Chapter 2 MICROBOLOMETER THEORY

In this chapter, the basics of microbolometer operation are covered. The general theory of bolometer-type detectors is discussed in some detail, as well as their performance limitations. A thermal model for predicting bolometer performance will be introduced. This model and the analytical theory will then be related to the experimental results for a bismuth microbolometer.

2.1 Operation

It was pointed out in chapter 1 that microbolometer operation depends on a change in resistance as a function of temperature. To perform as a detector, there must be some method for introducing the radiation. A biasing network is also required to maintain current through the bolometer and to sense changes in its resistance. Figure 2.1 shows a bow-tie antenna-coupled microbolometer. This is a quasi-optical system in which the radiation to be detected is focussed on the antenna with a hemispherical lens. The radiation passes through both the lens and the substrate before reaching the antenna, where it induces a current which is dissipated in the microbolometer. To sense changes in the detector temperature, the leads on either end of the antenna are connected to a biasing circuit. A stable current source is provided by a battery in series with a large bias resistor. A lock-in amplifier or an oscilloscope may be used to read the detected signal.

Whereas conventional bolometers operate by a particle or photoelectric phenomenon, radiation is *electromagnetically* coupled to the planar antenna. In conventional bolometers, orientation of the electromagnetic field is not important. In most planar antennas, however, the degree of coupling between the radiation and the antenna strongly depends on the electromagnetic field orientation. If the radiation is properly aligned with the antenna, it will induce a current, as shown in



Fig. 2.1: Bow-tie antenna-coupled microbolometer. In this quasi-optical receiver, the radiation is focussed on the bow-tie antenna with a hemispherical lens. The coupled radiation is dissipated in the bolometer element, causing a temperature rise. The resulting change in resistance is sensed by the biasing network.

Fig. 2.2. This current passes back and forth through the microbolometer at the radiation frequency, and is dissipated by Joule heating.

Most of the microbolometer work done so far has utilized the bow-tie antenna [¹]. However, coupling to the detector with a microstrip-fed twin slot antenna has also been investigated, and this is covered in chapter 4. Other planar antennas which may be mated with microbolometer detectors are the vee [²,³] and the spiral [⁴,⁵]. For the bow-tie antenna on a dielectric, electromagnetic radiation is confined to within about two dielectric wavelengths (λ_{diel}) of the apex [⁶]. By reciprocity, the antenna therefore only receives radiation within about 2 λ_{diel} of this apex. This property of the bow-tie antenna allows for the extension of the bow arms for use as dc leads.



Fig. 2.2: Electromagnetic radiation induces current in the bow-tie antenna. Notice that orientation of the electromagnetic field is important.

A characteristic of planar antennas is that they absorb radiation better from the substrate side than the air side. This is most easily understood by looking at the antenna as a current source, as shown in Fig. 2.3. A current source will deliver maximum power to a short circuited load. It will therefore deliver more power to the lower impedance substrate than the higher impedance air. It turns out that more power will flow through the substrate than to the air by about an $\varepsilon_{r}^{3/2}$ to 1 ratio [⁷]. Again invoking reciprocity, the antenna will therefore receive more radiation from the substrate side.

The signal voltage v generated by a change in temperature for a microbolometer detector is

$$\mathbf{v} = \mathbf{I}_{\mathbf{b}} \left(d\mathbf{R}/d\mathbf{T} \right) \Delta \mathbf{T} \tag{2.1}$$



Fig. 2.3: In the transmission line equivalent circuit, a bow-tie antenna sees both the substrate and air. The substrate has a higher dielectric constant, and therefore a higher impedance. Since the current source representing the antenna will deliver more power to the larger impedance, more power is coupled to the substrate.

where I_b is the bias current, dR/dT is the change in resistance of the bolometer for a given change in T, and ΔT is the temperature change. By an electrical analogy, $\Delta T = P |Z_t|$, where the change in temperature ΔT , absorbed power P, and magnitude of the thermal impedance $|Z_t|$ are analogous to voltage, current, and resistance, respectively. Thus, responsivity r is given by

$$r = v/P = I_b (dR/dT) |Z_t|$$
 (2.2)

