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REAL-TIME AND PORTABLE MICROWAVE IMAGING SYSTEM

by

MOHAMMAD TAYEB GHASR

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Approved by

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ABSTRACT

Microwave and millimeter wave imaging has shown tremendous utility in a wide variety of applications. These techniques are primarily based on measuring coherent electric field distribution on the target being imaged. Mechanically scanned systems are the simple and low cost solution in microwave imaging. However, these systems are typically bulky and slow. This dissertation presents a design for a 2D switched imaging array that utilizes modulated scattering techniques for spatial multiplexing of the signal. The system was designed to be compact, coherent, possessing high dynamic range, and capable of video frame rate imaging. Various aspects of the system design were optimized to achieve the design objectives. The 2D imaging system as designed and described in this dissertation utilized PIN diode loaded resonant elliptical slot antennas as array elements. The slot antennas allow for incorporating the switching into the antennas thus reducing the cost and size of the array. Furthermore, these slots are integrated in a simple low loss waveguide network. Moreover, the sensitivity and dynamic range of this system is improved by utilizing a custom designed heterodyne receiver and matched filter. This dissertation also presents an analysis on the properties of this system. The performance of the multiplexing scheme, the noise floor and the dynamic range of the receivers are investigated. Furthermore, sources of errors such as mutual coupling and array response dispersion are also investigated. Finally, utilizing this imaging system for various applications such as 2D electric field mapping, scatterer localization, and nondestructive imaging is demonstrated.
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1. INTRODUCTION

1.1. MICROWAVE AND MILLIMETER WAVE IMAGING

Microwave and millimeter wave imaging techniques have been successfully applied to a variety of applications such as nondestructive testing (NDT) of materials and structures [1]-[2], medical diagnosis [3]-[4], concealed weapon detection [5], and through-wall imaging [6] to name a few. Electromagnetic waves at microwave (300MHz – 30 GHz) and millimeter-wave (30 GHz – 300 GHz) frequencies, are able to penetrate a wide variety of optically opaque and non-conducting materials such as various composites, ceramics, concrete, wood, foam, and clothing and interact with their interior structure. Furthermore, signals at microwave and millimeter-wave frequencies are non-ionizing and are not considered to be sources of hazardous radiation, other than creating heat in an object at high power levels.

Microwave NDT techniques have been extensively used in the past for characterizing the dielectric and magnetic properties of a wide variety of materials as material characterization tools [1]-[2], [7]-[9]. Other applications include surface defect (e.g. cracks and corrosion) detection on metallic surfaces [10]-[11] and detection of surface and subsurface defects in dielectric composite structures [1]-[2], [12]. Imaging an object, for NDT purposes, is very desirable since images provide a detailed and comprehensive “view” into the internal structure of the object [1], [13]-[14]. Microwave and millimeter wave imaging techniques demonstrated the utility in producing images with high fidelity when other techniques failed to accomplish the same [15].

Millimeter wave imaging has also been used for security purposes [5]-[6], [16]-[21]. The ability of millimeter wave signals to penetrate clothing makes them suitable for detection of concealed objects and contraband. Near-field 3D holographic imaging has been implemented for personnel screening check points such as in airports. This system was developed by the Pacific Northwest National Laboratory (PNNL) [5]. This system is an electronically scanned 1-D array which is mechanically scanned in the second dimension to provide a 2-D image in approximately 1.5 seconds. While 1.5 seconds may be acceptable for some applications, in many other applications it is desirable to have a “real-time” imaging system that can provide the same information. To this end there are
several ongoing efforts towards developing high-speed and high-resolution imaging systems [19]-[21].

Microwave imaging techniques in the biomedical areas has included applications such as hyperthermia monitoring [3],[22]-[24], skin cancer [4], and breast cancer detection [25]-[27]. In biomedical applications, there has been limited success in developing practical systems [22],[25]. However, most investigations have concentrated on the theory of imaging based on reconstruction techniques employing simulated data or experimental data obtained using raster scanning.

Microwave imaging is based on collecting the coherent relative scattered field from an object over a known two-dimensional (2D) plane. To measure the electric field on a 2D plane in real-time, a 2D array of antennas is required. Currently, the widely used imaging technique is a basic imaging system consisting of a single antenna (probe) mechanically raster scanned over the scene or object of interest. This technique is very slow, yet it is the most accurate method available for the microwave and millimeter wave regime. The high accuracy is due to the absence of any source of interference such as mutual coupling between array element, array calibration inaccuracies, and limited spatial sampling in physical arrays. There are many systems that try to speed-up the imaging process using several types of 1D or 2D arrays of which some were successful for certain application. The following section gives a more detailed insight into the attributes of some of these techniques.

1.2. PREVIOUS WORK

Development of practical microwave and millimeter wave imaging systems have been limited by the available technology in various components such as sources, receivers, detectors, etc. Recently, the development toward microwave and millimeter-wave imaging has evolved from the two opposite ends of the frequency spectrum, high frequency millimeter wave and relatively lower frequency microwave systems. The higher frequency millimeter wave systems utilize techniques borrowed from the well-developed optical region where a physical lens is utilized to focus the 2D scattered field on a focal plane 2D array of millimeter wave detectors [19],[28]-[31]. These techniques are real-time and compact. However, they have a relatively small aperture and limited
field of view due to practical constraints. Furthermore, the resolution of these systems is low compared to the wavelength.

The microwave systems measure the coherent 2D scattered field directly using a multiplexed array of antennas. Subsequently, algorithms such as synthetic aperture is used to back-propagate the electric field to the source of scattering (i.e., form an image of scattering sources) [5],[32], or algorithms such as reconstruction techniques to obtain information about the geometrical and dielectric distribution of the object (or essentially its image) [25]-[27]. Microwave systems allow for possible real-time operation and they can provide high resolution. These systems are typically custom designed for specific applications. The high resolution is attained in the near-field of the array [5],[32]-[33], and the depth of focus is limited by the size of the array [5]. At far-field distances, these systems may be operated as phased arrays to provide for angular scanning of their field of view. In this case their resolution will be dependant on the beamwidth of the array, which depends on the array size. There are many challenges in designing such systems. These challenges come from the requirements of a tightly spaced measurement grid (λ/2 spacing) dictated by Nyquist sampling theorem. Two methods have been used for multiplexing array elements. The first method utilizes switching capabilities in the array (switched array) to route the signals from an array element to the transceiver. The second method utilizes modulation scattering techniques (MST) to distinguish the desired signal at the receiver. Both methods are explained in this section.

**1.2.1. Switched Array.** Switched array systems utilize a set of RF switches [5]-[6],[17] to multiplex the array antennas with the receiver that perform the coherent measurement. Switching may also be possible in the IF domain utilizing low frequency switches. This method would require a down-converting RF front-end on each array element. Either concept proves to be expensive, challenging, and bulky. Another challenge is to design small yet efficient array elements (antennas). These antennas must be much smaller than λ/2 in order to fit in the tightly spaced λ/2 grid. To date the development of large scale 2D imaging arrays have been limited due to these challenges. At high frequencies exceeding 20 GHz, only 1D successful imaging systems based on switched array have been developed [5].
1.2.2. Modulated Scattering Technique. Other arrays that perform the spatial multiplexing are those based on modulated scattering techniques (MST) which has been widely used for antenna pattern measurements [34]. The use of MST for electric field measurements was first introduced in [35]. Since its introduction, MST has been utilized using various linear scatterers, such as small dipole antennas, and has been extensively used for electric field mapping and imaging applications. Several MST-based 1D imaging systems have been designed and built with an array of sub-resonant dipoles at frequencies up to 12 GHz [22]-[24],[34],[36].

MST tackles some of the issues with switched arrays (bulkiness and isolation) by modulating “tagging” the received signal at a low frequency. Using MST, the measured signal is distinguished and spatially localized to the probe location by proper modulation and demodulation techniques. The advantage of MST is that multiplexing is performed at a very low frequency, typically 100’s of KHz. Another advantage of MST is that, it requires only a single RF front-end and receiver thus reducing the overall cost. Traditional MST has a couple of limitations [37]. The small dipole used provide for very small modulation depth. The typical modulation depth (ratio of power in modulated signal to the incident signal) of -40dBc [36] to -70dBc [24] limits the sensitivity and dynamic range of the overall system. Another limitation with MST is the signal transfer from the scatterer to the receiver. Spatial collection schemes or passive combiners are lossy solutions, lowering the overall system dynamic range [34]. Furthermore, the mutual coupling among the array elements (dipoles) can significantly limit system dynamic range. These problems become even more significant and challenging at higher frequencies such as those in the millimeter-wave region where currently there is a great need for rapid imaging systems at these frequencies.

1.3. CURRENT WORK

This work describes the design and development of a novel 2D microwave imaging system (i.e. Camera) capable of rapid electric field mapping, operating at 24 GHz. There are certain requirements to be optimized. The design should allow for adequate spatial sampling dictated by Nyquist theorem, adequate aperture size, high dynamic range, and video rate image acquisition. Moreover, coherent detection is a
must. Coherent sampling of the microwave field will allow for utilizing back propagation or other image reconstruction technique vital for meaningful image output. This prototype is 24 x 24 (576 element) array of slot antennas spaced by $\lambda_0/2$, where $\lambda_0$ is the free-space wavelength. The array covers an area of 6” by 6”. Other features of this design are the relatively low cost, coherent wave-front measurement, and real-time operation. This design is easily scalable for at least the next two bands in the millimeter-wave regime, namely the Ka- and Q-bands.

As mentioned earlier, an RF multiplexed array of antennas provide for the closest performance to a scanned system. Practical multiplexing would require high isolation between the array elements through the switching network, and very low signal loss (insertion loss) between the array element and the source/receiver. Any amplitude or phase dispersion between the array elements will consequently have to be measured and calibrated for. However, compromising between all of these requirements might not lead to desirable performance. The design described hereafter greatly improves on the limitations of available imaging systems. This design is a combination of an RF multiplexed system and a set of MST probe antennas arranged as array elements. This design utilizes a PIN diode controlled resonant slot antenna [37]-[38] to perform the switching directly on the array elements rather than using RF switching. Subsequently, using measurement techniques derived from MST, the isolation and consequently the measurement accuracy is increased. The backbone of this array is a low loss waveguide network which routes the modulated signal from each array element to a set of receivers. The receivers down-convert the modulated signals to an IF stage where a sensitive demodulator is used to match filter the modulated signal. This system is capable of video frame rate imaging with a high dynamic range.

Section 2 of this dissertation discusses measuring electric field by modulation (MST). A comparison between the slot and scattering dipole as electric-field mapping probes is performed. Furthermore, a description of a 2D 30-element array prototype is included along with some preliminary imaging results. Section 3 describes the design of the camera. Each component of this camera was optimized to be integrated in the final system. Detailed description, optimization, and design feature of various components in this design are presented. Section 4 provides an analysis on the overall performance of
the system. The performance of the multiplexing scheme, the noise floor and the
dynamic range of the receivers are investigated. Other investigations address the sources
of errors and error correction techniques. Section 5 describes examples of various
applications of this microwave camera. This camera is primarily a 2D electric field
mapping device. Experimental results of electric field measurement are presented along
with accuracy figures obtained from measurements. Utilizing this camera as an array for
localizing sources/scatterers is also demonstrated. Furthermore, using this camera for
imaging applications in through transmission and reflection modes is demonstrated.
Finally section 6 offers a summary and discussion of this investigation with suggestions
for future improvements.
2. REQUIREMENTS AND CONCEPT OVERVIEW

2.1. COHERENT 2-D ELECTRIC FIELD MAPPING OVERVIEW

One approach to microwave and millimeter wave imaging requires mapping of scattered electric field distribution on a surface which may take a planar, cylindrical or spherical shape. This electric field may be scattered from an object being imaged or in case of passive imaging, emitted by the object being imaged. When measuring the magnitude and phase of scattered field, a small antenna (probe) may be used to measure the field distribution as a function of location on the measurement plane. However, this approach requires mechanical scanning of the small antenna in the measurement plane which renders the approach cumbersome, time consuming and the scanning mechanism may also disturb the field of interest. Consequently, one may use an array of small antennas situated in the measurement plane and extract electric field distribution data from each individual antenna (i.e., no need for moving an antenna in the measurement plane). A general diagram of such an imaging system based on this concept is shown in Figure 2.1. Many individual antennas are used to collect the electric field distribution on a predetermined 2D measurement plane. As will be shown later through spatial multiplexing, the desired electric field information from each antenna in this array is coherently (i.e., magnitude and phase) measured by the RF circuit. The RF circuit also generates the incident electric field used to illuminate the object. The processor performs the data collections by synchronizing the RF circuit measurements with the spatial multiplexing. Further steps performed by the processor such as any required calibration or signal processing makes these data representative of the scattering object or the source being imaged. The basic designs associated with the RF circuitry and the processors are well established, yet they may be optimized for a specific application such as an imaging system (microwave camera). The array designed for this type of electric field distribution measurement is commonly referred to as a “retina”, for its similarity to the function of one’s eye retina [23]. The terms imaging system / microwave camera, array / retina, and antenna / probe may be used interchangeably in this dissertation.
The retina samples the scattered electric field distribution at the location of each antenna element. Thus, the spacing between individual antennas must be properly chosen so that the measured discrete electric field distribution is a close representative of the continuous electric field distribution on the retina. Consequently, Nyquist Sampling Theorem [39] dictates that the array element center-to-center interspacing not be larger than \( \lambda/2 \) (\( \lambda \) is the operating wavelength). Any larger interspacing results in errors associated with the reconstructed scattered electric field distribution. This interspacing requirement also affects the type (i.e., size) of antenna that may be used to construct the retina, as will be discussed later.

Finally, the imaging system must be capable of distinguishing electric field information for each antenna. One approach to accomplish this is to spatially "tag" or "multiplex" the scattered electric field. One attractive approach for doing so is based on modulated scatterer technique (MST). MST has been developed and used mainly for antenna pattern mapping applications. Direct measurement of the electric field requires connecting a small probing antenna through transmission lines (e.g. coaxial line) to the
RF receiver. In applications such as near-field antenna pattern measurements, these transmission lines were found to be prohibitively disturbing the field pattern. Furthermore, the coaxial transmission lines usually do not hold their wave propagation properties during the twists and turns associated with the mechanical scanning process, especially at high frequencies. Justice and Rumsy [40] introduced the concept of indirect electric field mapping using auxiliary scatterers. This method utilizes a small scatterer to locally perturb the desired electric field and re-scatter an electric field proportional to the desired electric field towards a receiver, thus eliminating the need for perturbing transmission lines. This efficacy of this method is limited by the instability of the source, presence of other unwanted scatterers, and other sources of clutter. To overcome these issues, Richmond [35] introduced the concept of modulating the scatterer (or probe) so that the desired electric field can be effectively distinguished from un-modulated fields due to clutter, etc. The concept of modulated scatterer technique is illustrated in Figure 2.2. MST utilizes an antenna which is usually a small dipole (for electric field measurement) loaded with a non-linear device, which is typically a fast switching PIN diode [34]. By switching the PIN diode ON and OFF (forward and reverse bias) through a modulation pulse train, the scattered field from the dipole becomes modulated. Subsequently, this modulated scattered field is picked up and demodulated, resulting in electric field information (magnitude and phase) at the location of dipole. Thus, if the dipole is placed in the 2D measurement plane (retina), then the electric field distribution over the retina is measured (either by scanning one dipole or using an array of them). Utilizing a coherent receiver followed by a lock-in amplifier results in information proportional to the magnitude and phase of the electric field at the location of the scattering antenna or dipole.

The electromagnetic formulations describing the scattering, effect of the non-linear element on the scatterer and propagation/coupling into the receiver has been extensively covered in literature [34]-[35], [41] and are not in the scope of this work. The upcoming subsection will describe the measurement technique from the system and signals point of view. Critical parameters dictating the performance of the measurement technique and improvement possibilities are investigated.
2.2. ELECTRIC FIELD MAPPING USING MST

Figure 2.3 shows a bi-static arrangement typically used for antenna pattern measurement using MST [34]. In this arrangement a modulated scatterer is used to measure the field pattern radiated by the antenna under test (AUT). Bi-static refers to the fact that an auxiliary pick up antenna is used to collect the modulated scattered signal. The received modulated signal is routed to a coherent homodyne receiver, which down-converts the scattered signal, using quadrature technique, resulting in an in-phase (I) and a quadrature (Q) output. The homodyne receiver, as shown in Figure 2.3, is typically used for this type of measurement since it is a low cost way to obtain magnitude and phase of a desired signal [34]. Finally, the I and Q are fed into a dual lock-in amplifier which is used to demodulate the received signal, resulting in a signal proportional to the real and imaginary parts of the desired electric field at the location of the scatterer.
What follows is a signal level formulation for this measurement process. The received electric field at the pick up antenna is a quadrature amplitude modulated signal, defined as:

$$E_r(t) = E_c(t) + Km(t)E_p(t)$$  \hspace{1cm} (2.1)

where $E_c(t)$ represents un-modulated electric field which may be due to other scatterers (i.e., clutter) or direct coupling into the pick-up antenna, $E_p(t)$ is the desired electric field at the location of the scatterer, $K$ is a proportionality constant accounting for gain of the
pick-up antenna and propagation from the scatterer to the pick-up antenna and \( m(t) \) describes the modulated scattering proportions of the scatterer, which is defined as

\[
m(t) = \frac{X_H + X_L}{2} + \frac{X_H - X_L}{2} s(t) \tag{2.2}
\]

where \( X_H \) and \( X_L \) are reflection or transmission coefficients described as the ratios of the scattered electric field to the incident electric field at the scatterer for the ON and OFF states of the PIN diode, respectively [42], and \( s(t) \) is the modulating square wave (i.e., pulse train) with fundamental frequency of \( f_m \) defined as:

\[
s(t) = \sum_{n=1}^{\infty} a_n \sin(n2\pi f_m t) \tag{2.3}
\]

The first term of \( m(t) \) is a constant which will result in an un-modulated scattered electric field, and the second term is time-varying which will result in the modulated part of the electric field. Ignoring the clutter, \( E_c(t) \), as it is assumed not to pass through the scatterer, the modulation depth, \( h \), can be written as:

\[
h = \frac{|X_H - X_L|}{|X_H + X_L|} \tag{2.4}
\]

For any particular scatterer and its non-linear modulating load (i.e. PIN diode), \( X_H \) and \( X_L \) can be described by the electromagnetic properties of the scatterer and its load, specifically their impedances. As will be shown later, these scattering parameters will have a direct impact on the overall sensitivity of the system. The parameters \( K, X_H, \) and \( X_L \) are all complex phasors having a magnitude and phase associated with them. These quantities are typically constant and independent of the incident field at moderate powers. Ultimately as it will be explained later, a reference measurement is used to correct for the magnitude and phase of these quantities.
When considering a single-frequency continuous-wave (CW) source with frequency $f_0$, equation (2.1) may be written as

$$E_r(t) = \text{Re}\left\{ A e^{-i(2\pi f_0 t + \phi)} + K m(t) A_p e^{-i(2\pi f_0 t + \phi_p)} \right\}. \quad (2.5)$$

In the homodyne receiver, $E_r(t)$ is mixed with the following reference signal from the source for the in-phase and quadrature mixers:

$$E_0(t) = \begin{cases} \text{inphase} & \text{Re}\left\{ A_0 e^{-i(2\pi f_0 t + \phi)} \right\} \\ \text{quadrature} & \text{Re}\left\{ -j A_0 e^{-i(2\pi f_0 t + \phi)} \right\} \end{cases} \quad (2.6)$$

