j\frac{3\theta}{2}} \cos(\frac{\theta}{2})}{2}
\end{bmatrix}^n \begin{bmatrix}
e^{-j\frac{\pi}{4}} \\
e^{-j(\frac{\pi}{4}+\theta)}
\end{bmatrix}
\]  \(11\)

While the solution to these equations would normally require numerical approximation, two values of \(\theta\) are of note for this design: 0 for right-handed operation and \(\pi\) for left-handed operation. Starting with right-handed operation, if \(\theta\) is equal to 0 then the above equations can be further reduced to \((12)\) and \((13)\).
\[ a_{1RH}^- = \frac{1}{4} \sum_{n=1}^{\infty} \left[ e^{-j \frac{\pi}{4}} \right]^T \begin{bmatrix} \frac{1}{2} & -\frac{1}{2} \\ \frac{1}{2} & \frac{1}{2} \end{bmatrix} \begin{bmatrix} e^{j \frac{\pi}{4}} + e^{-j \frac{3\pi}{4}} \\ e^{j \frac{\pi}{4}} + e^{-j \frac{3\pi}{4}} \end{bmatrix} \]  

(12)

\[ a_{4RH}^- = -j + \frac{1}{4} \sum_{n=1}^{\infty} \left[ e^{-j \frac{\pi}{4}} \right]^T \begin{bmatrix} \frac{1}{2} & -\frac{1}{2} \\ \frac{1}{2} & \frac{1}{2} \end{bmatrix} \begin{bmatrix} e^{j \frac{\pi}{4}} \\ e^{j \frac{\pi}{4}} \end{bmatrix} \]  

(13)

Inside the infinite summation of both equations is an identical matrix raised to the \( n \)th power; if \((-1)^n\) is factored from this matrix, however, it becomes apparent that it is idempotent. Utilizing this property the infinite summation can be evaluated as follows in (14) and (15), solving for \( a_1^- \) and \( a_4^- \) in the right-handed case.

\[ a_{1RH}^- = \frac{1}{4} \sum_{n=1}^{\infty} (-1)^n \left[ e^{-j \frac{\pi}{4}} \right]^T \begin{bmatrix} \frac{1}{2} & -\frac{1}{2} \\ \frac{1}{2} & \frac{1}{2} \end{bmatrix} \begin{bmatrix} e^{j \frac{\pi}{4}} + e^{-j \frac{3\pi}{4}} \\ e^{j \frac{\pi}{4}} + e^{-j \frac{3\pi}{4}} \end{bmatrix} \]

\[ = \frac{1}{4} \sum_{n=1}^{\infty} (-1)^n \left[ e^{-j \frac{\pi}{4}} \right]^T \begin{bmatrix} 0 \\ 0 \end{bmatrix} \]

\[ = 0 \]  

(14)

\[ a_{4RH}^- = -j + \frac{1}{4} \sum_{n=1}^{\infty} (-1)^n \left[ e^{-j \frac{\pi}{4}} \right]^T \begin{bmatrix} \frac{1}{2} & -\frac{1}{2} \\ \frac{1}{2} & \frac{1}{2} \end{bmatrix} \begin{bmatrix} e^{j \frac{\pi}{4}} \\ e^{j \frac{\pi}{4}} \end{bmatrix} \]

\[ = -j + \frac{1}{4} \sum_{n=1}^{\infty} (-1)^n \left[ e^{-j \frac{\pi}{4}} \right]^T \begin{bmatrix} 0 \\ 0 \end{bmatrix} \]

\[ = -j \]  

(15)
\[ S_{NC-RH} = \begin{bmatrix} a_{1RH}^- & a_{4RH}^- \\ a_{4RH}^- & a_{1RH}^- \end{bmatrix} = \begin{bmatrix} 0 & -j \\ -j & 0 \end{bmatrix} \] (16)

Examining the S-matrix in (16), it is apparent that is is equivalent to the S-matrix of a matched line of length \( \lambda/4 \); this is as anticipated as the physical structure is the same as that of two branched lines of the same length in parallel with an equivalent impedance of \( Z_0 \).