A characteristic of a microbolometer detector is its linear relation between resistance and dissipated power. This property assumes that dR/dT is constant over the temperature range considered, and that the temperature change in the device is proportional to changes in its dissipated power. This leads to an expression for dc responsivity [⁸], (the response of a detector to a step change in dissipated power),

$$\mathbf{r}_{dc} = \mathbf{I}_{b} \left(d\mathbf{R}/d\mathbf{P} \right) \,. \tag{2.3}$$

The dc curve of R plotted as a function of P therefore provides enough information to determine r and $|Z_t|$ for the device.

The microbolometer is a thermal detector; its thermal mass is too large to permit temperature changes from following the high frequency (>100 MHz) signal. However, the thermal mass is small enough for the microbolometer to follow a slower modulation frequency (<1 MHz). The next section shows how microbolometer responsivity is a function of modulation frequency.

2.2 General Theory

The general theory behind bolometer operation was first described in detail by Jones [9] in 1953. Other works have emerged since then which clarify the theory [10 , 11]. To aid in a discussion of this theory applied to a microbolometer, the variables used are summarized in Table 2.1.

The general case of a current biased bolometer will be treated. By choosing a current bias, current dependent resistance can be excluded from the derivation.

a	radius of hemisphere representing area of contact between the bolometer and the substrate	Т	temperature of the bolometer	
С	thermal capacity of bolometer	To	ambient temperature	
Cs	specific heat of the substrate	v	voltage across the bolometer	
G	thermal conductance between bolometer and surrounding thermal environment	Wo	radiation power dissipated by the bolometer	
Ib	bias current through the bolometer	ΔΤ	change in temperature	
Ic	critical current for thermal runaway	ΔTp	peak change in temperature	
r	responsivity	ρ _s	density of the substrate	
R	resistance across the bolometer	τ	time constant of the bolometer	
R _b	bias resistance of the bolometer	ω _m	2π times the chopping (or modulation) frequency	

Table 2.1: Variables associated with microbolometer theory.

Such a dR/dI term has been included in a similar derivation by Maul and Strandberg [¹²]. The voltage to be detected across the bolometer is

$$\mathbf{v} = \mathbf{I}\mathbf{b}\mathbf{R} \ . \tag{2.4}$$

The resistance will change as a function of temperature, and thus of absorbed power. With no radiation applied, temperature of the bolometer is governed by Joule heating from the bias current, and the conductance of heat out of the bolometer, as given in the heat balance equation

$$G(T-T_0) = I_b^2 R_b$$
 (2.5)

Here it is assumed thermal conductance accounts for all the heat loss. Now incident radiation is applied, and chopped at a modulation frequency ω_m , such that radiation dissipated by the bolometer W is

$$W = W_0 \sin \omega_m t \tag{2.6}$$

where W_0 is the unmodulated radiation power. For a conventional bolometer, this dissipated power is the product of radiation incident on the detector area and the detector absorptivity. For antenna-coupled microbolometers, W depends on the antenna cross section, the antenna gain, and the impedance match between antenna and detector [¹³]. The heat balance equation will now include heat from absorbed radiation, an accumulation term dependent on the change in the bolometer temperature and the thermal capacity, and heat from the change in resistance of the bolometer for change in T. The new heat balance equation is

$$C\frac{d(T + \Delta T)}{dt} + G(T + \Delta T - T_o) = I_b^2(R + \Delta R) + W_o \sin \omega_m t$$
(2.7)

where the first term on the left represents heat accumulation, the second term is the heat flow out of the bolometer, and the terms on the right represent heat flowing in. Equation (2.7) reduces to

$$C \frac{d\Delta T}{dt} + G\Delta T = I_b^2 \left(\frac{dR}{dT}\right) \Delta T + W_o \sin \omega_m t$$
(2.8)

or

$$\frac{d\Delta T}{dt} + \frac{G_e}{C} \Delta T = \frac{W_o}{C} \sin \omega_m t$$
(2.9)

where Ge is an effective thermal conductance,

$$G_{e} = \left(G - I_{b}^{2} \frac{dR}{dT}\right)$$
(2.10)

