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ANTENNA-COUPLED INFRARED FOCAL PLANE ARRAY

by

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M.S. University of Central Florida, 2000

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ABSTRACT

In this dissertation a new type of infrared focal plane array (IR FPA) was investigated, consisting of antenna-coupled microbolometers fabricated using electron-beam lithography. Four different antenna designs were experimentally demonstrated at 10-micron wavelength: dipole, bowtie, square-spiral, and log-periodic. The main differences between these antenna types were their bandwidth, collection area, angular reception pattern, and polarization. To provide pixel collection areas commensurate with typical IR FPA requirements, two configurations were investigated: a two-dimensional serpentine interconnection of individual IR antennas, and a Fresnel-zone-plate (FZP) coupled to a single-element antenna. Optimum spacing conditions for the two-dimensional interconnect were developed. Increased sensitivity was demonstrated using a FZP-coupled design. In general, it was found that the configuration of the antenna substrate material was critical for optimization of sensitivity. The best results were obtained using thin membranes of silicon nitride to enhance the thermal isolation of the antenna-coupled bolometers. In addition, choice of the bolometer material was also important, with the best results obtained using vanadium oxide. Using optimum choices for all parameters, normalized sensitivity (D^*) values in the range of mid 10^8 [cm $\sqrt{\text{Hz}}/\text{W}$] were demonstrated for antenna-coupled IR sensors, and directions for further improvements were

identified. Successful integration of antenna-coupled pixels with commercial readout integrated circuits was also demonstrated.

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CHAPTER 1

INTRODUCTION

1.1 Infrared Detectors

The discovery of infrared radiation occurred in 1800 when Sir William Herschel essentially repeated Newton's famous prism experiment and detected heat in a region where there was no visible radiation. The Planck radiation formula was derived in 1900 and quantitatively predicted the amount of energy radiated from a blackbody as a function of temperature and wavelength.

The thermometer was the first of three kinds of detectors that were to dominate the infrared detector field until World War I. The other two were the thermocouple, demonstrated by Seebeck after he discovered the thermoelectric effect in 1821, and the bolometer discovered by Langley in 1881. The years preceding and during World War II saw the origins of modern infrared imaging technology.

During the 1950's and 1960's infrared sensors were built using single-element cooled lead-salt detectors primarily for missile seeker applications. At the same time, rapid advances were being made in narrow bandgap semiconductors that would later

prove useful in extending wavelength capabilities and improving sensitivity. These developments paved the way for the highly successful forward-looking infrared (FLIR) airborne systems developed in the 1970's.

As photolithography became available in the early 1960's, it was applied to make infrared detector arrays. In the late 1960's and early 1970's, first generation linear arrays of intrinsic HgCdTe photoconductive detectors were developed. In these detectors an electrical contact for each element of a multielement array is brought off the cryogenically-cooled focal plane to the outside, where there is one electronic channel at ambient temperature for each detector element.

In 1970, the charge-coupled device (CCD) was invented and was immediately recognized as a means of obtaining a solid state imaging system that could replace vacuum tube imaging devices such as vidicons and plumbicons. Silicon CCD imaging devices operating in the visible spectrum have been intensively developed over the past 20 years and advances in VLSI technology have made feasible 600×600 pixel arrays for commercial applications. Specialized scientific CCD image sensors of 2048×2048 are currently being fabricated and even larger arrays being planned. Throughout this period parallel efforts in the infrared community have employed similar techniques to obtain infrared focal plane arrays (IRFPA's) which are integrated two-dimensional arrays of detectors on the focal plane with multiplexed readouts. Interest has centered mainly of the wavelengths of the two atmospheric windows from $3\text{-}5\text{ }\mu\text{m}$ and $8\text{-}14\text{ }\mu\text{m}$ (Fig.1.1), though in recent years there has been increasing interest in longer

wavelengths stimulated by space applications[2].

Figure 1.1. Transmission of the atmosphere for a 6000 ft horizontal path at sea level containing 17 mm of precipitable water. From [3].

IR detectors can be classified as photon detectors and thermal detectors [3]. Photon detectors absorb radiation by interactions with electrons, either bound to lattice atoms or impurity atoms, or with free electrons. The observed electrical output signal results from the changed electronic energy distribution. Photon detectors show a selective wavelength dependence of the response per unit incident radiation power (Fig.1.2). They exhibit good signal-to-noise performance and a very fast response. But to achieve this, the photon detectors require cryogenic cooling to prevent the thermal generation of charge carriers which will result in shot noise on the dark cur-

rent. The cooling mechanism which is bulky, heavy and expensive is the main obstacle to the more widespread use of IR systems based on semiconductor photodetectors.

Figure 1.2. Spectral Responsivity for a photon and thermal detector. From [3].

On the other hand, with thermal detectors the incident radiation is absorbed to change the temperature of the material (Fig.1.3), and the resultant change in some physical property is used to generate an electrical output. The signal does not depend upon the photon nature of the incident radiation. Thus thermal effects are generally wavelength independent (Fig.1.2), the signal depends upon the radiant power (or its range of change) but not upon its spectral content. Thermal detectors are typically operated at room temperature, they are usually characterized by modest sensitivity and slow response (because heating and cooling of a detector element is a relatively slow process) but they are cheap, easy to use and do not require cooling to suppress

dark current. They have found widespread use in low cost applications, which do not require high performance and speed.

Figure 1.3. Thermal detector mounted via lags to heat sink. From [3].

There are four principal categories of thermal detectors: resistive bolometers, pyroelectric detectors, ferroelectric bolometers and thermoelectric detectors[4].

- **Resistive bolometers** are temperature sensitive resistors, made from metals, semiconductors or superconductors.
- **Pyroelectric detectors** made from ferroelectric materials exhibit a polarization which depends upon the time rate change of the detector temperature.

- **Ferroelectric bolometers** are similar to pyroelectric detectors, but an electric field is applied to enhance the output signal. Because pyroelectric detectors and ferroelectric bolometers respond to the time rate of change of their temperature, they require the incident radiation to be temporally modulated or “chopped”.
- **Thermoelectric detectors** are junctions of dissimilar materials which exhibit the Seebeck effect. They are radiation-sensitive thermocouples. When several junctions are connected in series to enhance the signal voltage, the device is known as a “thermopile”. Like pyroelectric detectors, thermoelectric detectors do not require an electrical bias.

1.2 Characterization of Infrared Detectors

For detectors whose output consists of an electrical signal that is proportional to the radiant signal power, certain figures of merit have been defined.

1.2.1 Signal-to-noise ratio (SNR)

Signal-to-noise ratio, often written S/N or SNR , is a measure of signal strength relative to background noise and is given by

$$SNR = \frac{V_s}{V_n}, \quad (1.1)$$

where V_s is the signal voltage and V_n is the rms noise voltage. The three main types of noise mechanisms are thermal noise, low-frequency (1/f) noise, and shot noise.

1.2.2 Responsivity

The responsivity of an infrared detector is defined as the ratio of the rms value of the fundamental component of the electrical output signal of the detector to the rms value of the fundamental component of the input radiation power.

The voltage spectral responsivity is given by

$$\mathfrak{R}_v(\lambda) = \frac{V_s}{P_{in}(\lambda)}, \quad (1.2)$$

where V_s is the signal voltage due to $P_{in}(\lambda)$, and $P_{in}(\lambda)$ is the radiant incident power at a specific wavelength expressed in watts. If the signal is a current then the responsivity is expressed in amperes/watt.

An alternative to the above monochromatic quality is the blackbody responsivity which is defined by the equation

$$\mathfrak{R}_v = \frac{V_s}{\int_0^\infty P_{in}(\lambda) d\lambda}, \quad (1.3)$$

where the incident radiant power is the integral over all wavelengths of the spectral power distribution from a blackbody. The responsivity is usually a function of the bias voltage V_{bias} , the modulation frequency of the incident radiation f and the wavelength

$\lambda[5]$.

1.2.3 Noise Equivalent Power (NEP)

The noise equivalent power (NEP) is the incident power on the detector generating a signal output equal to the rms noise output. In other words, the NEP is the input power level that will produce a signal-to-noise ratio (SNR) of 1. It can be written in terms of responsivity:

$$NEP = \frac{V_n}{\mathfrak{R}_v}, \quad (1.4)$$

the unit of NEP is watt. As the responsivity was a function of wavelength and frequency, so is NEP . Also the NEP can be either spectral or blackbody, depending on the type of incident radiation.

The disadvantage of using NEP to describe detector performance is that it depends on the square root of the area of the detector (A_d) and the square root of the measurement bandwidth (Δf)[6].

1.2.4 Detectivity

The detectivity D is the reciprocal of NEP :

$$D = \frac{1}{NEP}. \quad (1.5)$$

It was found by Jones[7], that for many detectors the NEP is proportional to the square root of the detector signal that is proportional to the detector area. It means that both NEP and detectivity are functions of electrical bandwidth and detector area, so a normalized detectivity D^* (or D-star) suggested by Jones [7] [8] is defined as

$$D^* = D\sqrt{A_d\Delta f} = \frac{\sqrt{A_d\Delta f}}{NEP}, \quad (1.6)$$

the importance of D^* comes from the fact that it permits comparison of detectors of the same type but having different areas and different measurement bandwidths. D^* is defined as the rms signal-to-noise ratio in a 1 Hz bandwidth per unit rms incident radiant power per square root of detector area. D^* is expressed in unit $\frac{\text{cm}\cdot\sqrt{\text{Hz}}}{\text{W}}$, which is also known as Jones.

1.3 Bolometers

The bolometer is a resistive element constructed from a material with a very small thermal capacity and large temperature coefficient so that the absorbed radiation produces a large change in resistance. The device is operated by passing a bias current through the detector and monitoring the output voltage. In the case of bolometers, radiant power produces heat within the material which in turn produces a change in resistance, there is no direct photon-electron interaction.

The first bolometer was designed in 1880 by American astronomer S. P. Langley for

solar observations. This first bolometer used a blackened platinum absorber element and a simple Wheatstone-bridge to measure the output. Although other thermal devices have been developed since that time, the bolometer remains one of the most commonly used infrared detectors [3].

The analysis of the performance of any thermal detector begins with a heat flow equation. For a bolometric detector with an area A_d and an optical absorption coefficient η , a heat capacity C (J/K) and thermal conductivity of the main heat loss mechanisms G in W/K, the heat flow equation would be

$$C \frac{d(\Delta T)}{dt} + G(\Delta T) = \eta A_d W_{in} e^{j\omega t}, \quad (1.7)$$

where ΔT is the change in temperature due to the modulated incoming radiation, W_{in} (W/cm²) is the radiation intensity, and ω is the angular modulation frequency ($\omega = 2\pi f$, where f is the linear frequency).

The solution to Eq. 1.7 is

$$\Delta T = \frac{\eta A_d W_{in}}{G \sqrt{1 + \omega^2 \tau^2}}, \quad (1.8)$$

where τ is the thermal time constant in seconds of the detector and is given by

$$\tau = \frac{C}{G}. \quad (1.9)$$

The analysis of resistive bolometers assumes that the temperature increase ΔT

of the bolometer due to the absorption of IR radiation is small enough so that the resistance change ΔR is linear with ΔT , so that it is possible to express the change in resistance in terms of α , the temperature coefficient of resistance (TCR). Therefore

$$\Delta R = \alpha R \Delta T \quad (1.10)$$

where

$$\alpha = \frac{1}{R} \frac{dR}{dT}. \quad (1.11)$$

The temperature coefficient of resistance can be either positive or negative. For metals at room temperature it is positive, that is, the resistance increases with increasing temperature. For semiconductors at room temperature it is usually negative.

The signal voltage due to the radiation incident on the bolometer can be obtained from Ohm's law and Eq. (1.10) as

$$V_s = i_{bias} \Delta R = i_{bias} \alpha R \Delta T \quad (1.12)$$

From Eqs. (1.12) and (1.2) and assuming an absorbed power P_{in} given by $P_{in} = A_d W_{in}$ (where A_d is the area of the detector and W_{in} is the radiation intensity), then we can write the bolometer responsivity as

$$\mathfrak{R}_v = \frac{i_{bias} R \alpha \eta}{G \sqrt{1 + \omega^2 \tau^2}}. \quad (1.13)$$

Equation (1.13) shows that the responsivity is directly proportional to the temperature coefficient of resistance (α) and inversely proportional to the thermal conductivity associated with the heat conduction paths out of the bolometer (G). In the case of uncooled IR resistive bolometers, values of G can range over several orders of magnitude whereas the range of possible values of α is far less, therefore thermal isolation has a bigger impact on responsivity than the choice of bolometric material.

We can also use an electrical analogy to describe the behavior of bolometers, the change in temperature ΔT can be written as $\Delta T = P_{in} |Z_{th}|$, where the change in temperature ΔT , absorbed power P_{in} , and magnitude of the thermal impedance $|Z_{th}|$ are analogous to voltage, current, and resistance respectively. Using this electrical analogy on Eq. (1.12) and knowing that $V_{bias} = i_{bias} R$, we can write the voltage responsivity as

$$\mathfrak{R}_v = V_{bias} \alpha |Z_{th}|, \quad (1.14)$$

where the thermal impedance $|Z_{th}|$ is inversely proportional to the thermal conductivity. A perfectly isolated bolometer would have a very high thermal impedance which would translate into a high responsivity device. Following the electrical analogy of heat transfer the thermal impedance can be modeled using a thermal resistance R_{th} and a thermal capacitance C_{th} [9] as

$$|Z_{th}| = \frac{R_{th}}{\sqrt{1 + \omega^2 R_{th}^2 C_{th}^2}}, \quad (1.15)$$

where

$$\tau = R_{th}C_{th}, \quad (1.16)$$

is the thermal time constant. The thermal resistance is inversely proportional to the thermal conductivity out of the bolometer and the thermal capacitance is directly proportional to the mass of the bolometer[10]. If we plot the absolute value of the thermal impedance $|Z_{th}|$ as a function of the thermal resistance R_{th} and thermal capacitance C_{th} (Figure 1.4) we can see how the highest values are obtained for low values of thermal capacitance and high values of thermal resistance, therefore by increasing the thermal isolation and decreasing the size of the bolometer we can optimize its responsivity. A smaller bolometer will also have a smaller time constant which can be useful for high frame-rate applications. Figure 1.5 shows a bolometer used in commercial infrared imaging systems, the size of this bolometer is around $50 \times 50 \mu\text{m}^2$ which is a typical pixel area. A bolometer this big will have a typical time constant in the order of 15 ms which is slow for certain applications.

The problem of reducing the size of a bolometer is that its collection area also gets reduced since bolometers use their physical size to collect radiation. A way to increase the collection area of a small bolometer is to couple an antenna designed to resonate at the desired wavelength, this way we can have fast detectors without sacrificing collection area.

Figure 1.4. Magnitude of the thermal impedance as a function of the thermal resistance and thermal capacitance.

Figure 1.5. Bolometer used in a commercial infrared imaging system.

1.4 Antenna Theory

Microstrip and printed circuit antennas have gained prominence as viable antenna elements and arrays. These types of antennas have several advantages such as low profile, low cost, light weight, conformity to surface, mass production, variable-frequency operation possibilities and compatibility with integrated circuit technology. Their limitations are low gain and narrow bandwidth. The main difference between microstrip and printed circuit antennas is that the first ones have a ground plane which affects the radiation characteristics of the antenna significantly, this ground plane can be modeled using image theory, that is introducing virtual sources that will account for the reflections that occur at the ground plane.

A free-space antenna radiates equally on both sides because of its symmetry, the

presence of a dielectric substrate will break the symmetry and change the current distribution and wave velocity of the antenna. One of the main features of antennas on dielectrics is that they tend to radiate more power into the dielectric in the ratio $\varepsilon_s^{3/2} : 1$, where ε_s is the relative electrical permittivity of the substrate. Also waves propagate differently along metals at a dielectric interface, the waves tend to propagate at a velocity that is intermediate between the velocity of waves in the air and the velocity of waves in the dielectric. For antennas on a substrate in the millimeter-wave range, waves propagate with a velocity close to that characteristic of a material with a dielectric constant equal to the mean of the two dielectric constants, this is a quasi-static approximation to the effective permittivity [11] and is only valid for frequencies below a few GHz, above these frequencies the frequency dispersion of the effective permittivity must be taken into account. Hasnain et al. [12] derived an analytical expression for the dispersion of the effective permittivity which is given by

$$\sqrt{\varepsilon_{eff}} = \sqrt{\varepsilon_q} + \left(\frac{\sqrt{\varepsilon_s} - \sqrt{\varepsilon_q}}{1 + a\left(\frac{f}{f_{TE}}\right)^{-b}} \right), \quad (1.17)$$

where ε_q is the quasi-static value of the permittivity given by $\varepsilon_q = \frac{\varepsilon_s + 1}{2}$, f is the frequency of the wave in Hz and f_{TE} is the cut-off frequency for the lowest order TE mode, which depends on the substrate thickness d and is given by

$$f_{TE} = \frac{c}{4d\sqrt{\varepsilon_s - 1}}, \quad (1.18)$$

where c is the velocity of light in vacuum. The parameter a depends on the configuration and the dimensions of the transmission line, for the specific case illustrated in [12] $a \approx 51$, b was found to be always around 1.8 independent of the dimensions. Figure 1.6 shows the effective permittivity as a function of substrate thickness for waves at $10.6 \mu\text{m}$ wavelength (28.3 THz), we can see how for a substrate thickness below $1 \mu\text{m}$ the effective permittivity is equal to the quasi-static value, and for a thickness larger than $100 \mu\text{m}$ the effective permittivity is equal to the substrate thickness.

