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Gallium Arsenide Submillimeter Photoconductive Detection

by

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1. Introduction

The submillimeter region of the electromagnetic spectrum lies between the infrared and the microwave regions. It includes wavelengths from 1000 to 10 microns. The equivalent frequency interval is 300 GHz to 30,000 GHz; and the equivalent photon energy range is 1.24 millielectron volts to 124 millielectron volts. The techniques for generating, detecting, modulating and guiding submillimeter radiation are often comprised of both microwave and infrared techniques.

The submillimeter area is underdeveloped in comparison to its spectral neighbors. Presently, there are no wide-spread commercial uses of this radiation. One of the major reasons for this is the lack of practical, intense sources. Lasers have been built which operate throughout this part of the spectrum. However, they are not considered practical sources because of their size (often 3 meters in length) and maintenance problems. Another reason for the lack of commercial use is the fact that the most sensitive detectors for this region require cooling to liquid helium temperatures.

Thus, the major interest in submillimeter radiation has been limited to (a) astronomy; (b) spectroscopy; and (c) other basic research areas, e.g., plasma research.

The room temperature detectors used for this range are thermopiles, the Golay cell, solid state bolometers, pyroelectric detectors, and point contact diodes. Cryogenic detectors range from Josephson junctions to "hot electron" bolometers, to germanium bolometers, to extrinsic photoconductors. Figure 1-1 illustrates the relative performances of the currently developed detectors which are suitable for detection of submillimeter radiation [1].
Figure 1-1.

Comparison of Time Response and Sensitivities of Submillimeter Detectors
Extrinsic photoconductivity in high purity GaAs had been observed for wavelengths from 100 to 350 microns. It is one of the most sensitive detectors for this range. It is flanked by the boron-doped germanium extrinsic photoconductor and the indium antimonide free electron bolometer (Figure 1-2) [2]. Boron-doped germanium has a long wavelength cutoff of 130 microns. Indium antimonide is useful for wavelengths longer than 350 microns.

The response time of GaAs is very short. This results from the short carrier lifetime (approximately $10^{-8}$ seconds). The most common configuration in which a GaAs detector is used employs DC biasing techniques. Capacitances associated with this configuration impose limitations on the response time of the entire detection system. The short response time of the GaAs cannot be fully utilized.

The work presented here deals with the development and evaluation of a detection system which allows utilization of the fast response of GaAs. Two different detection systems were constructed. The first employed DC biasing, whereas the second employed microwave biasing. The performance of the two systems was theoretically analyzed and experimentally investigated.

Chapter II is a discussion of the extrinsic photoconductivity in GaAs. This discussion is a summary of the published work characterizing GaAs as a submillimeter detector.

Chapter III is a theoretical discussion of a DC biased photoconductive detection system. The figures of merit used in evaluating a detector's performance are introduced. The common sources of noise and the limitations on the bandwidth of a detection system are also presented.
Figure 1-2. Comparison of Spectral Response of Ge:B, GaAs, and InSb Photoconductive Detectors
Chapter IV presents an experimental investigation of a DC biased GaAs detector. The figures of merit describing the detector's performance are given. Conditions for optimum performance were determined. The sources of noise were identified and the bandwidth of the detection system was measured.

Chapter V describes the use of microwave biasing techniques for a photoconductive detector. The microwave cavity, its equivalent circuit, the conditions for optimum response, and the bandwidth limit are discussed.

The actual performance of the microwave biased detection system is presented in Chapter VI.
II. Extrinsic Photoconductivity in GaAs

II-A. Photoconductive Mechanisms

The extension of extrinsic photoconductivity beyond the 130 micron long wavelength limit of the boron-doped germanium detector requires a material with very shallow impurity levels. This requires a material with a low electron effective mass and a large dielectric constant. GaAs has an electron effective mass of $0.0665 \, m_0$ and a static dielectric constant of 12.5 [1]. Using the hydrogenic model of the impurities, the donor ionization energy is given by

$$E_d = \frac{e^4 m^*}{2(4\pi \varepsilon \varepsilon_0 h)^2} = 5.8 \text{ meV}$$

This donor ionization energy corresponds to a long wavelength cutoff of 214 microns.

Photoconductive response to submillimeter radiation in the range of 100 to 350 microns has been observed in high purity n-type epitaxial GaAs when cooled to liquid helium temperatures [2]. This photoconductivity is attributed to a two-stage ionization process. The impurity electrons are photo-excited from their ground state into excited impurity levels, and are then thermally excited to the conduction band. The thermal transfer of the electrons from the excited impurity states to the conduction band extends the long wavelength cutoff from 214 microns to beyond 350 microns. Because of this mechanism, the photon energy of the incident radiation must only be large enough to excite the electron to the first excited impurity level.

II-B. Hydrogenic Model of Impurity Levels

The energy level structure of the donor impurities is accurately described by the hydrogenic model for very high purity GaAs. A donor magneto-spectroscopy

Because of the small effective mass and large dielectric constant in GaAs, the donor wave functions have large radii, as can be seen by the following expression for the Bohr radius,

$$a_d = \frac{4\pi \varepsilon_0 \varepsilon_r \hbar^2}{m^* e^2}$$

Therefore, very pure material (\(N_D \leq 5 \times 10^{13} \text{ cm}^{-3}\)) is required to avoid the interaction of neighboring impurity atoms, and preserve the energy level structure of the hydrogenic model. When the donor concentration is increased, the interaction between the impurity atoms becomes significant, forming impurity bands. This decreases the effective donor thermal ionization energy. When the donor concentration reaches \(N_D \sim 10^{16} \text{ cm}^{-3}\), the donor ionization energy approaches zero. The GaAs is said to be degenerately doped.

II-C. Effects of Donor Concentration

The effects of donor concentration on the photoconductive responsivity spectrum have been studied [3]. The highest responsivity occurs for \(\lambda = 282\) microns. This peak does not shift as the donor concentration is varied from \(4.8 \times 10^{13} \text{ cm}^{-3}\) to \(1.1 \times 10^{15} \text{ cm}^{-3}\). This peak corresponds to the \((1s \rightarrow 2p)\) transition of the hydrogenic model of the donor impurity.

Responsivity peaks corresponding to the higher excited states depend quite heavily upon the donor concentration. This behavior is due to the interaction between different impurity atoms.
The highest overall responsivity occurs for donor concentrations in the range from $4 \times 10^{13}$ to $5 \times 10^{14}$ cm$^{-3}$. No photoconductive response occurs for $N_d \geq 10^{16}$. The acceptor concentration should be made as low as possible.

II-D. Effects of Background Temperature

There is no usable photoconductivity in GaAs until its temperature is decreased to 6.5$^{\circ}$K. Above 6.5$^{\circ}$K, the carrier concentration is determined by thermal generation. The response increases as temperature is reduced to 5.5$^{\circ}$K and remains fairly constant as temperature is lowered to 4.2$^{\circ}$K. Upon cooling to 1.93$^{\circ}$K, the responsivity drops by an order of magnitude.

It has also been shown that reduced background conditions can improve the responsivity by an order of magnitude. For a detector cooled to 4.2$^{\circ}$K, the thermal generation of carriers will be negligible compared to the generation caused by background radiation from 300$^{\circ}$K sources. Reducing background radiation increases the detector resistance and thus increases responsivity.

II-E. Response Time

The ultimate limit of the response time of a GaAs detector is determined by the extrinsic carrier lifetime. Observation of the decay of current resulting from impact ionization of the shallow impurities indicates that the carrier lifetime is $10^{-8}$ seconds [2]. The ultimate bandwidth of a detection system employing GaAs should be at least 100 MHz.

However, the capacitance of the detector element, the capacitance of the detector leads, and the stray capacitance of the system, in combination with the
high resistance of the detector and the load resistor, limit the bandwidth of a
detection system employing GaAs. Capacitive effects produce a decrease in the
response of the system at a frequency far below the frequency determined by the
reciprocal of the carrier lifetime.

The construction of a system utilizing the full potential of GaAs has not
yet been accomplished. GaAs detectors have been used to detect 337 micron
radiation modulated at a frequency of 300 KHz with a Ge impact ionization
modulator [4]. The present limit of bandwidth measurements on GaAs detectors is
5 MHz. This bandwidth was measured in a mixing experiment using two HCN lasers
tuned apart by 5 MHz [5].

II-F. Recombination Mechanisms

Recombination of the extrinsic electrons and ionized impurities occurs when
the energy of the electrons is emitted either in the form of photons, giving rise to a
radiative recombination, or in the form of phonons, giving rise to a nonradiative
recombination. Both photon emission [6] and phonon emission [7] have been experi-
mentally observed. Since the radiative recombination lifetimes are longer than
the time constants observed for the GaAs detector, the predominant recombination
mechanism is thought to be nonradiative multi-phonon emission.

The multi-phonon process is expected to be similar to that described by
Lax [8] for Ge and Si. The direct transition to the ground state of the impurity
atom by the production of phonons has a small capture cross section compared to
that of a multi-phonon transition. Recombination centers with energy states of
large radii possess large capture cross sections. Only the states having binding
energies less than kT can be included in the calculation of the total cross section of a recombination center. As the temperature is lowered, states having larger radii contribute to the capture process. Upon capture in an excited impurity state, the electron is no longer available for conduction. By a series of one-phonon transitions, the electron's energy is lowered toward its ground state energy. The last transition to the ground state may require a multiphonon transition.
III. Theory of DC Biased Photoconductive Detectors

III-A. Photoconductivity: Intrinsic versus Extrinsic

In a photoconductive material, illumination with radiation of the proper wavelength will generate excess free carriers by photoionization. The conductivity of the material, which is linearly dependent on the carrier concentration, is therefore increased.

For a pure semiconductor material, excitation of a valence electron by a sufficiently energetic photon produces an electron-hole pair. Both charge carriers contribute to the conductivity. There is a long wavelength limit to the intrinsic photoconductive response of a given material. The photons associated with radiation having a wavelength longer than this limit possess insufficient energy to produce an electron-hole pair. The cutoff wavelength is determined by the size of the energy gap, $E_g$, between the valence and conduction bands. Most intrinsic materials have band gaps greater than 0.18 eV, and therefore cannot be used for detection of radiation having wavelengths longer than 7 microns.