Equation (2.9) is a linear first order differential equation of the form and general solution

$$\frac{\mathrm{d}y}{\mathrm{d}x} + f(x)y = r(x) \tag{2.11a}$$

$$y(x) = e^{-h} \left[\int e^{h} r \, dx + c \right]$$
(2.11b)

where
$$h = \int f(x) dx$$
. (2.11c)

The integral term in (2.11b) has a solution found from the integral tables,

$$\int e^{gx} \sin \omega x \, dx = \frac{e^{gx}}{g^2 + \omega^2} (g \sin \omega x - \omega \cos \omega x).$$
(2.12)

Using these formulas, the solution for the change in temperature ΔT is

$$\Delta T = \frac{W_o}{G_e^2 + \omega_m^2 C^2} \left(G_e \sin \omega_m t - \omega_m C \cos \omega_m t \right) + c \exp\left(\frac{-G_e t}{C}\right). \quad (2.13)$$

This is somewhat simplified using a right angle identity to get

$$\Delta T = \frac{W_o}{\sqrt{G_e^2 + \omega_m^2 C^2}} \sin\left(\omega_m t - \tan^{-1}\left(\frac{\omega_m C}{G_e}\right)\right) + c \exp\left(\frac{-G_e t}{C}\right) . \quad (2.14)$$

The exponential term in (2.14) is important. If G_e is zero or negative, the temperature will rise indefinitely, leading to a thermal runaway condition known as 'bolometer burnout'. For materials with a negative α (such as Bi and Te), G_e as given in (2.10) will always be positive and ΔT will decay with time constant $\tau = C/G_e$. For superconducting materials, however, α is positive and can be quite large, and G_e could conceivably be negative. In this case, the temperature would rise until the device was no longer operating in the high α transition region. A critical current I_c that marks the onset of instability can be determined. Setting $G_e = 0$ in (2.10),

$$I_c^2 = \frac{G}{dR/dT}$$
 (2.15)

Thermal runaway can be avoided by limiting I_b to a fraction of the value of I_c . Typically, a value of 0.5 is used.

Assuming G_e is positive, ΔT is now a sinusoidal response to the chopped radiation, with a peak amplitude

$$\Delta T_{p} = \frac{W_{o}}{\sqrt{G_{e}^{2} + \omega_{m}^{2}C^{2}}}$$
(2.16)

The detected voltage across the bolometer will fluctuate by

$$v = I_b \frac{dR}{dT} \Delta T_p$$
(2.17)

so that

$$v = \frac{I_{b} (dR/dT) W_{o}}{\sqrt{G_{e}^{2} + \omega_{m}^{2} C^{2}}}.$$
(2.18)

The responsivity is this v divided by W_0 , or

$$\mathbf{r} = \frac{\mathbf{I}_{b} \left(\frac{d\mathbf{R}}{d\mathbf{T}} \right)}{\sqrt{\mathbf{G}_{e}^{2} + \boldsymbol{\omega}_{m}^{2} \mathbf{C}^{2}}} \quad .$$
(2.19)

From (2.2) and (2.19), the thermal impedance is determined to be

$$Z_{t} = \left[G_{e}^{2} + \omega_{m}^{2} C^{2}\right]^{-1/2} .$$
(2.20)

The thermal conductance term can be expanded as

$$G_e = G_s + G_a + G_m - I_b^2 dR/dT$$
(2.21)

where G_S , G_a and G_m correspond to thermal conductance from the bolometer into the substrate, air, and metal antenna leads, respectively [¹⁴]. For a typical 4 μ m x 4 μ m antenna-coupled microbolometer, G_a is much smaller than either G_S or G_m ($G_a < .1\% G_S$), so it is neglected. The change in Joule heating term, Ib²dR/dT, cannot be neglected since it is within an order of magnitude of G_m .

