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**FINAL REPORT ON:
AN AIRBORNE STUDY OF
MICROWAVE SURFACE SENSING AND
BOUNDARY LAYER
HEAT AND MOISTURE FLUXES FOR FIFE**

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Abstract

This is the final report on the project titled "An Airborne Study of Microwave Surface Sensing and Boundary Layer Heat and Moisture Fluxes for FIFE," Grant NAG 5-912. The objectives of this work were to perform imaging radar and scatterometer measurements over the Konza Prairie as a part of the First International land surface climatology project Field Experiments (FIFE) and to develop an mm-wave radiometer and the data acquisition system for this radiometer. We collected imaging radar data with The University of Kansas Side-Looking Airborne Radar (SLAR) operating at 9.375 GHz and scatterometer data with a helicopter-mounted scatterometer at 5.3 and 9.6 GHz. We also developed a 35-GHz null-balancing radiometer and data acquisition system.

Although radar images showed good delineation of various features of the FIFE site, the data were not useful for quantitative analysis for extracting soil moisture information because of day-to-day changes in the system transfer characteristics. Our scatterometer results show that both C and X bands are sensitive to soil moisture variations over grass-covered soils. Scattering coefficients near vertical are about 4 dB lower for unburned areas because of the presence of a thatch layer, in comparison with those for burned areas.

The results of the research have been documented in reports, oral presentations, and published papers. A list of reports and papers is provided at the end of this report.

Summary

The First International land surface climatology project Field Experiments (FIFE) were conducted on the Konza Prairie Natural Reserve Area (KPNRS) south of Manhattan, Kansas, during 1987. The goal of the FIFE was to collect data needed to interpret satellite observations of climatologically important land-surface parameters such as biomass, leaf area, and soil moisture [1]. As a part of this program we carried out a research project titled "An Airborne Study of Microwave Surface Sensing and Boundary Layer Heat and Moisture Fluxes for FIFE." The specific tasks associated with this program were the following:

1. Modify and operate The University of Kansas X-band SLAR for studying the usefulness of an X-band imaging radar in determining soil and vegetation parameters.
2. Perform backscatter measurements using a multifrequency scatterometer from a NASA helicopter.
3. Develop a null-balancing 35-GHz radiometer.
4. Develop a data acquisition system for collecting data with the radiometer

We developed a digital data acquisition system to record KU SLAR data to replace the analog recording system that we used prior to the FIFE. SLAR transmits a series of electromagnetic pulses periodically to illuminate the target being imaged and receives a reflected signal from the target. The amplitude of the received echo is controlled by the target-scattering characteristics, and time delay is proportional to the target range. The digital data acquisition system digitizes, averages and stores the received signals for further processing and provides a real-time display of data in a sequence that forms a radar image. The development of the hardware and software for the data acquisition system are discussed in the report by Ewbank, Rummer and Gogineni [2]. Fig. 1 shows a characteristic SLAR image of the Konza Prairie.

We collected backscatter data along selected flight lines at 5.3 and 9.6 GHz using a scatterometer mounted on the NASA helicopter during 1987. We acquired these data with HH polarization at incidence angles of 0°, 15°, 30°, 45° and 60°. We inverted radar data to estimate soil moisture profiles and compared these with in situ measurements made along the flight lines. The results are documented in papers by Gogineni, Ampe and Budihardjo [3], Wang, Gogineni and Ampe [4], and Gogineni et al. [5] (see Appendix I for these three publications).

A null-balancing dual-polarized radiometer was designed and built as part of a graduate student's master's project. Design, development and test results are documented in a report by Ampe [6] (see Appendix II).

In addition we also developed a data acquisition system to operate and collect data with this radiometer. The documentation on this is reported by Parikh [7] (see Appendix III).

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- [3] Gogineni, S., J. Ampe, and A. Budihardjo, "Radar Measurements of Soil Moisture over the Konza Prairie," *International Journal of Remote Sensing*, vol. 12, no. 11, pp. 2425-2432, 1991.
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- [5] Gogineni, S., A. K. Fung, K. S. Chen, and J. Wang, "Comparison of Measurements and Theory for Backscatter from Vegetation-Covered Soil on the Konza Prairie," *IGARSS'92 Digest*, IEEE 92CH3041-1, Houston, Texas, May 26-29, pp. 920-922, 1992.
- [6] Ampe, J. "Design and Construction of 37-GHz Noise Injection Radiometer," Appendix II, Radar Systems and Remote Sensing Laboratory Technical Report 7630-4, October 1995.
- [7] Parikh, M., "Data Acquisition System for 37 GHz Radiometer Experiment," Appendix III, Radar Systems and Remote Sensing Laboratory Technical Report 7630-4, October 1995.

Publications for Grant NAG 5-912

An Airborne Study of Microwave Surface Sensing and Boundary Layer Heat and Moisture Fluxes for the First ISLSCP Field Experiment

Journal Articles

Wang, J. R., S. P. Gogineni, and J. Ampe, "Active and Passive Microwave Measurements of Soil Moisture in FIFE," *J. Geophys. Res.*, vol. 97, no. D17, pp. 18,979-18,985, 1992.

Gogineni, S., J. Ampe, and A. Budihardjo, "Radar Measurements of Soil Moisture over the Konza Prairie," *International Journal of Remote Sensing*, vol. 12, no. 11, pp. 2425-2432, 1991.

Master's Project

Osman, M. S., "Step Sensitivity-Time-Control Design and Swath Expansion of The University of Kansas Sidelooking Airborne Radar," M. S. Project, The University of Kansas, 1989.

Conference Presentations

Gogineni, S., "Radar Measurement of Soil Moisture Over the Konza Prairie," Proceedings of Symposium on FIFE, American Meteorological Society, Anaheim, California, February 7-9, 1990, pp. 88-91.

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Ewbank, S. E., D. I. Rummer and S. Gogineni, "A Digital Data Acquisition System for Use with the University of Kansas Sidelooking Airborne Radar," Radar Systems and Remote Sensing Laboratory Technical Report 7630-1, January 1988.

Moore, R. K. and S. P. Gogineni, "An Airborne Study of Microwave Surface Sensing and Boundary Layer Heat and Moisture Fluxes for the First ISLSCP Field Experiment," Radar Systems and Remote Sensing Laboratory Technical Report 7630-2, February 1988.

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Gogineni, S. P., "Final Report for 'An Airborne Study of Microwave Surface Sensing and Boundary Layer Heat and Moisture Fluxes for FIFE,'" Radar Systems and Remote Sensing Laboratory Technical Report 7630-4, October 1995.

Ampe, J. "Design and Construction of 37-GHz Noise Injection Radiometer," Appendix II, Radar Systems and Remote Sensing Laboratory Technical Report 7630-4, October 1995.

Parikh, M. "Data Acquisition System for 37 GHz Radiometer Experiment," Appendix III, Radar Systems and Remote Sensing Laboratory Technical Report 7630-4, October 1995.

Technical Memoranda

Gogineni, S. P. and R. K. Moore, "Semi-Annual Progress Report on an Airborne Study of Microwave Surface Sensing and Boundary Layer Heat and Moisture Fluxes for FIFE," Radar Systems and Remote Sensing Laboratory Technical Memorandum 7630-1, November 1988.

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Gogineni, S. P., "Semi-Annual Progress Report on an Airborne Study of Microwave Surface Sensing and Boundary Layer Heat and Moisture Fluxes for FIFE," Radar Systems and Remote Sensing Laboratory Technical Memorandum 7630-3, September 1989.

Ampe, J. M. and S. P. Gogineni, "Semi-Annual Progress Report on an Airborne Study of Microwave Surface Sensing and Boundary Layer Heat and Moisture Fluxes for FIFE," Radar Systems and Remote Sensing Laboratory Technical Memorandum 7630-4, February 1990.

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Appendix I

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Radar estimates of soil moisture over the Konza Prairie

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Abstract. Radar backscatter measurements were made as a part of the First International satellite land surface climatology project Field Experiment (FIFE) to estimate soil moisture for use by other investigators. The helicopter-mounted radar was flown along selected transects that coincided with soil moisture measurements. The radar operated at microwave frequencies of 5.3 and 9.6 GHz and at selected incidence angles between 0° and 60°. Vertical polarization was used for two days in June of 1987 and horizontal polarization was used for three days in July and October of 1987.

The scattering coefficient data from different days were grouped by frequency and antenna angles and then related to soil moisture along the flight paths using linear regression. A measure of linearity for the regression, R^2 , ranged between 0.9 and 0.5. The larger coefficients were for X-band measurements made at large antenna incidence angles, and the smaller coefficients were for C-band measurements made at incidence angles near vertical.

1. Introduction

Active microwave measurements over land have been studied by researchers to determine the effects of soil and vegetation on the backscattered energy (Dobson and Ulaby 1986, Jackson and Schmugge, 1981, Ulaby 1974, Ulaby *et al.* 1982). The primary goal of these previous investigations was to determine the optimum set of system parameters such as frequency, polarization and incidence angle for maximizing the sensor response to soil moisture and minimizing the response to vegetation and surface roughness. The conclusions from these previous studies are that C-band radars operating over incidence angles between 7° and 17° are effective in estimating soil moisture. However, with the exception of the studies by Wang *et al.* (1983) these reported measurements were conducted on agricultural lands.

In this Letter we report the results of a helicopter-mounted scatterometer in estimating soil moisture along flight transects. The backscatter measurements were made over the Konza Prairie Natural Research Area near Manhattan Kansas. Our estimates were part of the First ISLSCP (International Satellite Land Surface Climatology Project) Field Experiment (FIFE) conducted from May to October 1987. The primary objective of this experiment was to collect data required to interpret satellite sensor observations in terms of climatologically significant land surface parameters. The objective of our measurement programme was to use a radar to estimate soil moisture on the Konza Prairie for use by other scientific investigations.

2. System description

A dual-frequency FM radar, termed HELOSCAT, was used for backscatter measurements over the Konza Prairie. This system collected backscatter data at both

C-band (5.3 GHz) and X-band (9.6 GHz). The antenna was mounted on the helicopter with a pivot joint that allowed an electric actuator to change the incidence angle on successive passes. The antenna incidence angles utilized were 0°, 15°, 30°, 45°, and 60°. Linear VV polarization was used for two days in June and HH polarization was used for two days in July and one in October. The system was operated at an average altitude of 30 m for incidence angles less than 45° and an average of 15 m for incidence angles greater than 45°. System specifications are shown in table 1. A more detailed system description is given in Gogineni *et al.* (1984). The radar antenna as installed on the helicopter is shown in figure 1.

The radar system was calibrated both internally and externally for estimating the backscatter coefficient from the measured power return. The internal calibration equipment consisted of a length of microwave cable short circuited at the far end. The transmitted microwave signal was directed to the delay line, and the reflected power was recorded. The internal calibration was performed before and after each data run over a particular flight transect. The external calibration consisted of measuring the power reflected by a standard radar target of known radar cross-section. The external calibration was performed before and after mounting the unit on the helicopter.

3. Site and experiment description

The Konza Prairie is in the western part of the tallgrass prairie region of the United States. The annual precipitation in this area is about 800 mm. The tallgrass grows to a height of about 500 mm, and flower stalks may reach a height of 1000 mm under favourable conditions (Schmugge *et al.* 1988). The vegetation cover is sparse with a leaf area index of approximately 1.0 for grasses and 0.25 for weeds (Zoughi 1988). The area has rolling hills with peak-to-valley height differences of about 15 m. The flight transects covered vegetation previously unburned, burned one year previously, and burned two years previously.

The flight paths were approximately 3500 m in length and covered areas where direct soil moisture samples were taken by other teams. The 15 surface zone (0–5 cm depth) soil measurements per transect covered approximately 800 m of the transects with samples spaced approximately 25 m apart. Volumetric soil moisture values were calculated from gravimetric measurements and soil bulk density measurements. The accuracy of the moisture measurement is approximately ± 10 per cent. The samples were taken within five hours of each backscatter measurement flight.

Table 1. Important system specifications.

Radar type	FM-CW
Radar frequency	4–17 GHz
Polarization	VV, VH, HV, HH
IF	
Centre Frequency	50 kHz
Bandwidth	13.5 kHz
Antenna	
Beamwidth	5° at 5.3 GHz 3° at 9.6 GHz
Calibration	
Internal	Delay line
External	Luneburg Lens

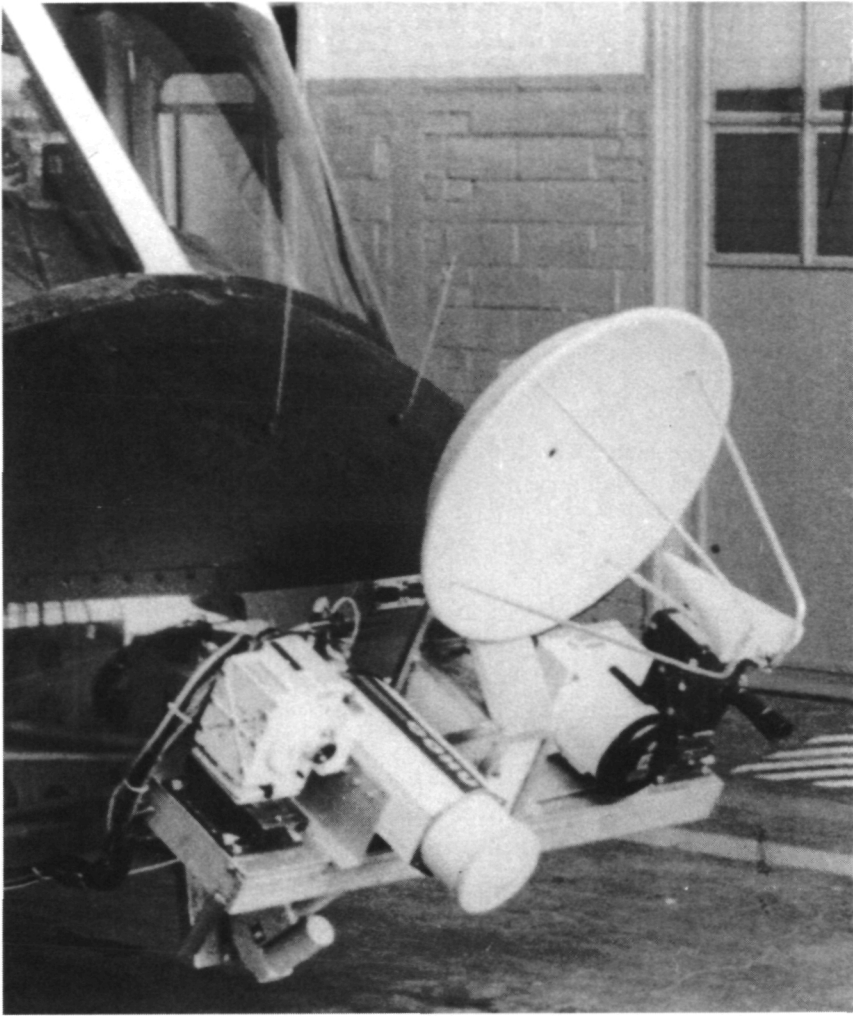


Figure 1. Heloscat radar mounted on NASA helicopter.

Range to ground was determined by the frequency difference between transmission and reception. The data recording system digitized the received microwave power level five times a second and the range to ground twice a second. The data in the time base were converted to position using the recorded range information and a digital terrain map of the area. The period between samples was compressed or expanded to align the recorded range information with known significant terrain contours. The resulting horizontal distance between backscatter samples averaged five metres.