Mixers perform the tasks of signal multiplication and the output is then low-pass filtered, yielding the following in-phase and quadrature signals:

$$i(t) = \text{Re}\left\{ A A_0 e^{-i\phi(t)} \right\} + \text{Re}\left\{ K \left( \frac{\chi_H + \chi_I}{2} \right) A_p A_0 e^{-i\phi(t)} \right\}$$

$$+ \text{Re}\left\{ K \left( \frac{\chi_H - \chi_I}{2} \right) A_p A_0 e^{-i\phi(t)} \right\} s(t) \quad (2.7)$$

$$q(t) = \text{Re}\left\{ -j A A_0 e^{-i\phi(t)} \right\} + \text{Re}\left\{ -j K \left( \frac{\chi_H + \chi_I}{2} \right) A_p A_0 e^{-i\phi(t)} \right\}$$

$$+ \text{Re}\left\{ -j K \left( \frac{\chi_H - \chi_I}{2} \right) A_p A_0 e^{-i\phi(t)} \right\} s(t) \quad (2.8)$$

The first and second terms of $i(t)$ and $q(t)$ are non-time varying signals due to the clutter ($1^{st}$ term) and the non-suppressed carrier ($2^{nd}$ term). The third term of $i(t)$ and $q(t)$ is a square wave at the frequency of $s(t)$ which also contains information about the incident field. The lock-in amplifier projects the down-converted signals $i(t)$ and $q(t)$ on the modulating signal $s(t)$, which results in two DC voltages proportional to the real and imaginary parts of $E_p$, as shown in the following equations.
\[ I = \langle i(t) \cdot s(t) \rangle = \text{Re} \left\{ K (\chi_H - \chi_L) A_0 A_r e^{-j(\phi - \phi_0)} \right\} \]  
\[ (2.9) \]

\[ Q = \langle q(t) \cdot s(t) \rangle = \text{Re} \left\{ -jK (\chi_H - \chi_L) A_0 A_r e^{-j(\phi - \phi_0)} \right\} \]
\[ = \text{Im} \left\{ K (\chi_H - \chi_L) A_0 A_r e^{-j(\phi - \phi_0)} \right\} \]  
\[ (2.10) \]

In equations (2.9) and (2.10), \( K \), \( \chi_H \), \( \chi_L \), \( A_0 \) and \( \phi_0 \) are constants and may be properly corrected for using a known (calibration) measurement, since they are the properties of the system and are independent of the electric field being measured, therefore the desired electric field vector is:

\[ I \propto \text{Re} \left\{ A_r e^{i\phi} \right\}, \]  
\[ (2.11) \]

\[ Q \propto \text{Im} \left\{ A_r e^{i\phi} \right\}. \]  
\[ (2.12) \]

For the ideal case of \( K = 1 \), which corresponds to when all of the scattered electric field by the probe reaches the receiver, the modulation depth will depend solely on the extent of change in probe scattering properties between the ON and OFF states of the PIN diode (i.e. \( |\chi_H - \chi_L| \)). This fact has direct impact on the overall system signal to noise ratio (SNR). The ideal scattering probe is one that modulates between an open and short (i.e. \( \chi_H = 1, \chi_L = -1 \)), albeit not being physically realizable with a direct antenna modulation scheme. This ideal scattering probe produces a double side-band suppressed carrier (DSB-SC) type modulation resulting in no wasted power in the carrier. Second best option is a probe that modulates between a total reflection and a matched load (i.e. \( \chi_H = 1, \chi_L = 0 \)) corresponding to 100% modulation depth, which is also not easily achieved in practice. Figure 2.4 shows comparison of simulated phase error and dynamic range of the system described in Figure 2.3 and equations (2.1)-(2.12). Three cases of scatterer modulation depth were simulated namely, the ideal DSB-SC case, the 100% modulation depth case and a \( \lambda/2 \) dipole case. Gaussian noise, with a power of -80 dB, was added at the input of the mixers and its impact may be noticed in the floor of the dynamic range for the three cases, shown in Figure 2.4(a). These figures also show a
one-to-one correlation between reduction in modulation efficiency of the scatterer and the degradation of the overall system performance in estimating the amplitude and phase of the signal. For example using a $\lambda/2$ dipole with a modulation depth of approximately 15%, reduces the useful dynamic range of the system by as much as 20 dB. Furthermore, when used in a 2D array, the $\lambda/2$ dipole does not fit the center-to-center spacing of $\lambda/2$, and the $\lambda/4$ dipoles which are commonly used in a 2D array presents a much smaller modulation efficiency, which further reduces the useful dynamic range of the system. The small dynamic range may be acceptable in some applications, such as antenna pattern measurement, where the user can compensate for lower dynamic range by increasing the radiated power. However, in imaging applications the scattered field most likely will present a relatively large variation in power over the retina domain [42]. Furthermore, targets of interest are usually weak scatterers (i.e., small, embedded in a dielectric host, etc.) and their associated power scattered power level will be near the noise floor of the system.

![Graph](image1.png)

**Figure 2.4.** Simulated system performance for various modulation efficiencies: (a) dynamic range, and (b) phase error.

Overall, the modulation depth of the scattering probe will greatly influence the performance of systems whose designs are founded on utilizing MST for measuring
electric field distribution. The above simulation intended to present the critical aspect of modulation depth and did not account for other issues such as distortions caused by the band-limited response of components, most importantly the non-linear components of the system such as amplifiers which tend to decrease the sensitivity of such systems especially when the power is high. Furthermore, having a strong carrier and the non-modulated clutter will limit the amount of gain these systems may have, thus further limiting the dynamic range of the system. The next subsection will present an alternative to dipole antennas that greatly improve the modulation depth, and clutter rejection capabilities of the scatterer in MST based electric field measuring systems.

2.3. RESONANT SLOT vs. DIPOLE SCATTERER

Shortcomings of the dipole scatterer when used in imaging arrays are multifold. As shown in Figure 2.4, the low modulation depth translates into poor overall system SNR and small dynamic range. Furthermore, when used in an imaging array with an element interspacing of $\lambda/2$ or smaller the dipoles are made sub-resonant (smaller than $\lambda/2$), which greatly reduces the modulation depth. One other critical issue is the strong coupling between these dipoles, which further degrades the overall system SNR and dynamic range. These issues become even more significant at higher microwave frequencies. The dielectric substrate that is typically used to build the array onto, will act as a low loss propagation medium to increase coupling between the array elements. Furthermore, parasitics due to the bias lines has a bigger contribution to reducing the modulation depth at higher frequencies.

Imaging arrays utilizing slot antennas provide many desirable features that improve most of the shortcomings of dipole antennas. The slot array is manufactured into a conducting sheet. The slots may be designed such as they are resonant and provide high modulation depth while maintaining dimensions smaller than $\lambda/2$. The slots can be operated such that they are closed and opened only to receive a signal. These attributes increase the power in the modulated signal, while reducing the power in the un-modulated carrier [38].

The design and efficacy of such a slot, for the purpose of using it in an imaging retina array, was proposed, described and analyzed in [38]. The schematic of this slot is
shown in Figure 2.5. This is an elliptically-shaped resonant slot loaded with a circular patch. Due to the elliptical shape, the slot becomes linearly polarized with the electric field being concentrated between the circular load and the surrounding conductor on top and bottom [38]. By shifting the circular load from the center of the elliptical slot to near its lower edge, the concentration of the electric field increases in that gap between the circular load and the elliptical slot, which becomes an ideal location for placing a switching PIN diode. When the PIN diode is turned OFF (reverse biased), the slot is open and passes the signal through it efficiently. Conversely, when the PIN diode is turned ON (forward biased), the slot is closed and almost no signal passes through it. These properties allow the slot to provide near 100% modulation depth when used in a transmission-through configuration. It has also been shown that the mutual coupling between two such slots is low compared to dipoles used for the same purpose [43]. Moreover, the presence of a ground plane allows for reducing the effect of the dielectric substrate and bias structure by properly shielding them from interacting with the electromagnetic field.

Figure 2.5. Schematic of the PIN diode loaded resonant elliptical slot.
There is one major difference between the slot antenna and the dipole in the way they are used, which limits the utility of the slot for general MST use. The slot is typically used in through transmission (i.e., the signal is passes through it) while the dipole is used in scattering mode. Furthermore, the slot is surrounded by a conducting sheet, which makes reflective when it is closed. These two properties of the slot, makes it unsuitable for mapping of near-field distributions for example mapping of antenna aperture electric field distribution.

This slot behaves as a resonant antenna and hence possesses a resonant frequency response. The resonant frequency, which is intended to coincide with the operating frequency of the imaging system, is dictated by the dimensions of the slot, the electrical properties of the dielectric substrate, and the capacitance of the PIN diode in the OFF state. The resonant frequency is determined through a 3D electromagnetic simulation of the overall slot structure. Figure 2.6 shows the simulated (using CST-MWS [44]) return-loss, looking into the elliptical resonant slot mounted on the aperture of a K-band (18-26.5 GHz) waveguide. In the simulation tool, the slot was built on a 0.5 mm thick Rogers4350 [45] dielectric substrate. The ellipse had a major radius of 2 mm and an axial

Figure 2.6. Simulated return-loss of the elliptical resonant slot.
ratio of 0.75. The load radius was set to 1 mm and the gap between the load and outer conductor was set to 0.2 mm (dictated by the PIN diode size). The PIN diode reverse state capacitance was set to 20 fF and its forward resistance was set to 5 ohms which are nominal values for commercially available PIN diodes [46]. For this structure the simulations results showed a resonant frequency of 24.63 GHz. When the PIN diode is turned OFF, the slot exhibited a return-loss greater than -20dB, i.e. the slot is open and transmits more than 99% of the signal through. When the PIN diode is turned ON (forward biased) the slot becomes closed except for a very small leakage (return loss of ~0.2 dB). This is translated into a modulation depth of approximately 78%.

Figure 2.7 shows the modulated transmission-through ($S_{12}$) response of similar resonant elliptical slot antenna designed to operate at 24 GHz measured using an HP8510C vector network analyzer (VNA). The measurement setup is shown in Figure 2.7a. The slot response transitions between a small leakage signal when the slot is closed, and a strong coupled signal when the slot is open. The modulation depth is greater than 20 dB. The modulation causes the signal through the slot to switch between two points in the complex domain. These two points constitute a vector. The length and angle of this vector corresponds to the magnitude and phase of the incident electric field.

Figure 2.8 shows a set of measured modulated signals in vector space (i.e., polar) representation. The measurement setup is similar to that shown in Figure 2.7. The left most red vector is the signal whose magnitude and phase are represented in Figure 2.7(b) and Figure 2.7(c) respectively. Shifting the phase of the incident signal by 20 degrees two times produce the other two red vectors. The black vectors represent signals which are 5 dB weaker in magnitude than the red vectors. The initial phase in the black vectors is arbitrary and caused by the attenuator. There is also a 20 degrees phase shift between these vectors caused by phase shifting the incident signal. Notice that these vectors follow the phase and magnitude of the incident field precisely. The vectors do not start from the origin due to the leakage in the slot when it is closed.
Figure 2.7. Modulated transmission-through ($S_{12}$) response of a resonant elliptical slot at 24 GHz; (a) measurement setup, (b) Magnitude, and (c) phase

Figure 2.8. Vector space representation of the modulated signal.
Figure 2.9 shows a comparison between the modulation depth of a resonant $\lambda/2$ dipole and the resonant elliptical slot both designed to operate at 24 GHz. Both antennas were used in a through transmission mode. For these measurements the slot was placed on the aperture of a K-band waveguide. The dipole was placed on the aperture of a small horn antenna connected to a K-band waveguide. Dipoles are not typically used directly on a horn or waveguide aperture. However, by placing the dipole on the horn aperture it is given similar advantage that the slot possesses, consequently larger modulation depth is obtained by collecting a larger portion of the scattered field.

![Graphs (a) and (b)](image)

(a) Measured Modulated spectrum ($f_m = 500$ KHz): (a) $\lambda/2$ dipole, and (b) resonant elliptical slot.

The spectra show the carrier at 24 GHz and a double side band square wave amplitude modulation with a fundamental frequency of 500 KHz. The $\lambda/2$ dipole (Figure 2.9a), generated a modulated signal of 23 dB below carrier, which corresponds to approximately 12% modulation efficiency. On the other hand the resonant elliptical slot (Figure 2.9b) generated a modulated signal of 7 dB below carrier corresponding to approximately 77% modulation depth.
2.4. ANTENNA PATTERN MAPPING

The following experiment demonstrates the capability of the resonant elliptical slot for electric field distribution measurement. In this experiment, a probing antenna is used to measure the radiated electric field magnitude and phase of an antenna under test (AUT) as a function of distance from that antenna. This experiment demonstrates the advantages of a matched probing antenna as well as the effect of modulating it. The experimental setup is shown in Figure 2.10 where each of three probing antennas is scanned perpendicular to the aperture of the AUT with a step size of $\lambda/4$. The AUT was a K-band open-ended rectangular waveguide operating at 24 GHz.

![Diagram of experimental setup](image)

Figure 2.10. Setup for measuring 1D electric field magnitude and phase variation vs. distance.

The probing antennas were a non-modulated open-ended rectangular waveguide (OEWG), a modulated $\lambda/2$ dipole mounted on a small horn, and a modulated resonant elliptical slot mounted on the aperture of a K-band waveguide. All antennas were
linearly polarized and the polarizations of the AUT and the probing antennas were matched in this experiment. The AUT and the probing antenna were placed inside a small anechoic chamber, while the rest of the test setup was outside the chamber. The positioning platform was placed on top of the anechoic chamber and covered by absorbing material to reduce the effect of unwanted reflections on the measurements.

The results of this experiment are shown in Figure 2.11. These measured electric field magnitude and phase are compared to the theoretical values obtained using the expression in equation (2.13). The derivation for this expression is shown in [2]. This expression defines the electric field when radiating into an infinite half-space, as TMz and TEz spectral functions $A^e(\xi, \eta)$ and $A^h(\xi, \eta)$. In the equation; $K$ is the propagation constant, $Z$ is the impedance of the medium, $\omega$ is the radial frequency, $\mu$ is the permeability of the medium, and $\xi, \eta, \zeta$ are the spectral wave-numbers. The spectral functions $A^e(\xi, \eta)$ and $A^h(\xi, \eta)$ are obtained by matching the boundary conditions as explained in [2].

$$\bar{E}_e(x, y, z) = \frac{1}{4\pi^2 K^2} \int \left[ -\xi \eta A^e(\xi, \eta) + \frac{\omega \mu \xi}{Z} A^h(\xi, \eta) \right] e^{-j(\xi x + \eta y + \zeta z)} d\xi d\eta \quad (2.13)$$

The average power received using the modulated slot and dipole antenna was lower than the OEWG case by 8.3 dB and 22.4 dB, respectively (not shown in the Figure 2.11). These figures are comparable to the modulation depth of the antennas as shown in Figure 2.9. The variation in the average received signals was accounted and corrected for to match the theoretical curve, at the distance of 100 mm (reference point). Figure 2.11(a) shows that all three measurements follow the theoretical curve on average. In the near-field of the antennas (distance < 50mm) the interaction (multiple reflections) between the AUT and the probe is noticeable as strong ripples in the measurement. Among the three probes, the OEWG shows the strongest multiple reflections, followed by the dipole. The Slot antenna shows very subtle multiple reflections due to the fact that it is a matched antenna with a return-loss of better than 20 dB as shown in Figure 2.6. Furthermore, the slot follows the theoretical curve better than the other two probes at the near-field region. In the far-field region where the received power becomes small, the
The advantage of modulating the probe is apparent. The modulated slot and even the modulated dipole antennas follow the theoretical curve closely. Even though the modulation depth associated with the dipole is not very high, given the large measurement system (i.e., receiver) dynamic range (>80 dB) the effect of this low modulation depth is not sensed. Conversely, the non-modulated OEWG probe measurement shows ripples due to the interaction of the received probe signal with signals coupled directly to the receiver from the source. This issue may be corrected by shielding the receiver, estimating the interference and subtracting it, or by modulating the probe as illustrated in this measurement. The non smooth amplitude ripples and the apparent phase error are due to the large step size associated with these measurements. The magnitude and phase errors compared to theoretical values are better presented in Figure 2.11 (b). The magnitude error in the far field is on the order of 0.001 and the phase error does not exceed 20 degrees everywhere. These errors may as well be due to alignment or positioning errors in the scanner. It must be noted that these measurements are performed at 24 GHz, where 20 degrees of phase error corresponds to less than 0.7 mm of position error.

Figure 2.11. Measured radiated electric field from an open-ended waveguide antenna. (a) Measured amplitude and phase vs. theory, (b) error compared to theoretical values.
2.5. INITIAL 2D IMAGING PROTOTYPE

To investigate the potential of using array of resonant elliptical slots in electric field mapping and imaging applications, a small scale prototype, operating at 24 GHz, was designed and built [37],[47] to demonstrate the proof-of-concept for designing such an imaging system. The design requirements and objectives were as follows:

- Adequate spatial resolution: The design should allow for sampling the electric field at steps of $\lambda/2$ or smaller for high resolution imaging.
- High overall system sensitivity and large dynamic range: The overall system dynamic range must be sufficient for capturing large variations in the electric field, yet the system noise floor must be low enough for the system to register signals from weakly scattering objects.
- Real-time operation: The electric field over the retina must be sampled rapidly, approaching video refresh rate.

2.5.1. Design. The schematic of this prototype is shown in Figure 2.12. Beside
the array elements (slots), many of this design’s attributes were similar to traditional MST arrays [34]. The RF receiver was a homodyne direct IQ down-conversion receiver followed by a lock-in amplifier, as explained earlier in Figure 2.3. The signal collection was based on a single antenna spatial collection scheme. A detailed description of the design of this prototype will be explained later.

This system is designed to operate mainly in the transmission mode as illustrated in Figure 2.12 where the retina acts only as a receiver, mapping the electric field incident on its aperture. It is also possible to operate this system in the mono-static reflection mode where the retina transmits the incident field and receives the transmitted/scattered field. In both modes mapping of the electric field is performed sequentially. When each slot in the retina is tagged, it is modulated by opening and closing the slot at a rate of 455 KHz. The detailed design attributes of this prototype may be found in [37],[47], the following subsection will provide a brief description of the components in this prototype.

2.5.1.1 Retina. The retina consisted of a 30-element array (6 rows by 5 columns), as shown in Figure 2.13. This array was manufactured using standard photolithographic techniques on a Rogers4350 [45] substrate. The largest dimension of the elliptical resonant slot was comparable to $\lambda/3$ which allows the array to be implemented with slot center-to-center interspacing of $\lambda/2$ (6.25 mm at 24 GHz). In this design, the DC bias lines, used for modulating the PIN diodes, were RF-coupled to ground at the edge of the

Figure 2.13. Pictures of the front and back of the retina.
slot using 5 pF SMT capacitors, as shown in Figure 2.13. The capacitors ensure that any RF signal coupled to the bias lines is shorted and does not propagate through the line creating unwanted resonances or signal losses. The vias encircling the slot prevent the electromagnetic waves from propagating into the dielectric substrate. These vias help reduce the leakage through the slots and coupling between the slots. The effect of these vias will be discussed in section 4.