The thermal conductivity through the metal leads, G_m , is considered frequency independent; a fairly valid approximation given the relatively high conductivity of the antenna metal. The term for thermal conduction into the substrate G_s is frequency dependent. Hwang *et al.* [14] calculated G_s as a function of frequency, modeling the microbolometer as a point source of heat, and found

$$G_{\rm S} = 2\pi k_{\rm S} a (1 + a/L_{\rm S})$$
 (2.22)

where k_s is the substrate thermal conductivity. The term a is the radius of a hemisphere representing the bolometer to substrate contact area. For convenience, $2\pi a^2 = A$, where A is the actual contact area. L_s is a complex heat diffusion length

$$L_{s} = \left(\frac{k_{s}}{j\omega\rho_{s}C_{s}}\right)^{1/2}$$
(2.23)

where C_S is the substrate heat capacity.

The complex thermal impedance term can be expressed as

$$Z_{t} = \left(G_{s} + G_{m} - I_{b}^{2} \frac{dR}{dT} + j\omega C_{b}\right)^{-1} .$$
 (2.24)

Now, following Neikirk [¹⁵], the frequency independent terms are lumped together into a single dc conductance term G_{dc} , and the magnitude of Z_t is

$$Z_{t} = \left[G_{dc}^{2} + 2G_{dc}\pi \mathbf{a}^{2}\sqrt{2k_{s}\rho_{s}C_{s}\omega} + 4\pi^{2}\mathbf{a}^{4}k_{s}\rho_{s}C_{s}\omega + 2\pi\mathbf{a}^{2}C\sqrt{2k_{s}\rho_{s}C_{s}}\omega^{3/2} + C^{2}\omega^{2}\right]^{-1/2} .$$

$$(2.25)$$

Equation (2.25) may be inserted into the responsivity equation (2.2) in order to estimate values of r as a function of frequency. The advantage of such an equation is the ability to quickly calculate the responsivity behavior as a function of substrate and bolometer material parameters, and device size. However, both G_{dc} and a are chosen as adjustable parameters to match the analytical model to experimental data. Also, this equation does not consider the effect of the thermal conductivity of the detector material. For the Bi microbolometers manufactured by Neikirk [8], values of a = 1.35 μ m and G_{dc} =1.5x10⁻⁵W/K gave best fit to data.

Figure 2.4 shows several r-f plots which highlight the sensitivity of device performance to several parameters. The material parameters for quartz (kO,CO,pO) and for bismuth (dR/dT, CB_i, and ρ B_i) are values taken from the literature [15-¹⁶¹⁷¹⁸]. Figure 2.4a shows that the dc conductance term Gdc effects the low frequency response of the device. Higher responsivities are obtained for minimal Gdc. This observation prompted Neikirk to construct an air-bridge microbolometer, where each end of the device contacts the antenna metal, but the detector is not in direct contact with the substrate [¹⁹]. Gdc is therefore minimized, and dc responsivities roughly 5 times higher than those from substrate-supported microbolometers were obtained. The air bridge structure also makes the substrate contact radius a essentially zero, resulting in a much sharper knee in the r-f plot. In Fig. 2.4b, reducing a makes the knee sharper. This plot also shows that high frequency performance can be improved by making the device, and therefore a, smaller. A smaller device will also have a lower heat capacity, resulting in a speed improvement as shown in Fig. 2.4c. This figure also shows that low frequency performance is not effected by the detector heat capacity.



Fig. 2.4: Responsivity vs frequency curves show sensitivity of performance on the parameters (a) G_{dc} , (b) a, and (c) C_{Bi} .