4. Results

The recorded power was converted to the backscattering coefficient (σ^0) using a calibration target and other previously described calibration factors. The recorded range information was then used to adjust the σ^0 values for local changes in terrain slope. The correction for slope used functions of average σ^0 versus antenna incidence

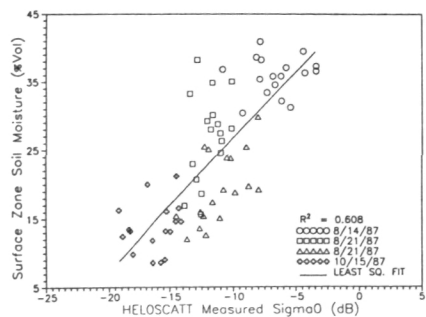


Figure 2. σ^0 versus measured soil moisture at C-band, 15°.

angle. The slope-corrected σ^0 data were then smoothed using a triangle weighting function. In addition to eliminating the fluctuations due to fading, the averaging reduced the error due to misalignment of actual soil measurement position with estimated soil moisture position. Orthogonal linear regression was then performed between the two variables of measured soil moisture and σ^0 . The σ^0 data were then transformed using the linear regression slope and offset as an estimate of the soil moisture.

The results of the analysis of data at C-band for antenna angles of 15° and 30° are shown in figures 2 and 3. Statistics of the linear regressions, coefficients of determination (R^2) and standard errors ($\sqrt{\text{MSE}}$) are given in table 2. The C-band 15°

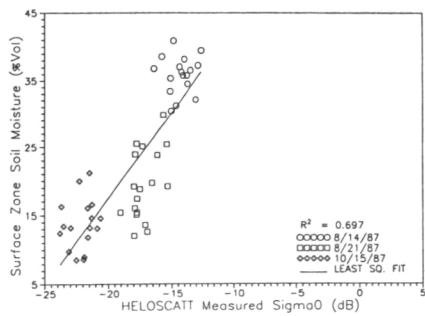


Figure 3. σ^0 versus measured soil moisture at C-band, 30°.

Table 2. Linear regression statistics between backscatter and soil moisture for both C-band and X-band.

	Antenna angle	Coefficient of determination, R^2	Standard error, $\sqrt{\text{MSE}}$
C-band	0	0.480	2.35
	15	0.608	2.40
	30	0.697	1.87
	45	0.729	1.67
X-band	0	0.5085	2.50
	15	0.798	1.88
	30	0.905	1.03

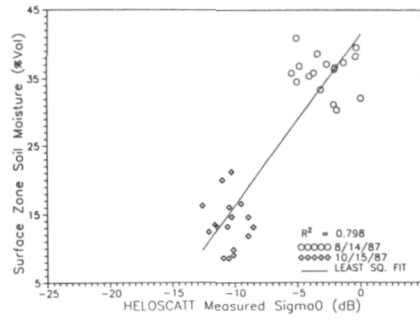


Figure 4. σ^0 versus measured soil moisture at X-band, 15° .

regression in figure 2 shows R^2 of 0.61 between σ^0 and soil moisture. This result is lower than that reported in earlier investigations by Ulaby *et al.* (1978) and Martin *et al.* (1989). The reason for the lower correlation observed during this experiment is that the scattering coefficient is strongly modulated by the terrain slope at antenna angles near vertical. Although we made corrections for the slope effect, we were not able to remove it completely. As an example, R^2 increased from 0.49 to 0.61 for C-band data at 15° incidence after slope correction.

The analysis for C-band data at 30° is shown in figure 3 with an R^2 for the correlation between backscatter and soil moisture of 0.70. Martin *et al.* (1989) also reported similar results from experiments conducted on the Konza Prairie using truck-mounted scatterometers. They reported that the correlation coefficient between backscatter and soil moisture decreased with increased antenna incidence angle. However, we observed an increased correlation with increasing antenna incidence angle. This effect is due to the negative exponential relation of σ° as a function of increased antenna incidence angle. Therefore the terrain slope modulation is smaller for the antenna positioned at large angles of incidence in comparison with that at angles near vertical.

The results of the analysis at X-band for antenna angles of 15° and 30° are shown in figures 4 and 5. The correlation between backscatter and soil moisture (R^2) is 0.80 for the 15° data and 0.90 for the 30° data. The increased correlation compared to that at C-band is partially due to using only two sets of data taken on days having a large difference of ground moisture. This is noticeable in figures 4 and 5 as two distinct

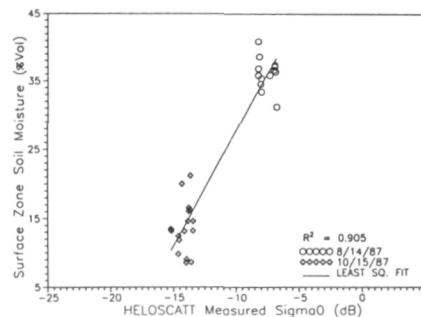


Figure 5. σ^0 versus measured soil moisture at X-band, 30° .

clumps of data. However, the data still indicate that radars operating at X-band frequencies can provide useful information on soil moisture over natural vegetation-covered soils. In a previous experiment at the Konza Prairie on unburned ground under similar conditions, a scatterometer with fine range resolution experimentally determined that roughly 30 per cent of the returned power was due to reflections from the soil and the remainder from the vegetation (Zoughi 1988).

The results of two soil moisture estimates along the flight transects are shown in figures 6 and 7. In the figures, the thinner soil line is plotted as the soil moisture estimate derived from σ^0 . The thicker line is the soil moisture estimate after corrections for slope modulation. The discrete circles in the figures are the *in situ* soil moisture measurements.

5. Discussion and conclusions

The radar measurement system and experimental site area have been presented. The description of the steps involved in processing the data obtained from a moving platform radar have been described. The moving platform enables more data to be obtained quickly, but also has a greater possibility of alignment errors with the soil samples. We believe our alignment is within 25 m or 1 moisture measurement distance. The correlation between measured moisture and σ^0 has also been shown to hold for both C- and X-bands over a range of incidence angles.

The conclusions from this experiment are:

1. C-band and X-band radars can be used to estimate soil moisture over grasslands. However, it is necessary to correct for terrain slope to obtain

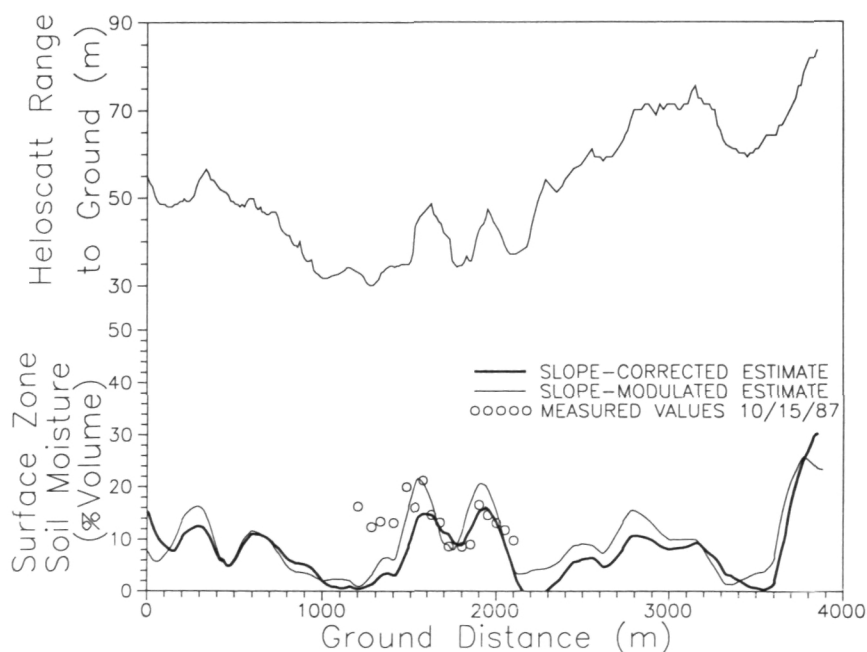


Figure 6. Measured soil moisture and estimated soil moisture along flight transect at C-band, 15°.

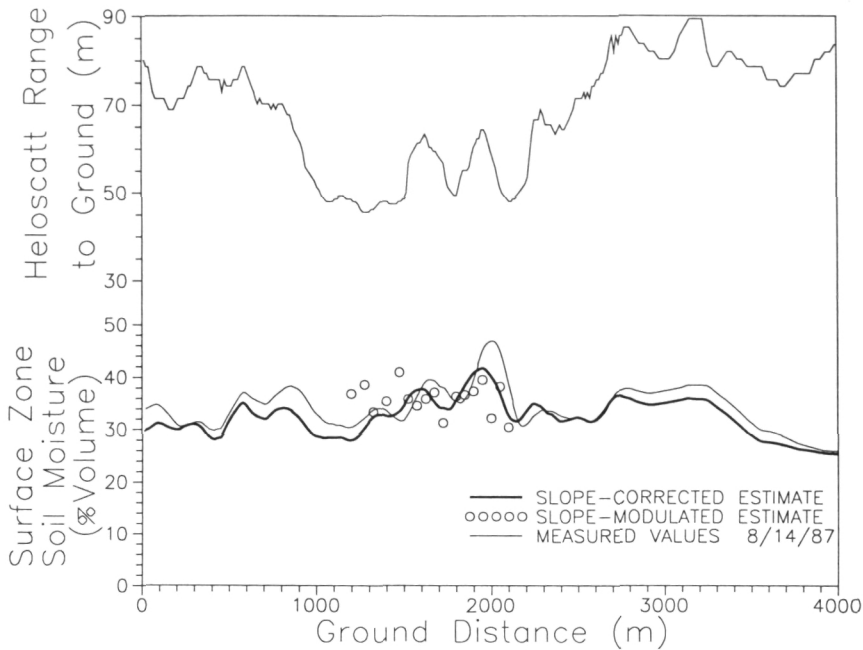


Figure 7. Measured soil moisture and estimated soil moisture along flight transect at C-band, 30°.

accurate estimates of soil moisture over areas with reasonable change in slope. A narrow antenna beam best determines the instantaneous ground slope.

2. Over grassland, C-band and X-band radars can be used for measurements over a wide range of incidence angles from approximately 15° to 45°. Other authors have shown that the power transmitted at larger antenna angles penetrate the vegetation less, but for the moving platform the larger angles see less change in terrain slope. The two effects may balance each other out.

Acknowledgments

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Active and Passive Microwave Measurements of Soil Moisture in FIFE

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During the intensive field campaigns of the First International Satellite Land Surface Climatology Project (ISLSCP) Field Experiment (FIFE) in May–October of 1987, several nearly simultaneous measurements were made with low-altitude flights of the L-band radiometer and C- and X-band scatterometers over two transects in the Konza Prairie Natural Research Area, some 8 km south of Manhattan, Kansas. These measurements showed that although the scatterometers were sensitive to soil moisture variations in most regions under the flight path, the L-band radiometer lost most of its sensitivity in regions unburned for many years. The correlation coefficient derived from the regression between the radar backscattering coefficient and the soil moisture was found to improve with the increase in antenna incidence angle. This is attributed to a steeper falloff of the backscattering coefficient as a function of local incidence at angles near nadir than at angles greater than 30°. Applying an initial slope correction to the measured data generally improves the correlation coefficient. At C band the improvement factor is found to be about 0.08 near vertical and gradually decreases to 0.04 at 45° incidence angle.

1. INTRODUCTION

Active and passive microwave signatures of land surface parameters, such as soil moisture, vegetation cover, and roughness, have been studied experimentally by a number of investigators in the past two decades [Schmugge *et al.*, 1974; Njoku and Kong, 1977; Wang *et al.*, 1980; Newton and Rouse, 1980; Ulaby *et al.*, 1974, 1978, 1979, 1982; Jackson *et al.*, 1981; Dobson and Ulaby, 1986]. A good understanding of the relationship between the microwave scattering coefficient or emissivity and the land parameters has been established as a result of these studies. Most of these research efforts, however, were performed with either active or passive sensor systems alone. On only a few occasions have measurements been conducted on common targets with both active and passive microwave sensors. For example, Wang *et al.* [1987] reported nearly simultaneous measurements over agricultural areas in Fresno, California, with Shuttle Imaging Radar B and an airborne microwave radiometer, both at L-band wavelengths. Brunfeldt and Ulaby [1984] measured agricultural targets with truck-mounted radar and radiometer systems at frequencies of 2.7 and 5.1 GHz. These authors have shown that a combined measurement, with both active and passive sensors, might improve the precision in the estimation of surface soil moisture.

In this paper we report results of nearly simultaneous measurements with a C-band scatterometer aboard the NASA helicopter and the L-band push broom microwave radiometer (PBMR) aboard the NASA C-130 aircraft. The measurements were made over a region of the Konza Prairie Natural Research Area (KPNRA) during three of the four intensive field campaigns (IFCs) of First International Satellite Land Surface Climatology Project (ISLSCP) Field Experiment (FIFE) in 1987 [Hall *et al.*, 1989]. The topogra-

phy and surface cover of KPNRA are markedly different from the agricultural targets reported previously. The region of measurements included areas with different surface treatments and thus provided a good opportunity for studying the effect of yet another type of vegetation cover on microwave emission and backscatter.

2. MEASUREMENTS

The overall experimental objectives and the test site description of the FIFE program have been given by Hall *et al.* [1989]. The entire area covered by the PBMR measurements has also been described in detail by Wang *et al.* [1990]. Therefore only a brief description of the common region that was shared by both PBMR and C-band scatterometer measurements is given below. Figure 1 shows a topographical sketch of this region located at approximately 39°04'N latitude and at 96°33'W longitude, some 8 km south of Manhattan, Kansas. The light solid curves in the figure indicate the graveled trails that provide access to various regions of the KPNRA. The flight lines for the C-band scatterometer and PBMR are given by the solid and dashed lines, respectively. The radiometer flights were made from east to west, and the scatterometer flights were made from both east to west and west to east. The 1D watershed was burned every year, whereas the 2D watershed was burned only every other year (it was unburned in 1987). The areas west of 1D and east of 2D have not been burned within 5 years. In association with the airborne microwave measurements, extensive soil moisture sampling in the top 5 cm was made in both the 1D and the 2D watersheds, at locations separated by about 50 m, along the three transects under the dashed flight lines [Wang *et al.*, 1989].

A multifrequency FM radar called HELOSCAT was used in measurements over the Konza prairie. This system is capable of collecting backscatter data at selected frequencies between 4 and 17 GHz over incidence angles from 0° to 70° with like and cross polarizations. The helicopter was flown

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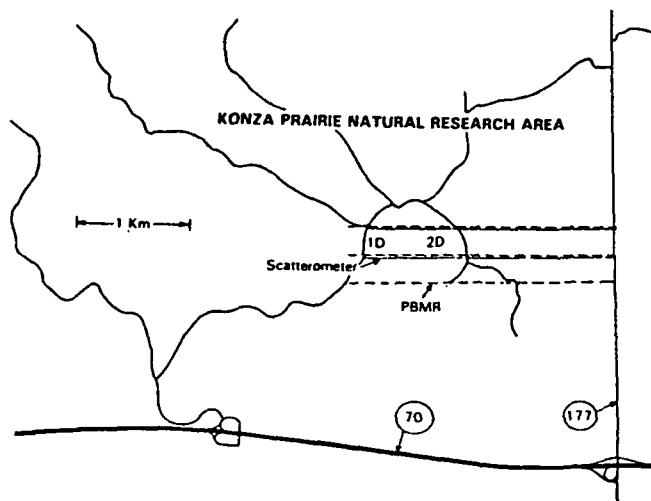


Fig. 1. Topographical sketch showing the flight lines for the L band push broom microwave radiometer (PBMR) and C and X band scatterometers over 1D and 2D watersheds.

at an altitude of approximately 300 m for measurements at incidence angles less than 30° and 15 m for measurements at angles greater than 30° . The important specifications of the system are given in Table 1. A more detailed description of the system is given by Gogineni *et al.* [1984]. The system was calibrated both internally and externally for estimating scattering coefficients from measured power returns. A short-circuited transmission line of known length and loss was switched in periodically in place of the antenna for internal calibration, which was performed before and after each data run over a selected flight line. The external calibration was performed by measuring the signal backscattered by a Luneburg lens of known radar cross section and was done before installation on and after removal from the helicopter.