Starting again from (10) and (11) with the left-handed value of \( \theta, \pi \), produces (17) and (18). These equations are trivial to solve and reveal the S-matrix of the left-hand operation of the NC-FSOC in (19).

\[ a_{1LH}^- = 1 + \frac{1}{4} \sum_{n=1}^{\infty} \begin{bmatrix} e^{-j \frac{\pi}{4}} & 0 \\ 0 & e^{-j \frac{3\pi}{4}} \end{bmatrix}^T \begin{bmatrix} 0 & 0 \\ 0 & 0 \end{bmatrix}^n \begin{bmatrix} e^{j \frac{\pi}{4}} + e^{-j \frac{3\pi}{4}} \\ e^{j \frac{3\pi}{4}} + e^{-j \frac{5\pi}{4}} \end{bmatrix} = 1 \] (17)

\[ a_{3LH}^- = 1 + \frac{1}{4} \sum_{n=1}^{\infty} \begin{bmatrix} e^{-j \frac{3\pi}{4}} & 0 \\ 0 & e^{-j \frac{\pi}{4}} \end{bmatrix}^T \begin{bmatrix} 0 & 0 \\ 0 & 0 \end{bmatrix}^n \begin{bmatrix} e^{-j \frac{3\pi}{4}} \\ e^{-j \frac{\pi}{4}} \end{bmatrix} = 0 \] (18)

\[ S_{NC-LH} = \begin{bmatrix} a_{1LH}^- & a_{4LH}^- \\ a_{4LH}^- & a_{1LH}^- \end{bmatrix} = \begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix} \] (19)

As predicted in the full wave cancellation model, the left hand operation of this circuit provides an electrical open at the port, effectively creating an open circuit at the edge of the FSOC.

Moving to the half-network of the NO-FSOC, its S-matrix can again be derived from basic properties and found to be (20).

\[ S_{NO-Half} = \frac{1}{\sqrt{2}} \begin{bmatrix} 0 & e^{-j\pi/4} & e^{-j(3\pi/4+\theta)/2} \\ e^{-j\pi/4} & e^{j\pi/2} & e^{-j(\pi+\theta)/2} \\ e^{-j(3\pi/4+\theta)/2} & e^{-j(\pi+\theta)/2} & e^{-j(\pi/2+\theta)} \end{bmatrix} \] (20)
Using the substitute variable of $\phi - \pi = \theta$, (20) may be rearranged as is shown in (21).

$$S_{NO-Half} = \frac{1}{\sqrt{2}} \begin{bmatrix} 0 & e^{-j\pi/4} & e^{-j(\pi/4+\phi/2)} \\ e^{-j\pi/4} & e^{j\pi/2} & e^{-j(\pi/2+\phi/2)} \\ e^{-j(\pi/4+\phi/2)} & e^{-j(\pi/2+\phi/2)} & e^{j(\pi/2-\phi)} \end{bmatrix}$$ (21)

This relation clearly displays the converse relation between nominally closed and nominally open FSOCs; as a $180^\circ$ phase shift on one results in an S-matrix identical to the other, the proofs for the NC-FSOC right-handed and left-handed operations are also valid for the left-handed and right-handed operations respectively of the NO-FSOC.

Intrinsic to the CLRH-TL operation is frequency dependent behavior; this is the core of its composite nature that enables FSOCs, but is also in opposition to FSOC’s wavelength dependent operation. The bandwidth of the frequency selective open circuit therefore requires aligning the crossover frequency of the propagation constant such that the line has fully transitioned to either left-handed or right-handed behavior at the frequency to which the physical lengths of the circuit correspond. For example, if a NC-FSOC was designed to be open at 9GHz and closed at 15GHz, the path lengths of the circuit would be most optimally designed for lengths of $\lambda/4$ at 9GHz in order to provide the best isolation at that frequency, and the resonance of the left-handed components would lean towards 15GHz to ensure a full left-handed transition by 9GHz.
3.3. Verification and Validation

To verify the theory behind FSOCs simulation in ANSYS HFSS is utilized. Beginning with the NC-FSOC, it is modeled as the connection between two 50Ω transmission lines using lumped RLC boundaries for the capacitors and inductors. Under the assumption that the bandwidth of 10.7GHz to 12.7GHz is relatively narrow, the design uses \( \lambda \) as the wavelength of 11.7GHz (the center frequency) in the substrate. For the substrate definition the simulation model of Rogers 5880 Duroid in the ANSYS database is used with a thickness of 31mil. Figure 3.7 depicts the 3D model used in the simulation of the NC-FSOC, with blue indicating capacitors, red indicating inductors, and yellow indicating the signal copper. RLC boundaries are sized such that in a manufactured design imperial 0402 sized SMT components may be used. In this model there are three CLRH unit cells present upon the NC-FSOC’s transmission line branch; this is in line with past work which has done so to induce a greater phase shift within the same space as compared to a singular cell\([10][9][7]\). In order to tune the transition frequency of the device to align with the center of the bandwidth, 11.7GHz, capacitor and inductor pairs that satisfy (1) for 100Ω are simulated in a parametric sweep, with the magnitude of the inductance set equal to 10,000 times the magnitude...
of the capacitance. The conclusion of this sweep reveals that a value of 0.2 pF capacitance and 2 nH inductance per a component results in a transition frequency close to 11.7 GHz; this combination would alone resonate at 8 GHz, however due to the electrically close proximity of the inductors and capacitors there is some reduction of magnitude, with the inductors acting in parallel and the capacitors acting in series.