In extrinsic semiconductors, excess carriers can be generated by the production of electron-hole pairs or by the ionization of impurity atoms. The ionization energy of the impurity atoms is much less than the band gap energy, $E_g$. In a p-type extrinsic semiconductor, the energy level associated with the impurities lies near the valence band. A photon of energy much less than $E_g$ is required to excite an electron from the valence band to an acceptor impurity level. A hole is left behind to contribute to the conductivity of the material. In an n-type extrinsic semiconductor, the energy level associated with the impurities lies near the
conduction band. A photon of energy much less than $E_g$ is required to excite an electron from a donor impurity level to the conduction band.

The use of extrinsic semiconductors extends the application of photoconductive detection to much longer wavelengths than was possible using intrinsic materials. Extrinsic germanium has been used to 130 microns; GaAs to 350 microns; and indium antimonide to 8000 microns. The small ionization energies necessary to extend photoconductivity to such long wavelengths requires cooling of the detector element to very low temperatures in order to limit the number of thermally-excited carriers.

III-B. Equivalent Circuit of Detection System

A photoconductive detector behaves as a variable resistor. When used in a detection system, the variations in resistance are monitored with an external bias supply and a load resistor. A typical DC biasing network is shown in Figure III-B-1.

Illumination of the detector with radiation decreases its resistance. This change in resistance alters the current flowing through it and the load resistor. The change in the voltage across the load resistor is proportional to this change in current, and provides the output signal of the detection system.

The voltage across the load resistor is

$$V_L = \frac{V_B R_L}{(R_D + R_L)}$$

The output signal, $V_S$, equals the change in $V_L$ for a given change in $R_D$. When $\Delta R_D$ is small in comparison to $R_D$, the output signal is given by

$$V_S = \Delta V_L = \left(\frac{1}{\frac{1}{R_D}}\right) \Delta R_D$$

$$V_S = -\frac{V_B R_L}{(R_D + R_L)^2} \Delta R_D$$
FIGURE III-B-1. BIASING NETWORK

FIGURE III-B-2. RESPONSE VERSUS LOAD RESISTANCE
For a given detector element, the output signal is affected by the choices of $V_b$ and $R_L$. The bias voltage influences both the signal and the noise. The signal increases linearly with bias voltage until heating, or some form of breakdown of the detector element, limits the response. The effect of bias voltage on the noise depends on the specific type of noise which is dominant in the system. Several sources of noise are discussed in III-D.

The choice of $R_L$ affects the signal amplitude, the noise characteristics, and the bandwidth of the detection system. The specific requirements of a detection system must be considered in the proper selection of $R_L$. The dependence of the signal amplitude on $R_L$ is shown in Figure III-B-2. The signal is maximized when $R_L$ equals $R_D$.

III-C. Figures of Merit

One of the most common figures of merit used in describing the performance of a detector is the responsivity, $R$. The responsivity is the detector output per unit input. Since most detectors are used to detect chopped radiation, the input power, $P_I$, and the output signal, $V_s$, are each specified in terms of the rms value of its fundamental component. The responsivity is given by

$$R = \frac{V_s}{P_I} \text{ (volts/watt)}$$

The responsivity of a detector is obviously not a single-valued function. It is a strong function of the bias conditions, the load resistor, the chopping frequency, and the wavelength of the incident radiation. The most meaningful value of responsivity of a detector is the value measured for the conditions under which the signal to noise ratio is a maximum.
The ultimate sensitivity of a detection system is determined by the minimum amount of incident radiation it can detect. This minimum value is determined not only by the responsivity of the detector, but also by the amount of noise present in the output signal. The noise equivalent power of a detector is defined as the incident power necessary to produce an output signal equal to the noise output of the system. Because of the difficulty of measuring a signal to noise ratio of unity, the usual technique used for measuring the NEP consists of measuring the signal to noise ratio \( \frac{V_S}{V_N} \) at a high signal level and calculating the NEP using

\[
\text{NEP} = \frac{P_I}{V_S/V_N} \text{ (WATTS)}
\]

\( P_I \) is the incident power which produced the signal \( V_S \). An equivalent expression in terms of the responsivity is

\[
\text{NEP} = \frac{V_N}{R}
\]

The noise voltage \( V_N \) is a function of the electrical bandwidth of the system used to measure the noise voltage. For a measurement of the noise voltage to be meaningful, the equivalent noise bandwidth must be specified. As a result, the NEP is often referenced to a 1 Hz bandwidth, or is expressed in terms of power per square root of the equivalent noise bandwidth.

"Many people inherently dislike a situation in which an improvement in the quantity of interest gives a lower value for the figure of merit" [1]. Therefore, the figure of merit known as detectivity, \( D \), was defined as

\[
D = \frac{1}{\text{NEP}}
\]

To account for the size of the receiving area of the detector, \( A_d \), and to eliminate differences in figures of merit resulting from the equivalent noise bandwidth of the circuitry used to measure the noise output of the system, a normalized detectivity is
defined.

\[ D^* = \frac{(A_d)^{\frac{1}{2}} (C_{EBW})^{\frac{1}{2}}}{\text{NEP}} \]

\(D^*\) can be thought of as the signal to noise ratio when one watt of radiation is incident on a detector having a receiving area of 1 cm\(^2\), and the noise is measured with a 1 Hz bandwidth.

The response time of a detector is determined by the time that it takes for the output signal to reach \((1 - \frac{1}{\tau})\) of its final value for a step change in the incident power. This time is referred to as the time constant of the detector.

III-D. Sources of Noise [2,3,4]

The ultimate sensitivity of a detection system is determined by the noise present in the output. The limiting noise of a system consists of random electrical fluctuations which are generated in the circuit elements of the detection system. A measure of the noise is given by the mean-square fluctuation in the noise voltage, \(\overline{V_n^2}\), or its square root \((\overline{V_n^2})^{\frac{1}{2}}\). In a system which contains two or more independent sources of noise, the net effect is found by adding the mean-square values of their noise voltages. That is, noise powers are additive, but noise voltages are not.

The most common sources of noise found in a photoconductive detection system are (a) thermal noise; (b) "1/f" noise; and (c) generation-recombination noise.

Thermal noise, or Johnson noise, is a result of the random motion of electrons in a solid. Any motion of an electron constitutes a current. For a material with no external bias, the average over a long period of time of the current contributions
resulting from the motion of electrons in the time between collisions is zero.

Considered over a short period, the sum of these currents is not zero. Since the
mean-square velocity of the electrons is proportional to the absolute temperature
of the material, this noise will vary with the temperature. The thermal noise of a
resistor, \( R \), at a temperature, \( T \), measured over a bandwidth \( \Delta f \) will have a mean-
square voltage of

\[
\overline{\mathbb{V}^2} = 4kT \Delta f
\]

where \( k \) is Boltzmann's constant. Thermal noise is independent of any external
current which may be flowing through the material.

For many devices, a sharp increase in noise is observed at low frequencies
when DC current is passed through them. The nature of the noise is not well under-
stood, even though it has been widely observed. It decreases with increasing
frequency. An expression of the mean-square noise voltage is of the form

\[
\overline{\mathbb{V}^2} = \frac{K T_0^{\alpha}}{f^\beta}
\]

where \( K \) is a proportionality constant, \( \alpha \) is usually close to 2, and \( \beta \) varies from
0.8 to 1.5. The values of \( \alpha \) and \( \beta \) must be empirically determined for a
particular device. If the performance of a device is limited by this "1/f" type
of noise, it is best to operate at as high a frequency as possible.

In semiconductor materials, noise results from fluctuations in the carrier
concentrations. The fluctuations in \( n \) and \( \rho \), resulting from the random processes
of generation and recombination, cause fluctuations in the resistance of the material.
When a current is flowing, the fluctuations in resistance show up as a noise voltage.
For an extrinsic semiconductor, the mean-square noise voltage is given by
\[ \frac{\Delta v^2}{\bar{N}} = \frac{4 R^2 I_0^2 \Delta f \tau}{\bar{N} (1 + 4 \pi^2 f^2 \tau^2)} \]

where \( \bar{N} \) = the average total number of free electrons
\( I_0 \) = the average current
\( \Delta f \) = bandwidth of measurement system
\( f \) = center frequency
\( \tau \) = carrier lifetime
\( R \) = resistance of the device

The frequency spectrum of generation-recombination noise is flat out to the reciprocal of the carrier lifetime, where it falls off rapidly.

III-E. Frequency Response Limitations

The sources of the limitations of the frequency response of a detection system can be separated into two main categories: (a) properties of the detector material; and (b) detection circuitry.

As shown in III-B, the output signal of a photoconductive detection system is

\[ V_S = \frac{V_b R_L}{(R_D + R_J)^2} \Delta R_D \]

The change in the detector resistance \( \Delta R_D \) is caused by the change in the carrier concentration resulting from the absorption of incident radiation. Expressing the change in the detector resistance in terms of the excess carrier concentration yields the following expression for the output signal of an n-type extrinsic photoconductor

\[ V_S = \frac{V_b R_L R_D}{(R_L + R_D)^2 N} \Delta n \]

where \( n \) is the steady state electron concentration and \( \Delta n \) is the excess electron concentration resulting from illumination.
The rate of change of the excess carrier concentration equals the difference between the generation rate and the recombination rate.

\[ \frac{d \Delta n(t)}{dt} = \eta F(t) - \frac{\Delta n(t)}{\tau} \]

where \( F(t) \) is the photon flux; \( \eta \) is the quantum efficiency; and \( \tau \) is the mean lifetime of the electrons. Assume the incident photon flux is sinusoidally modulated, so that \( F(t) \) is of the form

\[ F(t) = R_e \left\{ F_0 e^{j \omega t} \right\} \]

The particular solution of the above differential equation for \( \Delta n(t) \) will be of the form

\[ \Delta n(t) = R_e (\Delta N e^{j \omega t}) \]

Substituting and solving for \( \Delta N \) yields

\[ \Delta N = \frac{\eta F_0 \tau}{1 + j \omega \tau} \]

Thus, for a photon flux of

\[ F(t) = F_0 \cos \omega t \]

the excess carrier concentration is

\[ \Delta n(t) = \frac{\eta F_0 \tau}{(1 + \omega^2 \tau^2)^{1/2}} \cos (\omega t - \phi) \]

where

\[ \phi = \tan^{-1} \omega \tau \]

The resulting output signal of the detection system is

\[ V_s = \frac{V_e R_L R_D}{(R_L + R_D)^2} \frac{\eta F_0 \tau}{(1 + \omega^2 \tau^2)^{1/2}} \cos (\omega t - \phi) \]

For modulation frequencies small compared to the reciprocal of the carrier lifetime, the response of the system is independent of frequency. For modulation frequencies equal to or greater than the reciprocal of the carrier lifetime, the response of the system decreases rapidly with increasing frequency.
The bandwidth of a detection system, whose frequency response is limited by the carrier lifetime, is given by

\[ \Delta \nu = \frac{1}{2 \pi \tau} \]

In addition to the limitation imposed by the material parameters, the frequency response of a detection system may be limited by frequency-dependent characteristics of the biasing circuitry. The predominant sources which limit the frequency response are the capacitance of the detector element, the capacitance of the detector leads, and the stray capacitance of the system.