2.5.1.2 Collection scheme. As most traditional MST arrays, a spatial collection scheme was used in this design to pick up/collect the modulated scattered signal [34]. As shown in Figure 2.12, the collector antenna was chosen to be a waveguide mounted resonant slot antenna similar to the ones used in the retina array. This collector was place behind the array at a distance of 25 mm. Given that the half-power beamwidth of this particular resonant slot is approximately $60^\circ$ [38], this distance ensures that all elements of the array find adequate coupling path to the collector. The advantage of this scheme is its simplicity and low cost. One the other hand a major disadvantage of this scheme is the high loss associated with the signal transfer from the retina to the collector. In this design, the loss encountered was as high as 20 dB. This high loss makes this scheme not easily expandable to large arrays. However, this design can be used to in a small scale proof-of-concept.

2.5.1.3 Receiver. The RF receiver for this system was based on homodyne design, as shown in Figure 2.12. The homodyne receivers are desirable for their simplicity and low cost. In order to perform coherent detection of the electric field, this homodyne receiver utilizes a direct conversion IQ receiver to convert the high frequency (24 GHz) signal to a baseband signal (near DC). The 24 GHz source was a waveguide Gunn oscillator with a built-in heater for stability. The source signal is divided using a directional coupler to provide a reference signal for down-mixing. This reference signal is routed to the LO port of the IQ receiver. The remaining portion is sent to the transmitter, as the incident electric field used for illuminating the object being imaged. The received signal from the retina is fed to the RF port of the IQ receiver through a low noise amplifier (LNA) with a 20 dB gain. Having the LNA at the RF front-end ensures a low overall noise floor. The LNA has a 2 dB noise figure which dominates the overall noise figure of the system. Overall, the homodyne receiver had a front-end linear
dynamic range of approximately 60 dB, as shown in Figure 2.14. The IQ receiver used here was the state of the art in commercial technologies. However, it had a quadratic unbalance of 8° which furthermore affected the linearity in its response. This unbalance understandably has a larger effect on the phase accuracy.

![Figure 2.14. Dynamic range of the homodyne receiver.](image)

2.5.2. Results

2.5.2.1 Transmission-through mode. Several experiments were performed to verify the capabilities of this design in mapping electric field distribution. The setup for one of the experiments is shown in Figure 2.15. A small metallic sphere with a diameter of 4 mm was placed at a distance of 12 mm away from the retina. An open-ended K-band rectangular waveguide was used to illuminate the target and retina combination from a distance of 80 mm (far-field of the transmitter). The electric-field distribution without the presence of the target (i.e. metallic sphere) was measured and used as a reference. This setup was also simulated using CST-MWS [44].

Figure 2.16 shows the measured and simulated results for this experiment. Due to the limited number of measurement points on the retina, the measured results were up-
sampled by a factor of 6.25 and spatially interpolated resulting in a smoother image with an effective one pixel per millimeter resolution. The images in Figure 2.16 show good agreement between the measured and simulated data both in the shape of the electric field distribution and its associated magnitude and phase values. The distortion shown in the field map is due to the limited resolution of retina. Moreover, a simple linear interpolator was used to resample the images, which may not be optimum.

Figure 2.15. Initial prototype transmission through experiment.

2.5.2.2 Reflection mode. In another experiment the system was used in a mono­static reflection mode, where the retina both transmits and receives the signal. In this setup the source signal is fed to the collector of the retina through a directional coupler. The slots in the retina are sequentially tagged to illuminate the target. The reflected signal is then collected and routed to the receiver using the directional coupler. Figure 2.17(a)-(b) shows magnitude and phase of the back-scattered wave when imaging a 10 mm diameter metallic sphere from a distance of 10 mm using the imaging system (i.e., the camera) in its reflection mode. As a comparison, a single slot was used to raster scan the same 10 mm diameter metallic ball using HP8510C vector network analyzer (VNA)
as a transmitter/receiver, the results of which are shown in Figure 2.17(c)-(d). The images in Figure 2.17 show good agreement between the scanned and camera images both in magnitude and phase. Differences between the camera and scanned results may be seen in the background value of the images. The background in the camera images approximately has a magnitude of 0.9 with an angle of 54°, while the scanned image show a value of 0.85 with an angle of 25°. This difference may be attributed to referencing error in the camera. All of these images were obtained on a grid of $\lambda/2$ spacing and interpolated 6.25 times as explained before. It must be noted that this camera produced images at a video frame rate of 30 images per second, far surpassing any mechanically scanned systems.
Figure 2.17. Comparison between imaging using microwave camera and raster scanning for a 10 mm diameter metallic sphere at 10 mm distance; (a) camera magnitude, (b) camera phase, (c) scanned magnitude, and (d) scanned phase.

2.5.3. Limitations and Improvement Considerations. This small scale proof-of-concept camera prototype showed that utilizing the efficient elliptical resonant slot it is possible to measure the coherent electric field distribution. However, this design is not easily expandable to a large scale due to many limitations especially at high microwave frequencies. Improvements must be considered in many aspects of the design to produce a viable imaging system. The sub-systems that must be improved are: the RF transmitter/receiver section, the collection scheme, and the image acquisition rate.
The RF transmitter/receiver section is what dictates the overall dynamic range, phase accuracy, and sensitivity of the system. The homodyne receiver is a simple design which proves suitable for many applications. However, as a coherent receiver for high frequency microwave applications, exceeding the well-developed 2.5 GHz region, it suffers from major deficiencies in dynamic range, and phase accuracy. The most problematic issue associated with the homodyne receiver is the quadratic unbalance, which means the phase difference between the in-phase and quadratic outputs of the receiver are not exactly orthogonal. There are techniques to correct for this unbalance [24], yet even these techniques prove not feasible at high frequencies. This problem becomes much more prominent when the system is targeted to have some bandwidth. The best solution is to use a heterodyne receiver, which utilizes an intermediate frequency (IF) stage for processing and detection of the signal. The heterodyne receiver utilizes a second RF source, which is phase-locked to the transmitted signal, to translate the signal to the IF stage. At the IF stage, which is typically a low frequency stage (i.e., MHz range), the signal may be filtered and amplified before utilizing accurate analog or digital IQ receivers. Another advantage of the heterodyne receiver is the fact that the RF sources are phase locked to each other or better yet, phase-locked to a precise crystal oscillator which greatly enhances the stability and phase noise of the overall system.

The signal collection scheme is another critical aspect of this imaging system. The spatial collection scheme used in the initial prototype proved to be highly lossy and inefficient. The loss for this small array was on the order of 20 dB. Making the array larger would require the collector to be moved further away from the retina. The increased distance would result in an exponentially larger loss. Furthermore, the array elements in the middle of the retina will have a stronger link to the collector than the array on the sides of the retina. This spatial dispersion would reduce the useful dynamic range of the system. Correcting for this substantial spatial dispersion might also amplify the effect of noise in the system. Overall the dynamic range and sensitivity of the system would greatly depend on the extent of this dispersion. In a previous endeavor [47] microstrip line and waveguide collection schemes were also investigated. Both schemes proved to be more efficient than the single collector spatial scheme. The microstrip line scheme is wideband, low profile, and easily integrated with other electronics such as
amplifiers. Dielectric and to some extent radiation losses are the only drawbacks of using microstrip lines at high microwave frequencies. On the other hand, nearly loss free transmission is obtained using metallic waveguides. Furthermore, slot antennas may be easily integrated into the waveguide walls. The drawbacks of using metallic waveguide are the bulky size and the relative cost.

The image acquisition rate is another factor to be optimized. The initial prototype produced an image frame each 30 milliseconds. This rate translates to 1 ms spent on each slot (dwell time). This time is required to establish a good signal measure through averaging or passing through narrow filters. When using an array with hundreds of elements this dwell time would reduce to tens of microseconds. In other words, to obtain both a fast image acquisition rate and low noise performance many aspects of the overall design must be optimized. The RF sources should be stable and maintain a low level of phase noise. The IF filters and receiver response time must be less than the dwell time. The modulation rate should increase such that the number of modulation cycles in this dwell time is more than 10 (rule of thumb). The modulation rate is also limited by the switching time of the PIN diode and the bias network. Commercial PIN diodes currently have a switching time on the order of few nanoseconds, which result in modulation rate limit in the high hundreds of MHz. On the other hand, the bandwidth of the bias network and modulation circuitry must be considered when designing the camera. Last but no least, the analog to digital conversion must be performed at the required acquisition speed.

The following section will further discuss design aspects of a large aperture real-time microwave imaging camera. Proposed solutions to the aforementioned limitations are presented through new designs or variations to existing designs. Each element of the design is presented separately with test results. Finally the system is integrated and full system tests are presented.
3. IMAGING SYSTEM DESIGN

3.1. IMPROVED IMAGING SYSTEM DESIGN

The proof-of-concept design outlined in the previous section was founded on efficient spatial multiplexing of scattered electric field over the 2D space of a retina. The analysis showed the possibility of effective and rapid "electrical scanning" of the retina space. Employing modulated PIN diode-loaded resonant slots became the basis for an efficient spatial electric field probing antenna. However, the design associated with that system also exposed several areas that require significant improvement or redesign. Consequently, a new approach to this overall design was considered in such way to not only address those inefficiencies, but to also address other issues related to the extension of the collector design to a more viable imaging system (i.e., extension beyond proof-of-concept). This section outlines the design steps and required electromagnetic analyses for a portable, 2D (~6" by 6") electric field mapping system, capable of real-time imaging with video rate data acquisition and display speed, while possessing a reasonably large system dynamic range (> 70 dB). This last item directly addresses the necessary considerations that must be incorporated into designing the transceiver portion of the imaging system, while keeping the size small and significantly improving the speed of coherent electric field registration from a large number of sources (i.e., modulated slots).

As mentioned previously, it was determined that rectangular waveguides are well-suited for collecting and redirecting the electric field scattered by the modulated resonant slots. A rectangular waveguide is a low-loss transmission line, and can therefore improve the efficiency of the overall imaging system by reducing attenuation, and radiation loss of the desired electric field picked up by a resonant slot. In addition, the use of rectangular waveguides provides significant control over redirecting, redistributing, and adding signals from a large number of slots, as will be seen later. Figure 2.1 shows the overall system schematic for the collector/retina, signal combiner, modulation and the transceiver sections. The retina is machined in an aluminum frame, housing the entirety of the signal collection network. The waveguide are terminated into two combiner waveguides (one at each end), which in part terminate into four signal collection ports at each corner of the camera. The modulated resonant slot array is placed
on the front-side of the frame and secured using a rim. On the back-side, the various RF, IF and control circuit boards are mounted.

The design of this imaging system has a compact form factor. Integrating the high frequency transceiver components into the frame of the retina – beside its appealing compactness, eliminates errors caused by long and flexible coaxial transmission lines otherwise required. In basic operation the slots will be closed and sequentially tagged (modulated) to spatially multiplex the incident electric field. The signal picked up at each slot location will split in two halves traveling to each end of the collecting waveguide behind it. The collector waveguides are terminated at each side into combiners which transfer the signal into output ports that are connected to the receivers. Several aspects of this design were optimized through extensive simulations, experimentation, or a combination of simulation and experimentation. The design features and properties of each component of this imaging system are described in details in the following subsections.

Figure 3.1. Overall imaging system schematic.
3.2. SIGNAL COLLECTION SCHEME

As mentioned earlier, waveguides are preferable to transmission lines at microwave frequencies. Compared to coaxial or printed (e.g. microstrip, CPW) transmission lines, waveguides are virtually lossless. Furthermore, slot antennas may be easily integrated on waveguide walls with the possibility of optimizing the location for a given application. A schematic of rectangular waveguide is shown in Figure 3.2. The waveguide is a hollow metallic tube which has a rectangular aperture with dimensions $a$ and $b$ \((a > b)\) and the dominant mode electric-field is polarized parallel to $b$ [48].

![Figure 3.2. Schematic of a rectangular waveguide.](image)

When designing a waveguide based slot array, slots are placed along the side-walls of the waveguide by matching the polarization of the slot antenna to the direction of surface current flow on the waveguide walls. As shown in Figure 3.3, on the broad side-wall currents are along the $z$-direction in the middle of the wall and they curve towards the $x$-direction as they near the edge. On the narrow side-wall the currents are primarily in the $y$-direction [48]. Therefore, a linearly polarized slot antenna (like the one used here) may be placed on any side of the waveguide by matching its polarization to the direction of current flow.

The retina size of $-6'' \times 6''$ corresponds to an array of 24 rows by 24 columns with an interspacing of 6.25 mm corresponding to $\lambda/2$ at 24 GHz ($\lambda$ is the free-space wavelength). It was decided that each row of the array will be placed on a separate waveguide, as illustrated in Figure 2.1, since the loss in the waveguide is proportional to
the number of slots on its wall. Each PIN diode-loaded slot when closed contributes a small amount of unwanted leakage/radiation loss (0.28 dB per slot) which can add up to a high loss in the waveguide. Another issue to be considered when placing the slots on the waveguide walls has to do with space management given the array elements interspacing requirements ($\lambda/2$) and the variety of options for slot placement on waveguide walls. The broad dimension of the waveguide must be larger than $\lambda/2$ or else the waves will not propagate inside it [48]. Therefore, it is physically impossible to place two waveguides side-by-side with their broad side-walls in one plane and a center-to-center spacing of $\lambda/2$. For these reasons, it was decided that the slots will be placed on the narrow side wall of the waveguide. A standard K-band waveguide has a narrow dimension of 4.3 mm which leaves an adequate 1.95 mm of wall thickness between each two waveguides when placed at 6.25 mm distance from each other. This size is also slightly larger than the height of the slot.

Figure 3.4 shows the aluminum base of the retina including the 24 signal collection waveguides. The waveguides are machined in a single aluminum block. The retina slot array makes the fourth wall of the waveguides. Figure 3.4 also shows a picture of the assembled retina showing the slot array PCB mounted using an aluminum rim on top of the waveguide network. The PCB was connected to the waveguide array using
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conductive epoxy thus ensuring no signal leakage or coupling between adjacent waveguides. The design of the slot array PCB will be discussed in the upcoming subsections. The rim on top of the PCB serves two purposes. First, it provides for a secure mounting of the slot array PCB onto the base. The second purpose of the rim is to complete the flange on the side of the retina for terminating these waveguides into signal combiners. The design of the combiners will be discussed at later stages of this section. Next subsection will discuss the design of the slot array.

![Figure 3.4. Aluminum base of the camera (left), and the assembled retina (right): 24 parallel waveguides serve as a collection network.](image)

**3.3. SLOT ARRAY DESIGN**

**3.3.1. Resonant Slot.** As described in detail in the section 2, the resonant PIN diode-loaded elliptical slot of the initial prototype was found to be optimum for this application. This slot provides for very small signal leakage when "closed", and high radiation efficiency when "opened". These attributes make the slot a good candidate for a waveguide-fed, sequentially tagged imaging array. Furthermore, the slot can be easily integrated into a rectangular waveguide wall while maintaining efficient signal coupling. Independent of the location of the slot on a rectangular waveguide, an important issue that must be considered is the structure of the bias lines that feed the modulating PIN diode with which the slot is loaded. The initial design included a thin bias line reaching the inner circular portion of the slot, where the PIN diode anode is connected (See Figure
2.14, in section 2), on the same plane as the slot. This thin line broke the ground plane of the elliptical slot which required two small (0201) surface-mount capacitors to be mounted at that location between the bias line and the ground on both sides. A large array leads to an even larger number of capacitors resulting in additional complexity, cost, and potential for malfunction. The design used here, utilizes the second plane of the PCB to bring the bias line to the circular load through a via as shown in Figure 3.5. The via acts as a high impedance inductive load blocking the RF signal from propagating along the bias line. Furthermore, the bias line does not break the ground in the plane of the slot resulting in minimal change in the resonant characteristics of the slot. The slot is designed by simulating its response when placed on the aperture of a waveguide as explained in the previous section. Table 3.1 shows the final dimensions of this slot structure, which were optimized through extensive simulation and later tuned through experimentations. This final tuning was required due to the slight changes in the dimensions, dielectric substrate material property discrepancy from nominal values, and inaccuracies in placing the via-ring around the slot perimeter. This via-ring is essential in confining the waves within the slot structure. Extensive simulations showed that without the via-ring, wave propagate inside the dielectric substrate which affect the properties of the slot especially when the slot is closed (i.e., leakage increases). The optimum solution would be a solid ring around the slot, albeit not practical with commercial PCB manufacturing technologies. Simulations showed that a dense via-ring with spacing of $\lambda/20$ or smaller is sufficient to confine the wave in the slot and reduce leakage into the dielectric substrate.

![Figure 3.5. Resonant slot with the bias structure.](image)
One limiting factor in the design of the bias structure was the available space which was a square area of 6.25 mm-long ($\lambda$/2) sides dictated by the array interspacing requirements. When the slot is placed in the middle of that area, the end of the bias line reaches the border of that area. Therefore, the bias line does not extend on the top layer. Figure 3.6 shows the simulated peak current distribution on the slot structure at the resonant frequency. It is clear that the current is concentrated in the top layer of the slot structure. The bottom layer shows current that propagate on the bias line. This is understandable since the bias line is inside the feeding waveguide. However, very small amount of this current travels through the via to the top layer. Various simulations showed that this small current can be potentially troublesome if the bias structure on the top layer had any resonance response near the slot resonant frequency. To extend this bias line on the top layer a resistor, inductor, or a distributed filter structure must be placed inline with the bias line.

Figure 3.7 shows the simulated and measured reflection coefficient looking into a waveguide with an elliptical resonant slot mounted on its aperture during the ON and OFF states of the PIN diode. In the simulated cases, the presence or lack of bias line is considered. The results show that when the slot is open and radiating (i.e., the PIN diode

Table 3.1. Slot dimensions.

<p>| | |</p>
<table>
<thead>
<tr>
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<tbody>
<tr>
<td>Slot width (2x ellipse major radius)</td>
<td>4.2 mm</td>
</tr>
<tr>
<td>Slot height (2x ellipse minor radius)</td>
<td>3.15 mm</td>
</tr>
<tr>
<td>Load diameter</td>
<td>2.1 mm</td>
</tr>
<tr>
<td>Diode gap (gap between load and outer conductor)</td>
<td>0.2 mm</td>
</tr>
<tr>
<td>Via diameter</td>
<td>0.25 mm</td>
</tr>
<tr>
<td>Via cap diameter</td>
<td>0.5 mm</td>
</tr>
<tr>
<td>Bias line width</td>
<td>0.125 mm</td>
</tr>
<tr>
<td>Internal (back-plane) bias line length</td>
<td>3.125 mm</td>
</tr>
<tr>
<td>Clearance (in-plane gaps between various conductors)</td>
<td>0.125 mm</td>
</tr>
</tbody>
</table>
is OFF), the presence of the bias line has negligible effect on the resonant frequency of the slot. The reflection coefficient (return loss) shows a change of 5 dB at the resonant frequency which is also negligible (the scale on the graph is a non-linear logarithmic, and the signal level along with the change must be considered). The change from -25 dB to -20 dB in reflection coefficient corresponds to a very small change in the power transmitted through the slot. This observation is consistent with the intensity of the current distribution on the slot vs. the intensity of the current on the bias line structure. Moreover, when the slot is closed (i.e., the PIN diode is ON), a high (-0.2 dB) reflection coefficient was noticed as it can be seen by the dashed lines. The presence of the bias structure did not affect this reflection and therefore the leakage through the slot. The measured slot response shows a shift in the resonance frequency with an increase in
reflection coefficient when the slot is open. As explained earlier, the shift in the resonance frequency is due to slight variations in dimensions and material properties in the manufacturing processes. The increase in the reflection coefficient and the increased leakage (dashed black line) when the slot is open may also be attributed to the non-perfect contact between the slot PCB and the waveguide wall/flange. This issue was verified experimentally by increasing or reducing the pressure on the slot PCB.