Figure 3.8 depicts the collected data under the above inductor and capacitor combination. From the S-parameter plot it is apparent that the circuit displays the expected frequency controlled behavior, where below the threshold frequency there is high signal rejection and above the threshold frequency there is high signal transmission. Figure 3.8(b) and (c) agree with this conclusion, clearly
indicating an elimination of the E-field at 10GHz and a full transmission at 14GHz. Of concern is the rate of transition; from the S-parameter plot there is an approximately 1.5GHz (13%) band around the center frequency in which the behavior of the line is poorly aligned with either goal. Accounting for this switching overhead is necessary for implementing this circuit in a larger design.

A similar process is used with the simulation of the NO-FSOC; a key difference, however, is that some tuning is applied to the circuit before the addition of the CLRH cells to ensure total cancellation is occurring in right-handed operation. By utilizing a variable length transmission line (length $\lambda/4 + X$ in Figure 3.9) that variable can be parametrically swept to find the optimal length to be 4.1mm (Figure 3.10).
With the line lengths determined, three CLRH cells are again added to the branched lines with the same values of inductance and capacitance as the NC-FSOC simulation.
Figure 3.11 depicts the collected data of the NO-FSOC simulation. The S-parameter plot agrees with the expected outcome, where below the threshold frequency there is signal transmission and above the threshold frequency there is signal rejection. Figure 3.11(b) and (c) further support this, clearly indicating an elimination of the E-field at 10GHz and a full transmission at 14GHz. Similar
to the NC-FSOC the switching overhead is an issue with again a width of about 1.5GHz, however the NO-FSOC also appears to have the issue of the open operation being highly narrowband as well as relatively poor isolation and transmission. Both of these factors contribute to the circuit being notably more unwieldy than the NC-FSOC.

![Image of manufactured and populated FSOC designs](image)

**Figure 3.12: Manufactured and Populated FSOC Designs (Photograph)**

The design lengths used in the NC-FSOC and NO-FSOC simulations are utilized for the manufactured design in Figure 3.12. In order to retain similarity to the simulation, the board is etched on 31mil Rogers 5880 Duroid with half-ounce copper as well as populated with 0.2pF capacitors and 2nH inductors. Measurements for this board are taken with the E8363B PNA Series Network Analyzer.
Figure 3.13: Frequency Response of FSOCs (Measured)

Figure 3.13 depicts the measured frequency response of the NC-FSOC and NO-FSOC designs.
It is apparent from the overlay of the HFSS simulation S-parameters that the physical design correlates well, especially after taking the circumstances of their population into account (all 0402 components and ground connections are soldered by hand with an iron). As derived in (19) the FSOCs display an approximately $0^\circ$ phase angle at their open frequency, validating the theory. It is noted that responses at the higher end of the measured frequency band (approaching 15GHz) appear attenuated; the cause of this is likely the self-resonance of the inductors utilized. Examining the components populating the boards reveals that the 2nH inductors possess a self resonance of 15GHz, therefore as the measured frequency approaches 15GHz the inductors become more resistive until their power draw on the PNA’s measurement pulse becomes noticeable. Should more intensive implementation measures be taken (plated hole vias, SMT reflow techniques, and inductors with a higher self-resonance utilized), it can be expected that the measured S-parameters would more closely match the simulation.
4. Modified CLRH-TL Quadrature Hybrid

The modified CLRH-TL quadrature hybrid structure is a quadrature hybrid that, based on the input frequency, is capable of standard quadrature hybrid operation, phase inverted quadrature hybrid operation, or full transmission operation to either port 2 or port 3. This effectively grants it the ability to provide frequency selective polarization when connected to a dual linearly polarized antenna, with right-hand circular and left-hand circular polarization made possible through quadrature hybrid operation as well as vertical and horizontal linear polarization made possible through the transmission operation. This frequency selective behavior is created through the implementation of composite left/right handed transmission lines (CLRH-TL).
4.1. Quadrature Hybrid and Polarization