As an example, let the detector capacitance and the capacitance of the leads be represented by the capacitance \( C \) in the equivalent circuit shown in Figure III-E-1. The output signal, \( V_6 \), is now dependent on frequency, because shunting effects of \( C \) are a function of frequency. In order to determine the dependence of \( V_6 \) on the chopping frequency, consider the sinusoidal state case in which the detector resistance, \( R_D \), has a very small sinusoidal variation. The sinusoidal variation in \( R_D \) results from conductivity modulation by the incident radiation.

\[ R_D = R_{D0} + \Delta R_D(t) \]

\[ \Delta R_D(t) = r_0 \cos \omega t \]

The voltage across the load resistor has a similar form

\[ V_L = V_{L0} + V_6(t) \]

\[ V_6(t) = A \cos(\omega t + \phi) \]

Because \( R_D \) is no longer time invariant, the loop equations (Figure III-E-1) become nonlinear. However, if \( \Delta R_D(t) \) is very small compared to \( R_{D0} \), \( V_6(t) \) will be very small compared to \( V_{L0} \), and the second order terms can be neglected, resulting in a set of linear differential equations.
LOOP EQUATIONS

\[ V_B = \frac{1}{C} \int_0^t (I_1 - I_2) \, dt + I_1 R_L \]

\[ O = \frac{1}{C} \int_0^t (I_2 - I_1) \, dt + I_2 R_D \]

FIGURE III-E-I. EQUIVALENT CIRCUIT FOR DC BIASED DETECTOR
The solution for $V_s(t)$ yields

$$V_s(t) = \frac{V_B R_L}{(R_{D0} + R_L)^2} \frac{1}{[1 + \omega^2 c^2 R_p^2]^{1/2}} \Delta R_D(t)$$

$$R_p = \frac{R_{D0} R_L}{R_{D0} + R_L}$$

And so, the output signal $V_s$ has negligible frequency dependence until the chopping frequency approaches the break frequency of

$$f_B = \frac{1}{2\pi R_p C}$$

As the chopping frequency is increased above $f_B$, the output signal decreases rapidly. The break frequency is determined by $\sqrt{2\pi}$ times the reciprocal of the RC time constant of capacitance $C$ and the parallel combination of the detector and load resistances.

For an accurate prediction of the break frequency of an actual detection system, the equivalent circuit may have to be modified to include additional capacitances. The frequency dependence of the output signal will then be more complex than the situation considered above.

To see the combined effects of the limitations of the frequency response produced by the carrier lifetime and the capacitance $C$, the expression for $\Delta R_D$ derived for a sinusoidally modulated photon flux can be substituted in the above expression for $V_s$.

$$V_s = \frac{V_B R_L}{(R_{D0} + R_L)^2} \frac{1}{[1 + \omega^2 c^2 R_p^2]} \frac{R_D}{n} \frac{n F_0}{\tau (1 + \omega^2 \tau^2)} \cos(\omega t - \phi + \phi)$$

where $R_{D0}$ has been replaced by $R_D$. For a given detector material, the maximum bandwidth of the detection system is achieved when $\frac{1}{R_p C} > \frac{1}{\tau}$.
IV. Experimental Results for DC Biased GaAs Detector

IV-A. Detector Material

The detector used to study the performance of GaAs in a DC Bias mode of operation was obtained from Cayuga Associates. The dimensions and construction are shown in Figure IV-A-1. It consisted of a high purity n-type epitaxial layer on top of a semi-insulating GaAs substrate. The epitaxial layer thickness was 100 microns. The donor concentration was \( N_D = 4.61 \times 10^{14} \text{ cm}^{-3} \) and the acceptor concentration was \( N_A = 2.54 \times 10^{14} \text{ cm}^{-3} \). Electrical contact was made by means of Au-Ge-Ni contacts, to which wire leads were attached with low-melting point solder.

IV-B. Cryogenic Systems

The performance of the DC biased GaAs was studied using two different cryogenic systems. The first system was an open cycle, two-stage, Joule-Thomson refrigerator, known as a Cryo-Tip (produced by Air Products, Inc). The detector element was attached to a cold finger which had a small reservoir of liquid helium above it. The second system was a submersion cryostat. It consisted of an inner liquid helium dewar, and an outer liquid nitrogen dewar. In this system, the detector mount was suspended in liquid helium by a light pipe attached to the cryostat cover plate. The latter system was also used to investigate the performance of a microwave biased GaAs detector.

The results obtained using the Cryo-Tip are presented first (IV-C through IV-F). Then the results obtained using the submersion cryostat are presented (IV-G
FIGURE IV-A-1. GaAs DETECTOR CONSTRUCTION
through IV-J). Since the latter system was identical to that used for the microwave biased detector, the performances of the DC and microwave biased detectors can be compared.

IV-C. Cryo-Tip Experimental Configuration

The detector element was cooled to liquid helium temperatures by Indium soldering the detector substrate to a copper block, which was then threaded into the cold finger of the Cryo-Tip (Figure IV-C-1). One contact was grounded to the cold finger. A one-inch diameter gold coated mirror was mounted inside the radiation shield of the Cryo-Tip in order to reflect the submillimeter radiation, which entered a 1/4" diameter aperture in the radiation shield, onto the epitaxial layer.

Figure IV-C-2 illustrates the set-up utilized in evaluating the performance of the detector. An HCN laser was used as a source of submillimeter radiation. The laser line used had a wavelength of 337 microns. A pyroelectric detector was used to continuously monitor the output power of the HCN laser by splitting off a fraction of the beam using a wire mesh beam splitter.

The amount of submillimeter power incident on the detector element was measured with an Eppley thermopile. The thermopile was positioned a distance L from point P (Figure IV-C-2). A 1/4" diameter aperture was placed in front of it to simulate the aperture in the radiation shield of the Cryo-Tip. The receiving element in the thermopile was 3/8" in diameter. Thus, all of the submillimeter power passing through the aperture was detected by the thermopile. With the chopping wheel modulating (at 100 Hz) only the radiation incident on the pyroelectric...
FIGURE IV-C-I. CRYO-TIP
detector, the output power of the laser was varied. The thermopile emf and the pyroelectric detector signal were recorded. Using the calibration figures of the thermopile for responsivity (supplied by Eppley Laboratories), a plot was made of the pyroelectric detector signal versus submillimeter power passing through the 1/4" aperture (Figure IV-C-3). This calibration chart was used in the evaluation of the responsivity of the GaAs detector.

The biasing circuit is shown in Figure IV-C-4. \( V_B \) was supplied by two 7 volt mercury batteries. Only the GaAs was capable of being cooled to liquid helium temperatures. The load resistor was at room temperature.

IV-D. Voltage-Current Characteristics

The resistance of the detector under 300° K background conditions was measured as a function of temperature. The temperature was measured with a gold-copper thermocouple which was soldered to the cold finger. Figure IV-D-1 is a plot of the resistance versus temperature. The decrease in resistance from 400Ω at 300° K to 37Ω at 60° K is due to an increase in mobility of the electrons. The resistance begins to increase slowly below 60° K because the mobility begins to decrease from its maximum value at 60° K as a result of an increase in ionized impurity scattering. Below 20° K, the resistance increases sharply as the carrier concentration decreases because of the freeze-out of electrons in the shallow donor levels. For the temperature change from 20° K to 4.2° K, the resistance changes by four orders of magnitude due to the change of carrier concentration.
FIGURE IV-C-3. PYROELECTRIC DETECTOR CALIBRATION
FIGURE IV-C-4. BIASING CIRCUIT
Figure IV-D-1. Detector Resistance versus Temperature
The V-I characteristics of the detector at 4.2°K are shown in Figure IV-D-2. Voltage and current are linearly related until a bias current of 0.5 microamps or an electric field of 1.55 volts/cm is reached. A further increase in bias field decreases the resistance because of an increase in carrier concentration as impact ionization of the shallow impurities begins. At a bias current of 3.4 microamps, or an electric field of 6.57 volts/cm, breakdown occurs as a result of impact ionization.

IV-E. Responsivity

The Cryo-Tip was positioned so that the radiation window was at a distance L from point P (Figure IV-C-2). The window was made of a 1/8" thick white polyethylene plate and a 4 mil thick black polyethylene sheet. Transmission coefficient of the window was .625. With the vacuum shroud and radiation shield removed, the deflection mirror beneath the GaAs and a mirror at P were aligned to illuminate the detector element, using a He-Ne laser.

Upon illumination with submillimeter radiation, the shallow donor impurities in the GaAs were photo-thermally ionized. The increase in carrier concentration decreased the detector resistance, which produced a change in the current through the load resistor. The resulting change in \( V_L \) was monitored as the detection system output.

Figure IV-E-1 is a plot of the AC signal measured across \( R_L \) versus incident submillimeter power. The detector resistance was 2.9 M\( \Omega \); the load resistance was 1 M\( \Omega \); and the bias current was 0.1 microamps. The incident power, modulated at 100 Hz, was varied from \( 3.6 \times 10^{-8} \) watts to \( 6.94 \times 10^{-5} \) watts, using the
Bias Current = 0.1 μA

f = 100 Hz

R_L = 1 MΩ

Figure IV-E-1. Responsivity
variable output coupler of the HCN laser and three 10 db attenuators constructed from lava. The peak responsivity for this value of bias current was $2.9 \times 10^3$ volts/watt occurring at a low power level. The relation between the signal and incident power became significantly nonlinear at the higher power levels.

Saturation of the response at the higher power levels was due to the large change in the detector resistance upon illumination. The dark resistance of the detector was comparable to the load resistance. When illuminated with submillimeter power of the order of one milliwatt, the large number of excess carriers which were generated reduced the detector resistance, $R_D$, to a value small in comparison to the load resistance, $R_L$ ($R_D \ll 0.1 R_L$). Any further decrease in the detector resistance had no measurable effect on the current through $R_D$ and $R_L$. The current was limited by $R_L$. The maximum voltage that could appear across $R_L$ was the bias voltage setting of the 100 kΩ potentiometer, $V_{(Bias)}$. Therefore, the maximum change in $V_L$ was

$$\Delta V_L = V_{(Bias)} - V_L (Dark)$$

$V_L(Dark)$ was the value of $V_L$ when no submillimeter radiation was incident on the detector element.