The PBMR has previously been used for remote sensing of surface soil moisture on a few occasions [Jackson *et al.*, 1987; Wang *et al.*, 1989, 1990]. The details of the instrument description can be found elsewhere [Harrington and Lawrence, 1985]. It has four beams pointing cross track at $\pm 8^\circ$ and $\pm 24^\circ$ from nadir, each having a beam width of 16° . During the four IFCs of FIFE 1987 it was flown at both 300 and 600 m altitudes. The three flight lines in Figure 1 were

TABLE 1. Important System Specifications for HELOSCAT During FIFE 1987

Parameter	Specification
Radar type	FM continuous wave
Radar frequency	4–17 GHz
Polarization	VV, VH, HV, HH*
Intermediate frequency	
center frequency	50 kHz
bandwidth	13.5 kHz
Antenna	
beam width	5° at 5.3 GHz, 3° at 9.6 GHz
Calibration	
internal	delay line
external	Luneburg lens

*VV, vertical transmit and vertical receive; HV, horizontal transmit and vertical receive.

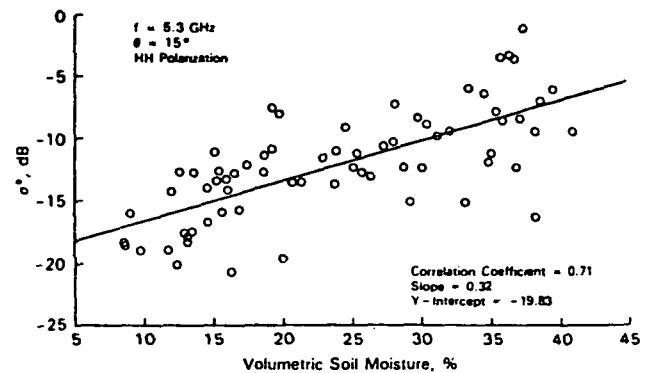


Fig. 2. Scatter plot of the measured σ^0 at 5.3 GHz versus soil moisture in percent volume. The measurements were made at 15° incidence angle and HH polarization.

flown at an altitude of 300 m, which resulted in a composite swath width of about 1 km, sufficient to cover both 1D and 2D watersheds. Five days of simultaneous PBMR and scatterometer measurements were made on July 2 and 7 of IFC 2, August 14 and 21 of IFC 3, and October 15 of IFC 4. The results from these measurements are the focus of the following discussion, although some other data sources are also used at times for comparison.

3. RESULTS OF SCATTEROMETER MEASUREMENTS

The measured data were converted into scattering coefficients (σ^0) and instantaneous range using internal and external calibrations. The procedure is similar to that discussed by Onstott *et al.* [1981]. The converted scattering coefficients were smoothed to reduce the effect of fading. The smoothing was accomplished by using triangular weighting functions. Linear regression analysis was performed on the scattering and in situ soil moisture data to establish a relationship between σ^0 and soil moisture.

The results of the analysis at the C band for incidence angles of 15° and 45° are shown in Figures 2 and 3, respectively. The results at X band for 15° incidence are shown in Figure 4. The data from the analysis are summarized in Table 2. The correlation coefficients between soil moisture (W_v) and σ^0 at C band are 0.71 and 0.82 at 15° and 45° incidence angles, respectively. At angles near nadir, Ulaby *et al.* [1978] reported a correlation coefficient of about 0.87 for bare soil, and Martin *et al.* [1989] found correlation

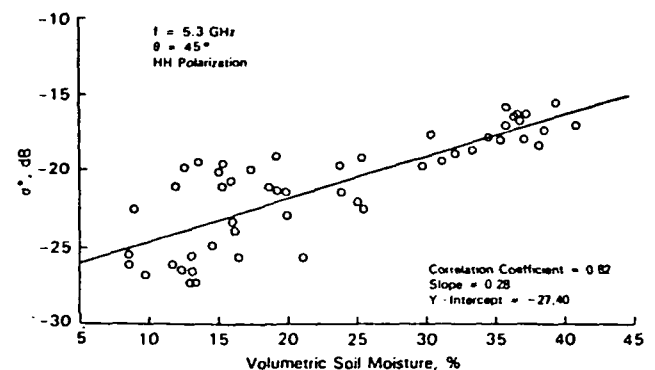


Fig. 3. Scatter plot of the measured σ^0 at 5.3 GHz versus soil moisture in percent volume. The measurements were made at 45° incidence angle and HH polarization.

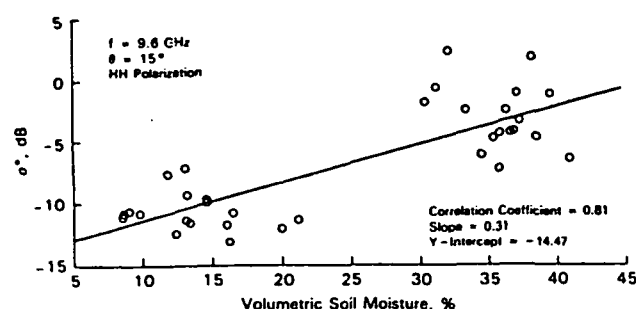


Fig. 4. Scatter plot of the measured σ^0 at 9.6 GHz versus soil moisture in percent volume. The measurements were made at 15° incidence angle and HH polarization.

coefficients of about 0.9 for burned areas and 0.81 for unburned areas. There are two reasons for the lower values of the correlation coefficient derived from the data acquired in FIFE. First, the measurements from both unburned and burned areas were used to generate the regression lines in Figures 2 and 3. On the other hand, Martin *et al.* [1989] separated the measurements from burned and unburned areas and determined separate relations for the two areas. Second, the scatterometer data were not corrected for the local slope effect in these analyses. The terrain had a maximum grade of 30%, calculated over 25-m $\times$ 25-m grids. When initial corrections for the slope effect were applied to the scatterometer data, improvements in the correlation coefficients were found at both frequencies, as shown in Table 2.

The procedure for the slope correction consists of three steps. First, from the scatterometer transect data the average scattering coefficient as a function of incidence angle is computed independently for each of three regions of different surface treatments. Second, for each observation point along the transect the instantaneous incidence angle, θ_i , is computed from the digital terrain map and radar observation angle θ_0 . Finally, the measured scattering coefficient for the point in question is replaced with a new value at θ_i determined from the relationship established in the first step. At C band the improvements were approximately 0.07 and 0.02 at incidence angles of 15° and 45°, respectively.

The higher correlation observed at incidence angles larger than 15° can be attributed to smaller terrain slope modulation at these angles. Martin *et al.* [1989] also observed high correlation between σ^0 and W_s at large incidence angles. Backscattered signal from vegetation-covered ground consists mainly of three terms; namely, the surface, the volume, and a surface-volume component resulting from the interaction between soil surface and vegetation volume [Fung and

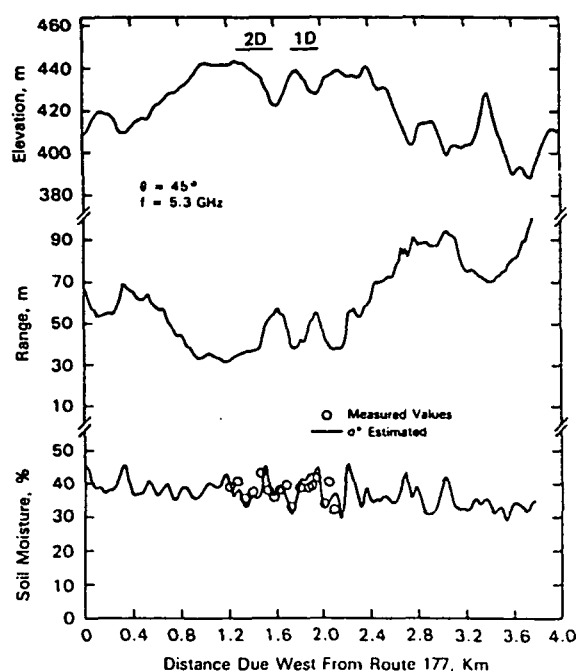


Fig. 5. Comparison of the estimated and measured soil moisture values. The estimated values are derived from the regression results in Figure 3. The elevation and range measured by the 5.3-GHz scatterometer are also shown.

Eom, 1985]. At angles near nadir the scattering is dominated by the soil surface. This component drops off rapidly with increasing incidence angle, giving way for the volume and surface-volume terms to become comparable to the surface term or dominant at large incidence angles. Fung and Eom [1985] reported that the surface-volume interaction term can become comparable to the volume term at large incidence angles for horizontal polarization. Therefore at large incidence angles an active sensor responds to changes in soil moisture partly due to the presence of the surface-volume term.

At the X band, the correlation coefficient between σ^0 at 15° incidence and W_s is 0.81. Although attenuation through the vegetation canopy generally increases with frequency, there is still a significant penetration at the X band. Backscatter measurements made by a fine resolution radar [Zoughi, 1987] indicated that roughly 30% of the return power over previously burned prairie was due to soil reflections. The rate of σ^0 decrease with increase of incidence angle is smaller at the X band than at the C band. Therefore the terrain slope modulation effect is reduced at the X band. The high correlation observed at the X band could be the results of reduced slope modulation at X band and the grouping of data into sets. We did not have the X-band measurements at intermediate values of soil moisture. More measurements are needed to substantiate the relation between σ^0 and W_s at this frequency.

The equations generated from these regression analyses were used to compute W_s values along the transect. Two examples of the results are shown in Figures 5 and 6 for the C band and the X band, respectively. In areas with a small change of slopes, both Figures 5 and 6 show good agreement between the values from in situ measurement and from radar estimation. In areas with a large change in slopes the

TABLE 2. Correlation Table for HELOSCAT During FIFE 1987

Antenna Angle, deg	C Band		X Band	
	Original	Slope Correction	Original	Slope Correction
0	0.6058	0.6931	0.6888	0.7131
15	0.7143	0.7799	0.8054	0.8933
30	0.7637	0.8349	0.9568	0.9514
45	0.8164	0.8538		

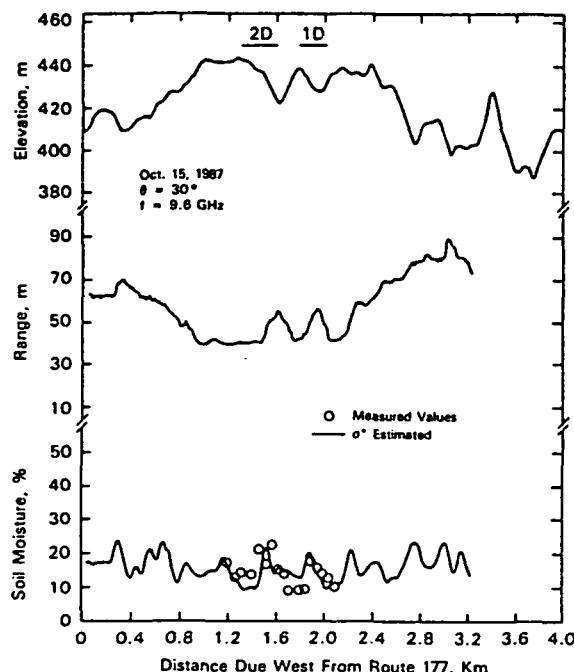


Fig. 6. Comparison of the estimated and measured soil moisture values. The estimated values are derived from the regression between soil moisture and σ^0 measured at 9.6 GHz and 30° incidence angle. The elevation and range measured by the scatterometer are also shown.

radar-estimated values can be larger or smaller than those from the in situ measurements. Lower backscatter is received for a negative slope in the direction of the flight, and higher backscatter is received for a positive slope. This results in the modulation of σ^0 by the terrain slope. For example, in Figure 5 the depressions in the terrain contour produce a noticeable increase in σ^0 . An increase in σ^0 is expected because the gullies retain higher moisture content resulting from rains. The σ^0 peak is shifted away from the center bottom of the gully, however. The direction of the peak shift is in the direction of the flight.

4. RESULTS OF PBMR MEASUREMENTS

Figures 7 and 8 show the variations of L-band brightness temperature, the surface temperature as measured by a PRT-5 sensor aboard the same aircraft, and W_s values. The distance in kilometers is measured from highway I-177, indicated in Figure 1. All of the measurements in Figures 7 and 8 are made along the middle transect. Figure 7 gives the results for 2 days of measurements in IFC 2, whereas Figure 8 shows the same for 2 days in IFC 3 and 1 day in IFC 4. A fair amount of rainfall was reported in the test area during both IFC 2 and IFC 3. No rainfall was recorded after September 15 and throughout IFC 4. Thus these 5 days of measurements covered a wide range of soil moisture variations.

Several features are noticed from an examination of Figures 7 and 8. First, the measured brightness temperature (T_b) in watershed 1D responds to change in W_s in a normal fashion, that is, a high W_s corresponds to a low T_b . This feature also holds for most of watershed 2D in Figure 8. Furthermore, the sensitivity of T_b to W_s variation is higher at 1D than at 2D, showing the effect of surface treatment.

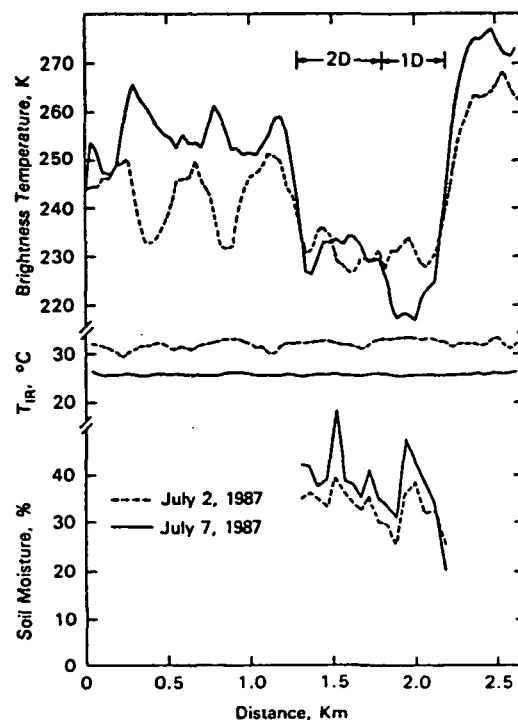


Fig. 7. Variations with distance of the measured L-band brightness temperature, thermal IR temperature, and soil moisture for 2 different days in July 1987.