In this dissertation printed antennas were fabricated on thick substrates and on a thin silicon nitride (Si_3N_4) membrane for detection of $10.6 \mu\text{m}$ radiation. The Si_3N_4 membranes used were 400 nm thick in which case the quasi-static value for the effective permittivity applies, substrates with thicknesses higher than $200 \mu\text{m}$ were also used, in this case according to Figure 1.6 the effective permittivity is equal to the permittivity of the substrate.

The most basic properties of an antenna are its radiation pattern, gain, impedance, and polarization. These properties are identical for linear passive antennas used either as a transmitter or receiver due to the reciprocity theorem. A radiation pattern is a graphical representation of the far-field properties of an antenna. It can be measured by rotating the antenna and plotting the response as a function of angular coordinates. Power gain is defined as 4π times the ratio of the radiation or detection intensity to the net power accepted by the antenna. The input impedance of an antenna is the impedance presented by the antenna at its terminals and is composed

of real and imaginary parts:

$$Z_{in} = R_{in} + jX_{in}. \quad (1.19)$$

Figure 1.6. Effective permittivity as a function of substrate thickness of a coplanar waveguide at 28.3 THz ($10.6 \mu m$)

The input resistance, R_{in} represents dissipation, while the input reactance X_{in} represent power stored in the near field of the antenna. To maximize the power transfer from a transmitter to an antenna the antenna impedance should be a con-

jugate match to the impedance of the load, meaning that input resistances should be the same and the reactances should have equal magnitude but opposite signs. Antennas that are electrically small (much smaller than a wavelength) have a large input reactance, in addition to a small radiation resistance. The polarization of a transmitting antenna is the polarization of the wave radiated by the antenna. In the case of receiving antennas the polarization of the antenna is the polarization of the wave that will maximize its response.

With each antenna we can associate a number of equivalent areas. These are used to describe the power characteristics of the antenna when a wave impinges on it. One of these equivalent areas is the effective collection area (aperture) A_{eff} which is defined as the ratio of the available power at the terminals of a receiving antenna (P_{in}) to the power flux density (irradiance) of a plane wave incident on the antenna (W_{in})

$$A_{eff} = \frac{P_{in}}{W_{in}}, \quad (1.20)$$

the wave being polarization matched to the antenna[1]. The effective collection area is related to the maximum directivity of an antenna by the following equation:

$$A_{eff} = \frac{\lambda^2 D_{max}}{4\pi}, \quad (1.21)$$

where the maximum directivity of the antenna is defined as the ratio of the maximum

radiation intensity to the radiation intensity averaged over all directions.

1.5 Antenna Coupled Microbolometers

The work by Matarrese and Evenson[13] showed that the polarization dependence observed in whisker diodes was due to the coupling of radiation at wavelengths as short as $3\text{ }\mu\text{m}$ through the whiskers, which acted like long-wire antennas. These findings made clear that the combination of antenna and junction can be used as an infrared radiation detector. Twu and Schwarz investigated the source impedance and the efficiency of the cat-whisker receiving antenna at $10.6\text{ }\mu\text{m}$ [14] and they also described the radiation patterns of these antennas at $10.6\text{ }\mu\text{m}$, which agreed with low frequency antenna theory[15].

Since then, infrared antennas have been used to collect infrared energy and apply it in the form of a voltage at the infrared frequency to metal-insulator-metal[16] or metal-semiconductor junctions[17]. The effect of the infrared frequency voltage on the junction is to produce a change in the dc voltage or current. This change is due to the rectification of the current at the infrared frequency by the nonlinearity of the junction[18].

In 1977 Schwarz and Ulrich performed a theoretical study on the use of antennas to couple radiation to bolometers[19], and they showed that for any detector characterized by a specific detectivity D^* , the noise equivalent power (NEP) can be reduced until the background blackbody fluctuation limit is reached. This improvement should

be on the order of λ/b , where b^2 is the area of the detector[19].

Because the cat-whisker-antennas are large, mechanically unstable, and unsuitable for replication into arrays Hwang et al.[20] were the first ones to fabricate a planar V antenna coupled to a bismuth bolometric detector for detection of $119\ \mu\text{m}$ radiation. The antenna was made of 65 nm of silver on a quartz substrate, was $650\ \mu\text{m}$ long (10 dielectric wavelengths) and the bismuth bolometer was $5 \times 4\ \mu\text{m}$ and 55 nm thick. This device was patterned using photolithography and liftoff. Radiation patterns measured on this device matched the expected patterns predicted by antenna theory.

Hwang *et al.*[21] were also the first ones to use the word “microbolometer” to describe bolometric detectors smaller than the wavelength, in their paper they discuss the frequency response of bismuth microbolometers and its dependence with size of the detector and substrate thermal conductivity. They found that the speed of the detector increased with smaller sized devices and higher thermal conductivity substrates. Substrates with high thermal conductivity reduced the response of the microbolometer.

An improvement in performance was suggested in [21] by fabricating microbolometers on low thermal conductivity substrates, which increases the thermal impedance of the device and this is proportional to its responsivity. Neikirk and Rutledge fabricated an air-bridge microbolometer using a photoresist bridge technique [22], which reduced the thermal conductivity of the device by a factor of 5, and gave a factor of 4 better sensitivity over the best previously reported bolometer.

Since then many different types of planar antennas have been coupled to infrared detectors: dipole [23], bow-tie [16], spiral [24], log-periodic [25], slot [26], and microstrip-patch [27], because of advantages in increased directivity, along with the possibility of polarization and wavelength selection obtained by using antennas. Figure 1.7 shows two different types of antennas designed to detect $10.6\text{ }\mu\text{m}$ radiation coupled to microbolometers, a log-periodic and a square-spiral antenna which were patterned using direct-write electron beam lithography and liftoff.

Figure 1.7. Log-periodic and spiral antenna coupled to a Nb microbolometer.

1.6 Focal Plane Arrays

The objective of focal plane array (FPA) technology is to satisfy the requirement for very large detector arrays by means of the integrated circuit (IC) approach. This

requirement is due to the fact that high density detector configurations lead to higher image resolution as well as greater system sensitivity [2]. The invention and development of the charged-coupled-device (CCD) was the technological breakthrough that initially made this possible. By the mid-1970's a number of concepts for IR-CCDs had been explored. Prior to the CCD, the only alternative for large arrays was to configure each detector connected to a single wire (and probably an individual preamplifier) which would all need to be packaged in a small dewar. For a large number of detectors this would obviously create an unmanageable maze of wires and processing electronics, and which would also require an unacceptable large cooler because of the thermal conductance of the wiring harness.

1.6.1 Focal Plane Array Architectures

The principal FPA functions are: photon detection, detector readout, signal processing and output multiplexing. In general an FPA may be classified according to its architecture as hybrid or monolithic (Fig. 1.8) [3]. In hybrid FPAs detectors and multiplexers are fabricated on different substrates and mated with each other by flip-chip bonding (Fig. 1.9(a)) or loophole interconnection (Fig. 1.9(b)). In this case the detector material and multiplexer can be optimized independently. Other advantages of the hybrid FPAs are near-100% fill factor and increased signal-processing area on the multiplexer chip.

By using flip-chip bonding, the detector array is typically connected by pressure

Figure 1.8. IRFPA ARCHITECTURES. For hybrid arrays: (a) flip-chip; (b) Z-technology. For pseudo-monolithic arrays: (c) XY-addressable. For monolithic arrays: (d) all-silicon; (e) heteroepitaxy-on-silicon; (f) non-silicon (e.g., HgCdTe CCD). From [3].

contacts via indium bumps to the silicon multiplex pads. The detector array can be illuminated from either the frontside (with the photons passing through the transparent silicon multiplexer) or backside (with photons passing through the transparent detector array substrate). In general the latter approach is most advantageous as the multiplexer will typically have areas of metallizations and other opaque regions which can reduce the effective optical area of the structure.

With loophole interconnection, the detector and the multiplexer chips are glued together to form a single chip before the detector fabrication. Then the photovoltaic detector is formed by ion implantation and loopholes are drilled by ion-milling. The loophole interconnection technology offers more stable mechanical and thermal features than that of the flip-chip hybrid architecture.

Figure 1.9. Hybrid IRFPA interconnect techniques between a detector array and silicon multiplexer: (a) indium bump technique, (b) loophole technique. From [3].

In the monolithic approach, some of the multiplexing is done in the detector material itself rather than in an external readout circuit. The electronics in charge of multiplexing and processing the detected signal is usually called the readout integrated circuit or ROIC.

1.6.2 Readout Integrated Circuits (ROICs)

The ROIC reads the photo-current from each pixel of the detector array and outputs the signal in a desired sequence that is used to form a two-dimensional image. A wide variety of submicron CMOS-based multiplexers have been designed, enabling fabrication of high performance advanced focal plane arrays with ultra-low noise suitable for a broad range of applications. The advantage of CMOS are that existing foundries which fabricate Application Specific Integrated Circuits (ASICs) can be readily used by adopting their design rules.

After the incoming photon flux is converted into a signal by the detector, it is coupled into the readout via a detector interface circuit. The input circuit is the most important part of the ROIC because it interfaces directly to the detector, this input circuitry generally requires that the impedance of the detector be in the order of $10 - 100 \text{ k}\Omega$ to reduce the power dissipation of the whole circuit and also to make the ROIC detector-noise-limited.

CHAPTER 2

DEVICE FABRICATION

The fabrication process of infrared detectors for imaging applications should be compatible with modern IC fabrication technology so that monolithic integration into commercially available readout integrated circuits (ROIC's) would be possible. Integration of IR detectors into an FPA would make an IRFPA which is the basic building block of an infrared imaging system.

2.1 Lithography

Lithography is the cornerstone of modern integrated circuit (IC) manufacturing. The ability to print patterns with submicron features and to position those patterns on a silicon substrate with better than $0.1\ \mu\text{m}$ precision is what makes integrated circuits possible. Figure 2.1 shows a schematic of a basic lithographic exposure system. The lithographic process starts by spinning a light-sensitive photoresist onto a wafer forming a thin layer on the surface. The resist is then selectively exposed by shining light through a mask (reticle) which contains the pattern information for the particular layer being fabricated, this exposure process modifies the resist making it more

(positive resist) or less (negative resist) soluble to a developer. After the development process resist will remain in some areas and be removed from some other areas resembling the mask pattern. The process of transferring the pattern to the wafer can be done by removing areas (etching), adding materials (deposition) or modifying the characteristics of the wafer (implantation or diffusion), the pattern transfer takes place by having some areas protected with photoresist and other areas exposed to these processes.

Figure 2.1. Schematic of a simple lithographic exposure system.

The photoresists used in IC fabrication normally have three components: a resin or base material, a photoactive compound (PAC), and a solvent that controls the mechanical properties, such as viscosity. In positive resists, the PAC acts as an inhibitor before exposure, slowing the rate at which the resist will dissolve when placed in a developing solution. Upon exposure to light, a chemical process occurs by which the inhibitor becomes a sensitizer, increasing the dissolution rate of the resist. The performance of a resist is measured in sensitivity and resolution, sensitivity refers to the amount of luminous energy (usually measured in mJ/cm^2) necessary to create the chemical change described above. Resolution refers to the smallest feature that can be reproduced in a photoresist. The most popular resists are referred to as DQNs, corresponding to the photoactive compound based on diazoquinones (DQs) and the matrix material novolac (N) which dissolves easily in an aqueous solution. Solvents are added to the resin to adjust the viscosity, which is an important parameter for spin-coating the resist to the wafer. Most of the solvent is evaporated from the resist before the exposure is done and so plays little part in the actual photochemistry. One of the great advantages of DQN resists is that they have very good resolution since the unexposed areas are essentially unchanged by the developer because it does not penetrate the resist. Another advantage is that novolac is fairly resistant to chemical attack, being a good mask for subsequent plasma etching. Negative photoresists swell during the development phase, broadening the linewidth. An after-develop bake will typically cause the lines to return to their original dimension, but this swelling and

shrinking process often causes the lines to be distorted. As a result, negative resists are generally not suited to features less than $2.0\ \mu\text{m}$.

The most common type of optical source for photolithography is the high-pressure mercury-xenon arc lamp. Arc lamps are the brightest incoherent sources available, they emit light from a compact region a few millimeters in diameter, and have total emissions from about 100 to 2000 W. A large fraction of the total power emerges as infrared and visible light energy, which must be removed from the optical path with multilayer dielectric filters. The useful portion of the spectrum consists of several bright emission lines in the near ultraviolet and a continuous emission spectrum in the deep ultraviolet. Because of their optical dispersion, refractive lithographic lenses can use only a single emission line, either the *g* line at 425.83 nm, the *h* line at 404.65 nm, or the *i* line at 365.48 nm. Each of these lines contains less than 2% of the total power of the arc lamp.

Because of diffraction effects there is a resolution limit in lithographic projection systems given by Rayleigh's criteria,

$$D = k_1 \frac{\lambda}{NA} \quad (2.1)$$

where D is the minimum dimension that can be printed, λ is the exposure wavelength, and NA is the numerical aperture of the optical system. The proportionality constant k_1 is a dimensionless number in an approximate range from 0.6 to 0.8. The resolution of optical lithography using mercury arc lamps is about $0.5\ \mu\text{m}$.

2.2 E-beam Lithography

In chapter 1 we saw that decreasing the size of a bolometer will optimize its responsivity, electron beam lithography (EBL) is a specialized technique for creating the extremely fine patterns required for antenna-coupled infrared detectors. It is also used to generate masks for optical lithography, and for low-volume manufacture of ultra-small features for high-performance devices[28].

Derived from the early scanning electron microscopes, the EBL technique consists of scanning a beam of electrons across a surface covered with a resist film sensitive to those electrons, thus depositing energy in the desired pattern in the resist film. The process of forming the beam of electrons and scanning it across a surface is very similar to what happens inside a common television or cathode ray tube (CRT) display, but EBL typically has three orders of magnitude better resolution. The main attributes of the technology are:

- It is capable of very high resolution;
- It is a flexible technique that can work with a variety of materials and an almost infinite number of patterns;
- It is slow, being one or more orders of magnitude slower than optical lithography;
- and
- The machinery required is expensive and complicated.

Figure 2.2 shows a block diagram of a typical electron beam lithography tool. The column is responsible for forming and controlling the electron beam.

Figure 2.2. Block diagram showing the major components of a typical electron beam lithography system. From [29].

Underneath the column is a chamber containing a stage for moving the sample around and facilities for loading and unloading it. Associated with the chamber is a vacuum system needed to maintain an appropriate vacuum level throughout the machine and also during the load and unload cycles. A set of control electronics

supplies power and signals to the various parts of the machine. Finally, the system is controlled by a computer, which may be anything from a personal computer to a mainframe. The computer handles such diverse functions as setting up an exposure job, loading and unloading the sample, aligning and focusing the electron beam, and sending pattern data to the pattern generator. The part of the computer and electronics used to handle pattern data is sometimes referred to as the data path[29].

One of the major areas of concern for electron beam lithography is pattern distortion due to proximity effects. This refers to the tendency of scattered electrons to expose nearby areas that may not be intended for exposure. There are several techniques to minimize the proximity effect, the most popular one is dose correction, where the dose is varied in such a way as to deposit the same energy density in all exposed regions of the pattern. Another way of correcting the proximity effect is shape correction, where the width of lines are decreased and spacings between them are increased to compensate for the widening of features due to the proximity effect. The magnitude of these corrections are obtained empirically from test exposures. For this reason this technique is generally applied only to simple, repetitive patterns or else this process would turn to be too time consuming. Also the use of beam energies much greater than 20 keV (e.g., 50 keV) reduces the proximity effect, because at higher energies the electrons are scattered into a considerably larger region giving rise to a lower concentration of scattered electrons in the pattern region.

The fabrication of the antenna-coupled infrared detectors described in this study

was done at the Cornell Nanofabrication Facility (Ithaca, NY), using a Cambridge EBMF 10.5 Electron Beam Lithography System at a 30kV accelerating voltage, which is capable of a resolution of about 150 nm.

2.3 E-Beam Resist Processing

Electron beam resists are the recording and transfer media for e-beam lithography. The usual resists are polymers dissolved in a liquid solvent. Liquid resist is dropped onto the substrate, which is then spun at 1000 to 6000 rpm to form a coating. After baking out the casting solvent, electron exposure modifies the resist, leaving it either more soluble (positive) or less soluble (negative) in developer. This pattern is transferred to the substrate either through an etching process (plasma or wet chemical) or by “liftoff” of material. In the liftoff process a material is evaporated from a small source onto the substrate and resist, as shown in Figure 2.3. The resist is washed away in a solvent such as acetone. An undercut resist profile aids in the liftoff process by providing a clean separation of the material. As a rule of thumb the thickness of the resist should be at least $3\times$ the thickness of the metallic film to get the best results using liftoff.