For a given detector resistance, $R_D$, and load resistance, $R_L$, saturation occurs at a specific power level. Saturation takes place at the power which reduces $R_D$ to a value small in comparison to $R_L$. Increasing the bias voltage increases the magnitude of the signal at saturation, but does not affect the power level at which saturation results.

The effect of bias current on the responsivity of the GaAs is shown in Figure IV-E-2. The submillimeter power level was $3.6 \times 10^{-8}$ watts. The bias
$P = 3.6 \times 10^{-8}$ WATTS \\
$f = 100$ HZ \\
$R_L = 1 \text{ M}\Omega$

**Figure IV-E-2. Responsivity Versus Bias Current**
current was varied from .05 microamps to 3.2 microamps, where breakdown due to impact ionization occurred. The photocurrent gain of an extrinsic photoconductor varies inversely proportional to the transit time of the carriers. \( \frac{1}{\tau_R} \) is given by
\[
\frac{1}{\tau_R} = \frac{\mu E}{L}
\]
where \( \mu \), \( E \) and \( L \) are the mobility, electric field, and length of the device, respectively. For operation in the linear portion of the \( V-I \) characteristic of the device, the responsivity was expected to increase linearly with bias current. In the nonlinear region of the \( V-I \) characteristic, the electric field strength no longer increased linearly with the bias current, and so the responsivity became nonlinear with bias current. For a bias current of 3.4 microamps, the electric field has reached its peak value of 6.57 volts/cm. The peak responsivity achieved at maximum bias current was \( 5.5 \times 10^4 \) volts/watt.

IV-F. Noise Measurements

The noise of the detection system was characterized by determining the rms noise voltage per square root of the equivalent noise bandwidth of the system centered about the chopping frequency. The rms noise was measured using the band pass amplifier in the signal channel of a PAR Model 120 Lock-In-Amplifier. With the band pass characteristics of the amplifier adjusted to a center frequency of \( f_c \) and a \( Q \) of \( Q_0 \), the equivalent noise bandwidth was given by
\[
\text{ENBW} = \left( \frac{\pi}{2} \right) \left( \frac{f_c}{Q_0} \right)
\]
For a center frequency of 100 Hz and a \( Q_0 \) of 50, the ENBW is 3.14 Hz.

The rms noise per \( \sqrt{\text{ENBW}} \) was measured as a function of the bias current (Figure IV-F-1). As the bias current was decreased to zero, the limiting value of
\[ f = 100 \text{ Hz} \]
\[ R_L = 1 \text{ M} \Omega \]

Figure IV-F-1. Noise versus Bias Current
the noise was $1.35 \times 10^{-7}$ volts/Hz$^{1/2}$. This value of noise equaled the measured value of the noise produced by the load resistance ($R_L = 1 \, M\Omega$) when open circuited. Using the expression given in III-D for thermal noise, the theoretical value of the rms noise voltage per $\sqrt{\text{ENBW}}$ for a $1 \, M\Omega$ resistor at room temperature was calculated to be $1.29 \times 10^{-7}$ volts/Hz$^{1/2}$. The measured and theoretical values of noise of the load resistance are in good agreement. At low values of bias current, the noise of the system was limited by the thermal noise of the load resistor.

As the bias current was increased above .02 microamps, the noise voltage approached a linear dependence on current. In order to determine the source of the noise, the spectral content of the noise must be investigated. In the process of this investigation, it was found that the mechanical vibrations of the Cryo-Tip, resulting from the boiling of liquid nitrogen, produced significant amounts of noise. There was a strong correlation between the noise voltage and the level of liquid nitrogen in its reservoir. As a result, meaningful measurements on the noise spectrum were not obtained using this cryogenic system. Without the noise spectrum, the source of the current-dependent noise cannot be identified. Generation-recombination noise and "1/f" noise both vary linearly with current. (The spectral content of the noise was determined using the second cryogenic system. The noise spectrum is presented in IV-1. From these results the current-dependent noise is identified as "1/f" noise.)

For optimum performance of a detection system, the bias current should be adjusted to produce maximum signal to noise ratio (S/N). For a given incident submillimeter power, the signal increased linearly with bias current until saturation
at 2.0 microamps. However, for bias currents above 0.1 microamps, increasing the bias current also increases the noise. The S/N is plotted against bias current in Figure IV-F-2. The maximum S/N occurred at a bias current of 0.75 microamps. For this value of bias current, an NEP of $3.83 \times 10^{-11}$ watts/Hz$^{1/2}$ and a responsivity of $2.11 \times 10^3$ volts/watt were achieved.

IV-G. Cryostat Experimental Configuration

The results presented in IV-H and IV-I will appear to be somewhat similar to the results presented in IV-E and IV-F. There were significant differences between the experimental configuration used for the DC biased detector with the Cryo-Tip cryogenic system, and the experimental configuration used in studying the microwave biased detector with a submersion cryostat (to be discussed in V and VI). The primary changes made for the microwave biased detection system were (a) the use of stainless steel light pipes to transmit the submillimeter radiation from the focal point of the output lens of the HCN laser to the microwave biased detector element; and (b) the constraint of having a small aperture in the top of the microwave cavity through which submillimeter radiation must be transmitted, requiring the use of the focusing properties of a lens or a cone inside the light pipe. Thus, there was enough uncertainty in the amount of submillimeter radiation illuminating the detector element that it was necessary to place the DC detector element in the same configuration in which the microwave biased detector element was positioned, in order to compare their performances.

The cryostat, used to cool the detector element, consisted of two large glass dewars. The outer dewar contained liquid nitrogen and the inner dewar contained
liquid helium. A brass cover plate was sealed with an O-ring to the glass flange at the top of the inner dewar. The cover plate possessed electrical feedthroughs, a light pipe feedthrough, and the provisions for a polyethylene pressure window in the light pipe.

The DC biased detector element, indium soldered to a copper block, was threaded into the detector mount shown in Figure IV-G-1. One contact was grounded to the detector mount. A twisted pair of teflon coated copper wires was used to make electrical contact to the detector. The aperture, lens, and its holder were easily removeable. The detector mount was suspended in the liquid helium by a light pipe, which was attached to the dewar cover plate.

A system of 0.5" O. D. stainless steel light pipes was utilized to transmit the radiation into the cryostat and to the Eppley thermopile for monitoring the power of the submillimeter radiation (Figure IV-G-2). The light pipe path leading to the Eppley was identical to that leading to the GaAs detector except for (a) the absence of the lens and aperture; and (b) a disc attenuator made of lava inserted in the GaAs light pipe.

The biasing network used was identical to that described in Figure IV-C-4. Again, only the GaAs was cooled to liquid helium temperatures. The load resistor was at room temperature.

IV-H. Responsivity Measurements

The performance of the DC biased GaAs was studied under three different illuminating configurations: (1) a polyethylene lens and 0.1" diameter aperture (Figure IV-H-1); (2) a 0.1" diameter aperture and a copper cone, tapering from
Figure IV-G-1. Cryostat DC Biased Detector Mount
0.5" to 0.1" in a height of 1.0" (Figure IV-H-2); and (3) no aperture and empty light pipe (Figure IV-H-3). The submillimeter radiation was chopped at 100 Hz, and the bias current in the detector was 0.1 microamps.

The copper cone was more efficient than the polyethylene lens at illuminating the detector element through a 0.1" diameter aperture. The peak responsivity for 0.1 microamps bias current was $9 \times 10^3$ volts/watt for the copper cone, and $4.5 \times 10^3$ volts/watt for the polyethylene lens. A factor of 2 improvement in responsivity can be achieved using the copper cone rather than the lens.

The responsivity for the case of no aperture and an empty light pipe was considerably lower than for the other two configurations. There were two principal reasons for this. (1) The area of the detector element was smaller than the cross-sectional area of the 0.5" O. D. light pipe. The detector area equaled 0.57 of the cross-sectional area of the light pipe. (2) The detector resistance was $8.0 \, M\Omega$ for the lens, $7.7 \, M\Omega$ for the copper cone, and $3.7 \, M\Omega$ for no aperture and empty light pipe. These decreases in resistance resulted from the increases in background radiation incident on the detector. The data plotted in Figures IV-H-1 through IV-H-3 were taken at a bias current of 0.1 microamps. As the detector resistance decreased, the electric field in the detector decreased because the bias current was held constant. The transit time of the carriers was increased which decreased the photocurrent gain, resulting in a lower responsivity. To produce an equivalent electric field in the sample, the bias current should have been increased to 0.21 microamps for the case of no aperture and an empty light pipe.

The responsivity achieved for no aperture and an empty light pipe was $2.52 \times 10^3$ volts/watt. If this responsivity is multiplied by $(0.21/0.1)$ to correct for the
$I_b = 0.1 \mu A$
$R_L = 1 \text{ M} \Omega$
$f = 100 \text{ Hz}$

**Figure IV-H-2. Responsivity with Copper Cone**
Bias Current = 0.1 \mu A
\[ R_\text{L} = 1 \, \text{M} \Omega \]
\[ f = 100 \, \text{Hz} \]

Figure IV-H-3. Responsivity with No Aperture
change in detector resistance, and by \((1/ .57)\) to correct for the detector size relative to the submillimeter beam size, a value of \(9.2 \times 10^3 \text{ V/w results, which is comparable to that obtained with the copper cone.}\)

In order to vary the submillimeter power from the HCN laser over the range from \(10^{-7}\) watts to \(10^{-3}\) watts, a set of attenuators was required. 0.475" diameter disc attenuators, capable of being inserted in the light pipe, were constructed from lava. The thickness required to give 10 db of attenuation was empirically determined using the Eppley thermopile. 20 db and 30 db attenuators were constructed by making discs twice and three times the length of the 10 db attenuator. The discontinuities in the responsivity plots resulted from errors introduced by reflections which occurred at the end surfaces of the attenuators.

The response of the detector varied linearly with bias current until the electric field began to saturate due to impact ionization (Figure IV-H-4). As the electric field approached its peak value, the response of the detector saturated at a responsivity of \(2.6 \times 10^4\) volts/watt.