![Figure 3.7. Measured and simulated response of the resonant elliptical slot.](image)

Mounting the slot on the aperture of the waveguide is a good method during the design phase for optimizing the slot in simulation or measurements. However, to make the retina array, the slots must be placed on a narrow side-wall of a rectangular waveguide, as mentioned earlier. Subsequent to establishing the slot dimensions as explained above, the signal coupling and radiation properties of the slot, when placed on the narrow side-wall of a waveguide was simulated using CST-MWS. The structure
simulated is shown in Figure 3.8 with the waveguide outlines in white. The waveguide has three solid metallic walls and the fourth wall was created by the two-layer PCB containing the slot. The bias structure was not taken into account in this simulation as it was decided that it does not adversely affect the properties of the slot. This structure may be considered as a three-port network, however since the radiation properties of the slot were desired, no port was placed on the slot, and only two ports on both sides of the waveguides were considered. Furthermore, due to the symmetry of the structure, only port 1 was considered active.

![Diagram of waveguide and slot](image)

**Figure 3.8. Simulated slot on the narrow side-wall of a waveguide.**

First, simulations were performed with a slot placed on a standard K-band rectangular waveguide \( a = 10.7 \text{ mm} \) and \( b = 4.3 \text{ mm} \), results of which are shown in Figure 3.9a. As seen in this figure, the slot presents minimum perturbation to the waveguide when it is closed (i.e., when the PIN diode is ON, solid lines) which are represented by the total signal transfer between ports 1 and 2 \( (S_{21} = 0 \text{ dB}) \) and very small reflection back to port 1 \( (S_{11} < -30 \text{ dB}) \). On the other hand, when the slot is opened, some power is lost through radiation which can be noticed by the slight dip in \( S_{21} \) at 24 GHz. Furthermore, the reflection at port 1 also increased. However, this radiated power
is not optimum and must be maximized. In the optimum case, $S_{11}$ and $S_{21}$ will both become $-6$ dB. In other words, half of the power reaching the slot from port 1 will be radiated, and the other half will be split equally, quarter going to port 2 ($S_{21} = -6$ dB) and the other quarter going back to port 1 ($S_{11} = -6$ dB). This phenomenon is better described by plotting the lost/radiated power $(1 - |S_{11}|^2 - |S_{21}|^2)$ as shown in Figure 3.9c. This figure shows that for a standard waveguide, the radiated power when the slot is open (i.e., when the diode is OFF) is less than $-6$ dB, and when the slot is closed (i.e., when the diode is ON) the radiated power (leakage) is less than $-20$ dB. Reducing the broad dimension of the waveguide effectively addresses the issue of low radiated power when using a standard waveguide. The reasoning behind this effect will be discussed later. The frequency of operation (24 GHz) is at the high end of K-band, therefore, the waveguide broad dimension may be reduced to 6.25 mm before reaching the cutoff frequency of the waveguide [48]. The waveguide broad dimension was reduced from the standard 10.7 mm to 7.7 mm which brings the cutoff frequency to $-19.5$ GHz, well below the frequency of operation. It is desirable to operate at a frequency which is approximately 25% greater than the cutoff frequency such that the propagation constant in the waveguide does not show strong dispersion properties [48]. The slot response on this modified waveguide is simulated and the results are shown in Figure 3.9b. For this modified waveguide, when the slot is closed (i.e., the PIN diode is ON), the waveguide is not disturbed (i.e., $S_{11} < -20$ dB and $S_{21} \approx 0$ dB). Furthermore, when the slot is open (i.e., the PIN diode is OFF) a more prominent dip (compared to the standard waveguide case) is seen in $S_{21}$ which corresponds to more radiated power. Looking at this structure from the radiated power point of view (Figure 3.9c), in the modified waveguide case the leakage when the slot is closed increases slightly, as it can be shown in Figure 3.9c. On the other hand, the radiated power when the slot is open reaches $-3.1$ dB which is twice as much (3 dB more) as in the standard waveguide case. In other words, approximately half of the power reaching the slot from port 1 is radiated. Again keep in mind that the results are shown in logarithmic scale where variations in the upper end correspond to more change than variations in the lower end. A $-3$ dB radiated power (50% total efficiency) is considered optimum in the sense that if the slot is used in the receiving
mode (as it will be used in the retina), half of the power reaching the slot will travel to each side of the waveguide and no power will be reflected back from the slot.

Figure 3.9. Simulated response of the resonant slot on waveguide narrow wall, (a) standard K-band waveguide, (b) modified waveguide, and (c) radiated power.

The increase in the waveguide-slot coupling when the broad dimension of the waveguide is reduced can be explained by looking at the current distribution on the
waveguide walls. Figure 3.10 shows a time snapshot of $\text{TE}_{10}$ current distribution on a standard K-band waveguide at 18 GHz, which is a frequency nearest to cutoff at the beginning of the band and at 24 GHz, the frequency of interest in this investigation which is at the end of the band. These current distributions were simulated using CST-MWS. The simulations at these frequencies and many other frequencies in between showed that,

![Figure 3.10. $\text{TE}_{10}$ current distribution on a standard K-band waveguide at frequency (a) 18 GHz, and (b) 24 GHz.](image)

the peak currents has more intensity at the narrow side-wall of the waveguide at the lower frequencies, and as the frequency increases, the peak intensity on the narrow side-wall decreases while the peak intensity on the broad side-wall increases. This phenomenon explains the week coupling to the slots encountered at 24 GHz when placing the slots on the narrow side-wall of a standard waveguide. By reducing the broad dimension of the waveguide, the frequency of interest becomes near the cutoff frequency which intensifies the currents on the narrow side-walls, as shown in Figure 3.11.
3.3.2. Retina Addressing Scheme. Electronically scanning the retina will require that each individual slot be tagged while the other slots are idle (i.e., sequentially scanned retina). This method of operation results in a time-multiplexed signals of the spatial electric field distribution over the retina space. Addressing each individual slot separately through a bias line (i.e., direct addressing) allows for such operation but it requires $24 \times 24 = 576$ bias lines and connectors to bring 576 bias lines to the retina and a control circuitry that can drive 576 simultaneous outputs. This solution is not feasible, in addition to the fact that there is not enough space on the retina to route this many bias lines. A solution to such problems is the matrix addressing scheme commonly used in electronic displays. Matrix addressing scheme is illustrated in Figure 3.12. In a matrix addressing scheme for an $M \times N$ array, only $M+N$ addresses (or bias lines) are needed. In this scheme to address a slot at the intersection of a row and a column, its corresponding rows and columns are addressed. To perform the electronic scan of the retina, first a row is enabled by properly biasing the line, and then the various elements in that row are scanned by sequentially addressing the columns which may also carry the modulating signal. Subsequently by going through all rows, the 2D electronic scan is performed.

A major disadvantage of using the waveguide-fed elliptical slot is that the slots have to be closed in normal operation (i.e. when idle and not tagged), and then sequentially opened (or modulated) to receive the signal. Closing the slots requires
forward biasing the PIN diode. Considering that each PIN diode nominally requires 5 mA to be turned ON or forward biased, the 576 PIN diodes of the retina will cumulatively require more than 2.8 Amperes of current. Furthermore, this current can go as high as 10 mA per diode when saturated as a result of a slight increase in the bias voltage. Moreover, when modulating a line, each row or column of the bias line should be able to switch currents higher than 120 mA (for 24 slots) at modulation frequencies of up to a couple of MHz. The design of the modulation and switching circuitry will be explained towards the end of this section.

![Matrix addressing scheme](image)

Figure 3.12. Matrix addressing scheme.

In a matrix addressing scheme such as those of displays, one of the address dimensions (e.g., columns) is connected to a voltage supply (positive) and the second dimension (rows) is connected to the ground. By pulling the column line high and row line low for a particular element in the array, that element is activated as illustrated with the loaded dipole in Figure 3.12. This is possible since both the cathode and the anode are floating. In the case of this particular elliptical slot antenna, the cathode of the slot PIN diode is always connected to ground. Furthermore, the slot must be idly closed (i.e., its PIN diode is ON). These two requirements of addressing the slots individually and
keeping the idle ones closed, was achieved by combining the column and row address lines at each junction using diodes in the bias lines before feeding them to the anode of the slot PIN diode as shown in Figure 3.12. These two bias diodes may be PIN diodes, schottky rectifier diodes, LEDs or any other device that provide a one-way path for the current. These two bias diodes perform an OR operation. In other words, the slot will become open (PIN diode turned off) if and only if both corresponding row and column address lines are pulled low (or turned off). In summary, to perform a scan using this design, all slots are closed (PIN diode ON) by pulling all row and column bias lines high, and then the slots are sequentially opened (PIN diode turned OFF) by pulling their corresponding row and column line low.

3.3.3. Retina Fabrication. The retina is shown in Figure 3.13. The retina was fabricated on a two-layer 0.020" thick Rogers4350 laminate [45]. Since the back layer is inaccessible in this design (because of the waveguide collection network), only the top layer is used to place components and addressing network. The 576 resonant elliptical slots, designed as described earlier, were placed in a 24 × 24 grid with center-to-center spacing of 6.25 mm (\(\lambda/2\)). The grid was laid out to match the positions of the waveguide collector array. LEDs were used in the bias structure to combine the row and column address lines for each slot as described in the previous subsection. The LED’s were low profile (0402 size) surface mount red and green LEDs for the column and row address connections, respectively. The LEDs serve multiple purposes. First, they perform the OR operation required for the proper switching, as explained in the previous subsection. Second, they act as voltage dividers, which ensures that the PIN diode is biased at the proper 1.3 to 1.5 V when the circuit is operated using 3.3V CMOS logic (1.8 to 2 V is dropped across the LED). Third, they serve as indicators when the PIN diode is forward biased. As indicators, the LEDs play an important role in troubleshooting the system when a PIN diode may malfunction. The LEDs turn ON when the slot PIN diode receives current through them. Any LED that does not turn on, indicates that the corresponding PIN diode is not working properly, or a modulation line is open-circuited, or the modulation circuitry is not providing enough current, or the LED itself is not functioning properly.
The row address lines were etched as 0.005" wide horizontal lines onto the top layer at midpoint between the rows. On the other hand, thin (0.005"-diameter) insulated wires were used for the column address lines and were soldered at each slot location to its corresponding red LED. The LEDs and the bias lines are expected to have a slight effect on the radiation properties of the slots. However, the low cost and ease of fabrication overruled any slight effect these components might produce. This retina was placed on top of the waveguide collector array and secured using the top rim. Conductive epoxy ensured direct electric connection between the waveguide walls and retina PCB.

Figure 3.13. Retina PCB showing the bias lines, the bias LEDs, and the slots,

3.4. COMBINER WAVEGUIDE

Several methods were investigated to combine the 24 rows of the signal collecting waveguide network into access ports. The first option considered was using waveguide bends to connect all rows together making one long waveguide, as illustrated in Figure 3.14a. This solution was investigated in [47] and it proved attractive for a small retina for
its simplicity and ease of manufacturing. However, given that each waveguide with 24 slots is anticipated to have as much as 7 dB of signal loss due to leakage from the slots (0.28 dB per slot), the signal received from the slots in the middle rows would experience more than 75 dB of attenuation before it reaches the nearest port, rendering this solution impractical.

![Figure 3.14](image.jpg)

Figure 3.14. Schematic of investigated combiner solutions: (a) one waveguide solution, and (b) external combiner solution.

The second solution investigated (Figure 3.14b), involved access ports on both sides of each row. This option has many advantages. The output of each collector waveguide may be amplified before entering the combiner. The combiner may include a switching functionality to route the signal to a single line or selected group of lines. A switching functionality enables the isolation of certain row of the array which can be used to transmit a signal in the case of reflection-mode operation. The most attractive advantage of this combining option is the ability to perform parallel modulation on one column of the retina, which can save significant amount of time in image acquisition.
when using large arrays. Since slots on each column do not share waveguides, by modulating all of them at the same time, they would not significantly interfere with each other. Therefore, one can modulate sets of slots (one column at a time) with well-established techniques such as orthogonal coding. With parallel modulation it is possible to reduce acquisition time by a factor of \( \log_2(N) \) where \( N \) is the number of slots being orthogonally modulated at the same time.

Since the waveguide collection network of Figure 3.4 share the broad walls of the waveguides, a simple coaxial side-feed was not feasible and an end-feed scheme was required. The option investigated was to utilize a finline structure to transfer the signal from a printed coplanar waveguide (CPW) to a rectangular waveguide [49]. As shown in Figure 3.15, the finline investigated utilizes cosine taper of length \( \lambda/2 \) due to its lowest return loss [50]. A Balun provides matching between the CPW and the finline transition [49]. This feed structure was optimized through extensive simulations in CST-MWS.

![Figure 3.15. Finline waveguide feed.](image)

Subsequent to optimizing the finline design, a test fixture as shown in Figure 3.16, was built and tested. This test fixture contained a 2" section of waveguide terminated at both sides with a CPW-Finline transition. The waveguide indicated by the dashed lines was machined into two aluminum blocks. The waveguide was mainly in the bottom block while the top block serving as a cover only, except for the finline location since the finline must be placed in the middle of the broad wall of the waveguide. The finline was
made on a PCB with a Rogers4350 substrate. The CPW side of the finline transition was terminated into an SMA connector which was rated to 26.5 GHz.

![Figure 3.16. CPW-finline-waveguide transition test fixture.](image)

Figure 3.16 shows measured and simulated response of the test fixture in Figure 3.16. While the simulated results show a desirable return-loss of 20 dB and a small insertion loss of 0.5 dB, the measured results are far from desirable. For this structure the measurements show a loss of up to 2 dB and increased reflection at the ports. Further investigation and measurement of the loss on a 1"-CPW line of similar properties and terminated in similar SMA connectors showed that the majority of the loss is encountered in the PCB portion of the finline transition. Furthermore, the imperfect contacts and gaps on the joints between different waveguide blocks and the finline PCB are believed to contribute to the higher reflections and loss, as well.

Overall, efficient coupling to a waveguide using an end-feed was not trivial and proved to be costly. Furthermore, the SMA connectors on a thin PCB are not mechanically robust. Other than the difficulties in building an efficient end-feed, the scheme of Figure 3.14b has few other drawbacks. For example a 24-to-1 low-cost combiner such as those utilizing Wilkinson design, will introduce a minimum signal loss of 12 dB not including the dielectric and radiation losses at these high frequencies. Due to the above reasons, this design — although desirable from the functionality point of view, was abandoned.
The combining method, which was finally used, is shown in Figure 3.18. This method utilizes a switchable iris at each end of each collector waveguide. The iris of choice was the resonant PIN diode-loaded elliptical slot placed in the wall between the waveguides. Using this method, the signal received from each slot is routed to four output ports – two ports on each combiner on each side of the camera, by properly switching the desired iris. Using this method, the maximum attenuation experienced by a signal before reaching the nearest port is less than 12 dB. This attenuation is mainly due to the leakage in the slots and the combiner irises.
To ensure maximum signal transfer from the retina to the output ports, only one of the combiner irises should be open at each time. In ideal situations when the irises have zero leakage, the connection between each collector waveguide and a combiner waveguide is considered as an E-plane Tee with a switchable iris. Figure 3.19 shows the Tee configuration simulated in CST-MWS. The switchable iris is the PIN diode-controlled resonant elliptical slot designed to work at 24 GHz, as described earlier. In this configuration, port 1 represents the collector waveguide side and ports 2 and 3 represent the combiner waveguide side. Ideally, it is required that the signal reaching the junction from port 1 be divided equally into ports 2 and 3 when the switch is open. On the other hand, when the switch is closed any signal in the collector waveguide (port 1) should not enter the combiner waveguide (i.e., there should be high isolation between port 1 and ports 2 and 3). Yet any signal in the collector waveguide should travel from one side to the other without experiencing any attenuation or distortion.

Figure 3.20a shows the simulation results for the case when the switch is open. It is established that a reciprocal and lossless three-port device cannot be simultaneously matched at all of its ports [48]. Furthermore, only the network properties (S-parameters)
at the resonant frequency of 24 GHz are of interest. The values of these S-parameters at 24 GHz are shown in the legend/marker. Figure 3.20 shows the response vs. frequency to confirm the resonant frequency of this switch. At the resonant frequency the power division in the Tee ($S_{21}$ and $S_{31}$) is a corresponding -3.9 dB which is close to the ideal value (-3 dB). On the other hand, the reflections ($S_{11}$, $S_{22}$, and $S_{33}$) are around -10 dB. To reiterate, it is impossible to create a matched power divider. Overall, these reflections are acceptable, given the difficulties associated with alternative designs, as mentioned earlier. On the other hand, when the switch is closed (Figure 3.20b), the isolation ($S_{21}$ and $S_{31}$) is at the level of 20 dB, while ports 2 and 3 show small reflections (i.e. matched). Moreover, the signal traveling through the collector waveguide experiences a very small amount of loss ($S_{23} < -0.1$ dB). This test setup of a single iris in a waveguide Tee (shown in Figure 3.19) was not manufactured and therefore the results were not verified experimentally.

For the final application of designing a combiner for the imaging system, 24 irises were manufactured on a single 0.020” thick 2-layered PCB with Rogers4350 laminate. This board was sandwiched between the combiner waveguide and the array of collector waveguides, as shown in Figure 3.21.
Figure 3.20. Simulation results of the switched waveguide tee: (a) switch is open, and (b) switch is closed.
Figure 3.21. Combiner waveguide assembly.

Figure 3.22 gives an indication on the switching capabilities of the combiners and the loss signals face while traveling through the waveguide collector network. The measurements were performed using the HP8510C VNA. The transmitted coefficient ($S_{12}$) between each two ports of the camera for opening the irises of the combiners, one at a time is shown. For this experiment, all the slots on the retina were closed. The transmission coefficient between ports 1 and 3 shows a semi-linear loss as the signal travels a larger distance through the combiner waveguides. The difference between the first row of the collector waveguides and the last row is approximately 25 dB. Similar response is noticed between the other two ports of the retina (i.e., ports 2 and 4). If it is assumed that the loss is equal in each combiner, the result is a loss of 12.5 dB per combiner waveguide which corresponds to approximately 0.5 dB loss per iris much
larger than the simulated value (0.1 dB). The 0.5 dB loss per slot was also measured using a single slot on a waveguide aperture, and is mainly attributed to the imperfect contact between the waveguide walls and the slot PCB.

The transmission coefficient between ports 1 and 4 shows an average of -25 dB. If it is assumed that the combiners switches -4 dB of the signal into the collector waveguide (-3.9 dB in simulation), subtracting the -4 dB once for each combiner results in -17 dB of the signal lost in the combiner and collector waveguides (-25 - 2(-4) = -17 dB). Consequently, if we assume that 12 dB of that loss is due to the combiners, the result is a 10 dB loss in the collector waveguide which is 3 dB more than the theoretical value. On the other hand, if it is assumed that the collector waveguide have a loss of 7 dB then the combiners must have switched -5.5 dB of the power only. In reality, this extra loss in the system may be distributed loss, not accounted for such as leakage through the gaps at points of mechanical interconnects such as flanges. Furthermore, reflections will also contribute to the loss seen in the transmission coefficient.