Default parameters (bold lines)

$$\begin{split} I_{bias} &= 1.0 \text{ mA} \\ G_{dc} &= 1.5 \text{ x } 10^{-5} \text{ W/K} \\ a &= 1.35 \text{ } \mu\text{m} \end{split}$$

 $a = 1.35 \ \mu m$ length = width = 4 μm thickness = 1000Å

$$\label{eq:Quartz:} \begin{split} &Quartz:\\ &k_Q=0.014 \ W/cm-K\\ &C_Q=0.18 \ cal/(g-K)\\ &\rho_O=2.20 \ g/cm^3 \end{split}$$

Bi: $dR/dT = -0.3 \Omega/K$ $C_{Bi} = 0.027 \text{ cal/(g-K)}$ $\rho_{Bi} = 9.8 \text{ g/cm}^3$

2.3 Noise

The true measure of a detector's performance is how well it extracts a signal from the surrounding noise. Sensitivity is therefore expressed as *noise equivalent power*, or NEP. NEP can be defined as that amount of power absorbed by the detector which gives a signal-to-noise ratio of 1. From (2.2), substituting noise voltage V_n for the signal voltage v, and NEP for the absorbed power P,

$$NEP = \frac{V_n}{r} \quad . \tag{2.26}$$

A receiver circuit is vulnerable to a host of noise sources. Amplifier noise is significant for the far-infrared spectral region, and in fact may limit performance in some applications [²⁰]. At the detector level, the most significant sources of noise are flicker noise, Johnson noise, and temperature noise.

Flicker noise, also called 1/f noise and contact noise, provides a motivation for chopping or otherwise modulating the incoming radiation at as high a frequency as possible. In general, 1/f noise depends on the cleanliness or quality of the contact interfaces. This is why multiple metal layer devices have better noise performance when they are fabricated *in-situ*, rather than in several vacuum evaporation steps [8]. This type of noise appears to be less a factor for antenna-coupled devices with low current bias and high responsivity [13].

Johnson noise is associated with bolometer resistance, and is a result of random movement of charge carriers in a resistor. Since electron movement increases with temperature, Johnson noise is also called thermal noise $[^{21}]$. Johnson noise voltage may be written

$$v_j^2 = 4kTR\Delta f \tag{2.27}$$

where v_j^2 is the mean noise voltage squared, k is Boltzmann's constant, T is the temperature, R is the bolometer resistance, and Δf is the bandwidth. This kind of noise decreases as the square of responsivity, and the NEP resulting from Johnson noise (NEP_{jn}) is given by

$$(\text{NEP}_{jn})^2 = \frac{\mathbf{v}_j^2}{\mathbf{r}^2}$$
 (2.28)

Like 1/f noise, Johnson noise may be relatively small for antenna-coupled devices with high responsivities.

The fundamental limit for bolometer detectors is temperature noise (also called phonon fluctuation noise), which arises from random thermal fluctuations between the bolometer and its surroundings $[^{22},^{23}]$. For devices where heat transfer is by conduction only, the standard form of the temperature noise is

$$\Delta \mathbf{T}^2 = \frac{4kT^2G\Delta f}{G^2 + \omega^2 C^2}$$
(2.29)

where ΔT^2 is the squared mean temperature fluctuation. For most microbolometers, $\omega C \ll G$ for modulation frequencies less than 1 MHz, and (2.29) becomes

$$\Delta \mathbf{T}^2 = \frac{4\mathbf{k}\mathbf{T}^2\Delta \mathbf{f}}{\mathbf{G}} \quad . \tag{2.30}$$

This temperature noise can be converted into a noise power by using the NEP relationship

$$NEP = \frac{\mathbf{v}_{n} \Delta f^{-1/2}}{\mathbf{r}}$$
(2.31)

where v_n is the temperature noise voltage for the device. This noise voltage is simply related to the temperature fluctuation by

$$\mathbf{v}_{n} = \mathbf{I}_{b} \left(\frac{\mathrm{dR}}{\mathrm{dT}}\right) \Delta \mathbf{T} \ . \tag{2.32}$$

Considering the responsivity relation

$$\mathbf{r} = \mathbf{I}_{b} \left(\frac{d\mathbf{R}}{d\mathbf{T}} \right) \frac{1}{\mathbf{G}} , \qquad (2.33)$$

the NEP resulting from temperature noise (NEPtn) can be written

$$(NEP)_{tn} = \sqrt{4kT^2G} \quad . \tag{2.34}$$

The squared NEP terms may be added arithmetically to obtain total squared NEP.

In section 2.5, the NEP_{tn} calculated from (2.34) will be compared with the NEP of an actual Bi microbolometer.