IV-1. Noise Measurements

The noise of the detection system was again characterized by measuring the rms noise voltage per square root of the equivalent noise bandwidth of the system, as described previously in IV-F. The noise measurements presented here were taken for the configuration of no aperture and an empty light pipe. The noise of the system as a function of bias current at a center frequency of 100 Hz is shown in Figure IV-I-1. The dependence of the noise on bias current in this plot is substantially different from that illustrated in Figure IV-F-1, where the noise was limited
No Aperture
\[ P = 6.3 \times 10^{-6} \text{ Watts} \]
\[ R_L = 1 \text{ M} \Omega \]
\[ f = 100 \text{ Hz} \]

Figure IV-H-4. Responsivity versus Bias Current
Figure IV-1-1. Noise versus Bias Current
by the thermal noise of the load resistor at low values of current, and approached a linear dependence on current prior to breakdown. The spectral content of the noise of the cryostat system is shown in Figure IV-I-2. There was a substantial amount of 60 cycle noise present due to the lack of electrostatic shielding of the detector leads in the cryostat. A twisted pair was used instead of coaxial cable in order to minimize the heat load on the liquid helium. Despite the contributions to the noise of the 60 cycle signal and its harmonics, it can be said that the noise voltage due to the detector element and load resistor has a \( f^{-3/2} \) dependence at a bias current of 0.1 microamps. The data in Figure IV-I-1 do not show the noise of the detector and load resistor at low values of bias current because the contribution of the 60 cycle noise is so great. It is not until breakdown is approached that the detector noise can be seen.

Based on Figure IV-F-1 and IV-I-2, the noise voltage limiting the performance of the DC biased GaAs has a linear dependence on current and a \( f^{-1/2} \) dependence on frequency. Thus, the power spectrum has an \( I^2 \) and a \( 1/f \) dependence. This type of noise is commonly observed in detectors and is known as "1/f" noise.

IV-J. Frequency Response

As discussed in II-E, measurements of Stillman, et al. [1] indicate that the lifetime of the carriers in the GaAs is as short as 10^{-8} seconds. This suggests that detectors constructed from GaAs should have a potential bandwidth limit of 100 MHz. The high impedance of the detector elements and the capacitances of a practical detection system introduce severe limitations, which prevent the achievement of this
Bias Current = 0.1 mA

Figure IV-1-2. Noise Spectrum
ultimate bandwidth.

The frequency response of the DC biased GaAs detector, in the configuration with no aperture and an empty light pipe, was measured by varying the chopping frequency. Figure IV-J-1 illustrates the response of the detector for a bias current of 0.1 microamps and a submillimeter power of $6 \times 10^{-6}$ watts. The response was flat out to a frequency of approximately 200 Hz.

Using a capacitance bridge, the capacitance looking into the detector leads at the cryostat cover plate yields a capacitance of 208 $\text{pF}$. With the detector leads disconnected from the biasing network, the capacitance measured looking into the load resistor terminals was 213 $\text{pF}$. For a detector resistance of 3.7 $\text{M} \Omega$ and a load resistance of 1 $\text{M} \Omega$, a break frequency of 480 Hz was predicted. Thus, the severe limitation in frequency response of the detection system was due to capacitive shunting of the high impedance load resistor and detector element.

IV-K. Summary

The performance of a DC biased GaAs extrinsic photoconductive detector has been experimentally characterized. The response of the GaAs increased linearly with bias current until the electric field in the detector saturated as breakdown from impact ionization occurred. A responsivity of $5.5 \times 10^4$ volts/watt was achieved. The response of the detection system saturated at a power level of $2 \times 10^{-3}$ watts of 337 micron radiation.

The noise limiting the sensitivity of the detection system consisted of thermal noise of the load resistor at low values of bias current and 1/f noise at high values of bias current.
Figure IV-J-1. Frequency Response of DC Biased GaAs
Optimum performance of the detection system occurred at an intermediate value of bias current, where the signal to noise ratio was a maximum. An NEP of $3.83 \times 10^{-11}$ watts/Hz$^{1/2}$ was achieved.

The response of the system to amplitude modulated signals with frequency content greater than 200 Hz was severely degraded. This was due to capacitive shunting of the high impedance detector and load resistor. The construction of a detection system using DC biasing techniques, which will achieve the potential bandwidth of the GaAs material itself, appears to be an extremely difficult task, if possible at all.

The next two chapters (V and VI) deal with the development and evaluation of a detection system using GaAs as the sensing element, whose bandwidth exceeds that of the DC biased detection system. It uses microwave biasing techniques. By proper design, the bandwidth of the system could be higher than 100 MHz, allowing full utilization of the potential of GaAs as a very fast submillimeter detector.
V. Theory of Microwave Biased GaAs Detection System

V-A. Introduction to Microwave Biased Photoconductors

The use of microwave bias of a photoconductor to detect low-level, broadband optical signals was reported by Sommers and Teutsch [1]. Their theoretical study of this type of detection showed that for an intrinsic photoconductor (a) increased photocurrent gain in the semiconductor material itself could be achieved if the microwave field reverses direction before the carriers are swept out; (b) current gain results from the impedance transformation from the high resistance of the detector to the amplifier input impedance; and (c) the gain-bandwidth limitations imposed by material parameters for a DC biased detector are removed because of the use of blocking contacts. Sommers and Gatchell [2] reported an experimental study of various microwave biased intrinsic photoconductive detectors. Detectors for the near infrared and shorter wavelengths were fabricated using germanium, indium arsenide, indium antimonide and silicon. Bandwidths of up to 100 MHz were achieved.

For a microwave biased intrinsic photoconductor, the reversal of the microwave field prevents the minority carriers from being swept out. This produces photocurrent gain in the semiconductor itself. Since there is no minority carrier conduction in extrinsic photoconductors, a microwave biased extrinsic photoconductor will not exhibit this advantage over a DC biased detector. However, as pointed out by Sun and Walsh [3,4], microwave biasing can still provide an advantage over DC biasing by supplying an impedance transformation from the high impedance characteristic of an extrinsic photoconductor, to the low impedance of the microwave
detection circuitry. This impedance transformation can be achieved by mounting the detector element in a resonant cavity. Thus, the microwave detection system is not susceptible to the stray capacitance shunting of the high impedance detector element as in the DC biased detection system. The bandwidth limit is determined by the resonant frequency and the loaded Q of the cavity, i.e., \[ B = \frac{f}{2Q_L} \]

As discussed previously, experimental evidence indicates that the lifetime of the electrons in GaAs submillimeter photoconductive detectors is approximately $10^{-8}$ seconds. Therefore, the ultimate bandwidth limit should be approximately 100 MHz. It is obvious that considerable improvement in the bandwidth limit of 200 Hz measured for the DC biased detector described in Chapter IV can be achieved if the stray-capacitance problem is avoided. Therefore, a microwave biased detection system using GaAs was developed. The following sections discuss:

(a) GaAs detector element fabrication; (b) design of a reflection cavity in which to mount GaAs; (c) an equivalent circuit model and parameters characterizing the cavity; (d) theoretical study of conditions for optimum responsivity of the microwave biased detection system to submillimeter radiation; and (e) the theoretical frequency response of such a system.

V-B. Detector Material

The GaAs used for a detector element was supplied by Cayuga Associates. The structure of the material is shown in Figure V-B-1. The thicknesses of the top contact layer and the epitaxial layer were determined by angle lapping and staining. The n-type epitaxial layer of donor concentration $7.6 \times 10^{14} \text{ cm}^{-3}$ (as quoted by Cayuga Associates) was grown on a degenerately doped substrate.
Figure V-B-1. Cross Section of GaAs Wafer

Metallization

n\textsuperscript{+} Contact Layer

n Epitaxial Layer

\[ N_D = 7.6 \times 10^{14} \text{ cm}^{-3} \]

200 \text{ \AA}

Degenerately Doped Substrate

n\textsuperscript{+} Contact Layer

6 \text{ \AA}

130 \text{ \AA}

6 \text{ \AA}
In order to produce a suitable detector element from this wafer, the top metallization and contact layer had to be removed. A small element was cut from the wafer (typical 75 mils X 75 mils). Mechanical lapping with 1.0 micron alumina polishing compound was used to remove the two top layers. The element was then polished with .05 micron alumina polishing compound. It was difficult to obtain a good surface finish due to breakage along the edges of the element, with subsequent scratching of the polished surface by the freed particles of GaAs. Finally, the detector element was etched in a 5% Bromine-Methanol solution to remove some of the surface damage introduced by lapping.

V-C. Reflection Cavity

The type of microwave cavity utilized in the submillimeter detection system to be discussed here had to meet certain fundamental, practical criteria: (a) Its resonant frequency must be in the X-band (8-12 GHz); (b) the GaAs detector element must be able to be removed and replaced; (c) it must be leak-tight to liquid helium; (d) some means must be provided for illuminating the GaAs detector element; and (e) the coupling of the cavity to the microwave circuitry must be variable.

Figure V-C-1 illustrates the design of the reflection cavity which was used. The cavity is a capacitively loaded coaxial cavity. With the center post completely removed, the field configuration is that of the TM_{0,1,1} mode (Figure V-C-2). With the center post fully inserted, a perfect coaxial cavity is formed which is resonant in the TEM_{0,0,1} mode (Figure V-C-3). With the center post partially inserted, the field configuration is a hybrid combination of these two modes [5]. The resonant frequency of this hybrid mode is a strong function of the height of the center post,
Figure V-C-1. Reflection Cavity
Figure V-C-2. Electric Field Lines for the $\text{TM}_{031}$ Mode of a Cylindrical Cavity

Figure V-C-3. Electric Field Lines for the $\text{TEM}_{031}$ Mode of a Coaxial Cavity
decreasing as the height is increased.

The GaAs element was lead-tin soldered to the top of the center post. The center post was threaded into a copper plate that formed the bottom of the cavity. The post height, and therefore the resonant frequency, were easily adjusted. The bottom of the plate was indium soldered to the cavity. After adjusting the height of the center post to the desired level, the threads were indium soldered to form a stable electrical contact and to prevent leakage of liquid helium into the cavity.

The cavity was coupled to the waveguide via a small wire probe inserted in the center of a teflon screw. The relative penetration of the coupling probe into the cavity and the waveguide was adjusted to provide the desired coupling coefficient. The cavity is supported from the top of the cryostat by .010" wall stainless steel waveguide.

Submillimeter radiation is transmitted into the cryostat and to the top of the cavity via a 0.5" O.D., .010" wall stainless steel light pipe. A 0.1" diameter aperture exists in the top of the cavity directly above the center post. Two techniques of coupling the submillimeter radiation through the aperture were investigated: (a) a polyethylene lens with a 0.5" focal length; and (b) a copper cone, 1.0" in height.