Figure 3.22. Measured signal loss through the waveguide collection network and the combiners.
Measuring the transmission coefficient through the other two opposite corners (ports 2 and 3) show an average of 5 dB more loss compared to the first case. This extra loss is attributed to the mismatch in port 2. This mismatch was caused by manufacturing errors which caused the monopole coax feed of port 2 to be slightly bent. This fact was verified by visual inspection. It may be also noticed in the transmission coefficient between ports 2 and 3 when the combiners were switched to collector waveguides at rows 23 and 24. This design as manufactured did not show ideal or even close to ideal performance in reflections and loss. However, it was desirable since it provided lowest overall signal loss compared to other alternative designs, as explained earlier. Furthermore, it has a compact form factor, since it adds less than 0.5" to the width of the camera on each side.

3.5. SIGNAL SOURCE AND RECEIVER

The transceiver design was based on a heterodyne scheme; where as the first harmonic of a single side-band of the modulated signal is demodulated. This scheme was first introduced by [51] in order to improve measurement accuracies by eliminating the quadrature error associated with high frequency components. In that design the mixing was performed down to baseband, such that the IF stage frequency is the fundamental frequency of the modulating signal. In the design presented here the mixing is performed in a way to make the IF stage frequency a fixed 10.7 MHz irrespective of the modulation frequency. This design has several advantages over the design introduced by [51]. First, the fixed IF frequency provides the flexibility of changing the modulation frequency without changing the IF stage hardware. Second, the higher IF frequency of 10.7 MHz used here compared to the low IF frequency which is the modulation frequency of less than 1 MHz used in [51] translates to lower flicker (1/f) noise. Third, the higher IF frequency translates to larger frequency difference between the RF sources reducing the design complexity associated with the sources. Finally, utilizing a standard IF frequency such as 10.7 MHz, allows for a greater selection pool of commercially available components. This scheme is illustrated in Figure 3.23. The main RF source generates a signal which is sent through the transmitter to illuminate the target. The received modulated signal (through the retina) is down-mixed to the IF stage where is it filtered
and amplified. Subsequently, the demodulator provides the estimate of the signal at the retina aperture. This imaging system requires four receiver channels. This section provides the design of the individual components of this system. A detailed description of the operation characteristics of this receiver and an analysis on its performance is provided in the next section.

Figure 3.23. Schematic of the sources and one channel of the receiver.

There is one fundamental difference between this design and the more traditional MST receivers [34]; namely, the lock-in amplifiers. In this scheme the IQ demodulator
serves as a quadrature lock-in amplifier which performs the match-filtering on the first harmonic of the square-wave modulated signal. The disadvantage of this scheme is that, capturing the first harmonic results in approximately 4.8 dB power loss, 3 dB of which is for ignoring one side-band and another 1.8 dB for ignoring the harmonics of the square-wave signal. To summarize, this design is selected for its high-accuracy and dynamic-range provided by a heterodyning scheme, a flexible frequency control provided by a fixed IF frequency, and finally an accurate IQ demodulator that also act as a matched filter. Overall, the signal power loss is compensated for by the high dynamic-range and sensitivity of the receiver and ability to filter and amplify the IF signal.

3.5.1. RF Sources. Figure 3.24 shows the schematic of the RF sources. A fractional synthesizer is used to control the frequency of each RF source. Using fractional synthesizers allows for the small and precise difference ($10.7 \text{ MHz} + f_{\text{mod}}$) in frequency between the two RF sources using a high phase detector frequency (PFD) in the PLL [52]. This high PFD frequency of 20 MHz used in the fractional synthesizer in this design – compared to a maximum of ~60.79 kHz that would have been used with an
integer synthesizer, results in little degradation in the phase noise of the VCO. Furthermore the fractional synthesizer allows for the use of conventional reference frequencies (i.e., 20 MHz). Since this system is designed to be operating at a single frequency, a narrow 10-KHz loop filter ensures a sufficient attenuation of the fractional spurious signals. Power amplifiers at the output of the VCOs ensure a high enough output power. In the LO VCO, the high power is needed to drive the mixers at the receiver. In the case of the RF VCO the high power ensures greater range. The two sources are also mixed together to create an IF test port. Finally, the RF sources are phase-locked to a crystal oscillator at 20 MHz.

Figure 3.25 shows a picture of the final assembled RF-source board. A 0.062" 4-layer FR4 board provides for sufficient mechanical stability. Industry standard stack-up is used which provides a thin dielectric (0.008") for the top layer. As a result of using this thin dielectric, 50-ohm coplanar line becomes as narrow as 0.014" which is better-suited for the surface mount MMIC RF components [53]. The Synthesizer selected for this design was ADF4154 fractional synthesizer from Analog Devices. This synthesizer accepts RF frequencies up to 4 GHz. The VCO used in this design was HMC533LP4 from Hittite Microwave Corp. This VCO provides the 24 GHz signal directly without the need for additional doublers. The VCO also provides a divide-by-16 output (24÷16 = 1.5 GHz) which can be directly fed to the synthesizer, thus eliminating the need for external frequency dividers. The power amplifier was HMC383LC4, which can provide up to 15 dBm of output power, and finally the mixer used was HMC292LC3B double-balanced mixer both from Hittite Microwave Corp.

Figure 3.26 shows the spectrum of the signal at the RF-out port. Due to the narrow PLL loop filter, the spectrum shows a single tone with no feed-through of the 20 MHz reference frequency signal. A slight shift of ~10 KHz is noticeable in the frequency of the source due to an even slighter shift in the frequency of the reference crystal. However, this shift is of no consequence as long as the frequency difference between the RF and LO sources falls inside the IF filter pass-band since both sources are phase-locked. Furthermore, the shift in their difference frequency is at approximately 4 orders of magnitude smaller than the shift in the actual frequencies for this design.
Figure 3.27 shows the RF and LO sources which are shifted from each other by 10.7 MHz. It also shows that when both the RF and the LO sources are ON, due to radiation cross-talk and the nonlinearity of the power amplifiers, the signal spectrum has spurs at their difference frequency. The radiation cross-talk is greatly reduced by shielding at least one of the sources. Eventually, due to the fact that the received signal through the camera will be modulated, the majority of these spurs and any other unwanted signal are not expected to pass through the IF filters. Finally, Figure 3.28 shows the spectrum of the difference (IF signal) which is set to 10.7 MHz. Using this figure, an estimate of the phase noise in both sources combined is obtained to be approximately -60 dBc at 10 KHz and -90 dBc at 100 KHz offset. These figures are close to the specification of the VCO which is -70 dBc at 10 KHz -95 dBc at 100 KHz.
Figure 3.26. Signal at RF out port, RBW=100 KHz, VBW=100 KHz.

Figure 3.27. Signals spectrum at RF out port and LO out ports, RBW=10 KHz, VBW=10 KHz.
3.5.2. Receiver Front-End. The RF receiver has four identical down conversion channels. Figure 3.29 shows the schematic of one of these channels. Each channel includes a down conversion mixer, a power amplifier to drive the mixer, and an RF low noise amplifier (LNA). The power amplifier and the mixers were similar to the ones used in the sources board. The LNA (HMC517LC4 from Hittite) has a noise figure (NF) of 2.5 dB and a gain of 20 dB. The LNA is most critical part of the receiver which dictates the overall NF of the system. The effect of the various components' NF on the overall noise floor of the system will be analyzed in section 4. The IF signal produced by the mixer goes through an IF amplifier with a 20 dB gain, a ceramic band-pass filter centered at 10.7 MHz and with a bandwidth of 100KHz, and then another IF amplifier with a gain of 20 dB.
As shown in Figure 3.30, each two channels are placed on a single board. Various test points are provided at the IF stage. First test point is at the output of the mixer, the second is at the output of the filter, and the third test point is at the receiver output. The noise power reaching the demodulator is directly proportional to the filter bandwidth. It is desirable to have a narrow filter bandwidth. On the other hand, the group delay in the filter is inversely proportional to the filter bandwidth. Therefore, speed associated with electronically scanning the retina puts a lower limit on the filter bandwidth. Furthermore, slight shift in the frequency must be anticipated and allowed for in the selection of filter bandwidth. Since this design is a prototype, the filter bandwidth was selected to be larger than the optimum value. Moreover, the system will have other filters at the output of the demodulator and as averaging in the acquisition stage. The response of all these filters must be cascaded to determine the overall system bandwidth. Section 4 will show the cascade of the filter and provide an analysis on the selection of an optimum filter bandwidth.
In order to test the dynamic range of these channels, a measurement setup as shown in Figure 3.31 was used. Two physically separate sources for the RF and LO were used to eliminate any possible cross-talk between them. A very stable RF source from an 8510C VNA was used as the source. This source output power was then attenuated using a precision variable attenuator (HP8495D) along with few fixed attenuators to extend the measurement range. A spectrum analyzer was then used to measure the output power. The result of this test is shown in Figure 3.32, which indicates that these receivers had a dynamic range of more than 100 dB. The linear dynamic range of this front-end has have a direct impact on the dynamic range, and sensitivity of the overall system since it represents the first stage (beyond the sources). The final system is not expected to have a dynamic range of more than 70 dB, although a large dynamic range in individual stages of the system contributes to a larger overall system dynamic range, which is desirable.
3.5.3. LO Divider. The LO signal is divided into 4 equal parts to drive the mixers in the four receiver channels. The front-end mixers require an LO power of +13 dBm. Taking the gain of the power amplifier and the loss of the board into account, the signal power at the input of the power amplifier should be larger than 0 dBm. A 1 to 4
power divider (Figure 3.33), utilizing a modified design of Wilkinson dividers [54] suitable for high microwave frequencies, were used to build the LO divider. This particular Wilkinson design was found to be suitable for high frequencies since it provides room for placing the required and very small (0201) surface mount 100 Ω resistor without sacrificing isolation. This divider was designed to operate at 24 GHz and its functionality was simulated using CST-MWS. The simulation results showed an insertion loss of -3.5 dB (0.5 dB beyond the ideal response) and a return loss better than 15 dB. Figure 3.34 shows the measured return loss and insertion loss of this particular 1 to 4 divider design. The best performance was obtained around 23.5 GHz with acceptable performance at 24 GHz. At 24 GHz the insertion loss is better than 10 dB at the input and 15 dB or better at the output ports. The insertion loss of this divider is between 9 to 10 dB (maximum of 3 dB beyond the simulated response). This additional loss is attributed to the dielectric losses and radiation losses which were found to be substantial at the connector microstrip point of contact. This divider is used for the LO port where frequency, amplitude, and phase of the signal are fixed. In other words, slight reflections at the ports are acceptable and do not affect the overall performance of the system. Furthermore, the output power of the LO source is also sufficient so that these additional losses do not adversely affect the overall system characteristics.

Figure 3.33. LO divider board.
3.5.4. IQ Demodulator. The final stage of the receiver is the IQ demodulator. Since the camera has four measurement ports, and there are four receivers, logically four IQ demodulators are required. The IQ demodulator board was based on Analog Devices' AD8339 chip which is a quad IQ demodulator. Since this chip is configured to be driven by a low noise preamplifier with differential outputs, a quad LNA (AD8334) was used. Furthermore, mixer outputs are current-based, and thus current-to-voltage transimpedance amplifiers (AD8021) were used for each I or Q output before applying them to the analog to digital convertors. These transimpedance amplifiers were also fitted with RC low-pass filters with a cutoff frequency of ~30 KHz. Figure 3.35 shows a picture of the assembled 4-channel IQ demodulator.

One drawback of using this chip is the requirement of 4LO, i.e. the LO input frequency of the chip must be 4 times the frequency of the IF signal. Consequently, for the 10.7 MHz signal input to the IQ demodulator, the 4LO input will require a signal at 42.8 MHz. To generate the 4LO, a synthesizer is locked to the system reference clock. The 4LO generator was built using an integer synthesizer and a VCO. Figure 3.36 shows a picture of the 4LO generator board, and its output spectrum. The 4LO is phase-locked to the RF sources ensuring coherent detection. The output spectrum of the 4LO source shows a phase-noise of approximately −100 dBc at 10 KHz offset.

Figure 3.34. LO divider test results: (a) Return loss, and (b) insertion loss.
Figure 3.35. Picture of the 4-channel IQ demodulator.

Figure 3.36. 4LO generator and the spectrum of its output (RBW = VBW = 1 KHz).
Figure 3.37 shows the setup used to test the performance of the 4-channel IQ demodulator. The IQ demodulator was tested at 10.7 MHz, which is the system IF frequency. Two Stanford Research generators were phase-locked, the first was set at a frequency of 10.7 MHz and its phase was swept from 0 to 360 degrees using the GPIB control command. The second generator was set to the 4LO frequency of 42.8 MHz. Using a variable attenuator, the signal power level was changed from -10 dBm (just below the saturation level) to -120 dBm. Figure 3.38 shows the dynamic range of the IQ demodulator, which shows a fairly linear response for a range of 90 dB. Figure 3.39 shows the phase estimation error for this IQ demodulator at various power levels. The average error in the detected phase of the signal was less than 0.3 degrees for signal power as low as -60 dBm. Overall, the performance of this IQ demodulator chip is unsurpassed by any commercial analog IQ demodulator in market today.
Figure 3.38. Dynamic range of the 4-channel IQ demodulator.

Figure 3.39. Phase error in the 4-channel IQ demodulator.
3.6. RETINA CONTROL & MODULATION CIRCUITRY

An off the shelf solution was used to control, or electronically scan, and modulate the retina. The off the shelf unit is the *Quicgate* controller from Dallas Logic™ based on Altera™ Cyclone-II FPGA shown in Figure 3.40. An FPGA solution gives the flexibility of easily changing the operation mode of the camera. Moreover, it is compact and cost-effective. This controller provides for 110 digital input/output lines, sufficient to address the retina's row and columns and the two switched combiners, while leaving few spare lines. The controller electronically scans the retina, while providing the modulation signal from its crystal oscillator. Furthermore, the reference signal needed to lock the RF sources and the 4LO source are provided from this controller. The controller also provides a synchronization signal to the ADCs that provides the clocks the data acquisition.

![Figure 3.40. FPGA controller stack-up.](image)

The combiner PIN diodes were driven directly by the controller, since each PIN diode draws maximum of 8 mA which is can be provided by the FPGA. On the other hand, the row and column lines of the retina can draw as much as 200 mA each. Therefore, a switching circuit was built to provide this high current directly from a power
supply, controlled by the FPGA. The switches used were the quad analog switch ADG841 from Analog Devices™. This analog switch was used over simple digital switches for its bandwidth characteristics, current handling capabilities, and most importantly its low resistance. Switching circuit and connector containing boards were stacked-up around the FPGA controller, as shown in Figure 3.40.

3.7. SUMMARY

This section presented a significantly improved design of a microwave camera based on a combination of a switched array and MST multiplexing. Several aspects of this design resulted in substantially improving the overall system performance. Improving several critical design features resulted in an increase in the dynamic range and sensitivity of this system. The first design feature was the improved RF transceiver. Each component of the RF system provided a dynamic range exceeding 90 dB. Furthermore, the improved array element (resonant slot) and the signal collection scheme, provide for stronger signals at the receiver thus enhancing the overall SNR and eventually the dynamic range of the system. Many elements of this design is designed by extensive full-wave simulation and verified experimentally. Some design elements are based on established design, engineered to provide for optimum performance given the requirements of this design. The next section of this dissertation will concentrate on full system test results and it provide for an analysis on the properties and features of the microwave camera.
4. SYSTEM TEST & CHARACTERISTICS

In the previous section, detailed design of various components of the camera were presented. Each component of this system was individually optimized to meet or exceed the design specifications. Figure 4.1 shows pictures of the front and back of the assembled camera. This system is compact, lightweight, and possibly handheld. This section provides a thorough investigation and analyses of the electrical characteristics of this system, sources of errors and their impact on the overall system performance and characteristics, and finally signal correction and characterization methods. Investigations are performed and the results are analyzed on: system noise level, isolation among the array elements, system range (i.e., distance of operation), slot mutual coupling, and array response.

![Figure 4.1. Pictures of the front and back of the camera showing the array at the front and the various RF and control components at the back.](image)

4.1. RECEIVER PERFORMANCE

The receiver design presented in the previous section is based on a heterodyne scheme, in which the first harmonic of a single sideband of the modulated signal is demodulated and analyzed. Figure 4.2 shows the received spectrum, at port 4 (bottom left corner) of the retina, due the modulating slot located at row 12 and column 12,
respectively. The irradiating source was an open-ended rectangular waveguide radiating towards the retina from a distance of 150 mm. The modulation frequency was 500 KHz. The spectrum shows the upper and lower sidebands (USB and LSB) of the modulated signal, respectively along with the carrier signal. This modulated signal contains even harmonics representative of a distorted square wave modulation. The distortion may be due to the limited bandwidth of the switching network, effect of switching high currents, or as a result of the non-linearity of the LEDs. Overall, the system only captures one of the first harmonics and all the other even and odd harmonics are filtered out. The carrier is more than 10 dB above the modulated signal rather than the 6 dB expected for a single slot. This extra power in the carrier is attributed to the leakage from the remaining slots in the retina. This leakage from various array slots can add-up constructively or destructively. Experiments showed that the power in the leakage depends on the electric field distribution on the retina and not the electric field on the specific slot being measured. However, the modulated part of the signal depends only on the field at the location of the slot being modulated.

![Figure 4.2. Measured signal at port 4 of the retina from to the modulating slot at row and column 12, respectively (RBW=1 KHz).](image-url)
The receiver front-end down converts the signal from 24 GHz to the IF frequency range by mixing the signal with another signal at a frequency of \((24 \text{ GHz} \pm f_{\text{IF}} \pm f_{\text{mod}})\) depending on which sideband is sought. Figure 4.3 shows the spectrum of the down-converted signal at the output of the mixer. In this figure, the LSB is mixed to the IF frequency of 10.7 MHz. The attenuation of the carrier and the USB is due to the effect of the IF filter, since the test port was not matched to 50 \(\Omega\) and only used to verify the presence and the frequency spectrum of the signal. Figure 4.4 shows the spectrum of the signal at the output of the IF stage (e.g., the input to the IQ demodulator). At this stage, the sought sideband is passed through the filter and amplified. Subsequently, the carrier and the other sidebands and harmonics are attenuated by the filter. The stopband rejection for this IF filter is approximately 45 dB, which is clearly evident by the measured results shown in Figure 4.4. The contribution of these unwanted signals will be further reduced by match filtering at the IQ demodulator.

Figure 4.3. Signal at the mixer output of channel 4 due to middle slot modulating (RBW = 3 KHz).
4.2. MODULATION FREQUENCY

While there are constraints on the modulation frequency, there are also several optimization opportunities. Traditionally, the lock-in amplifiers are operated at the modulation frequency, which produces a direct correlation between the modulation frequency and the flicker $(1/f)$ noise [34],[51]. In this design the lock-in amplifiers (i.e., the IQ demodulator) operate at the fixed IF frequency of 10.7 MHz. Therefore from the receiver point-of-view, the modulation frequency does not contribute to the performance of the receiver in the presence of flicker noise. On the other hand, the modulation frequency should be higher than the bandwidth of the IF filter such that the source leakage and other harmonics are filtered out. Furthermore, as shown in Figure 4.4 the IF filter attenuates the harmonics and leaked carrier in its stop-band. Subsequently, increasing the modulation frequency even higher gives the IQ demodulator (i.e., the matched filter) a better chance of rejecting these unwanted signals.