Polymethyl methacrylate (PMMA) was one of the first materials developed for e-beam lithography. It is the standard positive e-beam resist and remains one of the highest-resolution resists available. PMMA is usually purchased in two high molecular weight forms (496 K or 950 K) in a casting solvent such as chlorobenzene

Figure 2.3. Liftoff Process. (a) $PMMA$ is spun on top of copolymer $P(MMA-co-MAA)$ and developed in $MIBK:IPA$ giving a slight undercut. (b) Metal is evaporated and resist is removed using a liquid solvent, transferring the pattern to the substrate. From [29].

or anisole. PMMA is spun onto the substrate and baked at 170 to 200 °C for 1 to 2 hours. Electron beam exposure breaks the polymer into fragments that are dissolved preferentially by a developer such as methyl isobutyl ketone (MIBK). MIBK alone is too strong a developer and removes some of the unexposed resist. Therefore, the developer is usually diluted by mixing in a weaker developer such as isopropanol (IPA). A mixture of 1 part MIBK to 3 parts IPA produces very high contrast but low sensitivity. By making the developer stronger, say, 1:1 MIBK:IPA, the sensitivity is improved significantly with only a small loss of contrast.

The critical dose in PMMA scales with electron acceleration voltage, being roughly twice at 50 kV than at 25 kV exposures. Fortunately, electron guns are proportionally brighter at higher energies, providing twice the current in the same spot size at 50 kV. When using 50 kV electrons and 1:3 MIBK:IPA developer, the critical dose is around $350 \frac{\mu\text{C}}{\text{cm}^2}$ [29].

When exposed to more than 10 times the optimal positive dose, PMMA will crosslink, forming a negative resist. It is simple to see this effect after having exposed one spot for an extended time (for instance, when focusing on a mark). The center of the spot will be crosslinked, leaving resist on the substrate, while the surrounding area is exposed positively and is washed away. In its positive mode, PMMA has an intrinsic resolution of less than 10 nm. In negative mode, the resolution is about 50 nm. By exposing PMMA (or any resist) on a thin membrane, the exposure due to secondary electrons can be greatly reduced and the process latitude thereby in-

creased. PMMA has poor resistance to plasma etching, compared to novolac-based photoresists. Nevertheless, it has been used successfully as a mask for the etching of silicon nitride and silicon dioxide, with 1:1 etch selectivity[29].

A larger undercut resist profile is often needed for lifting off thicker metal layers. One of the first bilayer systems was developed by Hatzakis[30]. In this technique a high sensitivity copolymer of methyl methacrylate and methacrylic acid [P(MMA-MAA)] is spun on top of PMMA. A more common use of P(MMA-MAA) is as the bottom layer, with PMMA on top. In this case the higher speed of the copolymer is traded for the higher resolution of PMMA. The undercut of this process is so large that it can be used to form free-standing bridges of PMMA (Figure 2.4).

2.4 Thin Film Deposition Techniques

2.4.1 Evaporation

The metal layers for all of the early semiconductor technologies were deposited by evaporation, which has been displaced by sputtering in most silicon technologies for two reasons. The first is the ability to cover surface topology, also called the “step coverage.” Evaporated films have very poor ability to cover height discontinuities, often becoming discontinuous on the vertical walls. It is also difficult to produce well controlled alloys by evaporation. In some cases, the poor step coverage of evaporation can be used to advantage. Rather than depositing and etching metal layers, the film is deposited on top of a patterned photoresist layer. The films naturally tend to break

Figure 2.4. Resist bridge pattern used to fabricate airbridge microbolometers. From [22].

at the edges of the resist so that when the resist is subsequently dissolved the layer on top of the resist is easily lifted-off (Figure 2.3)[31].

In an evaporator the wafers are loaded into a high vacuum chamber that is commonly pumped with either a diffusion pump or a cryopump. Diffusion pumped systems commonly have a cold trap to prevent the backstreaming of pump oil vapors into the chamber. The charge or material to be deposited is loaded into a heater container called the “crucible”. It can be heated very simply by means of an embedded resistance heater and an external power supply or by using an electron-beam gun. As the material in the crucible becomes hot, the charge gives off a vapor. Since the pressure in the chamber is much less than 1 mtorr, the atoms of the vapor travel across the chamber in a straight line until they strike a surface where they accumulate as a film.

Evaporation systems may contain several crucibles to allow the deposition of multiple layers without breaking vacuum. To help start and stop the deposition mechanical shutters are used in front of the crucibles, and crystal monitors are used to control the thickness of the deposited metal film.

2.4.2 Sputtering

Sputtering is the primary alternative to evaporation for metal film deposition in microelectronics fabrication. First discovered in 1852, sputtering was developed as a thin film deposition technique by Langmuir in the 1920s. It has better step coverage than evaporation, induces far less radiation damage than electron beam evaporation,

and is much better at producing layers of compound materials and alloys. In the case of microbolometers sputtering provides better contact between the bolometric sensor and the antenna. This is important since it has been shown that bad contacts will affect the responsivity of the microbolometer[32] and will also increase its $1/f$ noise level[33].

A simple sputtering system consists of a parallel plate plasma reactor in a vacuum chamber where a high density of ions strike a target containing the material to be deposited. Atoms of this material are ejected and collected by the substrates that are to be coated with that material. In sputtering the target material (not the substrate wafers) must be placed on the electrode with the maximum ion flux. To collect as many of these ejected atoms as possible, the cathode and anode in a typical sputtering system are closely spaced, often less than 10 cm. An inert gas is normally used to generate the plasma. The gas pressure in the chamber is held at about 0.1 torr. This results in a mean free path in the order of hundreds of microns.

Due to the physical nature of the process, sputtering can be used for depositing a wide variety of materials. In the case of elemental metals, simple dc sputtering is usually favored. When depositing insulating materials, such as SiO_2 , an RF plasma must be used. If the target material is an alloy or compound, the stoichiometry of the deposited material may be slightly different than the target material[31].

2.4.3 Chemical Vapor Deposition (CVD)

Evaporation and sputtering are two types of “physical vapor deposition” where physical methods are used to produce the constituent atoms which pass through a low-pressure gas phase and then condense on the substrate. In the case of CVD, reactant gases are introduced into the deposition chamber, and chemical reactions between them on the substrate surface are used to produce the film. CVD has historically been used in the integrated circuit industry mainly for silicon and dielectric deposition, primarily due to its good quality films and good step coverage.

The materials usually deposited using CVD are silicon in the polycrystalline form (polysilicon), silicon nitride and phosphor silicate glass (PSG). Polysilicon has properties comparable to single crystalline silicon, and silicon nitride is a very hard, chemically inert and strong, but brittle material with small thermal conductivity, PSG is mainly used as a sacrificial layer for silicon micromachining.

There are three types of CVD, atmospheric pressure CVD (APCVD), low pressure CVD (LPCVD) and plasma enhanced CVD (PECVD). LPCVD has better step coverage and the mechanical and chemical quality of the film (in terms of impurities, pinholes and density) is much better than APCVD. Films that will be part of mechanical microstructures should be free of internal stresses or else bending and buckling will occur. The stress of a film grown on top of a substrate is indicated with respect to the underlying material. An expanding layer is then said to be under compressive stress, a contractive layer under tensile stress[34]. Bending and buckling will occur if

the film is under compressive stress. Low stress nitride films can be deposited in an LPCVD reactor at 835 °C. PECVD is used when the deposition needs to be done at low substrate temperatures (~ 300 °C) this deposition method has good step coverage but the films suffer from pinholes and a high hydrogen concentration making them less suitable for mechanical microstructures.

2.5 Etching

After thin films are deposited on the wafer surface, they can be selectively removed by etching to leave the desired pattern on the wafer surface. In addition to deposited films, parts of the silicon substrate itself may be etched, such as in creating trenches in isolation structures. The masking layer may be photoresist, or it may be another thin film such as silicon dioxide or silicon nitride. Oxide or nitride masks stand up better than photoresist to etching conditions and are often called hard masks. But they themselves must be selectively etched, usually using lithographically defined photoresist as the masking layer. The etching of a thin film is usually done until a different layer (known as “etch stop”) is reached underneath.

Etching can be done in either a “wet” or “dry” environment. Wet etching involves the use of liquid etchants. The wafers are immersed in the etchant solution and the exposed material is etched mostly by chemical processes. Dry etching involves the use of gas-phase etchants in a plasma. Here the etching usually takes place by a combination of chemical and physical processes. Because a plasma is involved, dry

etching is usually called “plasma etching” [35]. Both methods can be either isotropic, i.e., provide the same etch rate in all directions, or anisotropic, i.e., provide different etch rates in different directions (Figure 2.5). The important criteria for selecting a particular etching process are the material etch rate, the selectivity to the material to be etched versus other materials, and the isotropy/anisotropy of the etching process.

Figure 2.5. Schematic of: (a) Isotropic and (b) anisotropic thin film etching.

Wet etching provides a good etch selectivity and is usually isotropic with the exception of anisotropic silicon wet etch using potassium hydroxide (KOH). Dry etching is often anisotropic, resulting in a better pattern transfer, as mask underetching is avoided (Figure 2.5). Reactive ion etching (RIE) is a common form of dry etching where reactive ions are generated in a plasma and accelerated towards the surface to be etched, this process is anisotropic but has low etch selectivity.

2.6 Silicon Micromachining

Silicon micromachining is a process used to fabricate static and movable 3D microstructures, such as bridges, cantilevers and membranes on silicon substrates. There are two types of silicon micromachining, bulk micromachining and surface micromachining. In the case of bulk micromachining wet-etch and dry-etch techniques are used to remove parts of the silicon substrate and form the microstructure, whereas in the case of surface micromachining the microstructure is made of thin-film layers which are deposited on top of the substrate and selectively removed in a defined sequence.

Surface micromachining has become the major fabrication technology of microscale structures because it uses standard CMOS fabrication processes and facilities. The most commonly used surface micromachining process is sacrificial-layer etching. In this process a microstructure is released by removing a sacrificial thin-film material which was previously deposited underneath the microstructure (Figure 2.6).

Usually the sacrificial layer is made out of silicon dioxide (SiO_2), phosphorous-doped silicon dioxide (PSG), or silicon nitride (Si_3N_4) and the structural layers are then typically formed with polysilicon, metals and alloys. There are three key challenges in fabricating microstructures using surface micromachining: control and minimization of stress and stress gradient in the structural layer to avoid bending or buckling of the released microstructure; high selectivity of the sacrificial layer etchant to structural layers and silicon substrate; and avoidance of stiction of the suspended

Figure 2.6. Surface micromachining fabrication process. (a) Deposition and patterning of the sacrificial layer. (b) Deposition and patterning of the structural layer. (c) Release etch.

microstructure to the substrate[36].

Stiction is the tendency of the suspended structures to collapse due to the surface tension of liquids during evaporation. The liquid forms a droplet during drying between the microstructure and the substrate which generates an underpressure that will make the structure collapse if it is not stiff enough (Figure 2.7).

Figure 2.7. Capillary force during sacrificial layer etch.

Rinsing procedures like critical point drying or freeze drying can be used to avoid stiction. Critical point drying works by exchanging the rinsing liquid, after etching the sacrificial layer, with liquid carbon dioxide which is then removed in its supercritical state avoiding the liquid-gas phase transition. Another alternative to avoid stiction is to introduce some roughness to the structure or use an anti-stiction coating.

2.7 Typical Fabrication Process for Antenna-coupled

Microbolometers

Most of the antenna-coupled microbolometers in this study were patterned on 3-inch, 380 μm thick high resistivity ($\rho \approx 3 \text{ k}\Omega \cdot \text{cm}$) silicon substrates, with 200 nm of thermally or PECVD grown SiO_2 for thermal and electrical isolation. The substrates were spin-coated with a bilayer of copolymer (PMMA-MAA) and PMMA. A 300 nm layer of 11% copolymer diluted 3:1 in anisole was obtained by spin coating at 3000 rpm for 60 seconds and baking on a hotplate at 170 °C for 15 minutes. A second layer 150 nm thick of 4% PMMA was spun onto the substrate at 3500 rpm for 1 minute and baked afterwards for 15 minutes on a 170 °C hotplate. The total thickness of the bilayer was measured with a profilometer and was always close to 450 nm.

The antenna, microbolometer patch, bond-pads and bias lines were patterned using a Cambridge EBMF 10.5 Electron Beam Lithography System at a beam energy of 40 keV. The dose used for each exposure depended on the critical dimension of the pattern to write, the antennas and the bias lines were written at the same dose and with the same beam current, which varied from 150 $\frac{\mu\text{C}}{\text{cm}^2}$ for the square-spiral antennas to 250 $\frac{\mu\text{C}}{\text{cm}^2}$ for bowties. All the antennas were written using 1 nA of beam current. The bolometer patch was always written at a dose of 350 $\frac{\mu\text{C}}{\text{cm}^2}$ and 1 nA of beam current. The bondpads, being large structures (squares of 200 μm each side) that took a long time to write, were written with an e-beam current of 35 nA at a dose of 180 $\frac{\mu\text{C}}{\text{cm}^2}$ to decrease the exposure time. A wafer with 200 devices (antennas and bondpads)

took around 90 minutes to expose. After exposure, the devices were developed for 2 minutes on a 1:1 solution of MIBK:IPA, rinsed with IPA and blow-dried with a nitrogen gun.

The antennas were made out of 100 nm of e-beam evaporated gold over a 5 nm adhesion-layer of Cr and liftoff was done by soaking the wafers on methylene chloride for 2 hours. The microbolometers were made off a thin film (~ 70 nm) of RF-sputtered VOx or DC-sputtered Nb, and the liftoff process was done using methylene chloride also.

To increase the thermal isolation of the microbolometers some of them were fabricated on a silicon nitride membrane using surface micromachining. The antenna-coupled microbolometer was patterned on a silicon substrate with 3 μm of thermally grown SiO_2 and 400 nm of low-stress Si_3N_4 deposited using LPCVD. The silicon nitride membrane was patterned using CF_4 -based RIE and released by etching the SiO_2 sacrificial layer with hydrofluoric acid (HF 49% in water) and critical point drying.

CHAPTER 3

CHARACTERIZATION OF ANTENNA-COUPLED DETECTORS

Infrared radiation is collected by an antenna by the generation of current in its elements, this generated current has the same frequency as the incident radiation, in the infrared case the frequency would be in the 30 terahertz range. This generated current will flow through the sensing element, in this case the bolometer, and will increase its temperature by joule heating. The change in temperature will make the bolometer change its resistance, thus providing the detection mechanism. Other advantages of using an antenna as collection element are directivity, polarization dependence and tunability.

Antenna-coupled infrared detectors were fabricated using electron-beam lithography and liftoff on 3-inch silicon wafers, Figure 3.1 shows a dipole-coupled microbolometer fabricated using this method. Each processed wafer was scribed into 1 cm $\times$ 1 cm die with 10 devices each and bonded on chip carriers specially made to facilitate their testing.

Figure 3.1. Dipole-coupled Niobium microbolometer.

The test setup that was used to characterize antenna-coupled infrared detectors is shown in Figure 3.2. A CO₂ laser emitting infrared radiation at 10.6 μm focused using an F/8 optical train was used, the resulting spot can be seen in Figure 3.3. The diameter of the spot that encloses 84% of the flux in the diffraction pattern is approximately 200 μm , the power at the focal plane was set using a wire-grid polarizer, most of the measurements were made at 33 mW of optical power at the focal plane, which gives an approximate irradiance of 88 W/cm² at the focus. The optical train includes a half wave plate which is used to rotate the linear polarization of the CO₂ laser.

Figure 3.2. Test setup used to characterize antenna-coupled microbolometers.

The laser beam was modulated with a chopper at a frequency of 2.5 kHz. An electronics board was designed to bias the microbolometer, which allowed the bias to be set anywhere from -1 V to $+1$ V. A load resistor is connected in series with the microbolometer to limit the current that flows through it. This electronics board, which holds the chip carrier with the devices, is mounted on a micro-positioning stage with Melles-Griot nanomovers in the X and Y directions to make automated two-dimensional scans on the detectors. The position of the detector along the optical axis (Z position) is controlled manually. This stage can also be rotated manually in one degree increments to allow antenna pattern measurements. A low noise pre-amplifier gives the output signal a $10\times$ gain (or more) before being read by a lock-in

amplifier using the chopper frequency as reference.

Figure 3.3. Two dimensional scan of the F/8 beam used to characterize antenna-coupled bolometers. Contours are drawn at 20% intervals.