V-D. Equivalent Circuit Model and Analysis

The microwave cavity described in V-C was modeled by the equivalent circuit shown in Figure V-D-1. R, L, and C are the equivalent circuit parameters of the cavity with no sample present. \( R_2 \) and \( C_2 \) are the resistance and capacitance of the of the epitaxial layer of GaAs. \( C_1 \) is the capacitance of the gap between the surface
Figure V-D-1. Reflection Cavity Equivalent Circuit

\[ R_3 = \frac{1 + \omega^2 R_2^2 (C_1 + C_2)^2}{\omega^2 R_2 C_1^2} \]

\[ C_3 = \frac{C_1 \left[ 1 + \omega^2 R_2^2 C_2 (C_1 + C_2) \right]}{1 + \omega^2 R_2^2 (C_1 + C_2)^2} \]

Figure V-D-2. Detector Branch Equivalent Circuit
of the epitaxial layer and the top of the cavity.

Submillimeter radiation incident on the detector element photothermally ionizes donor impurities in the epitaxial layer. The generated electrons increase the conductivity of the GaAs, causing a decrease in \( R_2 \). The change in \( R_2 \) alters the loaded \( Q \) of the cavity. The resulting variations in the reflection coefficient of the cavity and the power dissipated in the cavity produce changes in the voltage and the power reflected by the cavity.

Consider the branch of the equivalent circuit which represents the effects of the detector element on the cavity impedance (Figure V-D-2). The branch admittance is

\[
Y = \frac{\omega^2 R_2 C_i^2 + j \omega C_i \left[ 1 + \omega^2 R_2^2 C_2 (C_i + C_z) \right]}{1 + \omega^2 R_2^2 (C_i + C_z)^2}.
\]

This branch may be replaced by an equivalent resistance, \( R_3 \), and an equivalent capacitance, \( C_3 \).

\[
R_3 = \frac{1 + \omega^2 R_2^2 (C_i + C_z)^2}{\omega^2 R_2 C_i^2}
\]

\[
C_3 = \frac{C_i \left[ 1 + \omega^2 R_2^2 C_2 (C_i + C_z) \right]}{\left[ 1 + \omega^2 R_2^2 (C_i + C_z)^2 \right]}
\]

In general, a change in \( R_2 \) will produce changes in both \( R_3 \) and \( C_3 \), which affect the power loss in the cavity, as well as the resonant frequency of the cavity. Figure V-D-3 shows the dependence of \( R_3 \) and \( C_3 \) on \( R_2 \). A minimum in \( R_3 \), as well as a minimum in the sensitivity of \( R_3 \) to variations in \( R_2 \), occurs at \( R_2 = \frac{1}{\omega (C_i + C_z)} \).

Maximum sensitivity of \( R_2 \) occurs at the extreme values of \( R_2 \), i.e., \( R_2 = 0 \) or \( R_2 \rightarrow \infty \). For \( R_2 \) very small, \( C_3 \) equals \( C_1 \) (for \( C_1 < C_2 \)). For large \( R_2 \), \( C_3 \) equals the series combination of \( C_1 \) and \( C_2 \). Maximum sensitivity of \( C_3 \) to
Figure V-D-3. $R_3$ and $C_3$ as Functions of $R_2$. 

\[ R_3 = (1 + \frac{C_2}{C_1})^2 R_2 \]

\[ R_2 = \frac{1}{\omega (C_1 + C_2)} \]

\[ C_3 = C_1 \]

\[ C_3 = \frac{C_1 C_2}{C_1 + C_2} \]

\[ R_2 = \frac{1}{\omega (C_1 + C_2)} \]
to changes in $R_2$ occurs at $R_2 = \frac{1}{\omega (C_1 + C_2)}$; and, minimum sensitivity at the extreme values of $R_2$. Since $C_3$ affects the resonant frequency of the cavity, maximum changes in the resonant frequency of the cavity for variations in $R_2$ occur when $R_2 = \frac{1}{\omega (C_1 + C_2)}$. Negligible changes occur for extreme values of $R_2$.

Whether a decrease in $R_2$, resulting from illumination of the GaAs with sub-millimeter radiation, increases or decreases the unloaded Q of the cavity, depends on the relative values of $C_1$, $C_2$, and $R_2$. The average power dissipated in the detector branch is

$$P_{av} = \frac{V^2}{2} \frac{\omega^2 C_1 R_2}{1 + \omega^2 (C_1 + C_2)^2 R_2^2}$$

$P_{av}$ as a function of $R_2$ is plotted in Figure V-D-4. For $R_2$ less than $\frac{1}{\omega (C_1 + C_2)}$, a decrease in $R_2$ decreases the power dissipated in this branch, resulting in an increase in the unloaded Q of the cavity. For $R_2$ greater than $\frac{1}{\omega (C_1 + C_2)}$, a decrease in $R_2$ increases the power dissipated in this branch, resulting in a decrease in the unloaded Q of the cavity.

The sensitivity of $P_{av}$ to small changes in $R_2$ is given by $\frac{\delta P_{av}}{\delta R_2}$ (Figure V-D-4).

$$\frac{\delta P_{av}}{\delta R_2} = \frac{\frac{1}{2} |V|^2 \omega^2 C_1 R_2}{1 + \omega^2 (C_1 + C_2)^2 R_2^2} \left[ 1 - \omega^2 R_2^2 (C_1 + C_2)^2 \right]$$

When $R_2 = \frac{1}{\omega (C_1 + C_2)}$, a small change in $R_2$ has little effect on the power loss in this branch. The maximum variation of power loss due to a change in $R_2$ occurs at very small $R_2$.

For the ideal detector, $C_1$, $C_2$, and $R_2$ would be selected to produce the greatest change in power loss in the cavity or the greatest change in the reflection
Figure V-D-4. $P_{av}$ and $\frac{dP_{av}}{dR_2}$ as functions of $R_2$. 

\[ R_2 = \frac{1}{\omega(c_1 + c_2)} \]
coefficient for a given \( R_2 \). However, in reality, the constraints of (a) available material, (b) the physical properties of the material, and (c) the cavity design limit the range of values of \( C_1 \), \( C_2 \) and \( R_2 \) from which one can choose. Based on the dimensions, the conductivity and the dielectric constants of the epitaxial layer of a typical detector element, approximate values of \( R_2 \) and \( C_2 \) are

\[
\frac{R_2}{\Omega} = 3.2 \\
C_2 = 6.7 \text{ pF}
\]

For a gap width of 1 millimeter, \( C_1 \) is

\[
C_1 = .07 \text{ pF}
\]

For a cavity resonant frequency of 9.4 GHz,

\[
\frac{1}{\omega (C_1 + C_2)} = 2.5
\]

Thus, \( R_2 \) is much greater than \( \frac{1}{\omega (C_1 + C_2)} \) for a typical configuration. The equivalent circuit parameters \( R_3 \) and \( C_3 \), become

\[
R_3 \approx R_2 \left(1 + \frac{C_2}{C_1}\right)^2 \\
C_3 \approx \frac{C_1 C_2}{C_1 + C_2}
\]

\( C_3 \) has no significant dependence on \( R_2 \), and so the resonant frequency remains constant for small changes in \( R_2 \). For the values of \( C_2 \) and \( C_1 \) given above, \( C_2 / C_1 \) is much greater than 1, and so

\[
R_3 \approx R_2 \left(\frac{C_2}{C_1}\right)^2 \\
C_3 \approx C_1
\]

For a smaller gap between the detector and the top of the cavity, \( C_1 \) is larger and the last approximation may not be valid.

The microwave system equivalent circuit of Figure V-D-1 can be replaced by that of Figure V-D-5, where
\[ L_0 = L \]

\[ C_0 = C + C_3 \]

\[ R_0 = \frac{RR_3}{R+R_3} = \frac{RkR_2}{R+kR_2} \]

\[ k = \left(1 + \frac{L_2}{L_1}\right)^2 \]

Figure V-D-5. Reduced Equivalent Circuit
\[ L_0 = L \]
\[ C_0 = C + C_3 \quad \text{and} \]
\[ R_0 = \frac{R R_3}{R + R_3} = \frac{R k R_2}{R + k R_2} \]
where \( k = (1 + a_r^2/c_1)^2 \).

The \( Q \) of the unloaded cavity, \( Q_0 \), is given by
\[ Q_0 = \frac{R_0}{\omega_0 L_0} \]

The maximum change in \( Q_0 \) occurs when the maximum variation of the power loss occurs. The power loss in the cavity, \( P_C \), is the sum of the wall losses and the sample losses.
\[ P_C = \frac{V^2}{R_0} = \frac{V^2}{R} + \frac{V^2}{k R_2} \]

Obviously, maximum sensitivity of \( P_C \) to changes in \( R_2 \) occurs when
\[ \frac{1}{k R_2} > \frac{1}{R} \]
such that the sample losses dominate the cavity losses.

In summary, the maximum sensitivity of the unloaded \( Q \) to changes in \( R_2 \), occurs for \( C_1 \) very large, \( C_2 \) very small, and \( R_2 \) less than \( R \). For the reflection cavity actually constructed, operation occurred far from these optimum conditions. Physical parameters were such that changes in \( R_2 \) were expected to change the losses in the cavity but not to affect the resonant frequency of the cavity.