2.4 Finite Element Thermal Model

Thermal models for the bolometer structures studied in this work are useful for understanding device operation, giving the experimentalist a good physical understanding of the device and a better understanding of its limitations. Further, a model might point the way to improvements in device design. A simple analytical model was used in section 2.2 to predict the performance of a substrate-supported bolometer [14]. The model was fairly practical since assumptions could be used to simplify device geometry. More complicated structures like the composite microbolometer cannot be simply modeled. Instead, a finite difference type of approach can be used.

The bow-tie microbolometer structure shown in Fig. 2.5a will be used to demonstrate the finite difference approach. Since the device is symmetrical, only one-half of the structure will be treated. The cut portion now rests against a perfect insulator (Fig. 2.5b). The thermal model device profile of Fig. 2.5b is shown in Fig. 2.6. Only a two dimensional model is considered; effects in the y-direction will be neglected. Figure 2.6 shows the device broken up into a grid of rows and columns, with nodes chosen at the center of each element. To reduce complication in the calculations, the detector and antenna portions are considered to occupy a single row.



Fig. 2.5: A bow-tie antenna-coupled microbolometer (a) is symmetrical about the center. The thermal model therefore considers only half of the device, where the 'cut' portion of the microbolometer rests against an insulator (b).

For a single element, properties of interest are the thermal capacitance, the thermal resistances in the x and z directions, and for the detector region only the electrical resistance in the x-direction (see Table 2.2). A further simplification assumes electrical resistance of the antenna is zero, and the substrate has an infinite electrical resistance. The heat generation in the detector elements must also be considered. An electrical analog to the thermal model is shown in Fig. 2.7 for several cases. In Fig. 2.7a, the thermal resistance between two nodes is just the sum of 1/2 the resistance of each of the two elements. Figures 2.7b and 2.7c show the cases for an element adjoining an insulator and a heat sink, respectively.

Using this model, the steady state (dc) operation is first examined in section 2.4.1. This is useful for predicting r_{dc} , and for generating isothermal plots. Then, in section 2.4.2, the time dependent case is discussed. The program for the finite element thermal model is listed in appendix A.



Fig. 2.6: The thermal model profile for the bismuth microbolometer shown cut in half in Fig. 2.5b. The grids define elements used in the Gauss-Seidel iteration.

Thermal Resistance:	$Z = \frac{\Delta x}{\Delta x}$	$Z = \frac{\Delta z}{\Delta z}$
(K/W)	$k \Delta y \Delta z$	$k \Delta y \Delta x$
Electrical Resistance:	$R = \frac{\Delta x}{\Delta x}$	$R = \frac{\Delta z}{\Delta z}$
(Ω)	σ Δy Δz	$\sigma \Delta y \Delta x$
Thermal Capacitance:	$mC = \rho C \Delta x \Delta y \Delta z$	
(cal/K)		

Table 2.2: Thermal properties of interest for a single element in the thermal grid.



Fig. 2.7: Electrical analog of the thermal model shows how resistance (electrical or thermal) is calculated between element-centered nodes. In (a), the resistance between nodes 1 and 2 is just the sum of half the node 1 resistance and half the node 2 resistance. In (b), element 2 is an insulator, so the total resistance between the two nodes is infinite. In (c), element 2 is a thermal (or electrical) ground, and the resistance between the two nodes is just half the resistance of element 1.

2.4.1 Steady State Case

At steady-state, the temperature at each node is calculated using the Gauss-Seidel iteration:

$$T_{i} = \frac{q_{i} + \sum_{j} \left(\frac{T_{j}}{Z_{ij}}\right)}{\sum_{j} \left(\frac{1}{Z_{ij}}\right)}$$
(2.35)

where T_i and T_j are the temperatures of the ith and jth element, q_i is heat generated in the ith element, and Z_{ij} is the thermal resistance between the ith and jth element. In our calculations, the heat generation term in a detector element will depend on the bias current applied (I_{bias}) and the electrical resistance of the element R_i , so

$$q_i = I_{\text{bias}}^2 R_i . \tag{2.36}$$

A bolometric detector works by changing resistance in response to changes in temperature. The dependence of the resistance of an element R_i to the temperature T_i is given by

$$R_{i} = R_{i0} (1 + \alpha (T_{i} - T_{0})) = R_{i0} + (\Delta R / \Delta T) \Delta T .$$
(2.37)

Thus, for a given ambient temperature and a chosen bias current, the iterative procedure for steady-state is as follows:

1) determine thermal resistances in both the x and z directions for all elements. These are assumed constant for our degree of temperature change.