3.1 Low Noise Amplifier Design

In order to have an accurate value for the signal-to-noise ratio of antenna-coupled microbolometers the noise level of the measurement system should be small compared to the noise of the device under test. Figure 3.4 shows a schematic representation of the amplification stages used to measure the noise and response of microbolometers. By using amplification stages the noise level of the measurement equipment relative to the noise of the detector gets reduced, for example amplifying the output signal $1000\times$ will make the noise of the measurement equipment appear $1000\times$ lower in comparison. This will give a more accurate reading, however the amplification stages introduce some additional noise to the system which needs to be small compared to the noise of the detector. Two amplification stages and a high-pass filter are needed if amplifications higher than $100\times$ are desired or else the dc-bias voltage (which is usually around 100 mV) will saturate the amplification stage. The noise contribution of the filter will be reduced if its placed after the first amplification stage. The noise analysis of the circuit shown in Fig. 3.4 proceeds as follows: the noise at point 1 (n_1) is given by the noise of the bias voltage (n_{bias}) divided by the voltage divider formed by R_L and R_b plus the noise of the bolometer (n_b) added in quadrature:

$$n_1 = \sqrt{\left(\frac{n_{bias}R_b}{R_b + R_1}\right)^2 + n_b^2}. \quad (3.1)$$

The noise at point 2 (n_2) is given by the noise at point 1 amplified by 10× plus the noise of the amplification stage ($n_{10\times}$) added in quadrature,

$$n_2 = \sqrt{(10 \cdot n_1)^2 + n_{10\times}^2}. \quad (3.2)$$

If more than one stage of amplification is required, a high-pass filter should be used between the amplification stages. For a first-order high-pass filter the noise at point 3 (n_3) will be given by the noise at point 2 (n_2) plus the addition in quadrature of the Johnson noise of the resistor used in the high-pass filter (R_{HP}),

$$n_3 = \sqrt{n_2^2 + 4kTR_{HP}}, \quad (3.3)$$

where k is the Boltzmann constant and T is absolute temperature. The noise at point 4 (n_4) is given by the noise at point 3 amplified by the gain of the second amplifying stage plus the noise of the amplification stage ($n_{100\times}$) added in quadrature,

$$n_4 = \sqrt{(100 \cdot n_3)^2 + n_{100\times}^2}. \quad (3.4)$$

The total noise of the system referred to the input is the total noise at the output divided by the total amplification, that is: $n_T = n_4/1000$.

The total noise in the circuit shown in Fig. 3.4 can be reduced by making the noise contributions of the bias voltage and the amplification stages as small as possible. Figure 3.5 shows the bias circuit used, the use of batteries and decoupling capacitors

Figure 3.4. Electronic setup to measure antenna-coupled microbolometers.

along with an ultralow noise op-amp (TL1028) helped reduce the noise of the bias voltage to $7 \frac{\text{nV}}{\sqrt{\text{Hz}}}$ at 100 Hz. The bias circuit in Fig. 3.5 gives a 1V fixed output which will be reduced to the desired bias voltage by resistor R_L from Fig. 3.4, a fixed bias voltage was chosen to avoid the high $1/f$ noise level of potentiometers. Decoupling capacitors are used to reduce the high frequency noise in the system. Batteries power the op-amp and are used instead of regulated power supplies due to their lower noise.

Figure 3.5. Low-noise bias source for microbolometers.

Figure 3.6 shows an amplification stage for microbolometers with a gain given by $(\frac{R_F}{R_1} + 1)$. The noise analysis for this circuit can be obtained by superposition and adding the individual noise contributions in quadrature. The noise contribution of the op-amp comes from two sources, its voltage noise and its current noise which are parameters than can be found in the op-amp's data sheet or measured

directly. The voltage noise contribution is given by n_{opamp} , which can generally be found in the opamp's datasheet, and the current-noise contribution is the voltage generated by the current noise flowing through external resistors. In the case of the circuit shown in Fig. 3.6, the current-noise contribution is given by $i_n(R_L \parallel R_b)$ and $i_n(R_1 \parallel R_F)$. The noise due to the external resistors R_L, R_b, R_F and R_1 is given by $\sqrt{4kTR_L(\frac{R_b}{R_L+R_b})}$, $\sqrt{4kTR_b(\frac{R_L}{R_L+R_b})}$, $\sqrt{4kTR_F(\frac{R_1}{R_F+R_1})}$ and $\sqrt{4kTR_1(\frac{R_F}{R_F+R_1})}$ respectively. All these expressions for noise voltages are referred to the input, therefore they have to be multiplied by the gain to get the noise at the output of the amplification stage. The total noise of the amplification stage, referred to the input, is given by:

$$n_{amp}^2 = n_{opamp}^2 + i_n^2 \left[\left(\frac{R_b R_L}{R_L + R_b} \right)^2 + \left(\frac{R_1 R_F}{R_1 + R_F} \right)^2 \right] + 4kT \left[R_L \left(\frac{R_b}{R_L + R_b} \right)^2 + R_b \left(\frac{R_L}{R_L + R_b} \right)^2 + R_F \left(\frac{R_1}{R_F + R_1} \right)^2 + R_1 \left(\frac{R_F}{R_F + R_1} \right)^2 \right]. \quad (3.5)$$

For a $10\times$ amplification stage with $R_F = 499 \Omega$ and $R_1 = 54.9 \Omega$ with a 200Ω -bolometer biased at 100 mV ($R_b = 200 \Omega$ and $R_L = 1.8 \text{ k}\Omega$) using a TL1028 ultralow noise op-amp ($n_{opamp} = 0.85 \frac{\text{nV}}{\sqrt{\text{Hz}}}$, $i_n = 1.6 \frac{\text{pA}}{\sqrt{\text{Hz}}}$ both measured at 100 Hz [37]) the total noise at the output will be given by the contribution of the bias voltage and the noise of the amplification stage:

$$n_{total} = \sqrt{\left(\frac{n_{bias} R_b}{R_b + R_L} \right)^2 + n_{amp}^2}. \quad (3.6)$$

Figure 3.6. Amplification stage for microbolometers.

Figure 3.7. Noise of one amplification stage.

The noise of the bias voltage was measured to be $7 \frac{\text{nV}}{\sqrt{\text{Hz}}}$ at 100 Hz and the theoretical noise contribution of the amplification stage is $2.1 \frac{\text{nV}}{\sqrt{\text{Hz}}}$ calculated using Eq. (3.5). The total noise referred to the input is $2.2 \frac{\text{nV}}{\sqrt{\text{Hz}}}$, calculated using Eq. (3.6), since a 200Ω resistor has a Johnson noise of $1.8 \frac{\text{nV}}{\sqrt{\text{Hz}}}$. The amplification stage shown in Fig. 3.6 will give detector-noise-limited measurements for microbolometers with resistances as low as 200Ω .

Figure 3.7 shows a graph of the total noise of the amplification stage compared to the Johnson noise of the microbolometer as a function of its resistance. The graph also shows the noise of the electronics which is obtained by subtracting in quadrature the total noise from the Johnson noise of the microbolometer ($n_{\text{electronics}} = \sqrt{n_{\text{total}}^2 - 4kTR_b}$). From the figure we can see that the noise of the electronics is around $1.5 \frac{\text{nV}}{\sqrt{\text{Hz}}}$.

Figure 3.8. Noise Floor of the measuring electronics.

3.2 Noise Measurements

Noise measurements were made using the setup shown in Fig. 3.4 and an HP3562A Dynamic Signal Analyzer which has a measurement range of 64 μHz to 100 kHz. The noise introduced by the signal analyzer was attenuated 1000 $\times$ referred to the noise of the detector due to the amplification stages of the test setup. The noise floor of the test setup was measured by shorting the bolometer. The result is shown in Fig. 3.8, and we can see that the noise floor is very close to the $1.5 \frac{\text{nV}}{\sqrt{\text{Hz}}}$ noise expected from the theoretical noise calculations. The spikes in the measurements are 60Hz power-line harmonics introduced into the system.

Figure 3.9. Noise spectrum for a 200Ω chrome microbolometer and fitting function.

The noise characteristics of microbolometers depend on the bolometric material used and its deposition process, Fig. 3.9 shows the noise spectrum of a 200 Ω e-beam evaporated chrome microbolometer. Devices that are sputtered show lower noise levels than evaporated bolometers, because sputtering provides a better contact between the bolometric material and the gold structures, which reduces $1/f$ noise. The noise spectrum of a 200 Ω chrome bolometer fits the following noise function:

$$n_b = \frac{100}{f} + \frac{70}{\sqrt{f}} + 6, \quad (3.7)$$

which shows two $1/f^k$ components. Figure 3.9 shows the noise spectrum of a 200 Ω chrome bolometer and the fitting function (Eq. (3.7)).

3.3 Response Measurements

Response measurements are taken using the test setup shown in Fig. 3.2. For signal-to-noise (SNR) calculations, noise and response measurements should be taken at the same level of amplification. The total response of the devices is measured in volts from the lock-in amplifier or can also be measured directly from an oscilloscope, in which case the output would look like a square wave with the same frequency of the chopper.

The maximum response of the device is measured at the polarization that best matches the antenna being measured (co-polarized response) and the minimum response is taken at the polarization that gives the lowest response (cross-polarized response), the co-polarized minus the cross-polarized response results in the pure antenna response. The ratio between the co-polarized and cross-polarized response gives the polarization ratio of the antenna. The cross-polarized response is usually due to heating of the substrate. Figure 3.10 shows the response of a dipole-coupled niobium microbolometer to copolarized and cross-polarized IR laser radiation as a function of bias voltage. The response is nonlinear because the bias voltage heats up the bolometer making its resistance decrease (in the case of a metallic bolometers), which from Eq. (1.12) will make the response decrease also.

Figure 3.10. Response of a dipole-coupled microbolometer as a function of the bias voltage.

3.4 Polarization Dependence

Polarization is a basic property of antennas. Since bolometers are not sensitive to polarization, if an antenna-coupled microbolometer shows a change in response due to a change in polarization then an antenna effect is taking place. This polarization dependence should match the polarization characteristics of the antenna used to couple radiation into the microbolometer.

The polarization of the linearly polarized laser beam shown in Fig. 3.2 can be rotated by using a half-wave plate, a quarter-wave plate can be used if circular polarization is needed. A wave plate is made of a slab of birefringent material that changes the phase between the two orthogonal components of a linearly polarized wave, if this phase change is equal to π then the linear polarization of the wave will be rotated,

if the phase change is $\pi/2$ then the resulting wave will be circularly polarized, any other phase change will give an elliptically polarized wave.

Figure 3.11 shows how the response of a dipole-coupled microbolometer changes with polarization. Dipoles are linearly polarized antennas and therefore have a cosine-squared polarization dependence. The measurements shown in Fig. 3.11 fit the following equation, $V = 215 \cos^2(x + 65^\circ) + 74$. The polarization ratio in this case is around 4.

Figure 3.11. Polarization dependence of a dipole-coupled microbolometer.

3.5 Time Constant Measurements

To measure the time constant of antenna-coupled microbolometers an acousto-optic (AO) modulator was incorporated into the optical setup shown in Fig. 3.2 instead of the mechanical chopper. The AO modulator can modulate the laser beam up to 10 MHz. Frequencies this high cannot be handled by the amplification stages therefore the output of the device under test has to go directly to a lock-in amplifier or spectrum analyzer. The time constant measurements were taken by varying the modulation frequency of the AO modulator with an HP 33120A function generator and measuring the response of the device with an HP 3585 spectrum analyzer. The measurements were automated using a computer with HPIB interface and software programmed in Labview.

Figure 3.12 shows the frequency response of an antenna-coupled microbolometer on a Si-SiO₂ substrate, the time constant of the device is obtained by fitting its frequency response to Eq. (1.13) or simply by finding the frequency that makes the response drop to 70.7% of its maximum value and then calculating the time constant using $\tau = 1/(2\pi f_{0.7})$ [6]. From Eq. (1.16) we can see that the time constant of a microbolometer will be reduced with higher thermal conductivity substrates and a smaller thermal mass. Since microbolometers are small compared to standard bolometers, if they are fabricated on high thermal conductivity substrates (Si-SiO₂) they will be very fast, having time constants in the order of ~ 130 ns which are several orders of magnitude faster than commercial bolometers which have time constants in

the ~ 10 ms range. There is a tradeoff between speed of the detector and responsivity, from Eq. 1.13 we can see that fast bolometers (small time constant τ) will have low responsivity, therefore the substrate material has to be chosen so that the maximum response can be obtained for the desired detector speed.

Figure 3.12. Time constant of an antenna-coupled microbolometer on a Si-SiO₂ substrate.

3.6 Radiation Patterns

Another basic property of antennas is their radiation pattern. In the case of antenna-coupled microbolometers its response as a function of angle of incidence will give the radiation pattern of the antenna. These antenna patterns are measured by rotating the stage with respect to the the optical axis in one degree increments and recording

the response of the device. After every rotation the X , Y and Z positions have to be readjusted to center the device at the focus of the laser beam. This is done by adjusting the position of the device until the maximum response of the device is obtained. The use of high $F/\#$ optics ($F/8$) to focus the CO_2 laser on the detector reduces the effect of the convolution of the antenna pattern with the angular distribution of the focused light cone.

Figure 3.13 shows the radiation pattern of an array of dipole-coupled microbolometers, it shows the normalized radiation pattern on a linear scale (Fig. 3.13(a)) and on a dB scale (Fig. 3.13(b)).

Figure 3.13. Radiation Pattern of a 2D-Array of dipole-coupled microbolometers. (a) Linear scale, (b) dB scale.

3.7 Spatial Response of Antenna-Coupled Detectors

The signal obtained by a detector is proportional to the irradiance distribution integrated over its collection area. Commercial bolometers are much bigger than the

wavelength and their collection area can be determined by scanning a probe beam across the detector and measuring the output of the detector as a function of the position of the probe. If the dimensions of the detector are large compared to the probe beam then the detector's spatial response will be very close to the scanned area that generated a response. If this method is used with microbolometers we are scanning a structure smaller than the probe beam, so that a two dimensional scan will result in the convolution of the beam with the spatial response of the detector. In order to extract the spatial response a deconvolution of the two-dimensional scan with the laser beam has to be made. Alda et al. [38] developed a deconvolution method to extract the spatial response of lithographic infrared antennas from two-dimensional scans by deconvolving the laser beam, which is modeled as a 2D Gaussian beam convolved with a slightly comatic Airy function. The analytical beam model is obtained by fitting the characteristics of the real laser beam obtained by knife-edge measurements.

The two-dimensional scan data is obtained by focusing a $10.6\text{ }\mu\text{m}$ CO_2 laser beam, using $F/1$ optics, on the detector. A serial scan is performed, moving the device in the X and Y directions using a motorized Melles-Griot Nanomover system, which is controlled by a computer that also records the response of the detector from the lock-in amplifier. Measurements were usually made by scanning a $100\text{ }\mu\text{m} \times 100\text{ }\mu\text{m}$ area at $1\text{ }\mu\text{m}$ steps. This scanning procedure was automated using software programmed in Labview which also collected the output data. Each measurement took around 1.5 h to acquire. The software saved the response in a two-dimensional matrix format

and a program made in Matlab performed the deconvolution using the algorithm described in [38]. Figure 3.14(a) shows the two-dimensional scan of a log-periodic-coupled microbolometer and Fig. 3.14(b) shows the results after deconvolution.

Figure 3.14. Spatial response of infrared antennas. (a) Two dimensional scan of a log-periodic antenna, (b) spatial response of log-periodic antenna after deconvolution.

CHAPTER 4

MATERIALS FOR ANTENNA-COUPLED IR DETECTORS

An antenna-coupled microbolometer consists of a metallic antenna, bias lines, a bolometric detector and a substrate. The materials used in the fabrication of antenna-coupled microbolometers play an important role in their performance. Materials that provide better thermal isolation, better matching of the IR radiation to the antenna along with a high-TCR microbolometer will make higher responsivity detectors.

4.1 Substrate Losses and Silicon Lenses

Antennas on a dielectric are most sensitive to radiation from the substrate side[39]. However for an antenna lying on a substrate (Figure 4.1), rays incident on the air-substrate interface at an angle larger than the critical angle are completely reflected and trapped as surface waves[40] reducing the efficiency of the antenna, by reciprocity this reduction in efficiency applies for receiving antennas also. These surface waves may show up experimentally as spikes in a receiving antenna pattern [11]. The sim-

plest way to solve this problem is to mount on the back side of the substrate a lens with the same dielectric constant, this way the rays are now incident nearly normal to the surface and do not suffer total internal reflection (Figure 4.2).

Figure 4.1. Trasmitting antenna on a dielectric substrate showing the rays trapped as surface waves. From [40].

The substrate lens takes advantage of the sensitivity of an antenna to radiation from the substrate side and eliminates the surface waves. The disadvantages are absorption losses and the difficulty of optically contacting and aligning the lens to a lithographic antenna. Hemispherical substrate lenses are particularly attractive because they are aplanatic, adding no spherical aberration or coma[41]. These lenses also do not refract the transmitted rays, (Fig. 4.2) a characteristic that makes them useful for antenna-pattern measurements. A hemisphere has a linear magnification equal to the index of refraction n of the material it is made off, increasing the effective collection area of an infrared detector by a factor of n^2 .

Figure 4.2. Antenna-coupled detector with a Silicon substrate lens.

A 15-mm diameter silicon hemisphere was attached to an antenna-coupled microbolometer on a Si-substrate. Measurements on these type of devices gave a $15\times$ increase in response, a bit more than the expected $n^2\times$ apparent increase in area of the detector ($n^2 \simeq 11.7$ for silicon at $10.6\ \mu\text{m}$). Figure 4.3 shows the radiation pattern of a bowtie antenna with and without a silicon hemisphere attached, the radiation pattern with the silicon hemisphere shows a sharp signal drop for angles of incidence higher than 15 degrees, this can be explained by total internal reflection due to an air gap between the substrate and the silicon sphere, the critical angle in a silicon-air interface is around 17 degrees, this effect can be avoided by making optical contact between the silicon sphere and the substrate or by fabricating the antenna-coupled detector directly on the silicon sphere.

Figure 4.3. Radiation Pattern of a Bowtie Antenna (a) on a silicon substrate, (b) with a hemispherical silicon lens attached.