Figure 4.4. Signal at the IF output of channel 4 due to middle slot modulating (RBW=10 KHz).
The speed of electronically scanning the retina also introduces a lower limit on the modulation frequency. The video frame rate of 30 images per second translates into approximately 50 microseconds spent measuring the electric field at each slot. If the rule-of-thumb of minimum 10 modulation cycles per measurement is applied, then this time translates to a minimum modulation frequency of 200 KHz.

Another limiting factor, on the modulation frequency, is the design of the modulation/addressing network. Furthermore, the current retina design requires higher bias current to be modulated or switched which leads to further constraints on the upper limit of the modulation frequency. Cross talk on the modulation line may also become an issue. However, given that the PIN diodes are forward biased and most probably saturated when not modulated, any small cross talk has no or negligible effect. Through experimentation and considering the above reasoning, a modulation frequency higher than 500 KHz was deemed necessary for favorable performance of this imaging system.

Figure 4.5 shows the influence of the leaked carrier presence on the noise level in the pass-band of the IF filter in the absence of modulation. These measurements were obtained using a spectrum analyzer at the IF output of the receiver while the retina was illuminated by a quasi-plane wave and all slots were closed. The presence of the leaked carrier through the retina, in conjunction with the IF amplifiers and filter, produces noise-like signals in the pass-band of the filter (Figure 4.5(a)). The power associated with this noise may either be reduced by decreasing the power in the leaked carrier (i.e. better isolation in the retina) as shown in Figure 4.5(b), or by increasing the modulation frequency thus moving the carrier further away from the filter pass-band, as shown in Figure 4.5(c). Better yet a combination of these two solutions yields lower noise level and higher improvement, as shown in Figure 4.5(d).

4.3. RECEIVER DYNAMIC RANGE

In section 3, it was shown that the receiver front-end and the IQ demodulator both exhibit a dynamic range exceeding 90 dB. However, the overall dynamic range of the receiver will be limited by signals passing through the IF filter such as the noise generated by the leaked carrier, as described above. The following experiment was performed to establish the overall dynamic range of this system. The experimental setup
is shown in Figure 4.6 where a modulated slot was used to modulate and pick up the signal transmitted by an irradiating open-ended rectangular waveguide. A precision variable attenuator was used to change the incident RF power on the slot. On the other hand precisely controlling the phase of the RF signal without affecting its magnitude is not an easy task. To change the phase of the RF signal the following property of modulated scattering was used. It is established that changing the phase of the modulating waveform changes the phase of the modulated RF signal [34]. With traditional receivers, this phase change may be lost if the signal is not decoded to extract a single sideband of the modulated waveform only. Typically, both sidebands are
measured and to extract a single sideband from those measurements, extra processing steps are required [34]. As explained earlier, with the current design of the receiver only a single sideband is measured. Therefore the measured phase is directly correlated to the phases of the RF and modulated signals combined. Since in this experiment the phase of the incident RF signal is constant, the receiver phase sensitivity may be measured directly by changing the phase of the modulating waveform. In this experiment two phase locked signal generators were used. One signal generator provides the 20 MHz reference clock for the sources on the camera (replacing the reference crystal) and the other is used to modulate the receiving slot at 1 MHz. The phase of this modulating signal is changed over one cycle for each incident power setting. The computer controls phase setting and the data acquisition.

![Diagram](image)

**Figure 4.6. Experimental setup for testing receiver dynamic range.**

The test results for this experiment are shown in Figure 4.7 and Figure 4.8 illustrating the receiver dynamic range and the accuracy by which phase is measured, respectively. Figure 4.7 shows the dynamic range of the receiver, indicating a linear
response in a range of approximately 70 dB. The receiver saturates on an input power level of -35 dBm. The saturation is due to the high IF stage gain, which saturates the IQ demodulator. Figure 4.7 shows the corrected signal output power after compensating for this IF gain and ADC gain. When analyzing this dynamic range these corrections must be taken into account. For example looking at the noise in Figure 4.5, a maximum noise level of -80 dBm in the IF is translated (i.e., subtracting a 35 dB front-end and IF gain) to -115 dBm RF signal level which is the floor of the receiver dynamic range.

![Figure 4.7. Dynamic range of the receiver.](image)

Figure 4.8 shows that the accuracy associated with measuring phase for this receiver is below 10 degrees error at the low end of its dynamic range and it is approximately 1 degree at the high end (i.e., ~50 dB) of the receiver dynamic range. This phase error seems large compared to the test performed on the IQ demodulator separately which yielded phase errors on the order of 0.1 degree. This large phase error may be due
to the low phase accuracy provided by the Stanford CG635 sources. These sources guarantee a phase error of less than one degree for frequencies higher than 200 MHz and it degrades by a factor of ten each decade below 200 MHz [55]. When testing the IQ demodulator, the Stanford sources were producing a 10.7 MHz signal, while in this test the modulation frequency was set to 1 MHz. Therefore, it was not expected that the Stanford test sources will maintain accurate phase setting. Overall, this receiver was optimized for when the retina captures scattered signals with corresponding relatively low powers. The saturation level of -35 dBm is acceptable since the retina is expected to measure small signals from far transmitters or scattering from week objects.

![Phase sensitivity of the receiver](image)

Figure 4.8. Phase sensitivity of the receiver.

4.4. ADDRESSING NETWORK PERFORMANCE

An issue of great importance in the design of the modulation addressing network is the ability to modulate one slot in a row/column while keeping the rest of the slots in
that row/column from modulating. As explained in the previous section, the idea behind the use of diodes (LEDs) to interface the PIN diode to the row and column modulation line is that if the PIN diode is supplied with a DC current from one of the LEDs (i.e., the control LED), it will become forward biased and saturated. Consequently, any variation (i.e., modulation) in the supplied current from the other LED (i.e., modulating LED) would not result in any impedance change to the slot. On the other hand, if the control LED is turned off, then the PIN diode on the slot would follow the modulating signal supplied through the modulating LED. An experiment was performed to check the performance of this switching scheme. In this experiment, the spectrum of the signal transmitted through a single slot was measured using a spectrum analyzer. Figure 4.9 shows the results for the carrier at 24 GHz and the first harmonics of the upper and lower sidebands of a square wave amplitude modulated signal for three cases of bias voltage. The control voltage of the retina is intended to be at 3.3 volts. However, due to the high current consumption, a slight voltage drop in the switches and the cables was expected and experimentally noticed. In this experiment, the slot exhibited the expected near 100% modulation efficiency when the control (green) LED was turned off. On the other hand when the green control LED was turned on, the slot did not completely close and a leaked modulated signal as high as -20 dB (normalized to incident power) was recorded. As the bias voltage increased, this leaked modulated signal decreased in amplitude since the PIN diode receives more current and becomes more saturated. As it was established earlier, any non-modulated leakage is filtered through the IF filters. However, this leaked modulated signal will pass through and will be measured. To summarize: when the modulation is passed through the column (red LEDs) to a slot, and the row address lines (through green LEDs) are used to close the rest of the slots, as much as -20 dB of the signal at the location of each slot from that column will contribute to the measured signal from a slot in that column. This translates to a poor isolation between the modulated array elements. One major reason for this modulated leakage phenomenon is the fact that the impedance of the red and green LEDs are not the same. The red LED requires 1.8 volts to turn on while the green LED requires 2 volts. Therefore, the green LED presents a larger impedance to the DC current reaching the PIN diode on the slot. Yet this impedance difference between the red and green LEDs may be used to fix the modulation
isolation problem. As shown in Figure 4.10 if the role of the red and green LEDs is reversed, the modulation isolation is improved. In this case, the modulation signal is passed through the row address lines (green LEDs) and the control signal is passed through the column address lines (red LEDs). Since the red LED presents lower impedance to the DC current, it takes priority in saturating the PIN diode especially at lower bias voltages where the retina is expected to be normally operating.

![Bias Red Modulating Red Modulating Voltage Green OFF Green ON](image)

Figure 4.9. Modulation isolation for various bias voltage: modulating signal through the Red LED and control signal through Green LED.
Figure 4.10. Modulation isolation for various bias voltage: modulating signal through the Green LED and control signal through Red LED.

Figure 4.11 shows the effect of this leakage on the measurements performed with the retina. A K-band open-ended rectangular waveguide antenna was used to illuminate the retina with a quasi-plane wave. When modulating the middle slot on the retina either through column (Figure 4.11a) or row (Figure 4.11c) address lines, the measured power at the IF output is approximately -35 dBm. When that middle slot is closed, only the cumulative leaked modulated signal is measured. In the case of column modulation (Figure 4.11b) the leakage was only 15 dB lower than the intended signal, while in the case of row modulation (Figure 4.11d) this leakage was 50 dB lower. Overall, the switching network and modulation creates a spatial multiplexing technique and the
leakage represents isolation in this multiplexing scheme. It is desirable to have as high of isolation as possible. On the other hand, it is difficult to design a compact array with very high isolation. Consequently a correction technique may be applied to correct for these unwanted contribution from untagged slots.

\begin{figure}[h]
\centering
\includegraphics[width=\textwidth]{figure411.png}
\caption{Signal and leakage measured at the output of IF when modulating a single slot in presence of plane-wave: (a) modulation on column; signal from middle slot, (b) modulation on column; cumulative leakage, (c) modulation on row; signal from middle slot, and (d) modulation on row; cumulative leakage.}
\end{figure}

4.5. RECEIVER DETECTION BANDWIDTH

The amount of noise affecting the measurement depends on the receiver detection bandwidth. As designed, the receiver has three filters. The first filter is the physical IF
filter with a bandwidth of 100 KHz. The second filter is the RC lowpass filter at the output of the IQ demodulator with a cutoff frequency of 30 KHz, and the third filter is obtained by the averaging process after the analog to digital conversion. These three filters can be cascaded to obtain the overall equivalent detector bandwidth. Assuming the maximum detection filter bandwidth is \( B = 30 \) KHz (i.e. the bandwidth of the cascaded IF and IQ filters), then the maximum possible sampling frequency \( f_s \) of the data becomes twice this bandwidth, as dictated by the Nyquist sampling theorem:

\[
f_s = 2B. \tag{4.1}
\]

Therefore, the time interval between two successive samples is:

\[
T = \frac{1}{f_s}. \tag{4.2}
\]

The time \( T \) (i.e. the dwell time) is the minimum time that should be spent tagging each slot in order to obtain meaningful non-aliased data. For \( B = 30 \) KHz, this time becomes \( T = 16.67 \) ms. Multiplying this time by the number of slots in the array (576 slots) results in a frame acquisition time of less than 10 ms.

Another method of looking at this problem is by setting the dwell time. If a video frame rate is required, for example a frame every 30 ms, the dwell time for each slot becomes \(~50\) \( \mu \)s. This dwell time is 3 times larger than the time calculated for the 30 KHz bandwidth or in other words, by increasing the sampling frequency and sampling time by a factor of 3, it is possible to take 3 measurements per slot or in a more general case:

\[
N = \frac{T}{f_s}. \tag{4.3}
\]

where \( N \) is the number of averaging samples. An average of \( N \) will reduce the effective filter bandwidth by a factor of \( N \). Furthermore, it is advantageous to sample at much higher sampling rate and take multiple samples to average out the noise of the analog to
digital convertors (ADC). For example using a sampling frequency of 2.5 MHz enables taking up to 125 samples per slot for averaging purposes.

Figure 4.12 shows the response of the three aforementioned filters, represented using Matlab filter visualization tool [56]. The IF filter is simulated using an order 3 Butterworth filter and the IQ filter is a first order (RC) Butterworth filter with a bandwidth of 30 KHz. The sampling frequency was assumed to be 2.5 MHz and an averaging length of 100 samples is used. The sampling frequency and number of samples satisfy a frame rate of 30 retina scans per second. This analysis helps in selecting the optimum filter. The effective cascaded filter shows a bandwidth of 10 KHz. Figure 4.13 shows the simulated group delay in these filters. The group delay in the cascaded filter is 30 μs.

![Graph of Figure 4.12](image)

Figure 4.12. Magnitude response of the various detection filters in the system and their cascaded response.
Figure 4.13. Group delay of the various detection filters in the system and their cascaded results.

4.6. CAMERA RANGE

The maximum range, $R_{\text{max}}$, of this camera can be obtained using the radar range equation,

$$ R_{\text{max}} = \left[ \frac{PG_tG_r\lambda^2\sigma}{(4\pi)^3 S_{\text{min}}} \right]^{1/4} $$  \hspace{1cm} (4.4) 

where $P_t$ is the transmitted power, $G_t$ is the gain of the transmitter antenna, $G_r$ is the gain of the receiver antenna (retina), $\lambda$ is the wavelength, $\sigma$ is the radar cross section of the target, and $S_{\text{min}}$ is the minimum detectable signal [57]. The minimum signal detectable ($S_{\text{min}}$) can be written as,

$$ S_{\text{min}} = (\text{SNR})_{\text{min}} (NF) kT_0 B $$  \hspace{1cm} (4.5)
where $SNR_{min}$ is the minimum required signal to noise ratio, $NF$ is the noise figure of the receiver, $k \approx 1.38 \times 10^{-23}$ is the Boltzmann constant, $T_0$ is the standard absolute temperature (usually 290 Kelvin), and $B$ is the bandwidth of the receiver. $NF$ may be found using the Friss' formula for noise (cascade rule) and the properties (noise figure and gain) of each component in the receiver chain [58]. Using the Friss' formula, the total noise figure for the receiver was found to be $\sim 4$ dB. As dictated by the Friss' formula, this noise figure is dominated by the low noise figure of the front-end LNA ($NF = 2.5$ dB) and therefore it does not increase much due to the losses in the mixers and filters.

Using (4.5), for $SNR_{min} = 0$ dB, $B = 10$ KHz, and $NF = 4$ dB, the minimum signal detectable $S_{min}$ (i.e., the noise power) becomes $-130$ dBm. This figure is 10 dB lower than the measured dynamic range floor of the receiver as shown in Figure 4.7. However, there are other factors and sources of error that can affect the minimum signal detectable. One factor is the noise due to the oscillators as shown in Figure 4.5. Other factors include the effect of dispersion of the array elements' response, mutual coupling in the array, and isolation in the multiplexing scheme as described above. The dispersion of the array elements' response is described as the variation between the array elements response from the receiver point of view [23]. This dispersion is mainly caused by the waveguide collection network and could be measured and corrected for. These sources of errors and corresponding correction methods will be described later.

For the current system configuration and assuming the noise level of $-120$ dBm (Figure 4.7), the range of the camera will depend on the target radar cross section and the required SNR. Figure 4.14 shows the maximum range for the camera for a metallic sphere with a diameter of 4 mm ($\sigma = 5 \times 10^{-5}$ m$^2$) vs. SNR. The transmitter was assumed to be a small horn antenna with a gain of $G_t = 10$ dB and a transmitted power was set to $P_t = 10$ dBm. The gain in the receiver $G_r$ was the gain of an individual slot (6 dB) multiplied by the array gain computed as $G_a = 4\pi L_2$ where $L_2 = 12$ is the size of the 2D array in wavelength [59]. Furthermore, an average of $-15$ dB loss was added to the gain of the array due to the attenuation in the waveguide network. This high range as Figure 4.14 suggests, is due to the high gain ($\sim 30$ dB) the array introduces. Without considering the array gain, the maximum range will drop by more than an order of magnitude.
Furthermore, this maximum range was found assuming an ideal array with no mutual coupling, loss, etc.

![Graph showing camera maximum range in free space plotted vs. SNR.](image)

Figure 4.14. Camera maximum range in free space plotted vs. SNR.

4.7. SOURCES OF ERROR

4.7.1. Slot Mutual Coupling. One of the advantages of using these modulated resonant elliptical slots for imaging arrays is the low mutual coupling they introduce when placed in an array [38]. A study on similar slots [43] showed that the mutual coupling between each two slots was at the level of approximately -16 dB at a distance of λ/2. Figure 4.15 shows the configurations used for simulating the mutual effect for two adjacent slots in the retina. Coupling was considered in the E-plane, where the slots are placed in the collinear configuration and in the H-plane where the slots are placed side-by-side. The slots were considered on a two layer PCB with Rogers4350 substrate as explained earlier. For the simulations, each slot was fed by the modified rectangular waveguide (a = 7.7 mm and b = 4.3 mm), as described in the previous section. The aperture of slots (top layer of PCB) was at z = 0, and the structure was extended to
infinity in the $x$- and $y$-directions. Various factors were anticipated to affect the coupling between two slots such as the dielectric substrate, the state of the slots (open or closed), and the distance between the slots.

![Diagram](image)

Figure 4.15. Configuration for simulating mutual coupling between two slots; (a) E-plane (collinear), and (b) H-plane (side-by-side).

When the slots are isolated through the dielectric substrate – by placing a solid conducting ring around them, the surface current on the top conductor becomes the only mechanism of coupling between them. This source of coupling was investigated on similar slots in [43] for the case when both “aggressor” and “victim” slots were open. As mentioned previously, when a slot is idle it is considered to be in the closed state. Figure 4.16 shows the tangential magnetic field – which is proportional to the surface current density [60], at the aperture of the slots at $z = 0^+$ when port 1 is active. A First glance shows that the surface current extends more in the $y$-direction (E-plane) than the $x$-direction (H-plane) which will result in a stronger coupling in the E-plane than the H-plane. When both slots are open (i.e., the PIN diode is OFF), the active slots produce a current distribution in the victim slot as shown in Figure 4.16a and Figure 4.16b. Once
again, this induced current is stronger in the E-plane than the H-plane. On the other hand when the victim slot is closed, the induced current on that slot becomes much smaller, as shown in Figure 4.16c and Figure 4.16d.

Figure 4.16. Tangential magnetic field (∝ surface current density) at \( z = 0^+ \) for the case of only port 1 active, (a) E-plane – both slots open, (b) H-plane – both slots open, (c) E-plane – aggressor slots open and victim slot closed, and (d) H-plane – aggressor slots open and victim slot closed.

Another mechanism of possible coupling is propagation through the dielectric substrate of the PCB. The via-ring around the slot, introduced in the previous sections, confines the signal within the slot structure, which results in a better resonance when the slot is open, and better isolation (lower leakage) when the slot is closed. This via-ring
also helps in lowering the mutual coupling between two slots. Figure 4.17 shows the
tangential magnetic field on the back-side of the slot plane at $z = 0^-$ (inside the dielectric
substrate). Without the via-ring to confine the fields in the slot area (Figure 4.17a and
Figure 4.17b), the dielectric substrate acts as a low-loss propagation medium. These field
mainly spread in the E-plane, with the field distribution stronger at the bottom of the slot
(-y-direction) starting from the location of the PIN diode, unlike the field distribution at $z
= 0^+$ where the distribution was symmetrical in the y-direction. By placing the via-ring,
the coupling through the dielectric substrate is greatly reduced, as shown in Figure 4.17c
and Figure 4.17d.