The quadrature hybrid is a special type of matched, lossless, reciprocal four port network that evenly divides power between ports opposite to the input while maintaining a $90^\circ$ phase difference (Figure 4.2); the S-matrix of an ideal quadrature hybrid is depicted in (22) [30].

$$S_{QH} = \frac{-1}{\sqrt{2}} \begin{bmatrix} 0 & j & 1 & 0 \\ j & 0 & 0 & 1 \\ 1 & 0 & 0 & j \\ 0 & 1 & j & 0 \end{bmatrix}$$

(22)

The length of each of its component transmission lines is $\lambda/4$, with one pair of lines possessing impedance $Z_0$ and the other pair of lines possessing impedance $Z_0/\sqrt{2}$. At the core of its design is path length based wave cancellation, with the power distribution determined by the impedance.
balance between these line pairs. The quadrature hybrid may be proven utilizing even-odd mode analysis.

![Diagram of Even and Odd Modes of Quadrature Hybrid](image)

**Figure 4.3: Normalized Even and Odd modes of Quadrature Hybrid**

In even-odd mode analysis the symmetry of the network is utilized to simplify the calculation of its S-parameters. This functions on the principal of even and odd functional decomposition; because there is an axis of symmetry across the network, all responses between the halves of that axis of symmetry can necessarily be decomposed into functions that possess either even or odd symmetry across that axis themselves. Open circuits possess a reflection coefficient of 1 and therefore provide even symmetry, while short circuits possess a reflection coefficient of -1 and therefore provide odd symmetry; by terminating the lines cut by the axis of symmetry with these loads the even and odd responses may be calculated. Following this setup, the simplest method to discover the response of each mode is to utilize the ABCD matrix representation of each half-circuit; cascading a shunt, followed by a transmission line, followed by another shunt accomplishes this for each mode in (23) and (24).

\[
\begin{align*}
\begin{bmatrix} A & B \\ C & D \end{bmatrix}_{\text{even}} &= \begin{bmatrix} 1 & 0 \\ j & 1 \end{bmatrix} \begin{bmatrix} 0 & j/\sqrt{2} \\ j/\sqrt{2} & 0 \end{bmatrix} \begin{bmatrix} 1 & 0 \\ j & 1 \end{bmatrix} = \frac{1}{\sqrt{2}} \begin{bmatrix} -1 & j \\ j & -1 \end{bmatrix} \\
\begin{bmatrix} A & B \\ C & D \end{bmatrix}_{\text{odd}} &= \begin{bmatrix} 1 & 0 \\ -j & 1 \end{bmatrix} \begin{bmatrix} 0 & j/\sqrt{2} \\ j/\sqrt{2} & 0 \end{bmatrix} \begin{bmatrix} 1 & 0 \\ -j & 1 \end{bmatrix} = \frac{1}{\sqrt{2}} \begin{bmatrix} 1 & j \\ j & 1 \end{bmatrix}
\end{align*}
\]  

(23)  

(24)
Converting the ABCD matrices into S-parameters provides (25), the even and odd reflection and transmission coefficients.

\[
\Gamma_{\text{even}} = \frac{A_e + B_e - C_e - D_e}{A_e + B_e + C_e + D_e} = 0
\]

\[
T_{\text{even}} = \frac{2}{A_e + B_e + C_e + D_e} = -\frac{(1 + j)}{\sqrt{2}}
\]

\[
\Gamma_{\text{odd}} = \frac{A_o + B_o - C_o - D_o}{A_o + B_o + C_o + D_o} = 0
\]

\[
T_{\text{odd}} = \frac{2}{A_o + B_o + C_o + D_o} = \frac{(1 - j)}{\sqrt{2}}
\]

(25)

Because the structure is symmetrical on both the vertical and horizontal axes, solving for the S-parameters of a single port provides the S-parameters of all ports. Solving for port 1 of the greater matrix then consists of then utilizing the even symmetry to calculate the upper half (even mode input coefficients) with odd symmetry being utilized for the lower half (odd mode input coefficients) leading to (26)

\[
S_{11} = \frac{1}{2} \Gamma_e + \frac{1}{2} \Gamma_o = 0
\]

\[
S_{21} = \frac{1}{2} T_e + \frac{1}{2} T_o = -j
\]

\[
S_{31} = \frac{1}{2} T_e - \frac{1}{2} T_o = -1
\]

\[
S_{41} = \frac{1}{2} \Gamma_e - \frac{1}{2} \Gamma_o = 0
\]