4.2 VOx microbolometers

From Eq.(1.14) we can see how the temperature coefficient of resistance α (TCR) of the bolometer is directly proportional to the responsivity of the detector, therefore the choice of the thin-film heat-sensitive material is an important factor in maximizing the response of microbolometers. A thin film of sputtered Nb, which has a reported TCR close to 0.3 %/K [42], was used as the bolometer in [43]. Vanadium is a metal with a variable valence forming a large number of oxides which have a very narrow range of stability[44], films of vanadium oxide (VOx) consisting of a mixture of various oxides present a $TCR \approx 2$ %/K and have been used in the past to fabricate microbolometers[45]. More involved deposition processes have also been reported, yielding films of stoichiometric VO₂ with TCR's greater than 5 %/K [46].

A comparison of the performance between Nb-based and VOx-based microbolometers was performed using two-dimensional arrays of log-periodic-coupled microbolometers with a $50 \mu\text{m} \times 50 \mu\text{m}$ pixel area (Figure 4.4). The Nb-based microbolometers

showed an average dc resistance of $1.2 \pm 0.1 \text{ k}\Omega$ and the VOx detectors presented $450 \pm 50 \text{ }\Omega$ average dc resistance. The measurements were made at a bias voltage of 300 mV.

Figure 4.4. Scanning Electron micrograph of a 2D array of Log-periodic antenna-coupled detectors.

The response of the antenna arrays to $10.6 \text{ }\mu\text{m}$ radiation was measured, the Nb-based detectors gave a co-polarized signal of $5.1 \pm 1 \text{ }\mu\text{V}$ while the measured response

of the VOx-based devices was $22.5 \pm 2.5 \mu\text{V}$, which corresponds to a $4.5\times$ increase in response. Figure 4.5 shows the noise frequency spectrum measured with an HP3562A dynamic signal analyzer. The Nb-based devices had a noise voltage spectrum of $120 \pm 10 \text{ nV}/\sqrt{\text{Hz}}$ at 100 Hz while the VOx-based devices presented a noise voltage spectrum of $60 \pm 5 \text{ nV}/\sqrt{\text{Hz}}$ at the same frequency. This represents a $9\times$ increase in signal-to-noise ratio of the VOx-based devices over the Nb-based ones.

Figure 4.5. Noise Frequency Spectrum for Nb and VOx based detectors.

Figure 4.6 shows the measured angular patterns of the Nb-based devices and the VOx based devices, the radiation characteristics for similar antenna array configurations show that the impedance at the feed does alter the electromagnetic characteristics of the antenna array, in this particular case the Nb-based array presents a more directive pattern than the VOx-based detector, which indicates that the impedance

of the Nb patch is a better match for the individual log-periodic elements of the array. The thickness of the VOx bolometer can be varied to better match its impedance to the antenna elements and get a further increase in response.

Figure 4.6. Radiation Patterns of (a) Nb-based 2D array, and (b) VOx-based 2D array.

This comparison of antenna-coupled microbolometers with different bolometric materials shows that the use of higher TCR materials will increase the response of the detector, this increase in response will be equal to the ratio of TCR's only if the bolometers have the same impedance so that the efficiency of the antenna remains the same. Using different materials will also change the noise characteristics of the detector.

An interesting characteristic of vanadium oxide films is that they exhibit a metal-semiconductor phase transition at which the resistivity and reflectance sharply changes

Figure 4.7. Experimental Resistance vs. Temperature dependence of VO_2 . From [47].

(Fig. 4.7). Single-crystal vanadium dioxide behaves like a semiconductor at temperatures below 45 °C and like a metal at temperatures above 67 °C, a change in the resistance by a factor of $10^3 \times$ is observed. Figure 4.7 shows a graph of the resistance of VO₂ as a function of temperature, TCR is proportional to the slope of the resistance-*versus*-temperature curve of the material, during its phase transition the TCR of VO₂ reaches values as high as 200 %/K [48]. This property can be used to increase the sensitivity of VO_x microbolometers. The moderate temperature of this transition (40-70 °C) can be reached by Joule heating of the sensitive element by passing a bias current through it with sufficiently good thermal insulation of the element. The difficulties in this method consist of the hysteresis character of the transition, variable biasing methods have to be implemented to use the large slope of the R(T) dependence [48] [46].

4.3 Thermal Isolation using Aerogel

From Eqs. (1.14),(1.15) and (1.16) we can see that an increase in the thermal impedance of the detector would increase the responsivity of the detector, but will also increase its time constant, slowing down the response of the detector and therefore decreasing its bandwidth. The highest thermal impedance would occur when the detector is completely isolated from the environment, therefore the use of high thermal conductivity substrates will make a bolometer faster but will reduce its responsivity.

From the above stated we can see that a trade-off exists between responsivity

and speed of the detector; in order to fabricate fast detectors low thermal isolation is required which yields low responsivity, therefore the thermal conductivity of the substrate has to be chosen so that it gives the maximum response for the required frame-rate.

Aerogels are materials that consist of pores and particles that are in the nanometer size range and have exceptional optical, thermal, acoustical and electronic properties[49] which depend on the porosity of the film. Thermal conductivity decreases linearly as porosity of the aerogel film increases, values lower than air have been measured on 70-99% porosity films[50]. Thin aerogel films with up to 98% porosity can be deposited on Si wafers by spin coating or dip coating, and can later be used as substrates for further lithographic processing[49]. Here the noise, response, time constant and radiation characteristics of two dimensional arrays of antenna-coupled microbolometers fabricated on aerogel-coated substrates are studied and compared to similar devices fabricated on Si-SiO₂ substrates. Two dimensional arrays of dipole-coupled [Fig. 4.8(a)] and bowtie-coupled [Fig. 4.8(b)] microbolometers were used in this study. A series configuration was selected to match the input impedance of typical commercial ROIC's which is in the kilo-ohm range[51], and cover a pixel area of $50\text{ }\mu\text{m} \times 50\text{ }\mu\text{m}$.

Silica aerogel thin films were deposited by spin coating onto thermally-oxidized silicon wafers using the process described by Clem et al. [50]. The resulting film was 700 nm thick and had a refractive index of 1.06, as measured by 632 nm ellipsometry, which indicates that the silica aerogel film had a porosity of 85 %[49]. Silica aerogels

Figure 4.8. (a) Dipole-coupled array of microbolometers, (b) bowtie-coupled array of microbolometers.

of this porosity have a thermal conductivity of around $15 \text{ mW}/(\text{m} \cdot \text{K})$ which is lower than air ($26 \text{ mW}/(\text{m} \cdot \text{K})$) and much lower than SiO_2 ($1200 \text{ mW}/(\text{m} \cdot \text{K})$)[50]. The two-dimensional arrays of microbolometers presented an average dc resistance of $3 \pm 0.3 \text{ k}\Omega$ and all of the measurements were made with a bias voltage of 300 mV.

The response of the antenna arrays as a function of the modulation frequency of the laser beam was measured using an acousto-optic modulator. The frequency response of the antenna arrays did not depend on the type of antenna measured but rather on the substrate the antennas were fabricated on. Antennas fabricated on aerogel presented time constants around $5 \mu\text{s}$ while antennas fabricated on SiO_2 showed time constants around 130 ns. Figure 4.9 shows a typical frequency response measurement on an antenna array fabricated on a SiO_2 substrate and on an aerogel substrate. The antennas on aerogel have a time constant around 40 times slower than

antennas on SiO_2 which is within a factor of 2 of the ratio of the thermal conductivities of the substrates[50].

Figure 4.9. Frequency response measurement of devices fabricated on SiO_2 substrates and on aerogel.

The response of the antenna arrays to $10.6 \mu\text{m}$ radiation was measured, the maximum signal was obtained for the polarization parallel to the antenna axis. The bowtie-coupled antenna arrays on SiO_2 gave a maximum response of $7.3 \pm 0.03 \mu\text{V}$, the same type of antennas fabricated on aerogel gave a maximum response of $142 \pm 2 \mu\text{V}$, a $20\times$ increase in response. The same measurements were made on dipole-coupled antenna arrays, resulting in a maximum signal of $3 \pm 0.7 \mu\text{V}$ for devices on a

SiO₂ substrate and $100 \pm 12 \mu\text{V}$ for the ones on aerogel, showing a $30\times$ increase in response. From Eq.(1.14) we can see that for an equal change in the thermal impedance of both type of arrays a similar increase in response should be obtained, however since we are working with antenna-coupled detectors a change in the permittivity of the substrate from 4.7 for SiO₂ to 1.1 for high-porosity silica aerogel[49] will change the electrical size of the antenna which would affect its radiation characteristics. Figure 4.10 shows the measured radiation patterns of both type of antenna arrays on SiO₂ and on aerogel. The radiation characteristics of the dipole-coupled arrays was more affected by the change in substrate compared to the bowtie-coupled array. This is due to the fact that bowtie antennas are broadband structures and a change in its electrical size does not significantly change its radiation characteristics[52].

Infrared detectors usually have large background noise pedestals, therefore figures of merit involving the signal-to-noise ratio (SNR) of the detectors are key to evaluate detector performance[2]. Figure 4.11 shows the noise versus frequency characteristics of detectors on SiO₂ and aerogel substrates. For a chopping frequency of 100 Hz the noise level of detectors on aerogel ($\sim 60 \frac{\text{nV}}{\sqrt{\text{Hz}}}$) is two times higher than for detectors on SiO₂ ($\sim 30 \frac{\text{nV}}{\sqrt{\text{Hz}}}$), which means that the increase in response of $20\times$ and $30\times$ due to the aerogel substrate will translate into a $10\times$ and $15\times$ times increase in D^* . The noise increase observed in detectors on aerogel can be attributed to lower quality contacts due to the roughness of highly porous material added to higher thermal fluctuations observed in metal films deposited on low thermal conductivity substrates[53].

Figure 4.10. Radiation Patterns for (a) Dipole-coupled microbolometer array on SiO_2 , (b) dipole-coupled microbolometer array on aerogel, (c) Bowtie-coupled array on SiO_2 and (d) bowtie-coupled array on aerogel.

Figure 4.11. Noise frequency spectrum for devices fabricated on SiO_2 and on Aerogel.

4.4 Heat conduction through the bias lines

From Eq. (1.14) we can see that the response of a bolometer depends on its thermal impedance $|Z_{th}|$, which accounts for all the heat conduction paths out of the bolometer. There are two main heat flow mechanisms out of a bolometer, the dominant one is through the substrate and the other one is through the electrical leads that bias the bolometer. By using low thermal conductivity substrates we can reduce heat conduction through the substrate, however since good electrical conductors are also good heat conductors the bias lines can become important heat conduction paths. The thermal impedance of the detector can be increased by selecting the material and size of the bias lines, the thermal impedance of a bias line can be calculated using Eq. (1.15), where the thermal resistance is given by [10]

$$R_{th} = \frac{L}{kA}, \quad (4.1)$$

where L is the length of the bias line, k is the thermal conductivity of the metal and A is the cross-section area of the bias line given by the product of the line's width and thickness. The thermal capacitance is given by [10]

$$C_{th} = \rho Vc, \quad (4.2)$$

where ρ is the mass density of the metal, V the volume of the bias line (given by the product of its length, width and thickness), and c is the specific heat of the

metal. Figure 4.12 shows the thermal impedance of a microbolometer whose only heat conduction path is a $35\text{ }\mu\text{m}$ long and 200 nm wide bias line as a function of the thickness of the bias line for different metals, there we can see how just by using titanium instead of gold a one order of magnitude increase in thermal impedance can be obtained.

Figure 4.12. Thermal impedance of a microbolometer as a function of the thickness of the bias leads for different metals.

4.5 Air-Bridge microbolometers

The thermal conductivity out of a device can be decreased by suspending the device on air above its substrate[22] (Figure 4.13), the standard procedure to make a suspended device is by fabricating it on top of a “sacrificial layer” which will later be selectively etched away just leaving the patterned structure suspended on air. Silicon dioxide (SiO_2) is widely used as a “sacrificial layer” because it is easy to deposit and can be etched away with HF which would not etch Silicon which can be conveniently used as an etch-stop.

Figure 4.13. Suspended Square-spiral-coupled microbolometer.

There are several problems associated with fabricating standing structures, two basic problems are thin film stress and stiction. There are two types of thin film stress, tensile or compressive lateral stress and stress gradients in the film[34], these forces make the suspended structures buckle (Figure 4.14). The second problem is stiction. When etching the sacrificial layer the free-standing structures have the tendency to collapse, the cause is the surface tension of the liquid during evaporation, the liquid forms a droplet during drying between the microstructure and the substrate which generates an underpressure that makes the microstructure collapse. To solve this problem other rinsing procedures should be used such as critical point drying or freeze drying.

To avoid buckling of the suspended antenna-coupled detectors a Silicon Nitride membrane was made that would free the detectors from any type of mechanical stress. The fabrication procedure consisted on thermally growing 3 μm of SiO_2 on Si substrates and then growing a 400 nm film of Low-stress Si_3N_4 using LPCVD, the detectors were patterned on such a substrate using electron-beam lithography and liftoff, after the detectors were fabricated windows were opened near the device on the Si_3N_4 film using optical lithography to pattern the windows and RIE to etch away the silicon nitride film, the SiO_2 layer was etched through the Si_3N_4 windows using HF leaving a silicon nitride membrane. Figure 4.15 shows the formed membrane and the undercut generated with the wet-etch process. After etching rinsing was performed with DI-water and then soaked in isopropanol without letting the liquids dry, the

Figure 4.14. Buckled bridge structure. From [54].

isopropanol was later dried using a critical point drier (CPD) to avoid stiction.

Figure 4.15. Window openings on Si₃N₄ to build a membrane.

Since HF attacks niobium and other bolometric materials the antenna-coupled detector was fabricated using gold for the antenna elements and bias line and chrome as the bolometric material (Figure 4.16).

The square-spiral detectors on membranes were tested under vacuum and without a vacuum, without a vacuum a responsivity of 144 V/W and a D^* of $3 \times 10^6 \frac{\text{cm}\sqrt{\text{Hz}}}{\text{W}}$

Figure 4.16. Square-spiral-coupled microbolometer on a silicon nitride membrane.

was obtained, under a vacuum the measured responsivity was 224 V/W and a D^* of $1.7 \times 10^7 \frac{\text{cm}\sqrt{\text{Hz}}}{\text{W}}$. Taking into account that Cr has a TCR 30× lower than VOx, by fabricating a VOx microbolometer on a membrane, values well above 1×10^3 V/W and $1 \times 10^8 \frac{\text{cm}\sqrt{\text{Hz}}}{\text{W}}$ for responsivity and D^* should be obtained when VOx is used in a membrane configuration.

This membrane fabrication process can be modified to allow the use of materials that are attacked by HF by masking them with chrome. After fabricating microbolometers using HF-sensitive materials a thin layer of chrome is deposited to protect them from the HF etch. Windows are opened through the chrome and the silicon nitride layers to allow for the silicon dioxide etch. The following step is to remove the chrome layer by using chrome etch. This process will allow the use of materials like titanium, vanadium oxide and niobium on a silicon nitride membrane using silicon dioxide as the sacrificial layer.

CHAPTER 5

COMPARISON OF DIPOLE, BOWTIE, SPIRAL AND LOG-PERIODIC IR ANTENNAS

Antenna-coupled microbolometers use planar lithographic antennas to couple incident radiation into a bolometer with sub-micron dimensions. The use of an antenna limits the throughput to one mode with one polarization. This limitation to one mode is potentially useful for bolometers used for diffraction limited observations over a broad spectral range. Planar antennas, which are built on a substrate, are quite different from ordinary microwave antennas mainly because they tend to radiate most of their energy into the substrate. For a planar antenna the power division in each medium varies approximately as $\epsilon^{\frac{3}{2}}$ [39]. Figure 5.1 shows the far-field polar diagram for a dipole on a dielectric/air interface for dielectric constants of 1, 4 and 12.

Surface wave excitation occurs in all substrate-based antennas because the lowest TM_0 surface wave mode has a zero frequency cutoff [55], by increasing the substrate thickness more surface wave modes appear which will reduce the efficiency of the antenna. At infrared frequencies (THz) the substrates are electrically thicker making antennas less efficient than their lower frequency counterparts. Antennas on grounded

substrates (microstrip antennas, Fig. (5.2)) are more efficient than printed antennas because they radiate in only one direction and the substrate thickness can be reduced to increase efficiency, this does not happen with printed antennas which can be viewed as microstrip antennas with very thick substrates. The radiation properties of printed antennas become sensitive to substrate losses as the substrate thickness increases[56].

Figure 5.1. Radiation Patterns for a resonant dipole on a substrate. (a) H-plane, (b) E-plane. From [39].

The performance of printed antennas depends on the substrate thickness (h) and the dielectric constant of the substrate (ϵ_s)[57], and there is a certain thickness that will maximize the performance of a printed antenna for a given dielectric constant. Figure 5.3 shows the efficiency of a half-wave dipole as a function of substrate thickness h (given in free space wavelengths) for a substrate dielectric constant of $\epsilon_s = 2.55$. The efficiency η is taken as the ratio of the power radiated into free space to the total

Figure 5.2. Printed Antenna on a grounded substrate.

radiated power and multiplied by 100. From Fig. 5.3 we can see that the maximum efficiency is obtained when $h \approx 0.2\lambda_0$ which is slightly below the cut-off thickness of the TE_0 substrate guided mode. Figure 5.3 also shows that the efficiency is close to 100% when the substrate thickness is close to zero this is because surface wave excitation is negligible for very thin substrates.