Figure 4.17. Tangential magnetic field ($\propto$ surface current density) at $z = 0^-$ for the case
of only port 1 active, aggressor slots open, and victim slot closed, (a) E-plane – without a
via ring, (b) H-plane – without a via ring, (c) E-plane – with a via ring, and (d) H-plane –
with a via ring.
The effect of slot-to-slot distance on the mutual coupling was investigated using the same numerical simulation tool. Figure 4.18 shows the normalized coupling coefficient for various cases of isolation through the dielectric substrate using a solid conductive ring, a via-ring, and when there is no isolation. The normalized coupling coefficient is defined as [43],

\[
C = 10 \log \left[ \frac{|S_{31}|^2}{\left(1-|S_{11}|^2\right)\left(1-|S_{22}|^2\right)} \right]
\]  

(4.6)

where \(S_{11}, S_{21},\) and \(S_{22}\) are the scattering parameters between ports 1 and 2 of the setup in Figure 4.15 computed at 24 GHz. This normalized coupling coefficient is plotted as a function of normalized, to free-space wavelength center-to-center interspacing. The smallest distance was limited by the dimension of the waveguide which is larger in the H-plane. When both slots are open (Figure 4.18a and Figure 4.18b), the coupling coefficient at the distance of \(\lambda/2\) is approximately -14 dB for the E-plane, and approximately -16 dB for the H-plane. Moreover, the coupling coefficient decreases faster in the H-plane compared to the E-plane. This difference between the H-plane and the E-plane is inline with the observation made on the surface current distribution results, shown in Figure 4.16. Furthermore, it can be noticed that the via-ring isolates the slots from the dielectric substrate to the point that the coupling coefficient for the solid ring and the via-ring cases are virtually the same. When no isolating ring is placed around the slot, the fields coupled through the dielectric substrate and the fields coupled from the outside add up constructively or destructively which directly affects the coupling coefficient. This interference is due to the difference in the propagation medium on top of the slots (air) and the dielectric substrate.

More important is the coupling coefficient when the second slot is closed as shown in Figure 4.18c and Figure 4.18d. In this case the coupling coefficient drops by 12 dB in the E-plane and by more than 20 dB in the H-plane compared to the case where both slots are open. Furthermore, the presence or lack of the via-ring has a substantial effect on the mutual coupling between the slots. Without the via-ring, the coupling coefficient stays almost constant irrespective of the distance between the slots. This
phenomenon is attributed to the dielectric substrate acting as a low-loss bounded medium (transmission line) coupling two slots to each other efficiently. To summarize, mutual coupling between the slots in the retina as designed and operated, is smaller than -30 dB. This small level of mutual coupling is not expected to affect the electric field measurements using this type of slots.

Figure 4.18. E- and H-plane mutual coupling between two slots, (a) E-plane – both slots open, (b) H-plane – both slots open, (c) E-plane – aggressor slots open and victim slot closed, and (d) H-plane – aggressor slots open and victim slot closed.
4.7.2. Isolation in the Switching Scheme. As it was shown earlier, the slots exhibit a small leakage when closed. The premise of modulation is that these leakages will be non-modulated and will be filtered out. However, when a slot is modulating, any signal passing through that waveguide, including the leaked signal will be partially modulated. This effect is more prominent between the slots in the same collector waveguide. The following simulated experiment shows the effect of modulation on this leakage. The simulated setup is shown in Figure 4.19. Two slots are considered on the narrow wall of a waveguide. The slots are fed with a waveguide in order to provide for an isolated port on each slot (ports 1 and 2). Two other ports (ports 3 and 4) were placed on each end of the collector waveguide. For this simulation, the slot at port 2 was closed and the slot on port 1 was active (open or modulated). The S-parameters between the various ports were obtained and the isolation is computed. Isolation is defined as the ratio in power between the leaked signal through port 2 and the actual signal through port 1. Since there are two measurement ports (port 3 and port 4), two isolation coefficients are obtained. For the non-modulated case, the isolation is defined as

![Figure 4.19. Simulation setup for obtaining the isolation in the multiplexing network.](image)
where \( j = 3,4 \) indicates the measurement port. For the modulated case, the isolation is redefined as

\[
I_j = 10 \log \left( \frac{|S_{j2}|^2}{|S_{j1}|^2} \right)
\]

(4.8)

where \( |\Delta S_{ji}|^2 \) is the power in the modulated signal received at ports \( j = 3 \) or \( 4 \), due to a signal at ports \( i = 1 \) or \( 2 \).

The simulation is performed for various distances between the two slots. The results are shown in Figure 4.20. As shown in Figure 4.20, the un-modulated isolation in

![Figure 4.20. Un-modulated and modulated isolation in the multiplexing scheme.](image-url)
port 3 is very poor. The isolation seen by port 4 in the un-modulated case is much smaller better, since the inactive slot (slot 2) is between the active slot and port 3. On the other hand, both modulated cases show an improved and comparable isolation of 16 dB. It is clear that modulation improves the isolation in array, yet this isolation is far from ideal. For imaging applications, this amount of isolation may be sufficient since in imaging a qualitative data is usually of concern. However, the accuracy of electric field measurement will be affected by this isolation. This isolation must be measured and corrected for in order to achieve high measurement accuracy.

4.7.3. Array Response Dispersion. Array dispersion is the largest contributor to errors in mapping electric field distribution and at the same time it is the easiest to correct for. Array dispersion is referred to the variation in the response of array elements to a uniform plane wave illumination [23]. In other words if the array is illuminated with a uniform plane wave, the measurement will be non-uniform due to distortions by the response of each individual array element. The variation may be due to manufacturing inaccuracies between the slots, due to inaccuracies in mounting the components on the array, or due to differences between nominal values of the component properties. However, experiments showed that these variations are typically very small. The largest variation in the response of the array elements comes from the link connecting each array element to the receiver. This link is the path that the signal travels through, from where it is picked up by a slot, until it reaches the receiver. This path takes the signal through the collector waveguide, the iris, the combiner waveguide, and finally through a short cable connected to the receiver. In each portion of this path the signal undergoes attenuation and reflection. For example, when the signal goes through the collector and combiner waveguides, it loses some of its power as leakage through each closed slot it passes by. It was anticipated that signal amplitude will depend linearly (in dB) on the distance between the slot and the output port. Using 4 ports, one at each corner of the array, alleviates some of this distortion since there is always a port for each quarter of the array. In this case the signal experiencing maximum attenuation comes from the middle slots.

The distortion can be modeled as

\[ S_{ij}^{\rho} = \gamma_{ij}^{\rho} E_{ij}^{\rho}, \]  

(4.9)
where \( S'_p \) are the measured signal at port \( p \) \((p = 1, 2, 3, 4)\) due to actual incident field \( E'_a \) at slot location \((i,j)\), \( \gamma'_p = \alpha'_p e^{j\theta'_p} \) is the complex error coefficient linking slot \((i,j)\) to port \( p \) (i.e., array response coefficients). Typically a response calibration is performed to find the elements of these distortions [61]. The response calibration is performed by illuminating the retina with a known electric field, which is typically a plane wave. A first-order approximation is to correct for each measurement port individually (i.e., invert equation (4.9)) and then average the corrected results. However, given that when the slot is far from a port \( \gamma'_p \) becomes very small (high attenuation), the contribution from that measurement is likely to introduce more noise than the measurement from the opposite port, which will be close to the slot. In this case one may consider using a weighted average, defined as

\[
E'_a = \sum_{p=1}^{4} w_p \frac{S'_p}{\gamma'_p},
\]

where \( w_p \) are the normalized weights and are found through the calibration process. The correction and characterization techniques will be described next.

4.8. CORRECTION TECHNIQUES

4.8.1. Aperture Dispersion. Two methods of calibration were investigated. In both methods it was assumed that mutual coupling is small and does not adversely contribute to the errors in measurement since it was shown through numerical simulations previously. Furthermore, mutual coupling is difficult to measure, since there is no access to individual array elements and therefore it is difficult to correct for it. The following are two techniques for estimating the array response and correcting for it.

4.8.1.1 Scanned localized excitation. Assuming that the isolation between array elements is high, it is possible to obtain the correction coefficients by exciting each individual array element separately. This is performed by placing a small transmitting antenna on the slot and measuring the signal transmitted through that slot. The through-transmitted signal is measured at all four ports. By moving the transmitter antenna using
a mechanical scanner over the array element, the response for all elements was obtained. Since the transmitted signal for each slot does not change, the variations in the measurements are due to characteristics of the retina only. By applying equation (4.9), the correction coefficients may be found, as shown in Figure 4.21. The transmitter was a resonant elliptical slot antenna. The sides of the slot antenna were covered by absorbing film to reduce multiple reflections between the retina and the transmitter. Furthermore, the characterization was made with several distances (one wavelength range) between the transmitter and the retina to average out the effect of any residual multiple reflections.

Figure 4.21. Picture showing the setup for characterizing the array using a mechanically scanned probe to provided localized excitation.

Figure 4.22 shows 2D maps of the normalized signal power measured at the four output ports of the retina. The process of mechanical scanning mimics a plan-wave excitation. These maps show the attenuation of the signal as it travels through the waveguide collectors and combiners towards the four output ports. They also show that when a port receives a highly attenuated version of a signal, there will be at least one other port that receives a strong version of that same signal.
Equation (4.10) can be rewritten as an optimum maximum ratio combiner (MRC) [62] where the weights are the complex conjugate of the array response coefficients $\gamma_p^r$ shown in Figure 4.22:

$$E_{\eta}^\eta = \chi_{\eta} E_{\eta}^\eta = \sum_{p=1}^{4} (\gamma_p^r)^* S_{\eta}^p$$  \hspace{1cm} (4.11)

where $E_{\eta}^\eta$ are the combined signals from all four measurement ports, and $\chi_{\eta}$ are the

Figure 4.22. Normalized measured signal power (dB) for uniform localized excitation, measured at the four ports of the retina.
array correction coefficients. Furthermore, the array coefficients may all be phase referenced to a single port (e.g., port 1) yielding:

\[
E_{ij}^m = \sum_{p=1}^{4} \alpha_{ij}^p S_{ij}^p e^{-j(\theta^i_p - \theta^i_1)}. \tag{4.12}
\]

Equation (4.12) describes the signal combination as a weighted average after phase aligning all signals with port 1. When performing the characterization, multiple measurements are performed to average out the effect of multiple reflections. When changing the height of the transmitter, the amplitude changes slightly, however the phase can go through a full cycle. By referencing the phases to the phase at port 1, the difference becomes independent of the phase of the transmitted signal and only as a function of the array properties as shown in Figure 4.23 for element \((i = 12, j = 12)\). A slight variation with a standard deviation of less than one degree is observed in these phase differences. These subtle variations may as well be due to multiple reflections between the transmitter and the retina.

![Figure 4.23. Referenced phases of the MRC coefficients for array element \((i = 12, j = 12)\) as a function of the transmitter height.](image-url)
After performing the MRC, the combined measurements $E^{m}_{0}$ are divided by $\chi_y$ to correct for the residual array dispersion and find the actual incident electric field $E^{a}_{y}$. The coefficients $\chi_y$ are found using

$$\chi_y = \frac{E^{m}_{y}}{E^{a}_{y}}$$

(4.13)

for a known incident electric field $E^{a}_{y}$. The known electric field in the case of plane wave excitation or in the case of scanned localized excitation is $E^{a}_{y} = 1$. Figure 4.24 shows the magnitude and phase distribution of $\chi$. After combining the four measured signals, the residual array correction coefficients shows a smaller variation in amplitude, except for the elements on the array edges which are stronger in response than the majority in the middle of the array as expected since they go through the least amount of loss before reaching the nearest output port. The phase of correction coefficients shows a trend moving towards port 1, since all measurements were phase referenced to that port. The large variation in phase is expected because of the array interspacing and the large guide wavelength.

Figure 4.24. (a) Magnitude and (b) phase distribution of the array correction coefficients $\chi$ obtained using the scanned localized excitation method.
Using a scanned transmitter to obtain the correction parameters is simple, yet upon testing the imaging system on known electric field distributions, errors as large as 3 dB in magnitude and up to 20 degrees in phase were noticed. These large errors were due to the inaccuracies in positioning the transmitter on top of the array elements while scanning. The scanner that was used to position and scan the transmitter was introducing a progressive positional error that resulted in as much as 3 mm position error towards the end of the array. This resulted in a skew in the phase of the array correction coefficients.

4.8.1.2 Plane wave excitation. Another viable method to characterize the imaging array is to illuminate it with a known electric field distribution. Typically a plane wave excitation is used since a plane wave has a constant magnitude and phase [61]. For this imaging array the far-field distance is more than 7 meters. In other words in order to produce a plane wave at its aperture, any simple and small transmitting antenna should be at least that far from the retina in an antenna test range. Not having this possibility, the second best solution was to use any other known electric field distribution. The schematic of the test setup is shown in Figure 4.25. The retina is placed in an anechoic chamber, with the transmitting antenna at a distance $h$ providing the known illumination. The transmitter was an open-ended waveguide antenna. The radiated electric field by the waveguide can be found using equation (2.13) and the expressions in [2].

![Figure 4.25. Schematic of the setup used for characterizing the imaging array.](image-url)
The measurements were performed at various distances, $h$, covering a range of two wavelengths from 387 to 411 millimeters with steps of 2 mm. At these distances the magnitude variations over the aperture of the retina does not exceed 1 dB. On the other hand the phase variations go through a full cycle. Figure 4.26 shows an example of the incident electric field on the retina aperture from a distance of 410 mm obtained using the theoretical formulation.

![Figure 4.26. Incident electric field ($E^o$) on the retina aperture at $h = 410$ mm, (a) magnitude (dB), and (b) phase (deg).](image)

The combining coefficients, $\gamma^p$, are obtained in a similar fashion as in the previous case of the scanned illuminator. The array dispersion correction coefficients, $\chi^v$ (Figure 4.27), are obtained by dividing the combined measurements by the actual electrical field distribution. In the case of the scanned localized excitation, the actual electric field was considered to be unity since it does not change from one array element to the next similar to a true plane wave excitation. In this case the illumination is not a true plane wave. The illuminator pattern ($E^o$) similar to that of Figure 4.26 at various distances was found using the theoretical formulation for an open-ended waveguide [2]. The array correction coefficients, $\chi$, shown here look similar to the previous case. The
differences are subtle and the ultimate effect is noticed when measuring an electric field distribution.

Figure 4.27. (a) Magnitude and (b) phase distribution of the array correction coefficients \( \chi \) obtained using a semi-plane wave illumination.

Figure 4.28 shows the corrected measurement vs. theoretical electric field distribution produced by an open-ended waveguide from a distance of 390 mm. This measurement was not part of the calibration set. The magnitude plot shows slight distortion. This distortion is very small and it is only noticed due to the small range in the actual field. The average absolute magnitude error in this measurement is \(-0.025\). The phase plot on the other hand shows a strong resemblance to the theoretical phase pattern. The absolute phase error was \(-3.5\) degrees. Overall, these accuracies are considered very high for such a high frequency system. Furthermore, for imaging applications and array processing techniques such as synthetic aperture techniques, the overall phase pattern is much more important than the absolute error. Section 5 of this dissertation provides other examples of measuring electric field distribution along with some imaging application.
4.8.2. Full Correction. A Complete error correction assumes that a signal received when tagging (modulating) a certain slot is a combination of the incident electric field on all slots. This assumption combines the effect of mutual coupling, isolation, and array dispersion into a set of error coefficients. The measured signal at each slot location is defined as,

$$ S_i^p = \sum_{j=1}^{576} h_{ij}^p E_j $$

(4.14)
where $S^p_i$ is the signal measured at port $p$ due to modulating slot $i=1,2..576$, $h^p_i$ are the error coefficients, and $E^p_j$ are the actual electric field values at slot location $j=1,2..576$. In matrix form equation (4.14) becomes

$$[S]_{576×1}^p = [H]_{576×576} [E]_{576×1}.$$  

(4.15)

To find the correction coefficients, the above equation in inverted. The matrix $H$ contains the array dispersion effect in its diagonal elements. The rest of the elements, describes the combined effect of mutual coupling and isolation in the multiplexing scheme. Finding $H$ is not trivial. Simulating the complete array is not practical due to the vast amount of details involved. Furthermore, accurate measurement of the matrix $H$ will require feeding each element separately. Feeding each element separately will require a small feed structure to be placed on each slot while isolating this feed from the rest of the array. If such feed is possible, when feeding each slot, the array is modulated sequentially, thus one column of $H$ is found. Repeating this process for all slots, matrix $H$ is obtained, and the array becomes fully characterized. However, such feed is not possible since isolating a slot would require a structure of waveguide and absorber material to be placed on the slot that will affect the resonance property of that slot or the neighboring slots.

Another solution is to illuminate the array with 576 diverse electric field distributions. Thus $H$ is obtained by inverting equation (4.15). This solution was tried using the electric field distributions from the experiment in section 4.8.1.2. However, the resulting matrix $H$ was singular and non-invertible. An approximation to $H$ from those measurements resulted in correction coefficients that were applicable to the limited cases of electric field distributions only. Diversified electric field distributions will yield accurate correction matrix.
5. APPLICATIONS

There are numerous potential applications for this microwave imaging system. Any application that requires 2D electric field mapping is a potential candidate. For example, this imaging system can be used to rapidly map the electric field pattern of an antenna in its far-field. Figure 4.1 shows the system displaying the far-field pattern of a K-band open-ended rectangular waveguide (the two images on the left side). Another application is localizing a scatterer in a 2D space. This is achieved by utilizing the system as a 2D synthetic array producing a narrow scanning beam. Furthermore, by increasing the frequency bandwidth of the system it is possible to obtain some range resolution making this system a true 3D target localization device.

This system is more suitable for imaging applications, where the overall distribution of electric field (or pattern) is critical for forming an image of a scattering object. This system can be readily used in through transmission mode, where the object

![Image of the imaging system displaying electric field distribution of an open-ended waveguide.](image.png)

Figure 5.1. The Imaging system displaying electric field distribution of an open-ended waveguide.
is placed in between the transmitter and retina. Reflection measurements are also possible, if the object can be properly illuminated using an external transmitter. Monostatic reflection mode whereas the retina is used both as a transmitter and receiver is not possible with this current design due to the limited isolation in the waveguide network. This section demonstrates examples of utilizing this imaging system for certain applications. The calibration model and parameters are those of section 4.8.1.2.

5.1. MEASURING ELECTRIC FIELD DISTRIBUTION

The primary objective of this system is to map 2D electric field distribution on a predetermined spatial grid. Since this retina is relatively large in size, and therefore not suitable for mapping near-field pattern of antennas since the presence of the retina causes significant perturbation in the field of interest. Consequently, several measurements were performed with the goal of mapping the far-field radiation pattern of a K-band open-ended rectangular waveguide antenna at distances ranging from 300 mm to 460 mm with a step size of 10 mm. These measurements were performed in a small anechoic chamber with the retina placed face up at the bottom of the chamber, and the open-ended waveguide antenna radiating from the top. Figure 5.2 shows normalized measured vs. theoretical electric field maps for the open-ended waveguide placed at a distance of 300 mm from the retina. The distortion in the measured amplitude plot is due to small errors overwhelming the small range of the actual amplitude. The measured phase shows more resemblance to the theoretical phase except for a slight skew. This skew was due to a slight tilt in the aperture of the waveguide compared to the aperture of the retina.