2) determine electrical resistances for all detector elements in the x-direction at temperature T_i using (2.37). Initially, all elements are at ambient temperature.

3) determine q_i from (2.36).

4) Use (2.35) to determine a new set of temperatures T_i.

5) Cycle through steps 2-4 until the changes in T_i are small.

This iterative technique has been applied to the Bi microbolometer detailed in Fig. 2.4. An isothermal plot of the device at steady state is shown in Fig. 2.8. How far the isothermal lines descend into the substrate depends directly on the substrate thermal conductivity. Ideally, the isotherms will be bunched near the surface, and not much heat will be lost to the substrate.



Fig. 2.8: Isothermal plot for the bismuth microbolometer detailed in Fig. 2.4. Ambient temperature is 300 K.

2.4.2 Time Dependent Case

To predict how the microbolometer would perform as a detector, a small rf power is applied. The thermal capacitance C_i of each element is now very important. The new equation is

$$T_{i}^{p+1} = \frac{q_{i} + \sum_{j} \left(T_{j}^{p+1} / Z_{ij} \right) + \left(C_{i} / \Delta \tau \right) T_{i} P}{\sum_{j} \left(\frac{1} / Z_{ij} \right) + C_{i} / \Delta \tau}$$
(2.38)

where $T_j p$ is the temperature of an element at a specific time, and $T_i p^{+1}$ is the element temperature a time $\Delta \tau$ later.

The heat generation term in each element now also depends on the rf power applied. Adding an rf current gives

$$I_{tot} = I_{bias} + I_{rf} \cos(\omega_c t)$$
(2.39)

where I_{rf} is the magnitude and ω_c is the carrier frequency, and the heat generation term becomes

$$q_i = I_{tot}^2 R_i \quad . \tag{2.40}$$

Inserting (2.39) into (2.40) and integrating the current over one rf cycle,

$$q_i = (I_{bias}^2 + \pi (I_{rf}^2))R_i$$
 (2.41)

In the time dependent case, we are really interested in determining the signal voltage produced by a given amount of input power, and how this responsivity is related to modulation frequency of the input power. Thus, the magnitude of I_{rf} will be modified by a modulation frequency ω for a sinusoidally varying applied power, $I_{rf}(1/2(1 + \cos \omega t))$. The voltage that can be measured across the detector is

$$V = I_{bias} R_{det}$$
(2.42)

remembering that the rf power is confined to within about 3 λ of the region of the detector for a bow-tie antenna. The detected signal is then V_{max} - V_{min} .

The time increment, fixed by stability arguments, is

$$\Delta \tau \leq \left(\frac{C_{i}}{\sum_{j}^{1/Z_{j}}} \right)_{\min}$$
(2.43)

Examination of this criterion for several worst cases are shown in Table 2.3. Notice that $\Delta \tau$ is on the order of picoseconds when the antenna metal is considered since the thermal conductivity is so high. Such a small time increment would require a very long run time for the time dependent program. However, steady state analysis shows the temperature of the silver next to the detector to be very

center	Bi	Ag	Quartz
element:			
C(µW-sec/°K)	984 x 10 ⁻⁹	295 x 10 ⁻⁹	487 x 10 ⁻⁹
Z _x	0.1724	775 x 10 ⁻⁶	0.256
Zz	0.0431	436 x 10 ⁻⁶	0.401
$\Delta \tau$ (nsec)	45(1)	76 x 10 ⁻³⁽²⁾	25(3)



Table 2.3: "Worst-case" situations used to determine the maximum time step allowed for stability with each element 0.2 μ m long. If the silver antenna lead is considered a perfect ground, then the quartz element at the corner of a thermal sink (#3) is the limiting case.

close to ambient. Thus, the antenna leads are treated as thermal grounds.