In this chapter four different types of microstrip antennas were fabricated on thin substrates and coupled to microbolometers. These IR antenna-coupled detectors were measured at $10.6 \mu\text{m}$ and their performance compared. Fabrication was done on 200 nm of SiO_2 ($\epsilon_s = 4.84$ at $10.6 \mu\text{m}$) deposited using PECVD, the thickness being much smaller than the wavelength in the dielectric ($h \approx 0.04\lambda_d$, where $\lambda_d = \frac{\lambda_0}{\sqrt{\epsilon_s}}$), and a 50 nm Cr ground plane. Antennas were made out of 100 nm-thick gold and patterned using electron beam lithography and liftoff, and 70 nm of dc-sputtered niobium was used as the bolometric element.

Figure 5.3. Efficiency of a printed dipole as a function of substrate thickness. From [55].

5.1 Dipoles

A half-wavelength dipole can be made to resonate and show a purely real input impedance, thus eliminating the need for tuning to achieve a conjugate impedance match. The resonant condition for a half-wave dipole is that the physical length must be shorter but close to half the wavelength and as the antenna cross-arm thickness is increased the length must be reduced more to achieve resonance.

Starting from the optimum dipole length on free space for a certain load impedance Mizuno *et al.* introduced a reduction factor β to find the optimum length of dipoles on an ungrounded substrate[58], this reduction factor is defined as:

$$\beta = \frac{\text{optimum_length_on_a_substrate}}{\text{optimum_length_in_free_space}}, \quad (5.1)$$

and was obtained experimentally for various dielectric materials at 10 GHz (Figure 5.4). Figure 5.4 shows how this reduction factor becomes constant and will converge to a dipole length close to $\lambda_{eff}/2$ ¹ for effective thicknesses larger than 0.2.

Figure 5.4. Reduction factor of the antenna length as a function of the substrate thickness for different dielectric materials. From [58].

For the case of antennas on grounded substrates Pozar calculated the resonant length of microstrip dipoles as a function of the substrate thickness for different substrate materials [59]. Figure 5.5 shows the required lengths for the first resonance of a microstrip dipole versus substrate thickness for different materials, this resonant

¹ λ_{eff} is the effective wavelength given by $\lambda_{eff} = \frac{\lambda_0}{\sqrt{\epsilon_{eff}}}$ and ϵ_{eff} can be obtained from Eq. (1.17).

length oscillates around a length equal to $\lambda_{eff}/2$. From Figs. (5.4) and (5.5) we can see that for antennas on grounded substrates the waves propagate at a velocity defined by the effective permittivity (Eq. (1.17)) and this value can be used to calculate the effective size of printed antennas, in the case of ungrounded antennas this effective permittivity can be used only when the effective thickness is higher than 0.2, for thin ungrounded substrates these antennas will behave more like free-space antennas.

Figure 5.5. Resonant length of a half-wave dipole versus substrate thickness. From [55].

Microstrip dipole antennas coupled to Nb microbolometers were fabricated on a ground plane with a 200 nm SiO_2 substrate using electron-beam lithography and liftoff (Fig. 5.6). The grounded substrate is thin enough to reduce losses due to surface waves and the quasistatic approximation of the effective permittivity can be

used for such thickness at $10.6\ \mu\text{m}$ (Fig. (1.6)).

Figure 5.6. Dipole-coupled microbolometer.

Figure 5.7 shows the response of dipole-coupled microbolometers as a function of dipole length, we can see how the maximum response is around $3 \mu\text{m}$ which is close to half the effective wavelength. From Fig. 1.6 we can see how for a 200 nm SiO_2 substrate at $10.6 \mu\text{m}$ the effective permittivity is equal to its quasi-static value ($\epsilon_{eff} = \frac{\epsilon_s + 1}{2}$), since the permittivity of SiO_2 is $\epsilon_s = 4.84$ at $10.6 \mu\text{m}$ then the effective permittivity is given by $\epsilon_{eff} = 2.92$ which gives an effective wavelength in the substrate equal to $\lambda_{eff} = \frac{\lambda_0}{\sqrt{\epsilon_{eff}}} = \frac{10.6 \mu\text{m}}{1.7} = 6.2 \mu\text{m}$ according to this, the first resonance of the microstrip dipole should be around $\frac{\lambda_{eff}}{2} = 3.1 \mu\text{m}$ which agrees with the results shown in Figure 5.7.

Figure 5.7. Response of a microstrip dipole as a function of its length.

The polarization dependence of dipole-coupled microbolometers was measured using a half-wave plate and is shown in Figure 5.8, the maximum response was obtained

when the polarization was parallel to the antenna axis. The response as a function of polarization angle presented a cosine squared dependence which agrees with antenna theory, the ratio of maximum to the minimum polarization response (polarization ratio) of these devices was around 4.

Figure 5.8. Polarization Dependence for a dipole-coupled microbolometer.

A two-dimensional scan over the dipole-coupled microbolometer was performed using an $F/1$ beam at $10.6\ \mu\text{m}$ and is shown in Figure 5.9(a), this figure represents a convolution between the detector's spatial response and the laser beam profile. The collection area of the detector is obtained by deconvolving the two-dimensional scan

with the laser beam profile. The deconvolution procedure developed in [38] was used to obtain the spatial response of the detector (Fig. 5.9(b)). The laser beam profile used to deconvolve the two-dimensional scan was modeled as a convolution of a 2-D Gaussian with a slightly comatic Airy function. A knife-edge scan was fitted to this model and the resulting beam profile was used to perform the deconvolutions.

Figure 5.9. Dipole spatial response (a) convolved with the laser beam, (b) after deconvolution. The contours represent 15% increments.

Figure 5.9(b) represents the spatial response of the detector to $10.6 \mu\text{m}$ radiation, if we approximate the spatial response to a 3D-Gaussian function we can determine that 85 % of the total volume of the Gaussian is enclosed under the 0.2 z -contour level. If we define the collection area as the area where 85 % of the collected energy falls, then from Figure 5.9(b) we can determine that the collection area of these

dipole-coupled microbolometers would be approximately $10 \mu\text{m}^2$.

Figure 5.10 shows the radiation pattern of a dipole-coupled niobium microbolometer. This power pattern is related to the effective area of the detector by Eq.(1.21), where the maximum directivity can be obtained from the power pattern as

$$D_{\max} = \frac{4\pi U_{\max}}{P_{\text{rad}}}, \quad (5.2)$$

where $U_{\max}$ is the maximum radiation intensity and P_{rad} is the total radiated power. If we want to calculate the directivity of a printed antenna in the normal direction, then $U_{\max}$ will be the value of the power pattern at 90° and the total power can be obtained by calculating the volume under the power pattern $P(\theta)$,

$$P_{\text{rad}} = 2\pi \int_0^\pi P(\theta) \sin(\theta) d\theta. \quad (5.3)$$

Equation (5.3) assumes that the power pattern is omnidirectional (it does not depend on the spherical coordinate ϕ). For a more accurate calculation of the total power radiated by an antenna a two-dimensional radiation pattern (θ, ϕ) needs to be measured.

By using Eqs. (1.21), (5.2) and (5.3) on the power pattern shown in Fig. 5.10 we find that the directivity in the normal direction for this dipole antenna is 1.7 and the calculated effective area is around $52 \mu\text{m}^2$, the difference between measured and calculated effective areas is due to reflection and dielectric conduction losses, therefore

the radiation efficiency of this antenna is the ratio of measured and calculated effective areas, which is around 20 %.

Figure 5.10. Radiation pattern of a dipole-coupled microbolometer.

These dipole-coupled niobium microbolometers presented an average dc-resistance of $80 \pm 5 \Omega$, were illuminated using a laser beam at $10.6 \mu\text{m}$ and $F/8$ optics, with a power at the focal plane of 43.4 mW and an irradiance of 117 W/cm^2 . The measured response to the incident infrared radiation was around $11 \pm 2 \mu\text{V}$ for the polarization parallel to the dipole antenna, the noise was around $29 \frac{\text{nV}}{\sqrt{\text{Hz}}}$, which gives a signal-to-noise ratio (SNR) of ~ 380 , a Responsivity of 0.8 V/W and a D^* of $8.6 \times 10^3 \frac{\text{cm}\sqrt{\text{Hz}}}{\text{W}}$.

5.2 Bowties

If an antenna is made using perfect conductors and dielectrics and its dimensions change, the characteristics of that antenna (impedance, polarization, radiation pattern, etc) will remain the same as long as the wavelength of operation is changed in the same amount. Therefore, if the shape of an antenna is determined only by angles, the performance of that antenna would be independent of frequency since it would be invariant to a change of scale [52], bowties, spirals and log-periodic antennas are examples of frequency independent antennas.

The main advantages of bowtie antennas are simple design and broad-band impedance. A bowtie antenna is made from a bi-triangular sheet of metal with the feed at its vertex. This type of antenna, which is only defined by the bow angle θ , would be frequency independent if it extended to infinity on both sides. To fabricate a practical bow-tie antenna we must have a finite gap between the feed points and

a finite size which would result in limited bandwidth, however typically the antenna can be terminated with a bow-arm length of $2\lambda_{eff}$ without a significant effect on the pattern or the impedance [11].

The radiation of a bowtie antenna is linearly polarized and has a bidirectional pattern with broad main beams perpendicular to the plane of the antenna. The impedance can be calculated accurately from transmission line theory[40] and is given by:

$$Z = \sqrt{\frac{2\mu_0}{\epsilon_{eff} + \epsilon_0}} \cdot \frac{K(k)}{K'(k)}, \quad (5.4)$$

where

$$K(k) = \int_0^1 \frac{dx}{\sqrt{(1-x^2)(1-k^2x^2)}}, K'(k) = K(k'), \quad (5.5)$$

and

$$k' = \sqrt{1-k^2}. \quad (5.6)$$

K and K' are elliptic integrals of the first kind, $k = \tan^2(45^\circ - \theta/4)$ and θ is the bow angle. Figure 5.11 shows the impedance of a bowtie antenna on a silicon substrate as a function of the flare angle, this impedance is purely real for any bow angle, this behavior has been demonstrated experimentally in [40].

Figure 5.12 shows an electron micrograph of a planar bowtie antenna similar to the one used in this study. The antenna is $4 \mu\text{m}$ long and has a bow angle of around

Figure 5.11. Impedance of a Bowtie antenna on a Silicon substrate as a function of the flare angle.

60 degrees which would yield an impedance close to 75Ω , intended to match the impedance of the sensing Nb patch.

Figure 5.13 shows the polarization response of a bowtie-coupled microbolometer, this response also follows a cosine squared dependence, characteristic of linearly polarized antennas and it presented a polarization ratio close to 17.

Figures 5.14(a) and 5.14(b) show a two dimensional scan over a bowtie-coupled microbolometer and the spatial response of the same device to $10.6 \mu\text{m}$ radiation after deconvolution. From Figure 5.14(b) we can see that the effective collection area of the bowtie-coupled microbolometer is close to $\sim 14 \mu\text{m}^2$. Figure 5.15 shows the radiation pattern of a bowtie-coupled niobium microbolometer, the directivity in the normal direction for this power pattern is 1.2 and the calculated effective area is

Figure 5.12. Bowtie-coupled microbolometer.

Figure 5.13. Polarization Dependence for a bowtie-coupled microbolometer.

around $37.5 \mu\text{m}^2$ the radiation efficiency of this antenna, comparing the calculated and the measured effective area, is around 37 %.

Figure 5.14. Bowtie spatial response (a) convolved with the laser beam, (b) after deconvolution. The contours represent 15% increments.

These bowtie-coupled niobium microbolometers presented an average dc-resistance of $90 \pm 5 \Omega$, were illuminated using a laser beam at $10.6 \mu\text{m}$ and $F/8$ optics, with a power at the focal plane of 43.4 mW and an irradiance of 117 W/cm^2 . The measured response to the incident infrared radiation was around $14 \pm 1 \mu\text{V}$ for the polarization parallel to the bowtie antenna, the noise was around $29 \frac{\text{nV}}{\sqrt{\text{Hz}}}$, which gives a signal-to-noise ratio (SNR) of ~ 482 , a Responsivity of 0.72 V/W and a D^* of $9.3 \times 10^3 \frac{\text{cm}\sqrt{\text{Hz}}}{\text{W}}$.

Figure 5.15. Radiation Pattern of a bowtie-coupled niobium microbolometer.

5.3 Square Spirals

The equiangular spiral is a geometrical configuration that can be described just by angles, therefore it fulfills the requirement for shapes that can be used to design frequency independent antennas. The lower limit of their frequency bandwidth depends on the outer circumference of the spiral and the upper frequency limit depends on the configuration near the feed point. Spiral antennas are circularly polarized with a sense that depends on the winding sense of the spirals. A square spiral antenna is regarded as a counterpart of a round spiral antenna with the advantage that it offers a size reduction of 22 % (based on the increase of perimeter from πD to $4D$)[60]. The radiating characteristics of this antenna can be explained by modelling it as a two-wire transmission line. A current wave traveling along the arms of the antenna radiates energy which decreases the amplitude of the current along the conducting antenna arms, and beyond a certain point the presence or absence of the conductor

makes little difference. The point of negligible current occurs about one wavelength from the feed point[61], therefore when the wavelength is shorter than the arm length, the performance is practically frequency independent. If the antenna arms become shorter than the wavelength the polarization becomes elliptical where its axial ratio increases by decreasing the arm length until the polarization becomes linear and the spiral becomes a dipole.

Figure 5.16. Geometry of a two-wire square spiral antenna.

Figure 5.16 shows the geometry of a two-wire square spiral antenna, the antenna arms A and B are composed of linear filaments wound in the X-Y plane where the length of the n_{th} -filament is given by:

$$L_n = \begin{cases} a & \text{for } n = 1 \\ 2a(n - 1) & \text{for } n = 2, 3, \dots \end{cases} \quad (5.7)$$

A planar spiral antenna with arms $17\text{ }\mu\text{m}$ long, 200 nm wide and 100 nm thick made out of gold with a Nb microbolometer in the feed point is shown in Figure 5.17. This antenna was fabricated on a 200 nm SiO_2 substrate and a Cr ground plane.

Figure 5.17. Spiral-coupled microbolometer.

Figure 5.18 shows the polarization response of a square-spiral-coupled microbolometer to linearly polarized radiation at $10.6\text{ }\mu\text{m}$, spiral antennas have circular polarization for wavelengths smaller than the arm length, this spiral-antenna presented a linear-polarization ratio close to 1.5 which indicates that the polarization of the an-

tenna is actually elliptical, this ellipticity might be due to contributions to the total response of the bias leads acting as long dipoles which are linearly polarized. The orientation of the principal axis of the elliptical polarization corresponds to the angle where the maximum response is found.

Figure 5.18. Polarization dependence of a Square-spiral-coupled microbolometer.

Figures 5.19(a) and 5.19(b) show the two-dimensional scan of a spiral-coupled microbolometer and its deconvolution with the laser beam to find the effective collection area of the detector. From Figure 5.19(b) we can see that $\sim 85\%$ of the maximum infrared radiation is collected on an area of $\sim 12.5\ \mu\text{m}^2$. Figure 5.20 shows

the radiation pattern of a spiral-coupled niobium microbolometer, the directivity in the normal direction for this power pattern is 1.6 and the calculated effective area is around $49 \mu\text{m}^2$ which gives a radiation efficiency close to 25 %.

Figure 5.19. Spiral spatial response (a) convolved with the laser beam, (b) after deconvolution. The contours represent 15% increments.

Figure 5.20. Radiation Pattern of a spiral-coupled niobium microbolometer.

The spiral-coupled niobium microbolometers were measured using a $10.6\ \mu\text{m}$ CO_2 laser and $F/1$ optics, the power at the focal plane was $8.7\ \text{mW}$ with an irradiance of $1506\ \text{W}/\text{cm}^2$, the detectors presented an average dc-resistance of $1.4 \pm 0.2\ \text{k}\Omega$. The measured response to the incident infrared radiation was around $1.8 \pm 0.2\ \text{mV}$, the noise measured was $160\ \frac{\text{nV}}{\sqrt{\text{Hz}}}$, which gives a signal-to-noise ratio (SNR) of $\sim 11,250$, a Responsivity of $9.5\ \text{V}/\text{W}$ and a D^* of $2.1 \times 10^4\ \frac{\text{cm}\sqrt{\text{Hz}}}{\text{W}}$.

5.4 Log-Periodic

If a special kind of repetitiveness in the physical structure of an antenna is introduced a repetitive behavior of the electrical characteristics will be obtained. In log-periodic antennas the geometry is chosen so that the electrical properties repeat themselves with the logarithm of the wavelength. Frequency independence can be obtained when variation of the electrical characteristics over one period is small.

The log-periodic toothed planar antenna is similar to the bowtie antenna except for the teeth. The teeth act to disturb the currents which would flow if the antenna were of bowtie-type construction. The radiation is linearly polarized parallel to the teeth edges, this is perpendicular to what it would be if there were no teeth, in which case the antenna would be a bowtie. This shows that the component of current that flows in the direction of the teeth dominates the radial component[52].