V-E Optimization of Cavity Coupling for Maximum Responsivity

The discussion in V-D dealt with optimizing the unloaded cavity equivalent circuit parameters in order to produce maximum changes in the unloaded \( Q \) of the cavity. The means of detecting the variations in the unloaded \( Q \), and the optimum coupling coefficient for maximum responsivity are considered next.
From the equivalent circuit of Figure V-D-5, the impedance of the unloaded cavity (impedance at plane c) is given by

\[ Z_c = \frac{R_0}{1 + \frac{j}{\omega_0} R_0 \left( \omega C - \frac{1}{\omega L} \right)} \]

\[ Z_c = \frac{R_0}{1 + \frac{j}{Q_0} (\omega - \omega_0)(\omega + \omega_0)/\omega_0} \]

where \( Q_0 = \frac{R_0}{\omega_0 L_0} \) and \( \omega_0 = (L_0 C_0)^{1/2} \)

For \( \omega \approx \omega_0 \), \( (\omega + \omega_0) \approx 2 \omega_0 \)

Therefore, the impedance of the unloaded cavity is

\[ Z_c = \frac{R_0}{1 + \frac{j}{2 Q_0} \delta} \quad \text{where} \quad \delta = \frac{\omega - \omega_0}{\omega_0} \]

Transforming this impedance to the generator side of the coupling transformer gives at plane b

\[ Z_b = Z_c / n^2 = \frac{R_0 / n^2}{1 + \frac{j}{2 Q_0} \delta} \]

The equivalent circuit of Figure V-D-5 reduces to the equivalent circuit of Figure V-E-1. The Q of this circuit is the loaded Q of the cavity, \( Q_L \).

\[ Q_L = \frac{R_0}{\omega_0 L_0} \frac{n^2 Z_0}{n^2 Z_0 + R} = \frac{Q_0}{1 + \beta} \]

where the coupling coefficient \( \beta \) is defined as

\[ \beta = \frac{R_0}{n^2 Z_0} \]

The external Q, \( Q_{\text{EXT}} \), which represents the effects of the external circuit coupled to the cavity, is given by

\[ Q_{\text{EXT}} = \frac{Q_0}{\beta} \]

Thus,

\[ \frac{1}{Q_L} = \frac{1}{Q_{\text{EXT}}} + \frac{1}{Q_0} \]

The optimum value of the coupling coefficient for maximum sensitivity of the microwave signal, which is reflected from the cavity, to changes in \( Q_0 \), depends
\[ Z_b = \frac{R_0 / n^2}{1 + j2Q_0 \delta} \]

At resonance:
\[ \delta = 0 \]
\[ Z_b = R_0 / n^2 \]

Figure V-E-1. Cavity Equivalent Circuit Transformed to Generator Side of Coupling Transformer
on the type of detector being used to monitor the reflected signal [6]. The effects of the coupling coefficient on the system responsivity, when using a square-law detector, and when using a linear detector, are presented next.

Upon illumination of the GaAs detector element with submillimeter radiation, \( R_2, Q_0, Q_L \), and, therefore, the microwave power reflected from the cavity are altered. For a square-law detector, it is desired to maximize the change in power reflected from the cavity for a given change in \( R_2 \). For a constant microwave power incident on the cavity, the change in the reflected power, \( P_R \), is the negative of the change in the power dissipated in the cavity. For the microwave source operating at the resonant frequency of the cavity, the power dissipated in the cavity, \( P_C \), is

\[
P_C = \frac{V_0^2 R_0 / \mu^2}{(z_0 + R_0 / \mu)^2}
\]

The signal of a square-law detector will be proportional to \( \Delta P_R \).

\[
\Delta P_R = -\Delta P_C = -[\frac{\partial P_C}{\partial R_2}] \Delta R_2
\]

Evaluating \( \frac{\partial P_C}{\partial R_2} \) yields

\[
\frac{\partial P_C}{\partial R_2} = \left[ \frac{\partial P_C}{\partial R_0} \right] \left[ \frac{\partial R_0}{\partial R_2} \right]
\]

\[
= \left[ \frac{V_0^2}{\mu^2} \frac{(z_0 - R_0 / \mu^2)}{(z_0 + R_0 / \mu)^3} \right] \left[ \frac{R_0^2}{kR_2^2} \right]
\]

Expressing in terms of the coupling coefficient \( \beta = R_0 / z_0 \mu^2 \)

\[
\frac{\partial P_C}{\partial R_2} = \frac{4 R_0}{kR_2} P_{\text{MAX}} \frac{\beta - \beta^2}{(\beta + 1)^3}
\]

\( P_{\text{MAX}} \) is the maximum power \( \left[ V_0^2 / 4z_0 \right] \) that can be transferred to the cavity under critically coupled conditions, \( \beta = 1 \). Since VSWR is the quantity most readily measured in the laboratory, the expression for \( \frac{\partial P_C}{\partial R_2} \) expressed in terms of VSWR becomes (Figure V-E-2),
Figure V-E-2. Optimum Cavity Coupling for a Square-Law Detector
\[ \frac{\partial P_0}{\partial R_2} = \pm 4 \frac{R_0}{k R_2^2} P_{\text{max}} \frac{V_{\text{SWR}} - (V_{\text{SWR}})^2}{(V_{\text{SWR}} + 1)^3} \]

where \(+\) sign is for an overcoupled cavity, \(\beta > 1\); and \(-\) sign is for an undercoupled cavity, \(\beta < 1\). Maxima of \(\pm 0.096 (R_0/k R_2^2) 4 P_{\text{max}}\) occur for a VSWR of 3.73, or a reflection coefficient of 0.58. Thus, when using a power sensitive detector to measure the signal reflected from the cavity, maximum responsivity to changes in \(R_2\) occurs when the cavity and waveguide are far from critically coupled.

For a linear detector, such as a mixer, it is desired to maximize the change in the voltage reflected from the cavity for a given change in \(R_2\). The voltage reflected from the cavity at resonance is

\[ V_R = V_0 \Gamma = V_0 \frac{R_0/n^2 - Z_0}{R_0/n^2 + Z_0} \]

A change in \(R_2\) due to incident submillimeter radiation will produce a change in \(V_R\) given by

\[ \Delta V_R = \left[ \frac{\partial V_R}{\partial R_2} \right] \Delta R_2 \]

Evaluating \(\partial V_R / \partial R_2\) yields

\[ \frac{\partial V_R}{\partial R_2} = \frac{\partial V_R}{\partial R_0} \frac{\partial R_0}{\partial R_2} = \frac{R_0}{k R_2^2} \frac{R_0}{n^2 Z_0} \frac{2 V_0}{(R_0/n^2 Z_0 + 1)^2} \]

Expressing in terms of coupling coefficient \(\beta\) gives

\[ \frac{\partial V_R}{\partial R_2} = \frac{R_0}{k R_2^2} \frac{2 V_0 \beta}{(\beta + 1)^2} \]

In terms of VSWR, \(\partial V_R / \partial R_2\) is (Figure V-E-3),

\[ \frac{\partial V_R}{\partial R_2} = \frac{R_0}{k R_2^2} \frac{2 V_0 \text{VSWR}}{(\text{VSWR} + 1)^2} \]

\(\partial V_R / \partial R_2\) has a maximum value of \(R_0/k R_2^2 V_0/4\) at a VSWR of 1.0, or a coupling coefficient \(\beta = 1\), and decreases monotonically as the cavity coupling to the waveguide is decreased or increased from a matched condition.
Figure V-E-3. Optimum Cavity Coupling for a Linear Detector
In summary, there is a significant difference in the optimum value of the coupling coefficient for maximum system responsivity to changes in $R_2$, when using a square-law detector as opposed to a linear detector. For a perfect linear detector, a cavity critically coupled to the waveguide should give maximum responsivity. For a perfect square-law detector, a cavity mismatched to the waveguide to produce a VSWR = 3.73 should give maximum responsivity.

V-F. Frequency Response

To investigate the fundamental limitations of the bandwidth of a microwave biased detection system, assume that the carrier lifetime is arbitrarily short so that the response time of the material is not limiting the bandwidth. The techniques used to detect microwave signals are sufficiently developed that a system for receiving the signal reflected from the cavity, with a bandwidth in the GHz range, is achievable. Most of the other miscellaneous microwave components used in the main signal path of a microwave biased detection system can be obtained with sufficiently broad-band capabilities.

If the radiation illuminating the detector element undergoes a step increase in power level, the excess carrier concentration increases and the detector resistance decreases. The unloaded Q of the cavity and, therefore, the loaded Q change. The steady state amplitude of the electric and magnetic fields inside the cavity change to account for the increase in power loss in the cavity. The decay time, $\tau_c$, required for a steady state to be reached determines the bandwidth of the cavity.
\[ \Delta f = \frac{1}{2\pi \tau_c} \]

For a resonant circuit, the decay time equals the real part of the natural frequencies associated with the circuit. For a parallel RLC circuit, the natural frequencies are

\[ s = -\frac{1}{\omega RC} \pm \sqrt{(\frac{1}{\omega RC})^2 - \frac{1}{\omega L}} \]

Since the loaded Q is given by

\[ Q_L = \frac{\omega_0}{RC} \]

\[ s = -\frac{\omega_0}{2Q_L} \pm \frac{\omega_0}{2Q_L} \sqrt{\left(\frac{1}{2Q_L}\right)^2 - 1} \]

Therefore, the decay time \( \tau_c \) is given by \( \frac{2Q_L}{\omega_0} \) and the bandwidth of the cavity is given by

\[ \Delta f = \frac{f_0}{2Q_L} \]

Thus, the fundamental limit of the bandwidth of a microwave biased detection system is determined by the resonant frequency and loaded Q of the cavity. By properly designing the cavity, a desired bandwidth can be achieved. The system can be designed to utilize the full bandwidth potential of a given detector material.
VI. Experimental Results for Microwave Biased GaAs Detector

VI-A. Microwave Biased Detection System

The microwave circuitry utilized in the detection system is shown in Figure VI-A-1. The main signal path consists of the X-13 Klystron, circulator, reflection cavity, circulator and crystal detector. Miscellaneous equipment is utilized for measuring the frequency and power level of the microwave signal incident on the cavity, and for measuring the VSWR in order to determine the degree of mismatch of the cavity to the waveguide. A slide-screw-tuner (SST) was inserted in the signal path just prior to the waveguide entrance into the LHE cryostat, in order to vary the effective coupling of the cavity and waveguide.

Stabilization of the operating frequency of the klystron to the resonant frequency of the cavity is necessary in order to minimize AM and FM noise. The fluctuations of the reflection cavity resonant frequency due to thermal changes and mechanical vibrations of the cavity submerged in liquid helium prevented stabilization with respect to an external reference cavity. Thus, it was necessary to use the reflection cavity as the reference cavity in the stabilization circuit. Using two 10 db directional couplers, the signal incident on the cavity and the signal reflected from the cavity were transmitted to ports 1 and 2 of a magic tee discriminator. The signals incident at ports 1 and 2 arrive at crystal detector A, terminating the E plane arm, shifted 180° from their original relative phase, whereas the signals incident at ports 1 and 2 arrive at crystal detector B with their original relative phase preserved. The phase of the signal reflected from the cavity is very frequency dependent, when the frequency of operation is close to resonance.
Combining this reflected signal with the incident signal (entering port 1) converts a signal whose phase is dependent on frequency to a signal whose amplitude is dependent on frequency. Subtracting the outputs of crystals A and B in a differential amplifier produces a discriminator characteristic. The discriminator output was amplified and used to drive an LED in an optically coupled isolator. The output of the optically coupled phototransistor was fed back to the reflector supply to correct for the frequency error.