Second, over the regions that were unburned for many years, the T_b values appear to be positively correlated with W_s . This effect has been qualitatively pointed out in a previous study to be caused by a thatch layer that was built

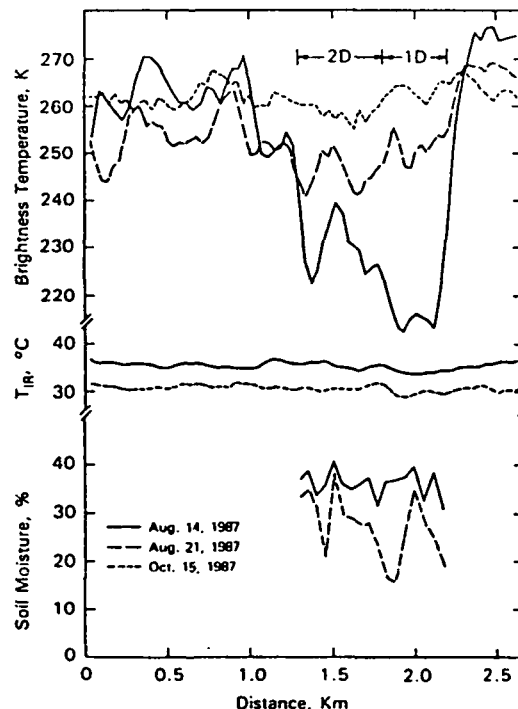


Fig. 8. Variations with distance of the measured brightness temperature, thermal IR temperature, and soil moisture for 3 different days in August–October 1987.

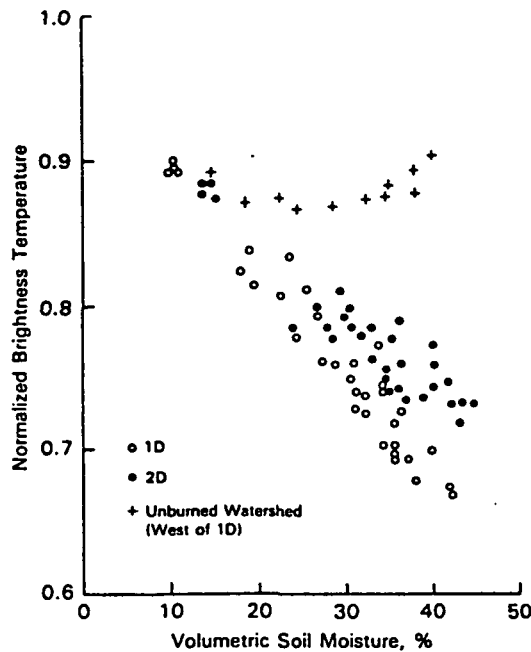


Fig. 9. Normalized brightness temperature plotted as a function of volumetric soil moisture for 1D and 2D watersheds and a region unburned for many years.

up by many years of dead vegetation [Schmugge *et al.*, 1988]. A satisfactory interpretation is not given, however, owing to the lack of data in the dry regime. During the four IFCs of FIFE 1987 the middle transect shown in Figure 1 was extended about 0.7 km into the unburned region west of watershed 1D. Soil moisture samples in this extended region were also acquired in some of the PBMR flights. These additional measurements over the unburned region provide a wide range of soil moisture variation enabling us to examine in detail the effect of a thatch layer on microwave emission.

Figure 9 shows the variation of normalized brightness temperature (T_n) with W_s for watersheds 1D and 2D, as well as part of the unburned region with soil moisture sampling. Data for 1D and 2D watersheds were derived from all three transects in Figure 1 and were analyzed previously [Wang *et al.*, 1990]. Each data point corresponds to a pair of average T_n and W_s values on a given day along the segment of a transect in each watershed. The correlation between T_n and W_s for 1D and 2D watersheds is markedly different. The regression slope for the 1D watershed is clearly steeper than that for the 2D watershed, effectively showing the difference between surface burning every year and every other year. For the region unburned for many years the variation of T_n with W_s is far from linear. The measured T_n values for this region are high when $W_s < 25\%$ and $W_s > 35\%$.

To qualitatively explore the characteristics of absorption due to three different surface treatments, we examine the equation giving the emissivity ϵ from the vegetation covered soils [Mo *et al.*, 1982; Wang *et al.*, 1987],

$$\epsilon = \epsilon_s L + (1 - \omega)(1 - L) + (1 - \epsilon_s)(1 - \omega)(1 - L)L \quad (1)$$

with

$$L = \exp(-\tau/\cos \theta) \quad (2)$$

where ϵ_s is the emissivity of the underlying soil surface, τ is the optical depth, and ω is the single scattering albedo, which

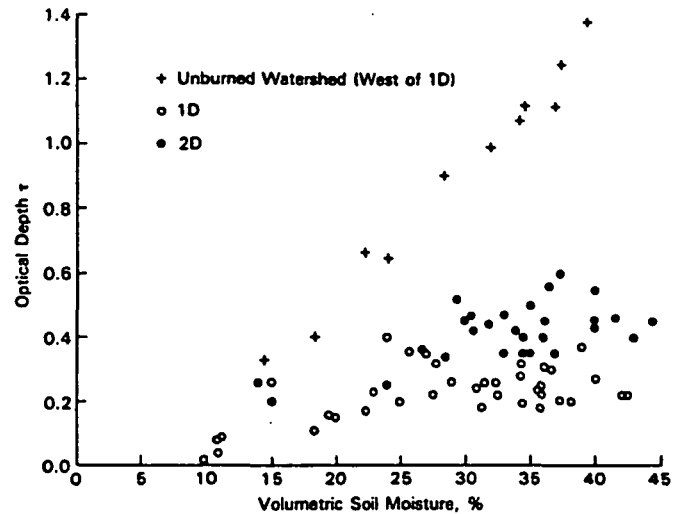


Fig. 10. Estimated variation of optical depth with top 5-cm soil moisture for three vegetation-covered regions.

can be assumed to be ≈ 0.04 for incidence angle $\theta < 50^\circ$ [Mo *et al.*, 1982]. Knowing W_s , ϵ_s for a relatively smooth surface can be obtained from the radiative transfer calculations [Wilheit, 1978]. These two equations could be used with the data in Figure 9 to estimate τ . The results are displayed in Figure 10 for the three regions of different surface treatments. Clearly, τ increases with W_s almost linearly for all three vegetation-covered surfaces. The moderate increase of τ with W_s for watersheds 1D and 2D results in the general T_n decrease with increase in W_s over the entire observed range of W_s shown in Figure 9. The steep increase of τ with W_s for the unburned region west of 1D, which is presumably caused by the thatch layer, gives rise to a more complicated variation of T_n with W_s .

5. COMPARISON OF ACTIVE AND PASSIVE MEASUREMENTS

A comparison of results from active and passive measurements for flight line 14 is shown in Figure 11. As can be seen from Figure 11, both types of measurements have good sensitivity to soil moisture over burned areas. The radiometer is not responding well to soil moisture over areas unburned for several years. The range of radar returns over these unburned areas is slightly lower than that from burned areas, but the σ^0 is still strongly modulated by the soil moisture. Even over the watersheds 1D and 2D the microwave responses to soil moisture are different between a radiometer and a radar, as shown in Figure 12, where the L-band emissivity is plotted against σ from the C-band scatterometer at both 15° (Figure 12a) and 45° (Figure 12b) incidence angles. For both plots the differences in the slopes between 1D and 2D watersheds are mainly caused by the different relationship between emissivity and vegetation cover for these two watersheds [Wang *et al.*, 1989].

The presence of vegetation cover reduces the sensitivity to soil moisture variation, to different extents, for both active and passive microwave sensors. For a passive sensor the loss of sensitivity is much larger than with an active system. The emission from vegetation-covered soils consists of three terms: radiation from soils, upwelling radiation from the vegetation medium, and downwelling radiation from the vegetation medium reflected from the soil surface and fur-

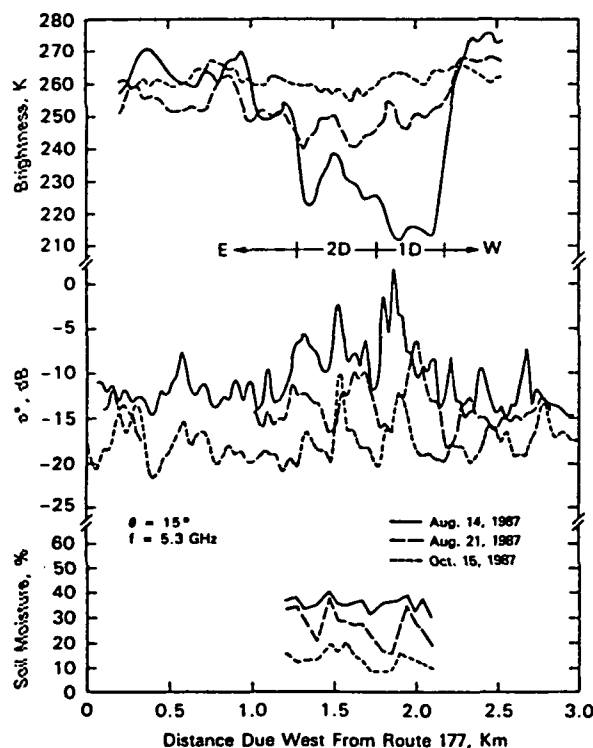


Fig. 11. Variations of brightness temperature, backscattering coefficient, and soil moisture measured along the middle transect shown in Figure 1 during 3 days in August–October 1987.

ther attenuated by the vegetation medium. Over the unburned areas where the thatch layer is thick and dense, the emission from the vegetation medium could become much larger than the other two terms. As a result, the microwave emission from the underlying soils is effectively masked and the sensitivity to soil moisture variation of a passive sensor is greatly reduced [Schmugge *et al.*, 1988; Wang *et al.*, 1990]. On the other hand, the backscattered signal from vegetation-covered terrain consists of three components, as described in section 3: surface scatter from terrain, volume scatter from vegetation, and surface-volume scatter from the interactions between soil surface and vegetation. The latter two are the dominant sources of scattering at large incidence angles. For horizontal polarization at large incidence angles the surface-volume term could become comparable to the volume term [Fung and Eom, 1985]. This could be attributed to the radar's sensitivity to soil moisture over unburned areas.

Over moderately vegetated areas a scatter plot between σ° and T_n with W_s as a parameter, like Figure 12, would help provide a more precise estimation of W_s than the measurements of either σ° or T_n alone. The figure may not illustrate this potential convincingly due to the scarcity of the data points. If more concurrent radar and radiometer measurements are available, one could establish contour curves of soil moisture on these plots. A pair of radar and radiometric measurements could in principle provide the estimation of soil moisture and the amount of vegetation cover at the same time.

6. CONCLUSIONS

Simultaneous measurements along two transects in the KPNRA by scattermeters at X and C bands and by an

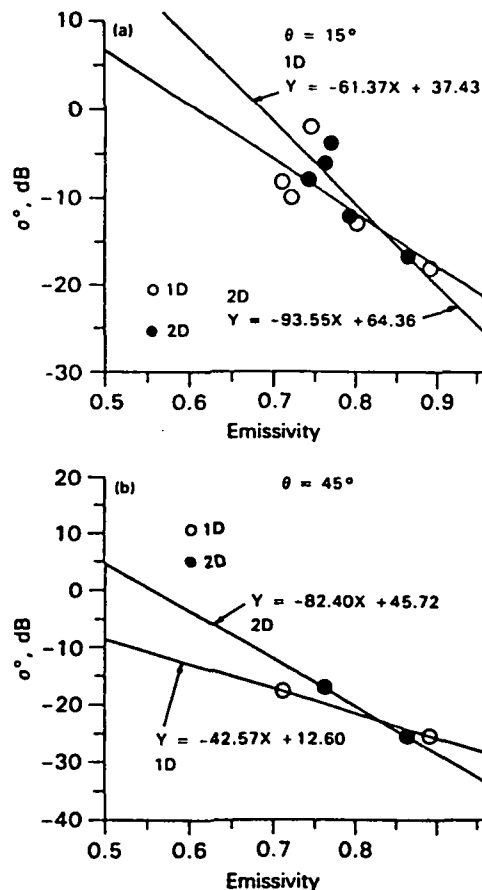


Fig. 12. Scatter plots of C band backscattering coefficient and L band emissivity for the incidence angle of the scatterometer at (a) 15° and (b) 45°.

L-band radiometer were conducted during three intensive field campaigns of FIFE in 1987. Concurrent soil moisture sampling in the top 5-cm layer was made along the transects to support the microwave measurements. Results of these measurements are summarized below.

1. Scattermeters at both the C and the X bands are sensitive to moisture variation in soils covered with dense and tall grass. Sensitivity derived from the regression of the backscattering coefficient with soil moisture is comparable to that derived from the truck-mounted scatterometer measurements reported in the literature.

2. The scattermeters show varying responses to different surface treatments. At incidence angles near nadir the average backscattering coefficient for the previously burned areas is approximately 8 dB higher than that for the unburned. However, at 30° incidence the difference of average backscattering coefficients is hardly noticeable, only 2 dB.

3. The correlation of the backscattering coefficient with soil moisture improves with the increase in incidence angle. This is due to a steeper falloff of backscattering coefficient as a function of local incidence at angles near nadir than at angles greater than 30°. Application of an initial slope correction generally results in an improvement of the correlation coefficients.

4. For areas unburned for many years the L-band emissivity first decreases and then increases with the increase in soil moisture. Adopting an emission model for vegetated fields used by Mo *et al.* [1982], the observation could be

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explained by a rapid increase with soil moisture of the optical depth of the vegetation layer (including the thatch layer).

5. A comparison of the variations in the backscattering coefficient and emissivity measured over areas burned every year and every other year shows that in probing the underlying soils for moisture change, a radiometer is more susceptible to vegetation cover than a scatterometer. There is a potential for improving the precision of soil moisture estimation by simultaneous measurements with a radiometer and a scatterometer.

Clearly, more work is needed to obtain a better understanding of the effect of a thatch layer on microwave backscatter and emission. This includes both theoretical modeling and experimental observations. In addition, a thorough analysis of the relationship between terrestrial slope variation and microwave backscatter is required to satisfactorily interpret the observed improvement with incidence angle of the correlation between backscattering coefficient and soil moisture.

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COMPARISON OF MEASUREMENTS AND THEORY FOR BACKSCATTER FROM VEGETATION-COVERED SOIL ON THE KONZA PRAIRIE

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Abstract: We made radar backscatter measurements over the Konza Prairie as a part of the First ISLSCP Field Experiment (FIFE) to determine soil moisture. We made these measurements using C- and X-band scatterometers in a helicopter. Nearly simultaneous radar and radiometer data sets were collected along two transects that coincided with direct soil-moisture measurements. The results from this experiment showed that radars can be used for soil-moisture estimation over the complete transect, whereas radiometer sensitivity to soil moisture was drastically reduced over regions left unburned for many years. A combined rough-surface/volume-scatter model was formulated. Here we compare calculated and measured scattering data to determine the sensitivity of the scattering coefficient to different surface treatments.

Introduction

Use of active and passive microwave systems for measuring soil-moisture and vegetation parameters has been a topic of considerable research. Many measurements have been made of backscattering and emissivity of bare and vegetation-covered soils [1-9]. The primary objective of most of these studies was to determine an optimum system for measuring soil moisture. These investigations showed that L-band radiometers operating at angles near vertical, and radars operating at about 4 GHz with incidence angles from 7° to 17°, were effective in estimating soil moisture. However, most earlier studies were conducted on agricultural fields, with the exception of those by Schmugge et al. [10] and Martin et al. [11].