(26)

Solving for the S-parameters of the full structure by applying the four way symmetry provides (22) again, proving its function.
While the quadrature hybrid may be utilized as an even power divider, the 90° phase differential between its quadrature and inphase ports may be exploited for polarization purposes as well. Consider a symmetrical dual linearly polarized patch connected to a $\lambda/2$ transmission line (Figure 4.4). Assuming a perfect cosine signal input from port 1, the edges of the patch that are aligned with that input feed reach opposing field maximums at 0° and 180° phase. As the field lines are necessarily continuous before a full TEM wave is formed, these field maximums produce a field vector that when viewed from the top of the patch appears as vertical line oscillating between pointing down and up, visible in Figure 4.5(a) and (c). As it is a perfect cosine input, at 90° and 270° phase the field vector produced by the feed from port 1 would be zero. Assume now that there is a sine signal input from port 2. As sine signals are equivalent to cosine signals delayed by 90°, the edges of the patch aligned with port 2 receive equivalent field maximums at 90° and 270° respectively due to symmetry, depicted in Figure 4.5(b) and (d); these edges then experience nulls at 0° and 180°.
As polarization is a vector quantity it is produced by the sum of all contributing elements, therefore the polarization vector for this antenna at any given phase is the field vector from port 1 added to the field vector from port 2. Because the antenna is symmetrical the maximum magnitudes of these vectors are identical, therefore they can be normalized to the same value. Summing their normalized magnitudes therefore produces (27).

\[ \vec{pol} = \hat{a}_x \sin(\theta) + \hat{a}_y \cos(\theta) \]  

(27)
This is recognizable as the equation of a circle, therefore the polarization of an antenna under these feeding conditions is considered circular. In the above example the polarization is rotating clockwise, or left-handed; if the phase delay were inverted, such as if port 2 was fed with a negative sine signal instead, then it would rotate in the opposite direction of counter-clockwise, or right-handed.

Applying this back to the quadrature hybrid, if the input ports of the example antenna were to be connected to the output ports of an ideal quadrature hybrid then each port of the antenna would be receiving an signal of identical magnitude with a phase difference of $90^\circ$; these are the same conditions that have been proven to produce circular polarization. The direction of this polarization is determined by which quadrature hybrid port is connected to which antenna port; port 2 of the quadrature hybrid leads, therefore connecting it to the left port produces right-hand circular polarization while connecting it to the right port produces left-hand circular polarization.

### 4.2. Design of Modified Quadrature Hybrid

To utilize the FSOCs in a practical application the baseline circuit of a quadrature hybrid is chosen. This is because its critical lengths are $\lambda/4$, fitting well with the requisite lengths of the FSOC’s, and because the primary purpose of a quadrature hybrid is power distribution, making it the ideal testing ground for a component that provides frequency selective isolation. The core design objective behind the modified quadrature hybrid is to reroute power in the permutations required to provide either right-hand circular, left-hand circular, horizontal linear, or vertical linear polarization to an antenna based on the input frequency. To accomplish this a dual-linearly polarized antenna is connected to what would be the in-phase and quadrature ports of a standard quadrature hybrid, and then power is routed either in the same fashion as a standard quadrature hybrid (even power split with $90^\circ$ lead on the quadrature port), inverted phase quadrature hybrid (even power split with $90^\circ$ lag on quadrature port), full transmission to quadrature port, or full transmission to in-phase port.