Figure 5.21 shows a log-periodic toothed planar antenna, if $\beta_1 + \beta_2 = 90^\circ$ the antenna becomes self-complementary in which case it will have a constant impedance of 189Ω at all frequencies[52]. Also for a log-periodic toothed planar antenna the ratio of edge distances is constant and is given by the following scale factor:

$$\tau = \frac{R_{n+1}}{R_n}, \quad (5.8)$$

the parameter τ gives the period of the structure, therefore a periodic pattern and impedance behavior with the same period would be expected.

Figure 5.22 shows an electron micrograph of a planar log-periodic antenna with a scale factor of $\tau \approx 2$ coupled to a niobium microbolometer, this antennas were fabricated on 200 nm of SiO_2 and a chrome ground plane, the antenna and bias lines are made out of 100 nm of evaporated gold and the microbolometer is made out of a 70 nm film of sputtered niobium, the microbolometer is placed at the feed point of the antenna. The antenna is resonant when the length of any of the arcs A_n (as shown in Fig. 5.21) is equal to $\frac{\lambda_{eff}}{2}$, the arc length can be calculated as:

Figure 5.21. Log-periodic toothed planar antenna.

$$A_n = \left(\frac{\pi}{2}\right) \left(\frac{R_n + r_n}{2}\right), \quad (5.9)$$

for $\beta_1 + \beta_2 = 90^\circ$. The corresponding resonant frequencies are:

$$f_n = \frac{2c}{\pi (R_n + r_n) \sqrt{\epsilon_{eff}}}. \quad (5.10)$$

The log-periodic antenna shown in Fig. 5.22 was designed to have a frequency coverage of 18 - 70 THz which corresponds to a wavelength coverage of 4-16 μm .

Figure 5.22. Log-periodic antenna coupled to a microbolometer.

Figure 5.23 shows the polarization response of a log-periodic-coupled microbolometer, this response also follows a cosine squared dependence characteristic of linearly polarized antennas, the maximum response is obtained for a polarization parallel to the dc leads which is perpendicular to the polarization of a bowtie and a dipole, the polarization ratio close to 1.4. This low polarization ratio is due to the cross-polarized response given by the bowtie-like part of the log-periodic structure which competes with its co-polarized response, this vector addition of polarizations does not affect the magnitude of the co-polarized response.

Figure 5.23. Polarization dependence for a log-periodic antenna-coupled detector.

Figures 5.24(a) and 5.24(b) show the two-dimensional scan of a log-periodic-coupled microbolometer and its deconvolution with the laser beam to find the effective collection area of the detector. From Figure 5.24(b) we can see that $\sim 85\%$ of the maximum infrared radiation is collected on an area of $\sim 21 \mu\text{m}^2$. Figure 5.25 shows the radiation pattern of a log-periodic-coupled niobium microbolometer, the directivity in the normal direction for this power pattern is 1.5 and the calculated effective area is around $45.6 \mu\text{m}^2$ which gives a radiation efficiency close to 46 %.

Figure 5.24. Spatial response for a log-periodic antenna (a) convolved with the laser beam, (b) after deconvolution. The contours represent 15% increments.

Figure 5.25. Radiation pattern of a log-periodic-coupled niobium microbolometer.

The log-periodic-coupled niobium microbolometers were measured using a $10.6\text{ }\mu\text{m}$ CO_2 laser and $F/1$ optics, the power at the focal plane was 8.7 mW with an irradiance of 1506 W/cm^2 , the detectors presented an average dc-resistance of $400 \pm 50\text{ }\Omega$. The measured response to the incident infrared radiation was around $0.9 \pm 0.2\text{ mV}$, the noise measured was $80\text{ }\frac{\text{nV}}{\sqrt{\text{Hz}}}$, which gives a signal-to-noise ratio (SNR) of ~ 11250 , a Responsivity of 2.84 V/W and a D^* of $1.6 \times 10^4\text{ }\frac{\text{cm}\sqrt{\text{Hz}}}{\text{W}}$.

CHAPTER 6

ANTENNA-COUPLED IR PIXELS

Antenna-coupled microbolometers can have collection areas as small as $10 \mu\text{m}^2$. Detectors this small are unsuitable for imaging applications where a typical pixel area ranges from $20 \times 20 \mu\text{m}^2$ to $50 \times 50 \mu\text{m}^2$ [62]. In this chapter two different types of detectors that can cover a typical pixel area are fabricated and their characteristics measured. The first type of IR pixel is a two-dimensional array of antenna-coupled microbolometers that will increase the collection area of a single element to cover a whole pixel area, the second IR pixel is a Fresnel Zone Plate (FZP) coupled microbolometer. A FZP is used to collect the radiation that falls on a pixel area and will focus it on a single element microbolometer. These FZPs are fabricated using optical lithography and liftoff and were aligned to the microbolometers by using a back-side aligner. Responsivity and radiation characteristics of these two types of pixels were measured and compared to single element microbolometers.

6.1 2D Array of Antenna-Coupled Microbolometers.

A two-dimensional array of antenna-coupled microbolometers that will increase the area of the detector without sacrificing the response and time constant characteristic of microbolometers is proposed, fabricated and its performance compared to single element antenna-coupled microbolometers.

6.1.1 Antenna Array Theory

An array of antennas is an assembly of radiating elements such that the radiation from the elements “adds up” to give a maximum field intensity in a particular direction or directions and cancels or very nearly cancels in others. In addition to placing elements along a line (to form a linear array), individual radiators can be positioned along a rectangular grid to form a rectangular or planar array (Figure 6.1). Planar arrays provide additional variables which can be used to control and shape the pattern of the array. Planar arrays are more versatile and can provide more symmetrical patterns with lower sidelobes. In addition they can be used to scan the main beam of the antenna toward any point in space by controlling inter-element phase.

Figure 6.1. Rectangular array of antennas. From [1].

The total field of the array is equal to the field of a single element positioned at the origin multiplied by a factor which is widely referred to as the “Array Factor” (AF) [1]:

$$AF_n(\theta, \phi) = \left\{ \frac{1}{M} \frac{\sin\left(\frac{M}{2}\psi_x\right)}{\sin\left(\frac{\psi_x}{2}\right)} \right\} \left\{ \frac{1}{N} \frac{\sin\left(\frac{N}{2}\psi_y\right)}{\sin\left(\frac{\psi_y}{2}\right)} \right\}, \quad (6.1)$$

where $\psi_x = kd_x \sin \theta \cos \phi + \beta_x$, and $\psi_y = kd_y \sin \theta \sin \phi + \beta_y$, M and N are the number of antennas in the x and y directions respectively, d_x and d_y are the distance between elements in the x and y directions and β_x and β_y are the progressive phase differences between elements in the x and y directions.

When the spacing between the elements is equal or greater than $\lambda/2$, multiple

maxima of equal magnitude can be formed. The principal maximum is referred to as the “major lobe” and the remaining as the “grating lobes”. To avoid grating lobes in the xz and yz planes, the spacing between the elements in the x and y directions, respectively, must be less than $\lambda/2$ ($d_x < \lambda/2$ and $d_y < \lambda/2$). From Eq. 6.1 we can also see that the maximum value of the array factor is obtained when the distance between the elements is $\lambda/2$.

In the case of antenna-coupled microbolometers, currents at THz frequencies are induced in the antenna elements by infrared radiation and are dissipated in the microbolometer which, being a power detector, converts that energy into a dc signal. Therefore any phase information is lost, however some of the energy flows out of the antenna through the substrate and can couple to adjacent antennas providing an “array effect”.

6.1.2 Element Spacing

In order to optimize the response of a rectangular array of antenna-coupled detectors the characteristics of a single element and the spacing between the elements that maximizes the response to normally incident infrared radiation has to be obtained. According to antenna-array theory optimum radiation characteristics in the direction perpendicular to the planar array are obtained with a spacing of $\lambda/2$, however Fumeaux et al. [63] also discovered that at infrared frequencies the surface impedance of metals play an important role in the behavior of THz currents shortening the wave-

length of the propagating current and therefore shifting the resonant frequency of dipole antennas. The fact that these effects are not taken into account in equations that appear in antenna theory books makes it necessary to find the optimum spacing and length of elements in an infrared antenna array experimentally.

Rana and Alexopoulos calculated the mutual coupling between two broadside resonant microstrip dipoles on thin substrates[64], Figure 6.2 shows the mutual impedance between two parallel broadside dipoles as a function of their separation, there we can see that the resonances (where the mutual impedance is purely real) are located at a distance close to 0.3, 0.85 and 1.3 effective wavelengths. The effective wavelength is given by $\lambda_{eff} = \frac{\lambda_0}{\sqrt{\epsilon_{eff}}}$, where $\epsilon_{eff} = \frac{\epsilon_{SiO_2} + 1}{2}$.

A distance study of dipole-coupled microbolometers was performed using gold for the antenna elements and niobium for the bolometric material on a 200 nm layer of SiO₂ ($\epsilon_{SiO_2} = 4.84$ at 10.6 μm) as substrate and a 50 nm chrome ground plane. Figure 6.3 shows the response of these two broadside dipoles to 10.6 μm radiation for different spacings. There is a peak in the response at 2.4 μm which is 0.38 effective wavelengths, close to the first resonance of the mutual coupling of two broadside dipoles found in [64]. Interestingly enough 2.4 μm is a half wavelength in the dielectric, as if the antennas were not at an air-dielectric interface ($\frac{\lambda_0}{2\sqrt{\epsilon_{SiO_2}}} = \frac{10.6\mu\text{m}}{2\sqrt{4.84}} = 2.4 \mu\text{m}$), this indicates that coupling between printed dipoles at an air-dielectric interface may be caused by surface-waves and there might be phase addition at the surface-wave level that increases the response at half the wavelength in the substrate, inducing an

Figure 6.2. Calculated mutual impedance between two broadside microstrip dipoles. From [64].

antenna-array effect.

Figure 6.3. Coupling of dipoles as a function of distance.

6.1.3 Response and Noise analysis

A series array of microbolometers was chosen to match the impedance of commercial readout integrated circuits (ROIC's), and to cover the typical pixel area used in commercial infrared imaging systems. The Johnson noise of a microbolometer is given by

$$V_j = \sqrt{4KTR\Delta f}, \quad (6.2)$$

where K is the Boltzmann constant, T is the Temperature in K, R is the resistance of the microbolometer and Δf is the bandwidth of the measurement. By making an $N \times N$ array of microbolometers the total resistance of the detector will increase by a factor of N^2 . From Eq. (6.2), we can see that this will result in a $N \times$ increase in Johnson noise. However from Eq. (1.13) we can see that the response is directly proportional to the resistance of the microbolometer, therefore a resistance increase of a factor of N^2 will increase the response in the same amount giving an anticipated increase of $N \times$ in signal-to-noise ratio of a serial array compared to a single element microbolometer.

6.1.4 Experimental Results

Four different types of serial arrays of microbolometers that can cover the area of a picture element in an infrared imaging system were fabricated and tested. Dipoles and bowties in two different configurations were used, those parallel to the bias current (Figs. 6.5 and 6.6) will be called parallel-dipole and bowtie arrays and those with elements perpendicular to the bias current (Figs. 6.7 and 6.8) will be called perpendicular-dipole and bowtie arrays respectively. All these devices were fabricated on a 200 nm SiO_2 substrate and a chrome ground plane, the antenna elements are made out of gold and niobium is used as the bolometric material. These ar-

Figure 6.4. Series array of bolometers.

rays cover an area of $50 \times 50 \mu\text{m}^2$ and the spacing between elements is $5 \mu\text{m}$, this spacing represents 0.8 effective wavelengths which is close to the second surface-wave resonance (Fig. 6.2).

Knowing that antenna-coupled microbolometers respond to visible frequencies [65], a two-dimensional scan was performed using a HeNe laser in the visible, these type of measurements are an effective way of characterizing the homogeneity of the detectors and the effectiveness of the fabrication process[43]. The beam of the HeNe laser was focused at the center of the array by moving the detector along the three axes and maximizing its response. The two-dimensional scan was performed by keeping the z axis fixed, moving the device in the x and y directions and recording the response of the device in each position. Figure 6.9 shows a two-dimensional scan of

Figure 6.5. Parallel Dipole Array.

Figure 6.6. Parallel Bowtie Array.

Figure 6.7. Perpendicular Dipole Array.

Figure 6.8. Perpendicular Bowtie Array.

a 5×5 array of microbolometers. From the scan the individual response of each element of the array can be observed and determine the uniformity of the single element microbolometers that form the array, this type of measurement can be used as a diagnostic tool to determine the quality of the array and detect possible defects in the fabrication process.

Figure 6.9. Two-dimensional scan in the visible for a 5×5 array of microbolometers.

A two-dimensional scan in the infrared was also performed on a single element, 5×5 array and 10×10 array of microbolometers to determine the detection area covered by each device. The two-dimensional scans were taken following the same procedure used for the scan in the visible, but using a CO_2 laser at $10.6 \mu\text{m}$ and $F/1$ optics. The $F/1$ optics presented an almost diffraction-limited focal spot of $12 \mu\text{m}$ in radius. Figures 6.10(a),(b) and (c) show $80 \times 80 \mu\text{m}^2$ scans on a single element, 5×5 array and 10×10 array respectively. The scan on the single element (Figure 6.10(a)) showed only the beam profile, which indicates that the single element microbolometer can be considered as nearly a point receiver at infrared frequencies. Figures 6.10(b) and (c) show how the detection area increases by going from a single element to a two-dimensional array of microbolometers. A uniform detection area of approximately $25 \times 25 \mu\text{m}^2$ and $50 \times 50 \mu\text{m}^2$ are shown in Figures 6.10(b) and (c) respectively, these areas correspond to the physical size of the receivers. From the results shown in Figure 6.10 we can see that two-dimensional arrays of microbolometers make uniform area detectors and can be used as pixel elements in infrared imaging systems.

Figure 6.11 shows a comparison of the response for the different types of two dimensional arrays of microbolometers to infrared radiation, the parallel arrays presented a higher response than the perpendicular array of detectors, and the bowties had a higher response than the dipoles. All the measurements were made with a CO_2 laser at $10.6 \mu\text{m}$ and $F/8$ optics at the optimum polarization for each type of detectors, the optimum polarization for all the detectors was the one parallel to the

Figure 6.10. Two-dimensional scan in the infrared (CO_2 laser at $10.6\mu\text{m}$), for (a) Single Element, (b) 5×5 array and (c) 10×10 array of microbolometers. The dimensions are given in microns.

antenna arms, therefore the optimum polarization for the parallel arrays is 90 degrees rotated with respect to the optimum polarization of the perpendicular arrays.

In order to test the performance of a single element compared to an array of detectors, a single element square-spiral antenna (Fig. 6.12(a)) and a 9×9 array of square-spirals (Fig. 6.12(b)) were fabricated and tested. Single element detectors had a dc resistance around $80\ \Omega$ and were biased at 100 mV (bias current of 1.25 mA) while the arrays had a resistance around $1.3\ \text{k}\Omega$ and were biased at 300 mV (bias current of $230\ \mu\text{A}$). Noise and response measurements at $10.6\ \mu\text{m}$ were made, single element square spiral antennas gave a response of $10 \pm 1\ \mu\text{V}$ and a noise figure of $3.5 \pm 0.3\ \frac{\text{nV}}{\sqrt{\text{Hz}}}$. The 2D-Array of square spiral antennas presented a response of $25 \pm 3\ \mu\text{V}$ with a noise figure of $5.5 \pm 0.8\ \frac{\text{nV}}{\sqrt{\text{Hz}}}$, however these devices were biased at a current $5.4\times$ lower than the single elements, if we normalize to the bias current of the single element detectors the response of the arrays would be around $135\ \mu\text{V}$. The

Figure 6.11. Voltage response of different 2D-Arrays of antenna-coupled microbolometers to infrared radiation.

SNR for a single element detector was ~ 2860 , while for the two-dimensional array was ~ 4550 (normalized to ~ 24500), however D^* which is a figure of merit that permits comparison of detectors of the same type but having different areas gave a value of $3.2 \times 10^4 \frac{\text{cm}\sqrt{\text{Hz}}}{\text{W}}$ for the single element detector and $5 \times 10^3 \frac{\text{cm}\sqrt{\text{Hz}}}{\text{W}}$ for the arrays (normalized to $2.7 \times 10^4 \frac{\text{cm}\sqrt{\text{Hz}}}{\text{W}}$), therefore going from single element detectors to two-dimensional arrays will increase the signal-to-noise ratio but not D^* , which is the main figure of merit for infrared detectors.

According to Eq. (6.2) the Johnson noise for the single element detectors tested is $1.1 \frac{\text{nV}}{\sqrt{\text{Hz}}}$ and for the detector array is $4.6 \frac{\text{nV}}{\sqrt{\text{Hz}}}$, from the noise measurements we can see how $1/f$ noise is dominant in single element detectors and by going to an array the $1/f$ noise contributions of every single element does not add in quadrature like Johnson noise does, this seems to indicate that the noise sources are correlated to each other through the fabrication process. The increase in response between single element detectors and arrays normalized to the bias current is around $13.5\times$ which is close to the ratio of dc resistances ($16.5\times$), this 20% difference is due to the contribution of the bias lines to the dc resistance but not to the bolometric response. It is also worth noting that the signal to noise ratio had an increase of around $8.6\times$ which is close to the $9\times$ theoretical increase ($N\times$ for an $N \times N$ array).