We made microwave measurements on prairie vegetation-covered soils as a part of The First International satellite land surface climatology project Field Experiments (FIFE). FIFE was at the Konza Prairie Natural Research Area (KPNRA) south of Manhattan, Kansas. We measured during three of the four Intensive Field Campaigns (IFCs) in 1987. The goal of FIFE was to collect data needed to interpret satellite observations of climatologically significant land-surface parameters such as leaf area, biomass and soil moisture. The study area consisted mainly of native tall-grass prairie vegetation with different surface treatments. Some areas of the study were burned and other were unburned in various annual rotations [12].

To understand better the effect of various surface treatments on measured backscattered signals, we formulated a model and compared the model predictions with experimental data. The purpose of this paper is to provide a brief overview of the model and to compare calculated and measured data.

Instrumentation

We used a helicopter-borne scatterometer called HELOSCAT. This system was designed for collecting data at selected frequencies between 4 and 17 GHz over incidence angles ranging from vertical to about 70° with all four linear polarizations. During FIFE we used this system at 5.3 and 9.6 GHz with HH polarization at incidence angles of 0°, 15°, 30°, 45° and 60°. We calibrated the system both internally and externally to aid in converting the measured returns to scattering coefficients. The internal calibration was done by

periodically connecting a short-circuited transmission line in place of the antenna. The external calibration used measurements of the signal backscattered by a Luneburg lens of known radar cross section. A more detailed description of the system is available in Gogineni et al. [13].

Experiment Description

The radar repeatedly overflew a selected transect to collect data at various incidence angles. The emphasis was at C band; however, whenever possible, we also measured at X band on the same lines. The helicopter flew at an altitude of 30 m for incidence angles less than 30° and at an altitude of 15 m for angles greater than 30°. Before and after each transect was overflown, the return from the delay line was recorded for internal calibration. The external calibration occurred before the system installation on and removal from the helicopter.

The Scattering Model

A vegetation-covered surface is modeled as a layer of Rayleigh scatterers above an irregular surface. Our modeling effort was not to recover properties of vegetation. Rather, it was to account for its presence as a volume-scattering layer. It is characterized by an albedo and an optical depth $\tau_d = k_e d$ where k_e is the extinction coefficient and d is the physical depth. For surface scattering we used the integral equation model (IEM) was developed by Fung, Li and Chen [14].

The total scattering consists of two major parts: (1) surface scattering attenuated by the vegetation and (2) volume scattering from the vegetation plus surface-volume interaction. Since volume scattering tends to be isotropic, we used a plane boundary approximation in calculating the volume scattering, while for the surface scattering we used the IEM model. Thus, the backscattering model is given by

$$\sigma_{pp}^0 = \exp[-(2\tau_d)/(\cos \theta)] \sigma_{pp}^s + \sigma_{pp}^v \quad (1)$$

where

$$\sigma_{pp}^s = \frac{k_z^2}{2} \exp(-2k_z^2 \sigma^2) \sum_{n=1}^{\infty} \sigma^{2n} |I_{pp}^n|^2 \frac{W^n(-2k_z, 0)}{n!} \quad (2)$$

with

$$I_{pp}^n = (2k_z)^n f_{pp} \exp(-\sigma^2 k_z^2) + \frac{k_z^n [F_{pp}(-k_z, 0) + F_{pp}(k_z, 0)]}{2}$$

and

$$f_{vv} = (2R_d)/(\cos \theta) \quad (3)$$

$$f_{hh} = (2R_s)/(\cos \theta) \quad (4)$$

$$F_{vv}(-k_z, 0) + F_{vv}(k_z, 0) =$$

$$\frac{2 \sin^2 \theta (1 + R_0)^2}{\cos \theta} \left[\left(1 - \frac{1}{\epsilon_r} \right) + \frac{\mu_r \epsilon_r - \sin^2 \theta - \epsilon_r \cos^2 \theta}{\epsilon_r^2 \cos^2 \theta} \right] \quad (5)$$

$$F_{HH}(-k_x, 0) =$$

$$-\frac{2\sin^2 \theta (1 + R_{\perp})^2}{\cos \theta} \left[\left(1 - \frac{1}{\mu_r}\right) + \frac{\mu_r \epsilon_r - \sin^2 \theta - \mu_r \cos^2 \theta}{\mu_r^2 \cos^2 \theta} \right] \quad (6)$$

In the above $R_{\perp}$, $R_{\parallel}$ are the Fresnel reflection coefficients; ϵ_r , μ_r are the relative permittivity and permeability of the ground; $k_z = k \cos \theta$; $k_x = k \sin \theta$; σ^2 is the variance of the surface heights; k is the wave number, and W^n is the Fourier transform of the n th power of the surface correlation function. The subscript p denotes the polarization state and may be either vertical or horizontal.

The volume-scattering term plus the volume-surface interaction part is

$$\sigma_{pp}^v = 4\pi k_s \left(\frac{(1 - \alpha) \cos \theta}{2k_e} \times \left[P_{pp}(\mu, -\mu) + \alpha R_p^2 P_{pp}(-\mu, \mu) + \alpha d R_p [P_{pp}(\mu, \mu) + P_{pp}(-\mu, -\mu)] \right] \right) \quad (7)$$

where $\alpha = \exp[-(2\tau_d)/(\cos \theta)]$, R_p is the Fresnel reflection coefficient, $\mu = \cos \theta$; P_{pp} is the phase function with p polarization, and k_s is the volume-scattering coefficient.

Comparison with Scattering Model

Figure 1 shows sample results from microwave measurements made over one of the transects that included various types of treatments. The area marked 1D is burned every year, the area marked 2D is burned every other year, and the areas east and west of 2D were left unburned for many years. The results showed that the C-band radar was responding to soil moisture over most areas, whereas the radiometer system lost its sensitivity in regions left long unburned as shown in Figure 1.

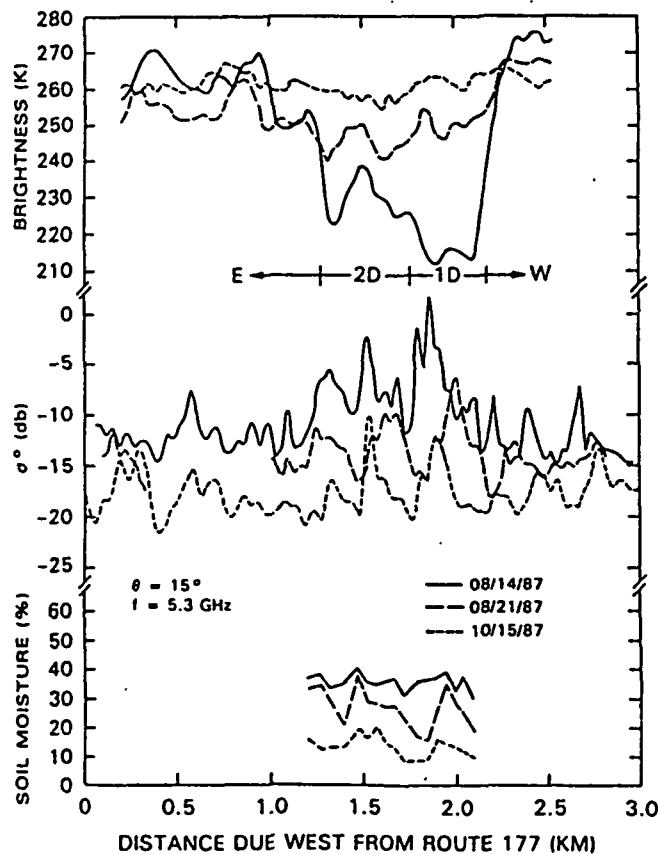


Figure 1. Brightness temperature, scattering coefficient and soil moisture profiles along one of the transects during 1987.

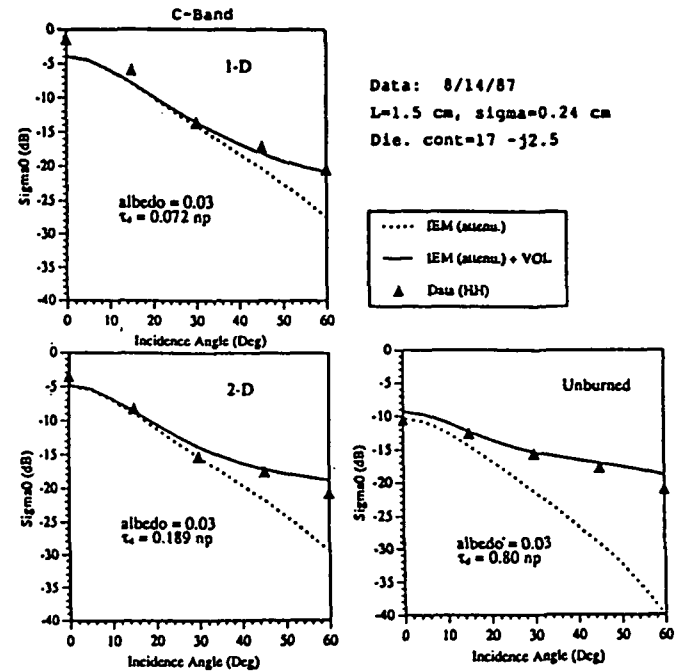


Figure 2. Comparison of measured and calculated data at C band for August 14, 1987.

Figure 2 compares data from August 14, 1987, with model predictions at 5.3 GHz. Backscatter data for each transect were segmented according to surface treatment, and we found average scattering coefficients for each case. Values of σ^0 at C band (5.3 GHz) from burned and unburned grass-covered surfaces and X band are compared with the theoretical model to indicate the extent of surface and volume contributions by plotting (1) and its first term alone. Solid curves are for equation (1), dashed curves are for its first term, and triangles are measured values. The difference between the solid and dashed curves shows the effect of volume and the volume-surface interaction effects. The August burned grass was of two types: burned one year ago and burned two years previously. We assume the surface properties on a given date are the same for both burned and unburned areas, as is the scattering albedo. The change in height of the grass corresponds to a change in optical depth. Both albedo and optical depth change when the frequency changes.

Table 1: Fractional Contribution from Soil Surface at C Band

	Angle (°)	% Surface Contribution*
Burned Area (1D)	0	100
	15	100
	30	89.1
	45	63.1
	60	25.1
Unburned Area (2D)	0	100
	15	94.4
	30	66.8
	45	35.5
	60	8.4
Area Unburned For Many Years	0	80.4
	15	63.1
	30	26.6
	45	8.0
	60	0.7

$$* \sigma_s^0 / \sigma_r^0$$

Table 1 summarizes the results. Over burned areas the surface dominates σ^0 at incidence angles less than 45°. Even at 60° incidence,

25% of the backscattered signal comes from the soil surface on recently burned areas (1D). For long-unburned areas at incidence angles exceeding about 20°, volume scatter dominates. At angles near vertical (<20°) the thatch layer attenuates the surface-scatter component, but more than 60% of the signal is from the soil surface at C band. Thus the radar is sensitive to soil moisture over unburned areas. However, different inversion algorithms must be used to estimate soil moisture over burned and unburned areas because of this attenuation effect.

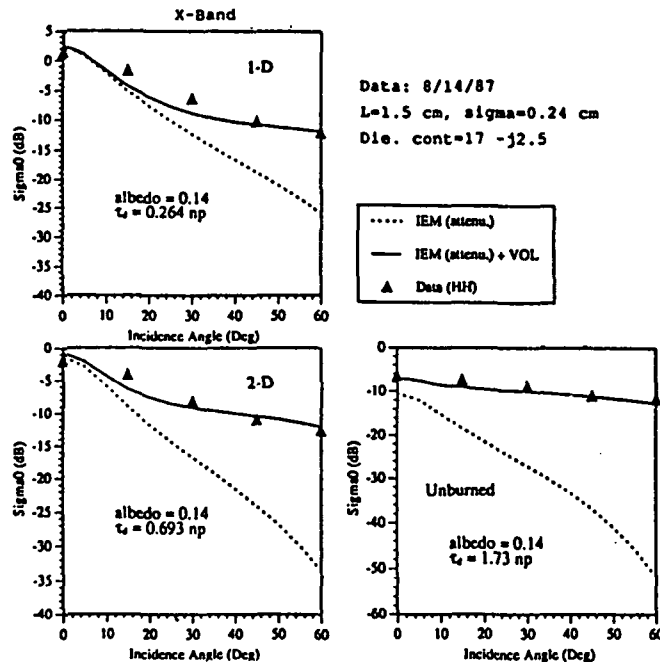


Figure 3. Comparison of measured and calculated data at X band for August 14, 1987.

Table 2: Fractional Contribution from Soil Surface at X Band

	Angle (°)	% Surface Contribution*
<u>Burned Area</u> (1D)	0	100
	15	79.4
	30	28.2
	45	16.8
	60	4.0
<u>Unburned Area</u> (2D)	0	89.0
	15	42.2
	30	17.8
	45	4.0
	60	0.6
<u>Area</u>	0	50.1
<u>Unburned</u>	15	12.6
<u>For Many</u>	30	2.0
<u>Years</u>	45	0.3
	60	0.01

* σ_s^0 / σ_r^0

Figure 3 shows the measured data and computations at X band (9.6 GHz). Table 2 summarizes the predicted results for the fractional contribution from the soil surface at different angles. Over burned areas (1D and 2D) there is a significant surface contribution near vertical. However, on areas unburned for many years the volume term dominates. At 30° incidence, the fractional contribution from the soil surface is only about 2%. Therefore, X-band radars may be

marginally useful in estimating soil moisture over grasslands, particularly those that are not burned.

Conclusions

These results show that both active and passive systems can be used to measure soil moisture over burned areas. The presence of a thatch layer greatly reduced the passive system's sensitivity to soil moisture over unburned regions. However, active systems operating in the lower part of C band seem to respond to soil moisture over the complete site. The presence of a thatch layer reduced the near-vertical scattering coefficient for unburned areas by about 4 dB relative to the burned areas. This implies that different radar inversion algorithms must be used to estimate soil moisture over grasslands with different surface treatments.

A combined rough-surface/volume-scatter model gave predictions that agreed fairly well with experimental results. The simulated results show that surface scatter is dominant over burned areas at incidence angles less than about 40° for C band. Over unburned regions, surface scatter dominates at C band for incidence angles less than about 25°, whereas volume scatter dominates for large angles. The results also suggest that X-band radars can be used at angles near vertical to measure soil moisture over burned areas, but a thatch layer greatly reduces the sensitivity of the X-band system to soil moisture.