To redirect power in four different ways there must necessarily be three different operation breakpoints. By implementing three distinct groups of CLRH-TLs grouped by resonant frequency
in the modified quadrature hybrid the four total operating modes are created: fully right-handed ($f_0$), one CLRH-TL group in left-handed operation ($f_1$), two CLRH-TL groups in left-handed operation ($f_2$), and all CLRH-TL groups in left-handed operation ($f_3$). As CLRH-TLs operate right-handedly above the crossover frequency, these frequencies are in descending order with $f_0$ being the highest frequency in the operational range. Because each CLRH group activating compounds upon the last each operating mode must include the effects of the previous.
Figure 4.6(a) depicts the partitioned block diagram of the proposed modified quadrature hybrid. As the basis structure is a quadrature hybrid, the simplest design choice is to assign right-handed operation to standard quadrature hybrid operation; the placement of the nominally closed and
nominally open FSOCs reflect this, with the equivalent circuit of the nominal state identifiable as a normal quadrature hybrid. In line with this design choice, the first CLRH-TL group shifting to left-handed operation brings the modified quadrature hybrid into inverted phase quadrature hybrid operation; this is accomplished by switching the transmission line to the in-phase port of the modified quadrature hybrid (physical length $\lambda/4$) from right-handed operation to left-handed operation. As the transmission line to the quadrature port is of equal length, this utilization of left-handed lines to produce a $180^\circ$ phase shift is identical to how a NC-FSOC operates; in right-handed operation the lines are “in phase” with each other (with the outputted phase difference being the result of the quadrature hybrid, not the transmission lines to the port), and in left-handed operation the lines are “out of phase” with each other (which with the phase difference produced by the quadrature hybrid results in a lead on the output rather than a lag compared to the standard quadrature hybrid).

In full transmission operation the modified quadrature hybrid begins to utilize the FSOCs to isolate connections as necessary. With the second group of CLRH cells activating comes several key alterations to the circuit: all paths to the quadrature port are blocked, the path to the terminated port 4 is blocked, and the impedance matching structure in the lower half comes into effect. As the FSOCs operate as an electrical open while they are in the open state, these blocked paths are electrically equivalent to removing them from the structure. The impedance matching structure functions through this principal as well, with the original path of $Z_0/\sqrt{2}$ impedance electrically opening and a new path of $Z_0$ beginning to allow transmission. The third group operates in a similar fashion, except that the path of $Z_0$ impedance to the quadrature port begins to allow transmission and the lower half of the structure is fully isolated by opening FSOCs. The equivalent circuits of these different states are displayed in Figure 4.6. In this figure all lengths between nodes are $\lambda/4$ unless stated otherwise.
To provide an initial proof of concept the modified quadrature hybrid is simulated using ANSYS Nexxim Circuit Simulator under ideal conditions (Figure 4.7); it is designed to operate within the same bandwidth as the stacked patch designed in Chapter 2.12, 10.7GHz to 12.7GHz. In this bandwidth, the four operating frequencies are evenly distributed across with $f_0 = 12.7\text{GHz}$, $f_1 = 12.05\text{GHz}$, $f_2 = 11.35\text{GHz}$, and $f_3 = 10.7\text{GHz}$. Because the Nexxim circuit simulator is incapable of simulating the CLRH-TLs the circuit is first designed to function in right-handed operation, then variables (CLRH1, CLRH2, and CLRH3) are added to the path lengths in which a CLRH-TL would normally be placed. To emulate each mode of operation of the modified quadrature hybrid these variables are then progressively switched between 0mm for right-handed
operation and 9.7mm (λ/2) for left handed operation.

![Graphs showing S-parameters for different modes: (a) Standard Quadrature Hybrid S-Params (dB), (b) Inverse Phase Quadrature Hybrid S-Params (dB), (c) Standard Quadrature Hybrid Phase Diff. (°), (d) Inverse Phase Quadrature Hybrid Phase Diff. (°), (e) In-Phase Port Transmission S-Params (dB), (f) Quadrature Port Transmission S-Params (dB).]

Figure 4.8: Ideal Case Modified Quadrature Hybrid S-Parameter Analysis

Figure 4.8 displays the relevant data to this ideal simulation. The key components of each operating mode are clearly visible: even power split and proper phase differences in the quadrature hybrid modes as well as nearly full transmission to the specified port in the transmission operating modes. There is some concern with the reflection in the quadrature port transmission operating mode, however it is not enough so to invalidate the design in this stage.
4.3. Verification and Validation of Modified Quadrature Hybrid

To design the modified quadrature hybrid it is first verified to function in right-handed mode without CLRH cells. The modified quadrature hybrid concept is divided into two operating modes, circular polarization (standard quadrature hybrid operation, right in Figure 4.9) and linear polarization (full transmission to quadrature port, left in Figure 4.9); path-length phasing is used to emulate the effects of the CLRH cells in a design for each of these, with the circular polarization design emulating the right-handed mode and the linear polarization design emulating the operation at $f_3$ (fully left-handed). NC-FSOCs are left as parallel lines in the circularly polarized designs and are altered to be out of phase transmission lines in the linearly polarized design; the converse is true for NO-FSOCs. The dimensions of the 3D model for each of these designs is defined fully through the use of variables, allowing for the rapid re-tuning of any component within it. The models for each of these two path-length phased designs are placed within the same HFSS design file and simulated simultaneously; this allows for some level of verification for both polarization outputs during the layout stage of the design rather than assuming linear function is the result of circular
function. Because the same variables are used to define the same lengths between each model all physical changes are propagated to both designs simultaneously, emulating a single design that utilizes CLRH-TLs to achieve the same effects as both path-length phased designs in one.