A better quantitative comparison is shown in Figure 5.3. In this figure the E-plane (along y-axis) and H-plane (along x-axis) cross-sections of the electric field map are plotted. Again, the error in the magnitude plot is on the order of the actual signal variations, while the measured phase is matching its theoretical counterpart precisely. The root mean squared error (RMSE) compared to theory, for this measurement was 0.075 for magnitude and 5 degrees for phase. Overall given the frequency of operation and the difficulties encountered aligning the measurement setup, these accuracies in magnitude and phase are acceptable.
Figure 5.2. Measured vs. theoretical field pattern at 300 mm distance of a K-band open-ended waveguide aperture, (a) theoretical magnitude (dB), (b) measured magnitude (dB), (c) theoretical phase (deg), and (d) measured phase (deg).

Figure 5.3. Measured vs. theoretical amplitude and phase of the E- and H-plane cuts of the field pattern at 300 mm distance from the K-band open-ended waveguide aperture.
Figure 5.4 shows the average amplitude and phase of the four middle slots vs. the distance between the waveguide and retina for the aforementioned set of measurement. The average magnitude and phase values from the four middle slots were taken due to the absence of a central slot (even number of rows and column), and the fact that from these distances the field pattern resembles that of a plane-wave in the relatively small area of these four slots. The magnitude measurement shows an RMSE of ~0.03 compared to the theoretical results, while phase shows a RMSE of ~13 degrees. These errors are quite acceptable for this experiment since the open-ended waveguide was positioned manually and the 13 degrees corresponds to less than 0.5 mm in position error at 24 GHz. For all practical purposes, this retina measures the electric field pattern accurately. Further improvements would require a better calibration of the retina that takes into account and corrects for the limited isolation in the waveguide network and the mutual coupling among the slots.

![Figure 5.4. Measured and theoretical amplitude and phase of vs. distance from the K-band open-ended waveguide aperture.](image-url)
5.2. TARGET LOCALIZATION

This imaging system with its relatively large array size of $12\lambda \times 12\lambda$ can provide a sharp beam (~5 degrees beamwidth) which can be used to localize a radiating source or scattering target within a 2D angular space. Furthermore, utilizing techniques such as synthetic aperture focusing (SAF) [5],[32] allows for localization of the target within the near-field of the array. For this array, the far-field distance ($\frac{2D^2}{\lambda}$) where $D$ is the largest linear dimension of the array, corresponds to more than 7 meters. SAF, a technique derived from Fourier optics allows for lens-like focusing abilities. Figure 5.5 shows the 3D spot size (3-dB level) simulated for the retina array placed in the $XY$-plane at $Z = 0$ for a point source target placed at $Z = 100$ mm. The height of this spot is called the depth of field, which is the range at which the target appears within 3-dB value and is considered to be an indication of range resolution. The diameter of this spot in the $XY$ plane at the focus distance (location of target), is an indication of cross-range resolution. The 3D spot size will depend on the distance of the target to the array.

![Figure 5.5. Simulated focused 3D spot size using the synthetic aperture focusing technique for the retina array.](image-url)
Figure 5.6 shows the simulated depth of field and the spot size diameter for a point target in the array’s broadside direction as a function of target distance to the array. The depth of field starts at approximately 12 mm for distances very close to the array, and exponentially increased with distance as shown in Figure 5.6(a). This is expected since the array operates at a single frequency and has no far-field range resolution. The depth of field in the array far-field will be infinity since a single frequency operation does not allow for any range resolution. The cross-range resolution starts at \( \lambda/2 \) [32], and then linearly increases with the focus distance (range). In the far-field of the array, this spot size will correspond to the angular beamwidth of the array. Overall, the SAF technique allows for limited 3D focusing capability, even at a single frequency operation. However, this capability is limited by the number of the scatterers and their scattering cross section. Strong scatterers near the array will not be easily defocused and will mask weak scatterers at larger distances. Utilizing larger frequency bandwidth allows for higher range resolution irrespective of the range.

![Figure 5.6. Simulated (a) depth of field, and (b) spot size; using the synthetic aperture focusing technique for the retina array.](image)

The experiment shown in Figure 5.7 demonstrates the ability of this imaging system in localizing scatterers in a 3D space. Two metallic balls were hung in front of the retina using thin nylon strings. A 4 mm-diameter ball was placed at a distance of 150
mm away from the retina, and a larger 12 mm-diameter ball was placed at a distance of 260 mm from the retina. A small horn antenna was used to illuminate the scene from the side of the retina.

Figure 5.7. Picture of the setup for target localization experiment.

Screen shots of the camera output are shown in Figure 5.8. This screen shot shows the raw magnitude and phase and a SAF image of the scene obtained in real-time. Figure 5.8(a) is the screen shot where the SAF image is focused at the distance of 150 mm corresponding to the location of the small ball. It is apparent that the small ball is in focus and the larger ball is blurred. On the other hand, if the SAF image is focused at 260 mm for the same raw data as shown in Figure 5.8(b), the large ball comes in focus and the smaller one is blurred. These two images also show the capability of the camera to localize the target in the 2D space relative to the retina spatial domain.
5.3. IMAGING

As it was demonstrated above, the camera has the capability to measure electric field with the accuracy that allows for back propagation techniques such as SAF to be implemented and reconstruct the scattering source. Using this process it is possible to create images of targets that interact with the electric field in many applications. One such application is in nondestructive testing where the image of an object under examination, in particular its inner structure, is of interest. Imaging may be performed in through transmission mode, where the microwave signal is passed through the object and recorded on the other side. Imaging may also be performed in the more desirable reflection mode, where the transmitter and receiver are on the same side of the object being imaged. The reflection mode is more desirable for nondestructive testing applications. As explained earlier, the current state of the camera does not readily allows only for one-sided measurement. This section gives examples of using the camera for this type of imaging applications, mainly in transmission mode. Furthermore, utilizing simple transmitters on the camera sides for reflection mode imaging is also demonstrated.

5.3.1. Through-Transmission Mode. In through transmission mode, the camera is placed in front of a transmitting antenna and the object is placed in between the transmitter receiver. Usually a small transmitter antenna is used to create a fairly broad pattern, such that the object and the retina will be somewhat uniformly illuminated. A

Figure 5.8. Screen shot of camera output, (a) SAF at 150 mm, and (b) SAF at 260 mm.
reference shot without the target is needed to correct for the variation in the pattern of the transmitter antenna [37]. Figure 5.9 shows the camera operation in the through transmission mode. In this case, the transmitter is a K-band open-ended rectangular waveguide antenna, and the target is a two-layer balsa wood composite with a small rubber inclusion. The images on the PC monitor show the raw magnitude, the raw phase and the SAF images obtained in real-time. In this configuration, the target either attenuates, or re-scatters the microwave energy towards the retina, and for this reason the rubber inclusion and the edges of the sample are clearly seen in the focused image while the majority of the balsa wood is seen transparent to microwave energy.

Figure 5.9. Real-time imaging in transmission through mode of a balsa composite with a small rubber inclusion.
5.3.2. External Illuminator Reflection Mode. Using the current camera, reflection mode is only possible utilizing an external transmitter to illuminate the target. However, the pattern of this external illuminator will play a role in the image of the scattering target. Figure 5.10(a) shows the phase pattern of scattered field off a long metallic rod with a diameter of 8 mm recorded at the retina aperture. The experimental setup is similar to that shown in Figure 5.7. The phase pattern shows elongated concentric circles instead of vertical strips expected of a long rod. This phase map is due to the effect of the transmitter pattern on the field scattering from the long rod. Applying SAF to obtain the image of the rod produces a rod indication which varies in intensity inline with the pattern of the transmitting antenna which is more concentrated in the middle as shown in Figure 5.10(b). For simple structures such as rod, it is possible to correct for the pattern effect, as long as there is sufficient power incident on the entire rod.

![Figure 5.10. Reflection mode imaging of long 8 mm-diameter rod; (a) phase map of the scattered field at the retina aperture when the illuminator is in the middle, (b) SAF image when the illuminator is in the middle, and (c) combined SAF images of three separate illumination images.](image-url)
(i.e. no nulls in the transmitter pattern). On the other hand, for complex structures the correction will be difficult due to the effect of the target on the pattern of the transmitter. Another solution is to illuminate the target from different angles and then combine the focused images. Figure 5.10(c) shows an image of the same rod obtained by illuminating the rod from middle left side, top left corner, and bottom left corner of the retina. The three focused images were averaged - as a first order approximation, to produce a complete image of the rod. It is not possible to combine the raw phase and magnitude data, or to have multiple transmitters working at the same time, since that will produce peaks and nulls which will degrade the image. Increasing the number of the transmitters will allow for illuminating the targets from various angles. Subsequently, switching between them and then properly combining the images may results in better images if the retina is intended to be used in reflection mode.

In another experiment a 4" × 2" × 0.5" balsa wood sample with a thin 0.25" × 0.25" copper tape inclusion was imaged in reflection mode, as shown in Figure 5.11. The

![Figure 5.11. Experimental setup for NDI of a balsa wood sample containing a thin copper tape inclusion in reflection mode.](image)
sample was placed at 150 mm distance from the retina. Figure 5.12(a) shows the SAF image when the sample was placed vertically in front of the retina. An indication of the copper inclusion can be seen in the middle of the image. The dashed line indicates the boundary of the sample. The image also shows a strong specular reflection from the left edge which is closer to the transmitter. Placing the sample horizontally and tilting it slightly towards the illuminator reduces the effect of this specular reflection as can be seen in Figure 5.12(b).

Figure 5.12. SAF images of the balsa wood sample with copper inclusion in reflection mode; (a) sample is vertical, and (b) sample is horizontal.

As it was shown in these two reflection mode experiments, the choice of transmitting antenna and the illuminator pattern on the target affects the fidelity of the obtained image. Non-uniform illuminations and specular reflections may cause some areas of the target not to be illuminated properly. When image reconstruction techniques such as SAF is used, the ultimate solution is to have as many transmitters as the number of receivers, which in the limit converges to the array elements being used in mono-static mode whereas each element is used as a transmitter and receiver. Mono-static mode produce high quality images since the target is illuminated from various angles which greatly reduced distortions due to specular reflections [5].
5.4. APERTURE REPETITION

An important property of using arrays for imaging is that, the signal strength and the resolution is proportional to the array gain and beamwidth, which in terms is proportional to the array size. Building very large arrays may not be practical or cost effective. However, when a handheld and fast array, such as this camera is available, one can take snapshot images of various side-by-side locations, and then synthetically combine the images to produce larger, high resolution, and high fidelity images. Figure 5.13 shows the results of such experiment, whereas the scattering off an 8 mm-diameter metallic ball is measured at a distance of 170 mm. The measurements were performed by using the camera to take a snapshot of the scattered field. Then four measurements were taken on adjacent areas around the axis of the ball to synthesize a larger aperture. Only the phase of the scattered field is shown, since at these distances, the magnitude does not show any substantial variations. However, when SAF was applied, both the magnitude and phase of the measurements were used. As shown in Figure 5.13, a larger aperture provides for a better focused image, and a stronger indication of the target. Intuitively, a larger array (synthetic in this case) provide for higher gain and smaller beam width as explained earlier. Furthermore, from image reconstruction point of view, this experiment shows that capturing more of the scattered field (i.e., using larger aperture) will result in a better indication of the target.
Figure 5.13. Effect of synthetically increasing the aperture area of the retina.
6. SUMMARY AND FUTURE CONSIDERATIONS

Microwave and millimeter wave imaging has shown tremendous utility in a wide variety of applications. The ability of microwave and millimeter wave signals to penetrate low loss dielectric materials and interact with their internal structures, makes them suitable for imaging many materials when other techniques may fail to accomplish the same. Moreover, microwave energy is non-ionizing and safe to the operator and the inspected material at moderate power levels. Microwave and millimeter wave signals undergo attenuation and phase change as they travel through or get scattered off of a structure. By producing a spatial map of these electric field variations, an image of the structure may be formed. The image may be produced from an amplitude map of the electric field (power measurements), a phase map, or more comprehensively a coherent map of the electric field vector. A coherent map enables the use of various well-developed image reconstruction techniques such as synthetic aperture focusing to obtain high resolution images.

Producing images using mechanical raster scanning remains to be the most prevalent method in microwave imaging to date. The low cost and simplicity of mechanically scanned systems makes them attractive for many applications. However, these systems are typically bulky and slow. Several technologies have been investigated as potential candidate for microwave imaging with a handful achieving success for their intended applications. The most successful systems which were based on microwave technologies utilized switched arrays. However, current commercial technologies result in bulky array which hindered the development of practical 2D imaging arrays.

This dissertation presented a design for a 2D switched imaging array that utilizes modulated scattering techniques for spatial multiplexing of the signal. The design objectives were to build a compact 2D collector capable of Nyquist rate sampling of spatial electric field distribution. Furthermore, the system was designed to be compact, coherent, possessing high dynamic range, and capable of video frame rate imaging. Various aspects of the system design were optimized to achieve the design objectives.

The 2D imaging system as designed and described in this dissertation utilized PIN diode loaded resonant elliptical slots antennas as array elements. The impedance of this
slot antenna can be rapidly switched between two impedance values (i.e., a short and a matched load) by properly biasing its PIN diode. When the slot antenna is opened, it efficiently receives the electric field signal. On the other hand, when it is closed, it blocks the signal from entering the system thus enhancing the overall isolation between the array elements. These slot antennas allowed for incorporating the switching into the antennas, rather than the transmission lines. Being able to switch the array directly, eliminates the otherwise used RF switching network and results in a compact array that satisfies Nyquist sampling criteria. Furthermore, not having an RF switching network reduces the cost and size of the array. In addition, utilizing measurement concepts borrowed from modulated scattering techniques enhanced the isolation in this switching scheme.

The backbone of the array is a network of waveguides that collect the signals from the slot antennas and route them to the receivers with minimal loss. Waveguides are known low loss transmission lines at microwave and millimeter wave frequencies. Furthermore, the slot antennas are directly integrated into the waveguide walls, a design feature that further reduces the complexity of the system. Finally, further switching capabilities built into the combiner waveguides keeps the signal loss at a minimum and enhances isolation between adjacent rows of the retina.

The RF transceiver design is a merge between a heterodyne receiver and match filtering scheme of modulated scattering technique. Utilizing modulation and heterodyning, the leakage signal due to non-perfect isolation is filtered through the IF stage. Furthermore, the demodulator is designed as a matched filter. Match filtering as implemented in this design, allows for synchronized (to modulator reference) coherent estimation of the electric field's magnitude and phase. The overall detection bandwidth of the system (match filter bandwidth) was set to satisfy a compromise between the video frame rate imaging and maximum possible averaging time. The RF system was custom designed, based on surface mount MMIC technology. The RF VCOs where frequency controlled and phase locked using programmable synthesizers. The RF sources were simple, low cost, and small yet they lacked good phase noise performance. Alternative sources with better phase noise characteristics may further enhance the overall system performance and must be considered in future developments. Other consideration for
future development and enhancement include implementing the IQ demodulator and match filtering utilizing digital receivers. Such a digital receiver will provide better dynamic range, sensitivity and more flexibility in filtering. In addition, the output of a digital receiver is more compatible for transfer to PCs, or standalone processors.

An investigation on the sources of errors showed that the largest contributor to the errors is the spatial dispersion of the array response, which is mainly due to losses and reflections in the waveguide collection network and combiners. The four receivers at the four corners of the array reduced this spatial dispersion. The outputs of these four receivers were combined using a maximum ratio combiner, which is optimum in the presence of white noise. An investigation into the effect of the mutual coupling between the resonant slot antennas showed that, the mutual coupling is small especially when the slots are idly closed. This small mutual coupling was not considered to be a concern for imaging applications.

Calibrating the system required measuring various system properties. It is very difficult to measure the mutual coupling and isolation effects on the overall array response. Consequently, the effect of mutual coupling and isolation were not calibrated out. The applied calibration, corrects for the array dispersion effect. The calibration data was obtained by illuminating the retina with a known electric field distribution from a K-band open-ended waveguide. After calibrating the system, the root mean square errors were approximately 0.07 in magnitude and 13 degrees in phase, when measuring electric field pattern of the open-ended waveguide compared to the theoretical values. Aside from calibration inaccuracies, this error may be as well due to reflections from the surrounding environment. This amount of error may be acceptable for imaging applications where the overall pattern of the electric field map is of interest. For higher accuracy, a more comprehensive calibration model must be used that takes into account the mutual coupling and the limited isolation in the system.

Besides electric field mapping, other applications investigated include source/scatterer localization and imaging. The localization may be performed in the near-field or far-field of the array using synthetic aperture techniques. In the far-field the array provides a gain of −31 dB and a beamwidth of degrees that can be used to localize distant targets/sources. With slight modifications such as utilizing dual frequency radar
techniques, it is possible to use this system for ranging applications as well. Dual frequency ranging is achieved by measuring the phase difference between the signals at two closely spaced frequencies, a feat quite achievable with the current design of the camera.

As a microwave camera, this system is intended for real-time imaging applications. In the through transmission mode, the produced images possessed of high fidelity despite the large wavelength. This is mainly due to the ease of target illumination, and the high power reaching the target and the camera. In the through-transmission mode, correction for the non-uniform illuminator pattern was achieved by taking a reference image without the presence of a target. However, in the reflection mode this luxury is not easy attainable. In practice the reflection mode is more desirable, however the design of this camera does not accommodate a true mono-static reflection mode imaging. Utilizing simple transmitters, placed around the camera helped in producing images of simple structures in the reflection mode. However, specular reflection and non-uniform illumination prohibited the formation of meaningful images of complex targets in the reflection mode. A more practical one-sided system would require sufficient number of transmitters, or a suitably designed transmitting antenna, to illuminate the target uniformly from various angles. In the limit, the optimum case is to have as many transmitters as there are receivers and all must be placed on the same grid. In other words, the system must be a mono-static system with every array element transmitting and receiving. Practically, a compact array such as designed here may not have sufficient isolation to be operated as a transmitter/receiver. Next best solution is to have two separate transmitter and receiver arrays interleaved. Figure 5.12 shows the schematic of such array design. In this design, the transmitter and receiver arrays have center-to-center inter-element spacing of one wavelength ($\lambda$). The spacing between the transmitters and receivers is $\lambda/2$. By properly switching transmitter receiver pairs a virtual measurement grid with $\lambda/2$ spacing is created. This design has been successfully implemented by Sheen et.al. [5] for a 1D array using external microwave switches. However, they report practical difficulties in implementing a compact 2D array due to insufficient real-state for bulky switches. I believe with the waveguide based resonant slot design along with modulation; sufficient isolation may be obtained to successfully
implement this design. Overall, this design will create a true microwave camera that is capable of active imaging in reflection mode. Understandably, this camera will have to be operated from distances at least larger than the far-field distance of each array element or larger than the spacing of the transmitter receiver pairs due to the offset in the transmitter and receiver arrays' position. For the resonant slot antennas, the far-field distance is on the order of one wavelength.

Aside from the transmitter issue, future developments of the microwave camera must address increasing the bandwidth of the system to enable 3D imaging. Increasing the frequency of operation will provide for a wider variety of imaging applications and higher spatial resolution. Overall, the frequency and bandwidth will be application dependant. The aperture size will also be application dependant. Yet real-time handheld

Figure 6.1. Proposed design for transmitter receiver interleaved arrays.
camera with a small aperture comparable to the design in this dissertation is adequate for imaging many large and small structures alike.
BIBLIOGRAPHY


VITA

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