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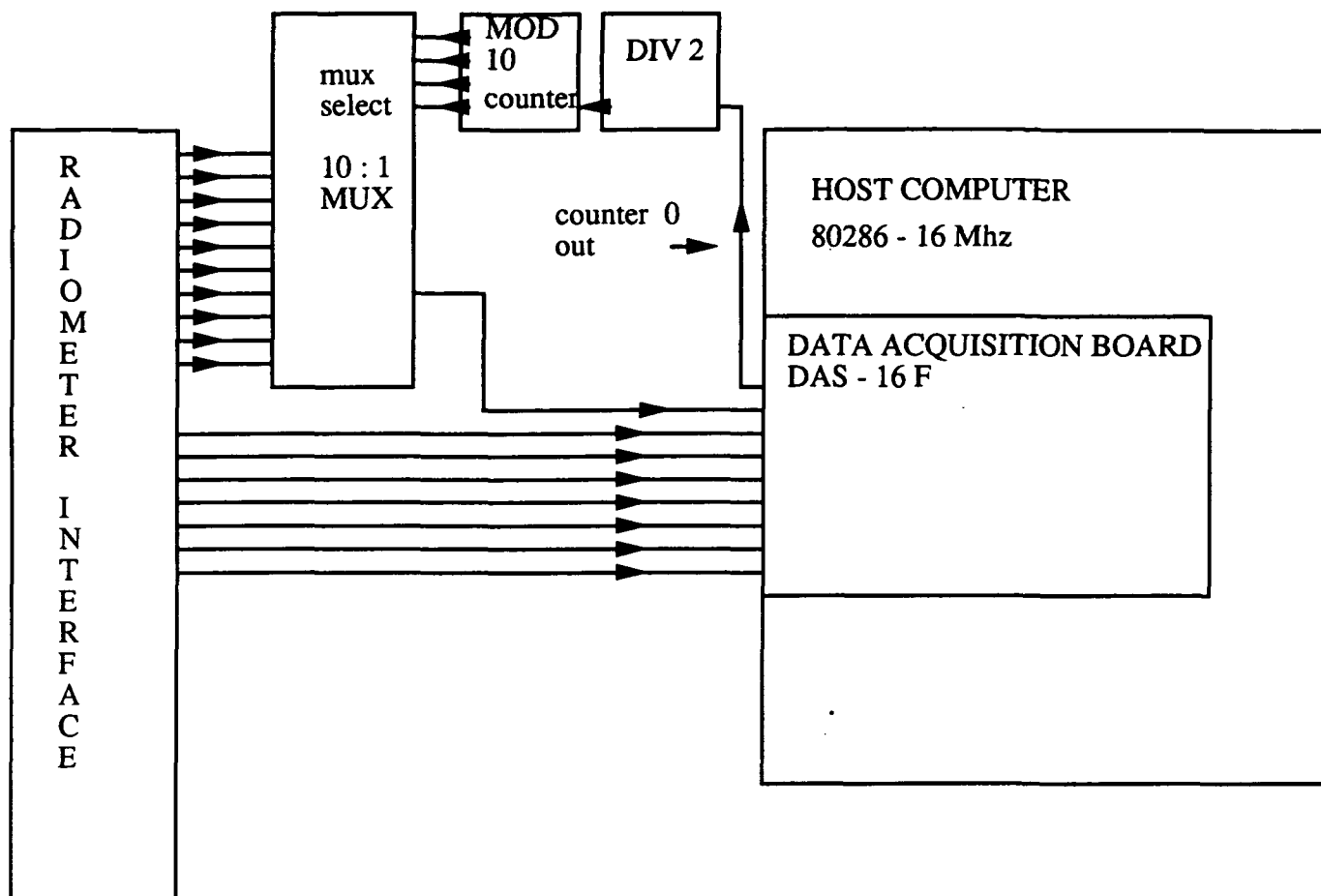
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Appendix II

***DATA ACQUISITION SYSTEM FOR 37 GHz
RADIOMETER EXPERIMENT***

PROJECT REPORT

***MEETUL PARIKH
RADAR & REMOTE SENSING LABORATORY
UNIVERSITY OF KANSAS, LAWRENCE
1991***



INTRODUCTION

The data acquisition system for the 37 GHz Radiometer Experiment samples the analog signals from the radiometer, stores the samples in a user specified file and displays up to four channels in real time. The data acquisition system is designed for eight differential input channels. A control signal is provided to multiplex ten inputs to a single differential ended input.

SYSTEM HARDWARE

The data acquisition system comprises of a plug in data acquisition card, DAS-16F from Metrabyte and a host computer. The host computer is an AT 286, 16 MHz rackmount computer from Texas Micro Systems. The DAS-16F is a 16 channel plug in card which can be configured as either 16 single ended inputs or 8 differential ended inputs. It's maximum sampling frequency is 100 KHz. The DAS-16F is configured as follows:

- 1) 8 differential ended input channels.
- 2) Bipolar input : range -10V to +10V.
- 3) DMA channel # : 1
- 4) Base address : \$300

For the case of eight differential ended input channels, a counter on the board is programmed to output a square wave at twice the sampling frequency entered by the user. For the Radiometer experiment this square wave is fed to a divide by two counter, whose output is fed to mod 10 counter. The mod 10 counter controls a multiplexer which multiplexes 10 input signals to one differential input of the data acquisition board.

SYSTEM SOFTWARE

The software for data acquisition and storage is written in Turbo Pascal version 5.5 and uses TTOOLS, Turbo Pascal drivers for DAS-16F from Quinn Curtis. The software requires that the monitor have a display size of at least 640 * 200. With this software the user can sample data at either 1, 2, 5, 10, 100 Hz or at a user defined frequency less than 100 Hz and can plot 4 channels of data real time on the screen and do real time logging to the storage device. The plots are updated every second and refreshed after every 500 samples. The channel shown on window

number 4 can be changed on the fly by pressing any key other than the <RETURN> key. The data acquisition run can be stopped by pressing the <RETURN> key. The data are stored in a user specified untyped (Turbo Pascal terminology) file. The user should not enter the file name extension. All data files will have a .dat extension. The parameters and the comments entered by the user will be stored in a file called *filename.par*. The parameter file will also contain the date of the data acquisition run, the time when data acquisition began, the time when data acquisition is finished and the number of samples collected. Each sample is stored as a raw binary value in a left justified word format, where the analog signal is stored in bits 4-15 of the word and the channel number of the analog data is stored in bits 0-3 of the word. To view the collected data in volts another program called POSTNEW is provided.

SOFTWARE GUIDE

The software comprises of 2 programs :

1) DAC4NEW:

This program comprises of the user interface module, data acquisition module, card driver module and the graphics module. The screen output for the opening banner, the data acquisition menu and the waveforms display were obtained by using the print screen command and are attached. To run this program type <DAC4NEW> at the system prompt. Note that the .BGI file (graphics driver) should be in the same directory as the program for the program to run. On running the program the main menu is displayed. The function keys F1 - F10 are used to change the data acquisition parameters. Pressing any function key will allow the user to change the parameter for data acquisition as follows :

F1 - Toggles between the type of data which can be either TARGET data or CALIBRATION data.

F2 - Toggles between the type of load which can be HOT load or SEMI HOT load.

F3 - This selects the sampling frequency for the data acquisition board. The user can choose from the following sampling frequencies : 1, 2, 5, 10, 100 (in Hz) or a user defined sampling frequency less than 100 Hz. The default is 100 Hz. To select a desired sampling frequency, press

F3 till the desired sampling frequency is shown in the parameter box. If the user want's to have some other sampling frequency less than 100 Hz, than the user should press the enter key on the "ENTER FREQ < 100" prompt and enter the desired frequency. Note that 500 should be an integer multiple of the user defined sampling frequency

F4 - On pressing F4, the user is prompted to enter the lower limit (0:7) and the upper limit (0:7) of the channel range to sample. The lower limit is to be entered on the CHANNEL LOW prompt while the upper limit is to be entered on the CHANNEL HIGH prompt. Note that the upper limit should be greater than the lower limit, otherwise an error signal with a beep is indicated.

F5 - This allows the user to modify the channels to be displayed. The default is channel 0, 1, 2 and 3. The user can plot up to 4 channels in real time. The user is prompted to enter the channels to be plotted. The channel should be within the channel range to be sampled, corresponding to F4. For example, if the channel range to be sampled is from 2 : 6 and the plot channel option has channels 0, 3, 5 and 7 to be displayed, the software will display 2 (in place of 0), 3, 5, and 6 (in place of 7). Note that channel shown by window 4 can be changed on the fly. While data acquisition is on, pressing any key other than <CR> will show the next channel on window 4 ie. if window 4 is showing channel 5, on pressing a key other than <CR> window 4 will now show channel 6.

F6 - This prompts the user to enter the file name where the acquired data are to be stored. On pressing F6 the parameter window will show the default(or the last modified) path and filename. The user should press the <BACKSPACE> to modify the path and the filename. Note that the user **should not enter the file name extension**. All data files will have a .dat extension and all parameter files will have a .par extension.

F7 - This prompts the user to enter the acquisition time in minutes. This should be an integer. The maximum acquisition time is 999999 minutes. Note that the acquisition time is limited by the amount of disk capacity.

F9 - To begin acquisition , press F9. On pressing F9, the user entered parameters are displayed. The user is prompted as to whether he want's to continue with the data acquisition or to go back to the main menu (default is 'Y'). On continuing, the user is prompted whether he would like to add any comments to the .par file (default is 'N'). The comments and the parameters are stored in the file *filename.par*. On entering the comments the user is asked whether he would like to start the run immediately (default 'Y'). If the user enters 'N' then he is prompted to enter the approximate time (in minutes) from that moment when he want's the data acquisition to start.

F10 - On pressing F10, the user exit's to the operating system.

2) POSTNEW :

The data is stored in a raw binary format for saving disk space. Each sample is stored as a two byte integer where bit 0-3 indicate the channel number and bits 4-15 represent the sampled analog voltage. This program converts the raw data to it's analog representation in volts and displays the data in a tabular format. To run this program enter <POSTNEW> at the system prompt. On running this program, the user is prompted to enter the filename which he would like to see. Note that the software does not check whether the file exist's or not. On entering a wrong filename the program exit's and gives a run time error. On entering a proper filename, this program senses the number of channels that were sampled. It displays seven channels on the screen at a time and twenty samples per channel. The samples are numbered in the order they were acquired. The window at the bottom of the screen shows the keys for looking up different channels and different samples. The control keys and their function are as follows:

< HOME > : Display samples starting from sample number one.

< END > : Display the last twenty samples that were acquired.

< PGUP > : Display the previous twenty samples.

< PGDN > : Display the next twenty samples.

- < CR > : Display the next twenty samples - same function as <PGDN>.
- < ↑ > : Go up one sample.
- < ↓ > : Go down one sample.
- < → > : Show channel to the right. If the channels displayed are channel 0 - channel 6, it will now display channel 1 - channel 7.
- < ← > : Show channel to the left. If the channels displayed are channel 1 - channel 7, it will now display channel 0 - channel 6.
- < G > : Move to a particular sample number. The user is prompted to enter a sample number where he want's to move.
- < C > : Convert to a text file. The user is prompted to enter the filename where he want's to save data in a text format to be used with a spreadsheet. The data from different channels are separated by spaces.
- < ESC > : Exit's from the program.
- < Q > : Exit's from the program.

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Appendix III

Design and Construction of
37-GHz Noise Injection
Radiometer

By

James Ampe

B.S.E.E., Iowa State University, 1986

Submitted to the Department of Electrical and Computer
Engineering and the Faculty of the Graduate School of the
University of Kansas in fulfillment of the requirements
for the Master of Science degree.

Dr. Prasad Gogineni

Dr. Richard K. Moore

Dr. Swapan Chakrabarti

April, 1992

Abstract

This paper describes the design and evaluation of a noise injection radiometer for measuring thermal radiation at 37 GHz. The instrument will be used by scientists in experiments to examine the relationship between detected thermal radiation and actual vegetation conditions at the earth surface. The problem is in detecting very small radiation quantities, changes of 1 Kelvin or less, so that it will have scientific use. A very sensitive microwave receiver in a controlled temperature environment is required to detect small thermal quantities. Radiometric theory is explained beginning with simplifier instruments to help reveal the advanced features of the noise injection radiometer. The design of the microwave assembly and detection circuits comprising the sensitive receiver are explained. The solution to the temperature control problem is also illustrated. The radiometer is then tested and shown to have a sensitivity of 0.5 Kelvin for aircraft use. The microwave radiometer does meet the desired performance requirements.

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1.0 Introduction

Radiometers are very sensitive passive microwave receivers. The microwave energy received in a particular bandwidth is commonly referred to as scene brightness temperature. According to black body radiation theory, an object radiates energy in a uniform spectrum. The amount of energy radiated is proportional to its temperature. However, the amount of energy detected in any bandwidth is dependant upon the emitter (earth in our applications) and the medium between the emitter and receiver (plants, atmosphere). Therefore, the energy received in particular bandwidths can give information on the earth surface, vegetation and atmospheric conditions.

The 37-GHz frequency is of interest to scientists who study remote sensing of vegetation because the wavelength, 0.8 cm, is comparable to geometries associated with plants. Electromagnetic scattering from leaves and branches for this wavelength are more significant than at the lower atmospheric window frequencies [1]. In addition, the difference in received energy between horizontal and vertical polarization is also important for scientific analysis.

Getting land and vegetation information from polarization difference data could be simpler than examining the magnitude of a particular channel alone. In taking the difference between polarizations, atmospheric effects in the data cancel out [1]. Water areas could first be separated from land areas because water has a much higher polarization difference. Polarization difference for smooth and dry soil (desert-like) is about 25-35 Kelvin. For dense vegetation though, the polarization difference is a maximum of 3 Kelvin. The small polarization difference for vegetation decreases further with dead vegetation present on the ground. Therefore the difference in received polarization can give a good indication of the amount of vegetation present over arid land [2].

Most of the previous radiometer research has been with soil moisture [1] and sea surface temperature [6]. Limited work has been done with vegetation. The purpose of this radiometer is to give accurate field data for 37-GHz radiometric research.

2.0 Comparison to Previous Radiometer Instruments

Radiometer implementation techniques are borrowed from early radio astronomy. Radio astronomy began after World

War II, and by the late fifties most of the basic principles that would be used for radiometry improvements had or were being introduced. In 1946, R. H. Dicke published a paper on a modulation principle that greatly reduces radiometer fluctuations due to receiver instabilities [3]. Then in 1952, Machin, Ryle and Vonberg published a paper on a null balancing receiver where noise is added to the reference of a Dicke receiver [7]. Although other circuits have been introduced that give slight improvements over these two methods, the Dicke switching and noise injection are the two most-often-used techniques employed in improving the basic radiometer.

One of the improvements was a different technique in noise injection. In 1967, W. Goggins published a paper on a null-balancing L-Band radiometer where injected noise is added to the antenna signal [4]. The paper compared two different methods for controlling the noise injection. The two methods were a mechanical attenuator and an electronic attenuator. The constant noise source was a gas discharge tube. The relation between scene temperature and attenuation was fairly linear from 90 to 200 Kelvin.

In 1968, Kuenzi and Schanda published a paper demonstrating an X-Band radiometer with noise injection in the reference [9]. In their circuit, noise injection is controlled by diode bias current. This gives a second-order relationship between diode current and scene temperature.

In 1974, Hardy, Gray, and Love published a paper on an S-Band noise injection radiometer for sensing sea temperatures [5]. This paper was based upon the detailed report for the instrument published in 1972. Their goal was good long-term absolute accuracy for space-borne applications. The absolute accuracy was achieved by adding the noise to the antenna signal. Thus the losses and reflections after the noise injection did not change the observed scene temperature. The noise source used is an avalanche diode. The injection was controlled by a PIN switch with constant pulse width.

This radiometer is different from previous published radiometers by the combination of all of the following:

- Dual Polarization.

- 37 GHz Noise Injection by Noise Diodes.

- PIN Switch control of noise injection.

- Selectable Hot or Warm reference loads.

Table 2.1 Compares previous published instruments which are similar to this radiometer.

Table 2.1 Comparison to Previous Published Instruments.