Following modeling and setting initial lengths based on the ideal simulation (Figure 4.7), key variables are chosen for use with the inbuilt genetic algorithm in HFSS and 1000 iterations of the design are run for fine tuning. Each of these iterations is evaluated against a cost function that describes their conformity to the ideal case: full transmission to the quadrature port in the case of the linear polarization design and even power split with 90 degrees phase difference in the case of then circular polarization design. The target frequencies for these objectives on their respective designs are identical to the initial frequency allocation in the ideal simulation (Figure 4.8). The genetic algorithm then semi-randomly searches within the defined parameters, using generations of similar iterations. In the case of this design the semi-random selection is more useful than the sequential nonlinear programming used for the antenna as it is capable of generating multiple different successful layouts to choose from rather than converging on a single optimal layout. At the end of the optimization cycle manual evaluation is performed on the lowest cost iterations to determine the most successful overall design, with that design chosen as the base layout to integrate the FSOCs with.
Figure 4.10: Path-Length Phased Designs S-Parameter Analysis (Simulation)

Figure 4.10 depicts the simulation data of the selected iteration from the genetic algorithm’s permutations (the same layout as in Figure 4.9). As a result of the genetic optimization the S-parameters are comparable to the ideal simulation in the Nexxim circuit, though losses from mismatching are approximately 0.5 to 1\,\text{dB} across the board. Some losses are expected, of course, as the Nexxim simulation does not take any bends, manifolds, or width junctions into account.
Figure 4.11: Manufactured Path-Length Phased Designs
Figure 4.11 displays results of manufacturing the boards of the path-length phased quadrature hybrid designs and their S-parameters as measured on the E8363B PNA Series Network Analyzer. The measurements align with the simulations at the target frequencies to an extent, with the circular polarization design displaying an even power balance and approximately 90° phase difference at 12.7GHz and 12.05GHz, however the reflection to the input port is larger than expected. Likewise, the linear polarization design exhibits transmission behavior to the intended port, but the reflection to the input is also greater than expected. A potential source of this discrepancy from the simulation is the terminations utilized; their behavior in the measured band of 9GHz to 15GHz offers poorer matching than desired as seen in Figure 4.11(f). To accommodate this issue the broadband termination of the calibration kit is utilized, however one port is still terminated with the aforementioned termination. The behavior of this termination, in conjunction with the good performance of the integrated antenna variation (Figure 4.14), suggests that the designs are more functional than is apparent from these collected measurements, though there is some concern over the reflection in the linear design. Future iterations should investigate the FSOC matching as a potential source of this reflection.
Figure 4.12 presents the simulation data from the integration of the circularly polarized path-length phased design with the antenna designed in Chapter 2. The results suggest good performance from the circuit, with the circular polarization at both design frequencies aligning well with the simulated circular polarization in Figure 2.13 (c) and (d). Compared to the ideally fed simulation the main lobe displays some distortion and loss of gain, however the results are still well aligned. The S-parameters of the integrated antenna also display sufficient magnitude, overall providing an optimistic outlook of the modified quadrature hybrid’s function in circular polarization.
Figure 4.13 depicts the simulation data from the integration of the linear polarized path-length phased design with the aforementioned antenna. The effects of the middle-of-band reflection present in the manufactured stand-alone design are clearly visible in this simulation, with the center of the band of the integrated design showing higher than expected reflection. Compared to the stand-alone antenna the radiation pattern is both more distorted and narrower with a higher cross polarization. The most likely cause of these distortions and cross polarization is improper isolation between the halves of the quadrature hybrid; if the halves are isolated as intended then the cross polarization should be on the level it is in the stand-alone antenna (Figure 2.13 (a) and (b)).
Figure 4.14 displays the S-parameter data collected after manufacturing the antenna-integrated designs. The S-parameters display good agreement with the simulation at the design frequency, confirming the inferences made. While the linearly polarized path-length phased design is less functional than desired, the circularly polarized path-length phased design operates as anticipated; this is more critical to the design of the modified CLRH-TL quadrature hybrid as it is the layout on which the CLRH cells will be implemented.
The radiation patterns of the integrated antennas are pictured in Figure 4.15. Data was collected with a vertically polarized linear antenna, therefore cross-polarization is unavailable for circular polarizations; instead the evenness between the E and H-planes is used to determine polarization purity. From the collected data the circularly polarized design aligns well with the simulations, displaying even response to linear polarization along both alignments. The linearly polarized designs display a more distorted radiation pattern than the base antenna, however their cross-polarization demonstrates some degree of linear polarization. With additional design tuning based on these results it is likely the radiation pattern could be further improved; the main source of cross-polarization is likely improper isolation between the halves of the quadrature hybrid. With the validation of the path-length phased designs, the CLRH-TLs are added to the circularly polarized
The full modified quadrature hybrid design consists of the dimensions of the circular polarized path-length phased design (as that is its intended right-handed operation) along with the implementation of the FSOCs as depicted in the initial design concept (Figure 4.6). To support the original design goal of providing four polarization types from one port as a reconfigurable antenna, port 4 is terminated by a lumped 50Ω resistor. Capacitors are indicated with blue, inductors are indicated with red, and the sole termination resistor on port 4 is indicated with green. Each set of three neighboring CLRH cells contain identical capacitance and inductance values leading to eight unique cell groups. The magnitude of the inductors of each cell group are set to a variable...
that reflects the appropriate multiple of their respective capacitance counterparts in order to satisfy
(1) for the transmission line width they are placed on. After setting initial capacitance values (and
therefore initial inductor values) based the individual FSOC’s simulation and validation, a genetic
algorithm optimizer is run again, in this case exclusively for the capacitance groups with no layout
changes. The cost function displays the same priorities as the path-length phased designs, however
they have been combined into a single layout.