The approach is to determine a 'dc' response by finding the steady state voltage difference between a steadily applied rf current and a totally off rf current. Since frequencies below 100 kHz would require an exorbitant amount of computer time, data will be limited to only a few data points above 100 kHz.

In Fig. 2.9, the Bi microbolometer default case of Fig. 2.4 is compared with thermal model data in a responsivity-frequency plot. For the thermal model, the parameters σ_{Bi} and k_{Bi} are adjusted to give a steady state resistance of 100 Ω , and a dc responsivity of -16 V/W.



Fig. 2.9: Comparison of analytical model (solid line) and thermal model (black squares) for the bismuth microbolometer detailed in Fig. 2.4. In addition to the properties listed in Fig. 2.4, from the thermal model $\sigma_{Bi} = 980 \ (\Omega\text{-cm})^{-1}$, and $k_{Bi} = .015 \ W/(\text{cm-K})$.

2.5 Bismuth Microbolometer

Bismuth microbolometers have been fabricated and tested for comparison with the theoretical models. The use of bismuth as a detector material as well as microbolometer fabrication will be discussed in more detail in chapter 3.

Figure 2.10 shows the resistance of the Bi element plotted versus power dissipated in the element. From the slope of this plot, the dc responsivity (from (2.3)) is calculated as -6.3 V/W at a bias voltage of 50 mV.

Figure 2.11 shows the theoretical responsivity compared with the measured responsivity for the Bi microbolometer with the properties listed. The solid line is theoretical responsivity using the analytical model, and the square points are theoretical points found using the thermal model. The experimental points were obtained by reading the peak-to-peak signal voltage across the bolometer resulting from a modulated 220 MHz signal fed to the detector. This technique is discussed in chapter 3 for a composite microbolometer. The number of thermal model points are rather limited because of computation time constraints. Also, at high



Fig. 2.10: Bismuth microbolometer resistance is plotted versus power dissipated. The dR/dP slope is used to calculate r_{dc} .

frequencies, the thermal model predicts significantly larger responsivities. The most likely reason for this disparity is because the finite-element model is twodimensional, and does not consider the extra thermal mass that exists in the ydirection.

The noise voltage is plotted in Fig. 2.12. Noise measurements over a bandwidth of 10% of the selected center frequency were made using a PAR 124A lock-in amplifier with a 117 preamp. The experimental NEP points in Fig. 2.13 are calculated by dividing the noise from Fig. 2.12 by the responsivity from Fig. 2.11. A comparison is made with the theoretical NEP determined from the analytical equation (2.34) assuming only temperature noise. This theoretical NEP sets the upper limit for microbolometer sensitivity. Best measured performance was obtained at 5 kHz, where NEP = $4.3 \times 10^{-10} \text{ W/}\sqrt{\text{Hz}}$. In Fig. 2.13, it is seen that at lower frequencies the NEP is limited by 1/f noise, while at higher frequencies the NEP increases because of a drop-off in responsivity.



Fig. 2.11: Responsivity as a function of frequency for a Bi micro-bolometer with the properties listed at a bias of 0.05V. Analytical theory (solid line) is compared with the thermal model (stars), and with experiment (squares).

2.6 Conclusions

In this chapter, the general theory of microbolometer operation has been reviewed, leading to an analytical model. A thermal model has also been introduced. Unlike the analytical model, the thermal model can be used to analyze more complicated device geometries. Also, the fitting parameters are all basic material properties, whereas the analytical model requires the rather nebulous thermal conductance and effective contact radius terms. The drawback to this model is the lengthy computation time required, compared to the very quick results available from the analytical model.



Fig. 2.12: Noise voltage measured for the Bi microbolometer biased at 0.05 V.



Fig. 2.13: Frequency dependent NEP for the Bi microbolometer of Fig. 2.5. NEP from the analytical model (solid line) is compared with experiment.

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