Summary of Pertinent Radiometer Instruments						
Author	Year	Frequency	Type	Noise Source	Noise Injection	Noise Control
Goggins	1967	L-Band	NIR	Gas discharge	Added to antenna	PIN atten.
Kuenzi	1968	X-Band	NIR	Diode	Added to reference	Diode bias current
Hardy	1974	S-Band	NIR	Diode	Added to Antenna	PIN switch
Ampe Thesis	1992	Ka-Band	NIR, Dual Polar	Diode	Added to antenna	PIN switch

3.0 Radiometry Theory

Radiometers detect naturally emitted radio noise, which can be related to temperature. The detected power radiated by an ideal black body is related to the objects temperature by:

$$P = kTB \quad (3.1)$$

where T is the temperature in Kelvin, k is Boltzmann's constant ($1.38\text{E-}23$ J/K) and B is the predetection noise bandwidth.

However, perfect black body conditions are not found in nature, so the deviation from ideal is described by an object's emissivity.

$$\text{Emmissivity} = \frac{\text{Actual Power}}{\text{Black Body Power}} \leq 1.0 \quad (3.2)$$

Emissivity is related to reflectivity by:

$$\text{Emissivity} = 1 - \text{Reflectivity} \quad (3.3)$$

A perfect blackbody does not reflect incoming energy. However, the black body will re-radiate absorbed energy to maintain equilibrium.

Another term that describes an object's radiated power is brightness temperature. The brightness temperature for a detected bandwidth is:

$$\text{Brightness} = 2kTBHxVER\lambda^2 \text{ units } \frac{\text{Watts}}{\text{m}^2 \text{ sr Hz}} \quad (3.4)$$

This equation for brightness uses the Rayleigh-Jeans approximation to Plank's radiation law. The Rayleigh-Jeans approximation is good up to approximately 1000 GHz for our range of brightness temperatures [17].

Brightness temperature and emissivity are also interrelated by:

$$\text{Brightness Temp} = \text{Emissivity} \cdot \text{Molecular Temp} \quad (3.5)$$

where the molecular temperature is often approximated by the physical temperature.

One problem with measuring the brightness temperature is that the electronic system inherently adds noise to the input signal. The electronic system is characterized with a noise temperature value that describes the amount of added noise. The added noise describes the ability of the system to detect small changes in the scene's brightness temperature. The system noise can be factored out by proper calibration, but the sensitivity of the device is still reduced. For example, using a 250 K scene brightness temperature and a 1500 K system noise temperature, a 1% change in the brightness temperature corresponds to a 0.14% change in the system output noise power. The amount of a scene brightness temperature change that the radiometer can detect is its sensitivity.

One factor that contributes to system noise is front-end losses and the physical temperature of the losses. The loss absorbs an amount of the incoming noise power and adds back another amount of noise power. This is shown graphically in figure 3.1.

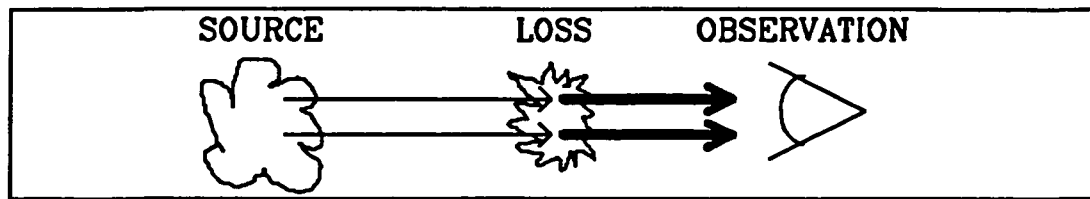


Figure 3.1 Representation of Effect of Loss.

Thus, depending upon the temperature difference between the source and the loss, the observed noise power can be greater or less than the transmitted noise power. This effect occurs when the signal travels through the atmosphere, as well as when traveling through the front end of the radiometer.

The exact observed temperature [13] is:

$$Temp_{obs} = (1 - Loss) \text{ Scene Bright Temp} + (Loss) \text{ Physical Temp} \quad (3.6)$$

Reflections have a similar effect:

$$Temp_{obs} = (1 - Refl) \text{ Scene Bright Temp} + (Refl) \text{ Microwave Temp} \quad (3.7)$$

where the loss or reflection is expressed as a fractional value. The microwave temperature in the reflection equation is the effective microwave temperature looking at the reflected power. Generally the losses have much more effect than reflections.

For effective interpretation of the received power, the losses and temperature thereof must be known and stable.

Changes in these values affect the absolute accuracy of the radiometer.

Calculations used in determining the effects of the lossy devices in the radiometer use equations for multiple input port lossy devices. For a three-port device, having two temperature inputs each with separate losses, the combined noise temperature referred to the output of the device [8] is:

$$T_{out} = T_o \left(1 - T_{in1} 10^{-\frac{L1}{10}} - T_{in2} 10^{-\frac{L2}{10}} \right) + T_{in1} 10^{-\frac{L1}{10}} + T_{in2} 10^{-\frac{L2}{10}} \quad (3.8)$$

where T_o is the physical temperature of the lossy device and losses, L , are in dB.

Derived from equation (3.8), the two-port equation for the noise temperature added by the device referred to the device input is:

$$T_{in} = T_o \left(10^{\frac{L}{10}} - 1 \right) \quad (3.9)$$

4.0 Noise Injection Radiometry Theory

To best understand the noise injection radiometer, simpler systems from which it was derived will first be

examined. The basic system is the total power radiometer. The block diagram is shown in figure 4.1.

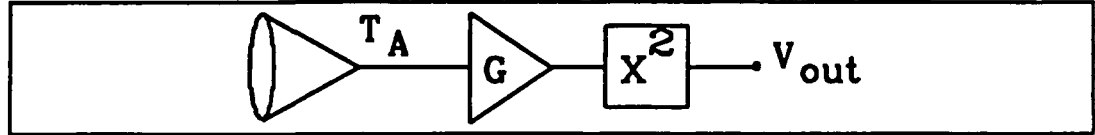


Figure 4.1 Total Power Radiometer.

The system consists of an antenna, amplifier, and square-law detector. The output is simply given by:

$$V_{out} = c(T_A + T_N)G \quad (4.1)$$

where T_N is the noise added by the system and c is the detection conversion factor (mV/mW). The corresponding ideal sensitivity is:

$$\Delta T = \frac{T_A + T_N}{\sqrt{B\tau}} \quad (4.2)$$

where B is the predetection noise bandwidth and τ is the integration time.

One of the difficulties of using total power radiometers is that the output varies with changes of gain in the system. The gain changes have the effect of changing the sensitivity according to:

$$\Delta T = (T_A + T_N) \sqrt{\frac{1}{B\tau} + \left(\frac{\Delta G}{G}\right)^2} \quad (4.3)$$

The system gain is calibrated during use by the antenna periodically looking at known temperature loads. However, to some extent, the gain changes will reduce the sensitivity from that of the ideal value. The reference by Thomsen stated that "Low-frequency gain fluctuations have the most deteriorating effect on the radiometer performance" [15 p.148].

The Dicke radiometer [3] is a total power radiometer that alternately switches the radiometer front end between the antenna and a known temperature load. The block diagram is shown in figure 4.2.

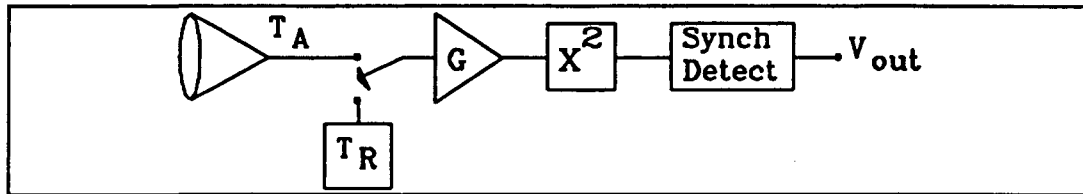


Figure 4.2 Dicke Radiometer.

The added components from a total power radiometer are a low-loss switch, a reference load, and a synchronous detector circuit. The synchronous detector separates the two time multiplexed temperatures. The output is given by:

$$V_{out} = c(T_A - T_R) G \quad (4.4)$$

Equation (4.4) shows that the output is some factor of the difference in temperature between the antenna scene, T_A , and the reference load, T_R . Since the system noise is added equally to T_A and T_R , the system noise temperature effectively cancels by the subtraction of T_R from T_A .

The corresponding ideal sensitivity is:

$$\Delta T = \frac{2 (T_A + T_N)}{\sqrt{B \tau}} \quad (4.5)$$

The Dicke radiometer has half the ideal sensitivity of the total power radiometer because the antenna scene is only looked at half of the time. However, one benefit is that the system noise cancels at the output. The second and greater benefit is that the output is less susceptible to gain changes. The sensitivity due to gain changes is given by [16]:

$$\Delta T = \sqrt{\left(\frac{2 (T_A + T_N)}{\sqrt{B \tau}} \right)^2 + \left(\frac{(T_A - T_R) \Delta G}{G} \right)^2} \quad (4.6)$$

Equation (4.6) shows that changes in gain are only multiplied by the difference in antenna scene and reference temperature. Therefore the actual sensitivity is closer to the ideal sensitivity than with the total power radiometer.

The noise injection radiometer (NIR) is a Dicke radiometer where the antenna temperature and reference temperature are always equal. The block diagram is shown in figure 4.3.

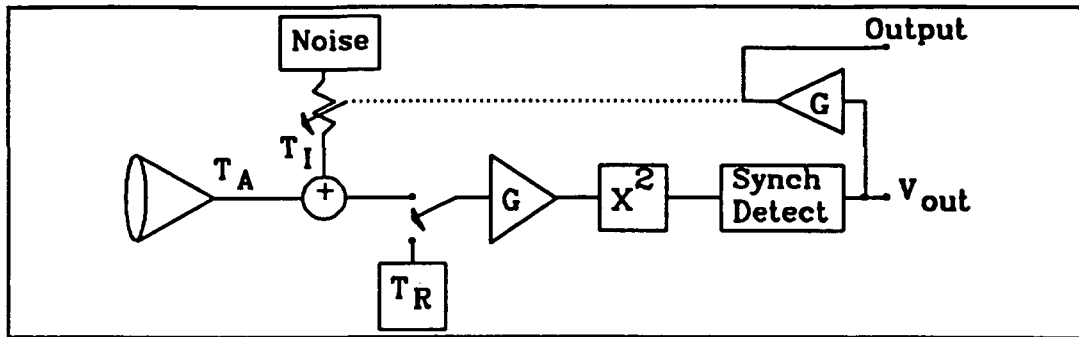


Figure 4.3 Noise Injection Radiometer.

The added components from a Dicke radiometer are a noise source, attenuation control and a coupling component. In this case, the noise is added to the antenna scene to equal the reference temperature. The output is the noise-controlling voltage and is given by:

$$V_{out} = C((T_A + T_I) - T_R) G \approx 0 \quad (4.7)$$

The output will not exactly be zero in normal operation. The ideal sensitivity is:

$$\Delta T = \frac{2(T_R + T_N)}{\sqrt{B \tau}} \quad (4.8)$$

The noise injection radiometer has similar sensitivity as the Dicke radiometer. The difference is the T_R term in the numerator, replacing the T_A term in the Dicke sensitivity equation (eq. 4.5). The effect is that if the system noise temperature is low, then the temperature of the reference, T_R , will give the NIR noticeably less sensitivity than the Dicke. The sensitivity with respect to gain changes for the NIR system is the same as the ideal sensitivity. Looking back to equation (4.6), the term T_A for the NIR is equal to T_R , thus the gain change effects are zeroed out.

Another advantage of the NIR occurs when the reference temperature is the same physical temperature as the waveguide components inside the radiometer. The microwave temperature within the components after the noise injection will be the same temperature as the reference load. Then, according to the basic radiometry theory, if the loss is at the same temperature as the microwave temperature, then the lossy device will emit the same power that it absorbs (eq. 3.6). The same is true for reflections (eq. 3.7). Therefore the output is not dependent upon both variations in microwave component loss and reflections between the point of injection and the mixer. This improves the long term-accuracy.

5.0 RF Section Analysis

The RF block diagram is shown in figure 5.1.

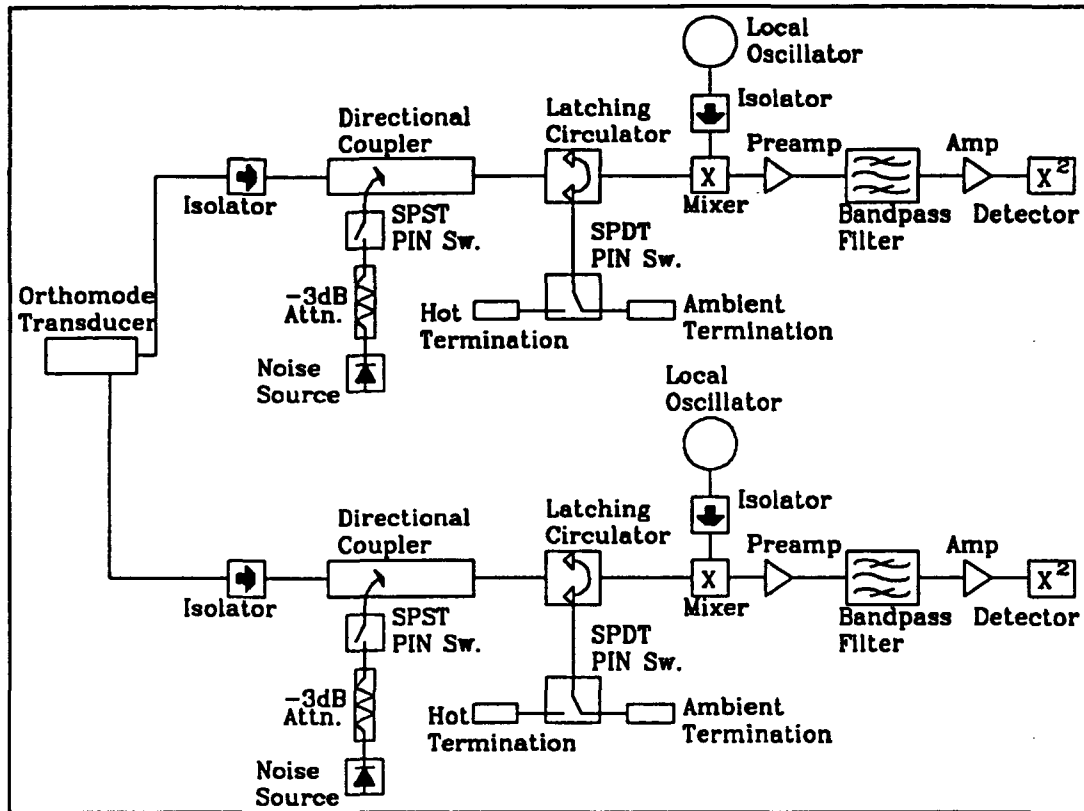


Figure 5.1 RF Diagram.

The RF system description begins with an orthomode transducer (OMT), which separates the received antenna energy to vertical and horizontal polarizations. The waveguide lengths after the OMT are the same to help equalize losses for the two polarization channels. The RF circuitry for each polarization is identical after the OMT. An isolator is located at the front end to give a better match to the input of the radiometer. The isolator also attenuates internal reflections going

toward the antenna. These reflections would decrease the long-term accuracy.

Following the isolator, injected noise power is combined with the antenna signal in the 20-dB directional coupler. The 20-dB coupler presents only a small loss for the antenna signal.