To study the performance of this microwave biased detection system, the HCN laser was used as a source of submillimeter radiation. A system of 0.5'' O. D. stainless steel light pipes was utilized to transmit the radiation into the liquid helium cryostat and to the Eppley thermopile for monitoring the power of the submillimeter radiation. The experimental set-up was identical to that described in IV-G, Figure IV-G-2.

VI-B. Responsivity Measurements

The analysis of V-E predicted that the system responsivity was a function of VSWR and that the peak responsivity should occur at different cavity coupling coefficients when using a square-law detector and when using a linear detector to monitor the microwave signal reflected from the cavity.

Figure VI-B-1 is a plot of experimental data showing the relative response of the microwave biased detection system versus VSWR. The incident submillimeter radiation, at a power level of 2.2 milliwatts measured by the Eppley thermopile, was chopped at 100 Hz. Ten milliwatts of microwave power were incident on the cavity at a frequency of 9.4 GHz. The AC signal from the crystal detector was
Figure VI-B-1. Responsivity versus Cavity Coupling
normalized to the peak value which was obtained at a VSWR = 1.85.

The square-law region and the linear region of the crystal detector were
determined. From Figure VI-B-2, it can be seen that the crystal detector behaves
as a perfect square-law detector for microwave power levels less than $10^{-5}$
watts, or a corresponding DC level of 18 millivolts. From VI-B-2 it can also be seen
that the crystal detector can act as a perfect linear detector only for small deviations
in power around the power level of $5 \times 10^{-3}$ watts, or a corresponding DC level
of 1.6 volts.

The DC level of the crystal detector was recorded as the data for Figure
VI-B-1 was taken. Figure VI-B-3 shows the DC level of the crystal detector as
a function of VSWR. It can be seen that the crystal detector behaved as a square-
law detector for VSWRs between 1.0 and 1.2, and that for VSWRs around 3.0 it
was approaching the linear detection region of its characteristics. Since the
behavior of the crystal detector was not characteristic of a square-law detector
or a linear detector throughout the entire range of VSWRs which was covered in
Figure VI-B-1, the peak responsivity occurred at a value of VSWR between the
value of 3.73 predicted for a square-law detector and the value of 1.0 predicted
for a linear detector.

Operating at a VSWR of 1.85, the crystal detector signal was measured as
a function of incident submillimeter power for a chopping frequency of 1000 Hz
(Figure VI-B-4). The system response was linear with incident power over the
entire range of measurements from $5 \times 10^{-6}$ watts to $10^{-3}$ watts. The responsivity
was 5.28 volts/watt.
Figure VI-B-2. Linear and Square-Law Regions of Crystal Detector
Figure VI-B-3. DC Level of Crystal Detector versus VSWR
$f = 1000 \text{ Hz}$

Figure VI-B-4. Responsivity of Microwave Biased Detector
VI-C. Noise Measurements

The noise of the detection system was determined by measuring the rms noise present in a given bandwidth about the chopping frequency, and then calculating the rms noise per square root of the equivalent noise bandwidth. The rms noise was measured using the band pass amplifier in the signal channel of a PAR Model 120 Lock-In-Amplifier. With the band pass characteristics of the amplifier adjusted to a center frequency of $f_0$ and a Q of $Q_0$, the equivalent noise bandwidth was given by

$$ENBW = \left(\frac{1}{V_2}\right)^{\left(f_0/Q_0\right)}$$

The rms noise per $\sqrt{ENBW}$ was measured as a function of the cavity coupling coefficient, or equivalent VSWR (Figure VI-C-1). The noise increased monotonically as the VSWR was increased.

Experimental information was gathered which indicated that the noise of the microwave biased detection system was due to system noise and not to detector element noise. There were four principal sources of noise: (a) the crystal detector; (b) microphonics; (c) frequency locking; and, (d) circulator isolation.

The crystal detector has inherent sources of noise. The predominant types of crystal noise are shot noise and thermal noise. Shot noise arises from the discreteness of electronic charge and the random arrival of electrons at the contacts. Shot noise is current dependent. Thermal noise arises from the Johnson noise of the spreading resistance of the contacts. However, the minimum noise achieved in the microwave biased detection system was at least an order of magnitude greater than the crystal detector noise.

Mechanical vibration of the waveguide components introduces microphonic
Figure VI-C-1. Noise versus Cavity Coupling
noise. The vibration of the flange joints produce reflections as well as changes in the electrical length of the line. Vibration of the cavity due to boiling of liquid helium may introduce microphonic noise as a result of vibration of the coupling probe. Microphonic noise takes the form of phase noise as well as amplitude noise. However, microphonic noise did not appear to be the predominant source of noise.

Relative fluctuations between the cavity resonant frequency and the klystron oscillating frequency arose primarily from (a) variations in the reflector supply causing the klystron frequency to change; (b) thermal expansion and contraction of the cavity causing resonant frequency to be altered; and (c) frequency noise inherent in klystron operation. The relative fluctuations between the cavity resonant frequency and the klystron frequency produce fluctuations in the magnitude and the phase of the reflection coefficient of the cavity.

With no frequency locking, the crystal detector signal contained significant amounts of 60 cycle noise and a certain amount of broad band noise. Frequency locking with a Pound stabilizer decreased the overall noise by a factor of 10, eliminating the 60 cycle noise and the low frequency components of the broad band noise. Since the cavity had a relatively low Q ($Q_L = 1160$), frequency locking eliminated amplitude noise due to fluctuations in the magnitude of the reflection coefficient. The gain of the stabilization circuit that was required in order to eliminate the 60 cycle noise originating in the reflector supply produced high frequency oscillation in the feedback loop. This required filtering the high frequency components out of the feedback loop. Thus, only relative frequency fluctuations of the cavity and klystron at a rate of 200 Hz or less were satisfactorily eliminated with the frequency locking circuit.
The circulator, utilized to separate the microwave signal incident on the cavity from the microwave signal reflected by the cavity, possessed 20 db of isolation between adjacent decoupled ports. Mixing of the cavity reflected signal and the leakage signal in the crystal detector was observed. As a result of this mixing, the output of the crystal detector contained a term which depended on the product of the amplitudes of the two signals incident on the crystal and on the cosine of their relative phases. The presence of this term was experimentally observed using a phase shifter to vary the phase of the cavity reflected signal.

Fluctuation of the relative phase of the two signals incident on the crystal detector introduces noise. There are several sources of fluctuation of this relative phase. (1) FM noise of the klystron cannot be completely eliminated. (2) Assume the klystron is perfectly locked to the cavity. Any shift in cavity resonant frequency results in a shift in the frequency of the klystron. Therefore, the electrical length of the signal path is altered. Hence, a fluctuation results in the relative phase of the two signals arriving at the crystal detector. (3) If frequency locking is not perfect, the resulting fluctuation of the phase of the reflection coefficient of the cavity produces fluctuations in the relative phases of the two signals arriving at the crystal detector.

Insufficient circulator isolation is believed to be the predominant source of noise. It was found that the noise of the system could be minimized by adjusting the relative phase of the two signals incident on the crystal detector until it was 0° or 180°. Complete elimination of this type of noise requires a circulator with better isolation.
VI-D. Best Performance

Based on the data of Figures VI-B-1 and VI-C-1, the relative signal to noise ratio as a function of VSWR was calculated (see Figure VI-D-1). To obtain best performance from the detection system, coupling to the cavity should be adjusted for a VSWR of 1.5. This will give maximum signal to noise ratio.

The maximum responsivity achieved was 5.28 volts/watt; the corresponding noise voltage was 0.8 microvolts/Hz$^{1/2}$. This gives an NEP of $5 \times 10^{-7}$ watts/Hz$^{1/2}$.

VI-E. Frequency Response

The frequency response of the microwave biased detection system was measured by varying the chopping frequency. The response was flat out to 1 kHz, which was the maximum chopping frequency attainable. This performance exceeds that of the DC biased detector which had a bandwidth of approximately 200 Hz.

The impedance of the cavity at the plane of the detuned short was measured as a function of frequency. The loaded Q of the cavity was determined from a Smith chart plot of this impedance by using graphical techniques. The loaded Q had a value of 1160. The resonant frequency of the cavity was 9.4 GHz. The bandwidth limit of the system determined by the cavity was

$$\Delta f = \frac{f_0}{2Q_L} = 4 \text{ MHz}$$

Currently, modulators for 337 micron radiation which would allow verification of this bandwidth do not exist. If two sources of submillimeter radiation of the approximate wavelengths were available, a mixing experiment could be performed to determine the bandwidth of the system.
Figure VI-D-1. Signal to Noise Ratio versus Cavity Coupling
VII. Conclusions

For the wavelength range from 100 to 350 microns, high purity GaAs is one of the most sensitive, as well as one of the fastest, materials available for constructing detectors. DC biasing techniques allow one to use a very simply constructed system. Responsivities of $10^4$ volts/watt and NEPs of $10^{-11}$ watts/Hz$^{1/2}$ can quite easily be achieved. However, severe limitations in bandwidth occur because of capacitive shunting. The DC biased GaAs detection system is ideal for applications where a large bandwidth is not required.

Microwave biasing allows the construction of an extremely wide band detection system. The bandwidth limit is determined by the ratio of the resonant frequency to twice the loaded Q of the cavity. It is possible, then, to design the microwave cavity to have a bandwidth limit which exceeds the reciprocal of the carrier lifetime. The full potential of the time response of the material can then be utilized.

At the present stage of development, the microwave biased detection system has a much lower responsivity than that achieved with DC biasing. Improvements in responsivity can be made in two categories: (a) the receiving system used to monitor the microwave signal reflected from the cavity; and (b) detector configuration. The use of heterodyne detection should provide increased responsivity and signal to noise ratio over the crystal detector employed in the present system. The detector element configuration was far from optimum. The thickest possible epitaxial layer should be used to decrease the capacitance of the epitaxial layer. A detector element with an insulating substrate appears to have some advantages over an element with a degenerately doped substrate. If the cavity
design was altered so that the submillimeter radiation entered through a side wall of the cavity, a detector element with an insulating substrate could be mounted atop the center post so that the epitaxial layer surface was parallel to the axial direction of the cavity. This would minimize the shunting effects of the epitaxial layer capacitance.

Despite its low responsivity, a microwave biased GaAs detector has definite promise as a wideband submillimeter detector.
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