6.2 Fresnel Zone Plate Lens

A Fresnel Zone Plate Lens (FZPL) was fabricated to collect infrared energy and focus it on a single antenna-coupled microbolometer, keeping the noise level low and still having a large collection area. The performance of these devices is compared to that of a two-dimensional array of detectors. The FZP lenses were patterned using optical lithography and aligned to single element square-spiral-coupled microbolometer on the backside of the wafer using an EV620 backside aligner. The FZP lenses were made out of 100 nm of e-beam evaporated Au over a 5 nm layer of Cr. The FZP lens was designed to work at a wavelength of $10.6\text{ }\mu\text{m}$ with a focal length of $380\text{ }\mu\text{m}$, close to the substrate thickness so that the antenna-coupled detector would be at its focus(Figure 6.13). The main function of the zone plate is to increase the gain of the spiral antenna and also reduce the energy loss due to guided waves in the substrate by altering the boundary conditions of the dielectric slab waveguide[66].

Fresnel zone plates work by concentrating energy of an incident plane wave by blocking the portions that would add destructively at the focal point. This effect can be explained using the Huygens-Fresnel principle where the wavefront is replaced by point sources in the open region of the zone plate. The geometry of the zone plate is such that these point sources emit waves that arrive at the focal point with a relative phase between 0 and π . In contrast, if point sources are placed in the adjacent blocked regions of the zone plate, their waves would arrive at the focal point with a relative phase between π and 2π . Since these waves would add destructively to the field at

Figure 6.12. Square Spirals. (a) Single Element, (b) Two dimensional array.

Figure 6.13. Fresnel Zone Plate Lens in the transmissive configuration coupled to a microbolometer.

the focal point they are blocked by an opaque zone.

The determination of the open and opaque regions is cylindrically symmetric about the zone-plate/focal-point axis. This cylindrical symmetry gives the zone plate its characteristic light and dark rings. The boundary between the open and opaque rings can be determined by geometry, by calculating the phase difference that will result from light traveling different distances. First the focal length f of the zone plate is chosen, and then the radii of the zone plate rings is determined. The equation for the n^{th} radii r_n is given by [67]

$$r_n = \sqrt{nf\lambda + \left(\frac{n\lambda}{2}\right)^2}. \quad (6.3)$$

This zone plate is called a transmission zone plate since it concentrates the wave as it passes through the open rings (Fig. 6.14(a)).

A variation of the zone plate is made by placing a reflecting surface in the opaque regions. In this case another focal point is formed on the incident-wave side of the zone plate (Fig. 6.14(b)), this variation is called a reflection zone plate since it concentrates the wave by reflecting it from the opaque rings.

Two different types of Fresnel Zone Plate Lenses were fabricated, the traditional ones which consist of concentric rings (Figure 6.15) and an approximation to these made using concentric squares (Figure 6.16). Eight circular and five square FZPs were fabricated, these lenses vary in the number of zones each one has, circular FZPs go from 1 to 8 opaque zones while square FZPs go from 1 to 5 opaque zones. Figure

Figure 6.14. Illustration of (a) transmission zone plate, (b) reflection zone plate.

6.17 shows the results obtained by using circular FZP lenses in the transmissive and reflective configuration (the reflective configuration would be with the radiation coming through the detector first and then reflecting off the FZP), these results are normalized to the response obtained with a detector without an FZP, in this case the transmissive circular FZP gave a higher response than the reflective one, the response does not increase more with lenses above the sixth opaque zone because the lenses become bigger than the $F/8$ spot used to test them.

By increasing the number of Fresnel zones the response of the detector increases. However the collection area of the detector, which is the area covered by the largest zone plate, also increases. Figure 6.18 shows the gain of a circular FZP in the transmissive mode as a function of the number of opaque Fresnel zones, next to each of

Figure 6.15. Circular FZPs.

Figure 6.16. Square FZPs.

Figure 6.17. Results for Circular FZPs.

the measured points is the ratio of the calculated D^* for the FZP-coupled detector and a detector without an FZP, from these results we can see that we get an increase in D^* after the fifth opaque zone which is around $200\ \mu\text{m}$ in diameter. Figure 6.18 also shows the expected gain increase for a transmissive circular zone plate, which is proportional to N^2 where N is the number of zones. From these results we can see that a $50 \times 50\ \mu\text{m}^2$ FZP-coupled infrared pixel gives a low D^* , making them not the best option for infrared imaging systems.

Figure 6.19 shows the results obtained by using square FZP lenses in the transmissive and reflective configuration, these results are also normalized to the response obtained with a detector without an FZP, in this case the reflective FZP gave a higher response than the transmissive one, since these square FZPs are approxima-

Figure 6.18. Gain of a circular FZP in the transmissive mode and normalized D^* as a function of Fresnel zones.

tions to the circular ones the higher response in the reflective case could be due to the fact that the gold structure is more efficient as a simple metal reflector than as a diffraction-based concentrator. The square FZPs are designed by making squares that have the same area as its circular counterparts, this approximation works better for some square dimensions than for others, that is why the gain oscilates and does not increase monotonically.

Figure 6.19. Results for Square FZP's.

A comparison was made between a $50 \times 50 \mu\text{m}^2$ 2D-array of square spirals and a single antenna with a circular FZP with a diameter of $223 \mu\text{m}$ and consisting of 6 opaque fresnel zones. The 2D-Array of square spiral antennas presented a response of $25 \pm 3 \mu\text{V}$ with a noise figure of $5.5 \pm 0.8 \frac{\text{nV}}{\sqrt{\text{Hz}}}$, while the FZP-coupled detectors presented

a response of 1.1 ± 0.1 mV and the noise figure was $3.5 \pm 0.2 \frac{\text{nV}}{\sqrt{\text{Hz}}}$, which translates into an SNR of ~ 455 and D^* of $5 \times 10^3 \frac{\text{cm}\sqrt{\text{Hz}}}{\text{W}}$ for the 2D array and an SNR of $\sim 33 \times 10^3$ and D^* of $9.5 \times 10^4 \frac{\text{cm}\sqrt{\text{Hz}}}{\text{W}}$ for the FZP-coupled detector.

The angular response of these two detectors was also measured, Fig. 6.20(a) shows the radiation pattern for a 2D array of square spirals and Fig. 6.20(b) shows the radiation pattern for an FZP-coupled square spiral microbolometer. The directivity of the 2D array is 1.6 with a $50 \times 50 \mu\text{m}^2$ area and the FZP has a directivity of 4.1 for an area of $31,415 \mu\text{m}^2$ (100 μm -radius zone). With the FZP-coupled detector we gain $2.5\times$ in directivity with an area increase of $12.5\times$, which indicates that the 2D array is a more efficient area receiver than the FZP-coupled detector.

Figure 6.20. Radiation Patterns. (a) 2D Array of Square Spirals. (b) FZP-coupled square spiral microbolometer.

CHAPTER 7

INTEGRATION TO COMMERCIAL READOUT

INTEGRATED CIRCUITS

Commercial Readout Integrated Circuits (ROIC's) were provided by Raytheon to integrate antenna-coupled pixels monolithically onto them and make an Antenna-coupled Infrared Focal Plane Array (IR-FPA). The ROIC's used had a $1.2\text{ }\mu\text{m}$ layer of SiO_2 and a 500 nm layer of Si_3N_4 as passivation layers (Fig. 7.1). In order to avoid a high step profile between the detectors and the ROIC this passivation layer was thinned down to 250 nm using CF_4 -based RIE. One of the most involved fabrication tasks was finding the way to align the pattern of the antenna-coupled pixels to the ROIC's so that the contact-pads of the detectors would match the location of the contact openings on the ROIC. A CAD file in GDSII format was provided by Raytheon with the top-level metal layer of the ROIC (similar to the CAD file shown in Figure 7.2), with this file we could locate the exact coordinates of distinctive features on the ROIC (like letters, numbers or previous alignment marks) and align to those features during the e-beam patterning process.

Figure 7.1. Cross-section of a ROIC showing passivation layers.

The monolithic integration of antenna-coupled pixels to the ROIC started with the thinning of the passivation layer. Global and local alignment marks were then placed using e-beam lithography and liftoff, by aligning to existing structures on the ROIC. Openings on the passivation layer were made to uncover the ROIC's contact pads by using CF_4 -based RIE, the contact pad openings were patterned using e-beam lithography and PMMA was used to mask the RIE process. After contact openings were made on the ROIC the standard fabrication process was used to pattern the antenna-coupled pixels which consisted of a two-dimensional array of log-periodic and square-spiral antennas. Figure 7.3 shows one of the 8×8 pixel arrays fabricated on the Raytheon ROIC, and Fig. 7.4 is a pixel element of that 8×8 array.

After the monolithic integration of antenna-coupled pixels to the ROIC was performed, the antenna-coupled IRFPA was bonded and mounted in a dewar custom-made for that particular FPA (Fig. 7.5). Each IR-FPA had an 8×8 array of log-

Figure 7.2. Sample ROIC CAD file that can be used as an alignment aid in the fabrication of monolithic IRFPAs.

Figure 7.3. 8×8 Pixel array of antenna-coupled infrared detectors.

periodic and square-spirals pixels integrated (Fig. 7.6).

The IR-FPA was tested, using the same camera emulator used to test commercial infrared imaging systems based on the same ROIC, with a black-body at 100°C as the hot source. Figure 7.7 shows the image obtained with the 8×8 array of log-periodic antennas looking at the 100°C black-body, which shows that the integration of antenna-coupled infrared pixels to make an antenna-coupled IR-FPA was successful.

Figure 7.4. Antenna-coupled area receiver.

Figure 7.5. Dewar used to mount the antenna-coupled IRFPA.

Figure 7.6. Antenna-coupled IRFPA.

Figure 7.7. Hot source imaged by an 8×8 array of antenna-coupled area receivers.

CHAPTER 8

CONCLUSIONS AND FUTURE WORK

In this dissertation a new type of infrared focal plane array was fabricated and tested. The main detector array consisted of antenna-coupled microbolometers, integrated into a commercial Readout Integrated Circuit (ROIC) by using e-beam lithography and conventional microfabrication techniques. Several issues such as the use of different types of IR antennas, materials and fabrication processes to increase the responsivity of the detectors were investigated.

Single element antenna-coupled microbolometers were fabricated on Silicon substrates and tested for polarization dependence, radiation patterns, frequency response and collection efficiency. Four different types of IR antennas were investigated, dipoles, bowties, square-spirals and log-periodics. These single-element detectors presented a responsivity in the order of $1 - 2 \text{ V/W}$ and a D^* in the $10^4 \frac{\text{cm}\sqrt{\text{Hz}}}{\text{W}}$ range with time constants in the nanosecond range ($\sim 130 \text{ ns}$). Dipoles, bowties and log-periodic antennas showed a linear polarization dependence with polarization ratios of 4, 17 and 1.4 respectively, square spiral antennas showed an elliptical polarization dependence, this ellipticity could be due to the bias lines acting as long-wave dipoles, converting a

circularly polarized device into an elliptically polarized one. A relation between the radiation patterns of these antenna-coupled detectors and their collection area can give us an approximation to the efficiency of IR antennas, these collection efficiency ranged from 20 % for the dipole antennas to 46 % for the log-periodic antennas. The collection areas for these antenna-coupled detectors went from $10 \mu\text{m}^2$ for the dipole antennas to $21 \mu\text{m}^2$ for the log-periodic antennas, which are far from the $25 \times 25 \mu\text{m}^2$ to $50 \times 50 \mu\text{m}^2$ pixel areas used in commercial infrared imaging systems.

In order to cover the pixel area required by commercial readout integrated circuits, two different types of antenna-coupled pixels were designed and fabricated. The first one consisted of a two-dimensional array of microbolometers, it was determined by measurements that these arrays can cover any area without sacrificing time response which was measured to be in the nanosecond range for devices on a high thermal conductivity substrate. Taking advantage of their response to visible frequencies, a two-dimensional scan in the visible was performed on these detectors. These measurements proved to be an effective way of characterizing the homogeneity of the detectors and the effectiveness of the fabrication process.

An increase in response was measured on two-dimensional arrays compared to single element microbolometers, an additional increase in response can be obtained by adjusting the distance between elements to get electromagnetic coupling which will allow vector addition of the collected radiation by the individual antenna elements (antenna array effect). Even though an increase in signal-to-noise ratio was achieved

(the increase depended on the number of elements in the array), D^* which is the main figure of merit for infrared detectors and depends on the area of the detector did not increase from going from single element detectors to arrays, this is due to the difference in areas and to the fact that the collection area of a single element detector is smaller than its physical size.

A second type of antenna-coupled pixel was designed and fabricated by placing a Fresnel Zone Plate Lens (FZPL) aligned on the back side of the wafer to a single antenna-coupled detector. This FZPL acts as a concentrator, gathering the infrared energy that falls over its area and concentrating it at its focal point where the single element detector is located, by using a FZPL we increase the collection area without increasing the noise figure of the detector. These detectors gave a two-order of magnitude increase in signal-to-noise ratio compared to single element detectors, however an increase in D^* was observed only for FZPLs with more than $200\text{ }\mu\text{m}$ in diameter, which is larger than what commercial ROICs allow, the observed increase was of a factor of 2.

Several different materials and fabrication techniques were investigated in order to increase the responsivity of antenna-coupled detectors. Vanadium oxide (VOx) which consists of a mixture of various oxides has a high temperature coefficient of resistance (TCR). Microbolometers were fabricated using VOx as the sensing element and compared to devices made out of niobium (Nb), the response, noise and radiation patterns of Nb-based and VOx-based 2D arrays of log-periodic antenna-coupled

microbolometers were measured. The VOx-based devices showed a response $4.5\times$ higher and a $5.5\times$ better signal-to-noise ratio than the Nb-based devices. Measured radiation patterns showed that the gain in response and in signal-to-noise ratio could be further increased by better matching the impedance of the bolometric detector to the antenna elements which would yield an increase in response closer to the $\sim 10\times$ expected due to the better TCR of VOx compared to Nb thin films.

Thermal isolation was found to be the method that will enhance the response of antenna-coupled microbolometers the most. The use of silica aerogel, which has a thermal conductivity lower than air, resulted in a gain in responsivity and a noise increase that gave a one order of magnitude overall increase in signal-to-noise ratio as compared to the same type of devices fabricated on SiO₂ substrates, by using aerogel a value of D^* in the $10^5 \frac{\text{cm}\sqrt{\text{Hz}}}{\text{W}}$ range was measured. One of the interesting characteristics of silica aerogels is that its thermal conductivity can be controlled by varying its porosity, therefore it can be used as a substrate for antenna-coupled microbolometer arrays to maximize the responsivity of the detector for a specific frame-rate.

The maximum increase in response was obtained by fabricating antenna-coupled detectors on silicon nitride membranes, here the thermal isolation can be further increased if the device is put under vacuum, by using this procedure a value for D^* in the $10^7 \frac{\text{cm}\sqrt{\text{Hz}}}{\text{W}}$ range was obtained. It is worth noting that these detectors were fabricated using chrome (Cr) as the bolometric material, since VOx has a TCR $30\times$

higher by using it as the bolometer a value of D^* in the $10^8 \frac{\text{cm}\sqrt{\text{Hz}}}{\text{W}}$ range should easily be reached.

An 8×8 array of antenna-coupled pixels were fabricated on a commercial ROIC supplied by Raytheon, measurements on this antenna-coupled infrared focal plane array showed that the integration of antenna-coupled detectors to a commercial ROIC was possible and also resulted in the first image obtained with antenna-coupled detectors.

This dissertation shows how antenna-coupled microbolometers are a viable option for uncooled infrared imaging systems. State of the art uncooled detectors usually have D^* values in the $10^9 \frac{\text{cm}\sqrt{\text{Hz}}}{\text{W}}$ range, values in this order of magnitude can be obtained with antenna-coupled microbolometers by improving thermal isolation and using higher TCR materials. The thermal isolation of microbolometers can be increased by using titanium (Ti) for the bias lines, by doing this an increase in response of one order of magnitude should be obtained. In order to reach detectivities close to the ones measured for commercial devices a new membrane fabrication process has to be developed that will allow the use of Titanium and VOx as materials for microbolometers.

Also higher TCR materials could be used to further increase the response of antenna-coupled detectors, YBCO under certain conditions can reach TCRs higher than 50 %/K and VOx used in its metal-semiconductor transition region can reach TCRs as high as 200 %/K, this is a huge improvement considering that the VOx used

in this dissertation has a TCR of 3 %/K.

The use of antenna-coupled microbolometers for infrared detection has the big advantage of having the flexibility of tailoring the antenna element for a certain polarization or wavelength. A lot of work can be done in the design of novel antennas for infrared detection like multiple-band antennas and wavelength-agile antennas to name a couple. Figure 8.1 shows a picture of fractal-coupled microbolometers designed for infrared wavelengths.

Figure 8.1. Fractal antennas for infrared detection.

Another important issue that needs to be addressed is the need for a reliable simulation software or procedure to design antennas at THz frequencies and on a substrate, this will reduce design time and fabrication costs, simplifying the design cycle.

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