(a) Modified Quadrature Hybrid S-Parameters (dB)

(b) Modified Quadrature Hybrid Output Ports Phase Diff. (°)

Figure 4.17: Modified Quadrature Hybrid (Simulation)
Simulation data for the modified quadrature hybrid is unfortunately highly indicative of design infeasibility; it is, however, in line with the results of the performed FSOC verification and validation (Figure 3.13). It is discovered through this design that considering the switching overhead of the FSOCs as 1.5GHz (13%) is an optimistic interpretation with the actual value lying closer to 3GHz (25%); examining the S-parameter plot of the modified quadrature hybrid simulation in Figure 4.17 (a) reveals that at 10.2GHz the design is operating well in full quadrature transmission and at 12.9GHz the design maintains an even power split between the quadrature and inphase ports. This behavior reflects the intended outcomes of the two edge frequencies of $f_3$ and $f_0$, states in which all of the CLRH-TLs are in left-hand operation or right-hand operation respectively. Upon even closer examination, it can even be seen that at the full inphase port transmission frequency of 11.35GHz the inphase port actually is receiving the majority of the power. It can be concluded from this, then, that the primary issue with the modified quadrature hybrid in its current state is that it attempts to combine too many FSOC transitions in one bandwidth. If something could be done to either improve the switching bandwidth of the FSOCs, or widen the bandwidth of the right-hand operation of the quadrature hybrid, then the design would see noticeable improvement.
5. Conclusion and Future Work

The proposed frequency selective open device provides the expected function, however its switching overhead is too high for its proposed use. With this limitation now proven, alternate implementations may be considered, such as a modified quadrature hybrid that provides only two modes of polarization from a single port operating under generally the same mechanism. This modified quadrature hybrid would utilize two frequencies separated by a band of approximately 25% width around the center, with the lower frequency designed to provide transmission from port 1 to port 2 and from port 4 to port 3 and the upper frequency designed to function as a standard quadrature hybrid. Instead of utilizing NC-FSOCs for impedance matching, this modified quadrature hybrid would instead investigate an implementation of a variation of the open/short-circuit stub demonstrated in [8] to provide stub tuning. In linear operation this modified quadrature hybrid would still utilize NC-FSOCs to isolate the halves, with additional investigation into this functionality performed to ensure good isolation.

Figure 5.1: Triangular Array Layout

Following the development of a functional modified quadrature hybrid and its integration with the antenna investigation into array integration, specifically triangular array[31][32] integration, may be performed with a focus on beam steering. Previous designs have successfully utilized CLRH-TLs for beam steering in rectangular arrays[9] making triangular array steering a natural next step.
References