A PIN diode switch modulates the noise diode output, effectively controlling the amount of injected noise. Using a modulated PIN diode switch gives better attenuation stability than a PIN diode attenuator. The noise is generated by a noise diode, set for an approximate 23-dB excessive noise figure. A resistive card attenuator was placed after the noise source for maximum dynamic range of the noise modulation duty cycle. The attenuation is adjusted from its set attenuation of 3 dB. The attenuator also helps to attenuate multiple reflections between the noise source and the PIN switch. The ENR rating of the noise source is available noise power; the actual power out of the noise diode depends upon the match of its load.

A latching circulator acts as the low-loss Dicke switch. The circulator is a three-port device that passes signals between the ports in either a clockwise or

counterclockwise direction. The latching circulator only consumes power when switching. The circulator directs either the antenna + injected noise or the reference to the mixer. A cavity-tuned Gunn oscillator is the local oscillator and is matched to the mixer with an isolator. The mixer is operated in the double sideband mode for lower noise figure. The IF preamplifier is in a separate housing from the mixer, but both are a supplied pair.

The bandpass filter sets the detection bandwidth and filters out unwanted signals which could saturate the following amplifier. The filter has a 150- to 450-MHz 3-dB passband. Assuming relatively stable oscillators, the low-frequency cutoff is high enough to filter out the difference frequency term produced by the V and H local oscillators intermixing. The filter also rejects the harmonics of the Dicke switching frequency, which are greater than the noise level below 50 MHz. Harmonics of the switching frequency present in the IF bandwidth would reduce the sensitivity of the radiometer [15].

The amplifier following the filter boosts the noise signal for detection. The noise power at the input of the detector must be above the diode's minimum detectable signal but not beyond the square-law region. The back diode detector then converts the IF power to a voltage.

5.1 Calculated System Noise Temperature

The RF system losses are evaluated in table 5.1 to determine the expected system noise temperature. Losses used are manufacturer's specified maximum value. The system noise temperature is a factor in determining the radiometer sensitivity.

Table 5.1 Worst- Case System Noise Temperature.

Expected Worst-Case System Noise Temperature	
Component	Manufacturer Spec. Value
Orthamode Transducer	0.5-dB Loss
Isolator	1.5-dB Loss
Directional Coupler	0.7-dB Loss
Latching Circulator	0.4-dB Loss
Waveguide, 13 Inches	0.34-dB Loss
Total Losses	3.44 dB
Mixer-Preamplifier Noise Figure	F= 5.5 dB
Total Noise Figure, F	F= 5.5 + 3.44 = 8.94 dB
System Noise Temp, $T_{\text{sys}} = 290(F-1)$	$T_{\text{sys}} = 1982$ Kelvin

5.2 Calculated Detected Power

Next, the RF section is analyzed for determining expected noise power to the detector. The noise power to the detector is set by the reference load temperature. This is because the antenna + injected noise temperature is forced equal to the reference temperature.

Because of the nulling technique, the power at the detector will be either that seen looking at the ambient reference or the hot reference. The ambient reference is at the expected radiometer temperature of 313 K and the hot load is held to approximately 343 K. A PIN switch selects one of the two references. The actual noise temperature at the output of the PIN switch is computed as the output of a lossy three-port device. The specified losses involved are:

SPDT PIN switch on	loss	1.4 dB
SPDT PIN switch off	loss	22 dB
Waveguide bend		0.061 dB

The noise temperature at the output of the PIN switch for each reference is:

$$\begin{aligned}
 T_{amb} &= 313(1 - 10^{-0.146} - 10^{-2.206}) + 313 \cdot 10^{-0.146} + 343 \cdot 10^{-2.206} \\
 T_{amb} &= 313.187 \text{ K} \\
 T_{hot} &= 313(1 - 10^{-0.146} - 10^{-2.206}) + 313 \cdot 10^{-2.206} + 343 \cdot 10^{-0.146} \\
 T_{hot} &= 334.430 \text{ K}
 \end{aligned}
 \tag{5.1}$$

Then the losses between the SPDT PIN switch and latching circulator are considered.

1 waveguide bend	loss	0.061 dB
------------------	------	----------

The noise temperature at the output of this waveguide, is the sum of the input noise temperatures divided by the loss.

$$\begin{aligned}
 T_{amb} &= \frac{313.187 + 313(10^{0.0061} - 1)}{10^{0.0061}} = 313.184 \text{ K} \\
 T_{hot} &= \frac{334.430 + 313(10^{0.0061} - 1)}{10^{0.0061}} = 334.132 \text{ K}
 \end{aligned}
 \tag{5.2}$$

This is the noise temperature into the circulator. Notice that the ambient reference temperature is very close to its starting value, 313 K, but the hot reference temperature has decreased by 8.9 K.

The noise temperature out of the circulator is computed as a lossy three-port device. The two input temperatures are equal under balanced conditions. The specified losses involved are:

Circulator insertion	loss	0.4 dB
Circulator isolation	loss	20 dB

The reference temperatures out of the circulator are then:

$$\begin{aligned}
 T_{amb} &= 313(1 - 10^{-0.04} - 10^{-2.0}) + 313.184(10^{-0.04} + 10^{-2.0}) = 313 \\
 T_{hot} &= 313(1 - 10^{-0.04} - 10^{-2.0}) + 334.132(10^{-0.04} + 10^{-2.0}) = 332
 \end{aligned}
 \tag{5.3}$$

The noise temperature of the 5.5-dB NF mixer-preamp, referred to the mixer input is:

$$T_{mxi} = 290(10^{0.55} - 1) = 738.959 \text{ K}
 \tag{5.4}$$

The noise temperature out of the preamp is the sum of the temperature at the mixer input multiplied by the gain.

$$\begin{aligned} T_{\text{preamp amb}} &= (313.170 + 738.959) 10^{2.7} = 527300 \text{ K} \\ T_{\text{preamp hot}} &= (334.132 + 738.959) 10^{2.7} = 537000 \text{ K} \end{aligned} \quad (5.5)$$

The filter after the preamp determines the IF predetection bandwidth. The bandwidth used in the equations is the noise bandwidth, defined by:

$$B_N = \frac{\left(\int_0^\infty G(f) df \right)^2}{\int_0^\infty G(f)^2 df} \quad (5.6)$$

where $G(f)$ is the power transfer function.

The filter used in the radiometer was closely approximated by the combination of a bandpass and lowpass filter. For the model, the lowpass cutoff frequency is set near the bandpass upper cutoff frequency, giving the bandpass a sharper cutoff for the higher frequencies. The -3-dB bandwidth of the filter is 300 MHz, centered at 300 MHz. The transfer function used is:

$$G(f) = \frac{1}{1 + \left(\frac{f_o}{300E6} \left(\frac{f}{f_o} - \frac{f_o}{f} \right) \right)^8} \frac{1}{1 + \left(\frac{f}{480E6} \right)^{12}} \quad (5.7)$$

where f_o is the center frequency defined by the -3-dB points f_1 and f_2 as:

$$f_o = \sqrt{f_1 f_2} \quad (5.8)$$

The resulting modeled predetection noise bandwidth is 320 MHz.

The power out of the preamp is calculated using kTB with the preamp output noise temperature.

Preamp Output Power, ambient ref.	2.329 nW
Preamp Output Power, hot ref.	2.371 nW

The power to the detector is boosted by a 27-dB IF amplifier gain.

Detector Input Power, ambient ref.	1.167 uW
	-29.33 dBm
Detector Input Power, hot ref.	1.188 uW
	-29.25 dBm

This is an acceptable power level to the detector. The power should be within the detector square law range of -51 dBm to -17 dBm.

5.3 Calculated Injection Noise Power

Next, the amount of noise power needed for injection is determined. Assuming equal loss through the circulator for both the antenna signal and reference, the matching

temperature for the antenna signal into the circulator is that of the reference into the circulator. Starting at the circulator input, the matching temperatures for the two references are (eq. 5.2):

Circulator input, ambient reference	313.184 K
Circulator input, hot reference	334.132 K

The temperature at the output of the coupler is then determined. The losses between the coupler and circulator are:

Two Waveguide Bends	loss	0.122 dB
---------------------	------	----------

This loss results in the waveguide bend noise temperature:

$$T_{bend} = 313 (10^{0.0122} - 1) = 9.2 \text{ K} \quad (5.9)$$

The waveguide noise temperature is subtracted from the circulator input noise temperature. This results in the antenna + injected noise temperature at the output of the directional coupler as shown:

$$T_{c \text{ amb}} = 313.184 (10^{0.012}) - 313 (10^{0.012} - 1) = 313.189 \text{ K}$$

$$T_{c \text{ hot}} = 334.132 (10^{0.012}) - 313 (10^{0.012} - 1) = 334.733 \text{ K}$$

Then the maximum and minimum noise temperatures originating from the antenna scene are determined at the directional coupler input. Losses external to the radiometer housing are treated separately from losses internal to the radiometer housing because of the difference in physical temperature. The external temperatures used are:

Antenna Scene, Min Temp	20 K (sky at nadir)
Antenna Scene, Max Temp	300 K (earth)
External Waveguide Min temp	250 K (-23 Celsius)
External Waveguide Max temp	300 K (+27 Celsius)

The specified losses of the external waveguide are:

Orthamode Transducer	0.5 dB
Two Inches of Waveguide	0.053 dB

External losses	0.553 dB

The noise temperature of the losses for both high and low waveguide temperatures are:

$$\begin{aligned}
 T_{ext\ low} &= 250 (10^{0.0533} - 1) = 33.97\ K \\
 T_{ext\ high} &= 300 (10^{0.0533} - 1) = 40.76\ K
 \end{aligned}
 \tag{5.11}$$

The temperature into the radiometer is the sum of the scene temperature and the loss noise temperature, divided by the loss. The lowest input temperature is for low scene temperature and cold external temperature. The largest input temperature is for high scene temperature and warm external temperature.

$$\begin{aligned} T_{in\ low} &= \frac{20 + 33.97}{10^{0.0553}} = 47.51\ K \\ T_{in\ high} &= \frac{300 + 40.76}{10^{0.0553}} = 300.0\ K \end{aligned} \quad (5.12)$$

Next the losses inside the radiometer are considered. Specified losses are:

Isolator	loss	1.5 dB
Two Waveguide Bends	loss	0.122 dB

Front-End Internal Losses		1.622 dB

The noise temperature of the losses are:

$$T_{int} = 313(10^{0.162} - 1) = 141.706\ K \quad (5.13)$$

The minimum and maximum noise temperatures at the input to the directional coupler are then:

$$\begin{aligned} T_{cin\ low} &= \frac{47.51 + 141.70}{10^{0.162}} = 130.251\ K \\ T_{cin\ high} &= \frac{300.0 + 141.70}{10^{0.162}} = 304.051\ K \end{aligned} \quad (5.14)$$

The noise to be injected into the directional coupler is the amount needed to balance the scene temperature at the input of the coupler (eq. 5.14) with the matching reference temperature at the output of the coupler (eq. 5.10). Modeling the coupler as a three-port lossy device, the injected noise temperatures for minimum and maximum scene temperatures are:

Ambient Load:

$$T_{amb\ low} = \frac{313.189 - 130.251 \cdot 10^{-0.07} - 313(1 - 10^{-0.07} - 10^{-2.0})}{10^{-2.0}}$$

$$T_{amb\ high} = \frac{313.189 - 304.051 \cdot 10^{-0.07} - 313(1 - 10^{-0.07} - 10^{2.0})}{10^{-2.0}}$$

(5.15)

Hot Load:

$$T_{hot\ low} = \frac{334.733 - 130.251 \cdot 10^{-0.07} - 313(1 - 10^{-0.07} - 10^{-2.0})}{10^{-2.0}}$$

$$T_{hot\ high} = \frac{334.733 - 304.051 \cdot 10^{-0.07} - 313(1 - 10^{-0.07} - 10^{2.0})}{10^{-2.0}}$$

(5.16)

Using the maximum and minimum temperatures for computing the limits for the excess noise ratio needed at the coupler input:

$$ENR\ Cin_{max} = 10 \log \left(\frac{18040}{290} - 1 \right) = 17.86\ dB$$

$$ENR\ Cin_{min} = 10 \log \left(\frac{1094}{290} - 1 \right) = 4.42\ dB$$

(5.17)

With the range of injected noise powers known, the noise injection components are evaluated.

Noise ENR	23 dB
Attenuator Loss	4 dB
Waveguide Loss,	
1 bend + 1 inch straight	-0.087 dB
PIN Sw Loss, (full on)	-1.0 dB

Max Available Noise at Input to Coupler	17.91 dB ENR
PIN Sw Loss, (full off)	26 dB

Min Available Noise at Input to Coupler	-7.09 dB ENR

Examining the results, the maximum available noise to the input of the coupler, 17.8 dB, is slightly more than the maximum needed, 17.5 dB. Also, the minimum available noise to the input of the coupler, -7.1 dB, is less than the minimum needed, 4.4 dB. Therefore, the noise injection setup has the dynamic range to balance out expected scene temperatures to either the ambient or hot reference load.

5.4 Calculated Radiometer Transfer Function

The duty cycle of the noise PIN switch is then examined to determine the relation between duty cycle and scene

temperature. The antenna's physical temperature is assumed to be 290 K, and the reference is the ambient load. The matching temperatures at the output of the directional coupler (eq. 5.2) is:

Coupler output, ambient reference 313.184 K

The scene temperature is denoted by the variable "s" in the following equations. Accounting for the external front-end losses, 0.522 dB, the noise temperature into the radiometer is:

$$T_{in}(s) = \frac{s + 290(10^{0.055} - 1)}{10^{0.055}} \quad (5.18)$$

Then the internal front-end losses, 1.62 dB combined, are used to compute the noise temperature into the coupler.

$$T_{cin}(s) = \frac{T_{in}(s) + 313(10^{0.162} - 1)}{10^{0.162}} \quad (5.19)$$

Then modeling the directional coupler as a lossy three-port, the injected noise temperature needed for balancing is computed from the specified losses:

Coupler insertion loss 0.7 dB

Coupler isolation loss 20 dB

$$T_{inj}(s) = \frac{313.189 - C_{in}(s) 10^{-0.07} - 313(1 - 10^{-0.07} - 10^{-2.0})}{10^{-2.0}} \quad (5.20)$$

From a 23 dB-ENR noise source, the losses required for the noise temperature injection is:

$$L_{inj}(s) = \frac{313(10^{2.3} - 1)}{T_{inj}(s) - 313} \quad (5.21)$$

The fixed losses between the noise source and coupler are:

Card attenuator	4 dB
Waveguide, 1 bend 1 inch straight	0.087

The specified PIN switch loss characteristics are:

Attenuation, full on	1 dB
Attenuation, full off	26.5 dB

A duty cycle of 1.0 corresponds to the PIN switch being full on. Continually halving the duty cycle adds 3-dB attenuation until the full-off attenuation is reached. Therefore, the equation for the PIN switch loss in dB is:

$$LOSS_{PIN} = 10 \log \left(\frac{1}{(duty) 10^{-0.1} + (1-duty) 10^{-2.65}} \right) \quad (5.22)$$

The equation for duty cycle as a function of the scene temperature is then:

$$duty(s) = \frac{1 - 10^{-2.65} L(s) 10^{\frac{-4.0 - .087}{10}}}{10^{-0.1} L(s) 10^{\frac{-4.0 - .087}{10}}} \quad (5.23)$$

Equation (5.23) is the expected radiometer transfer function. The function is linear as shown in figure 5.2.

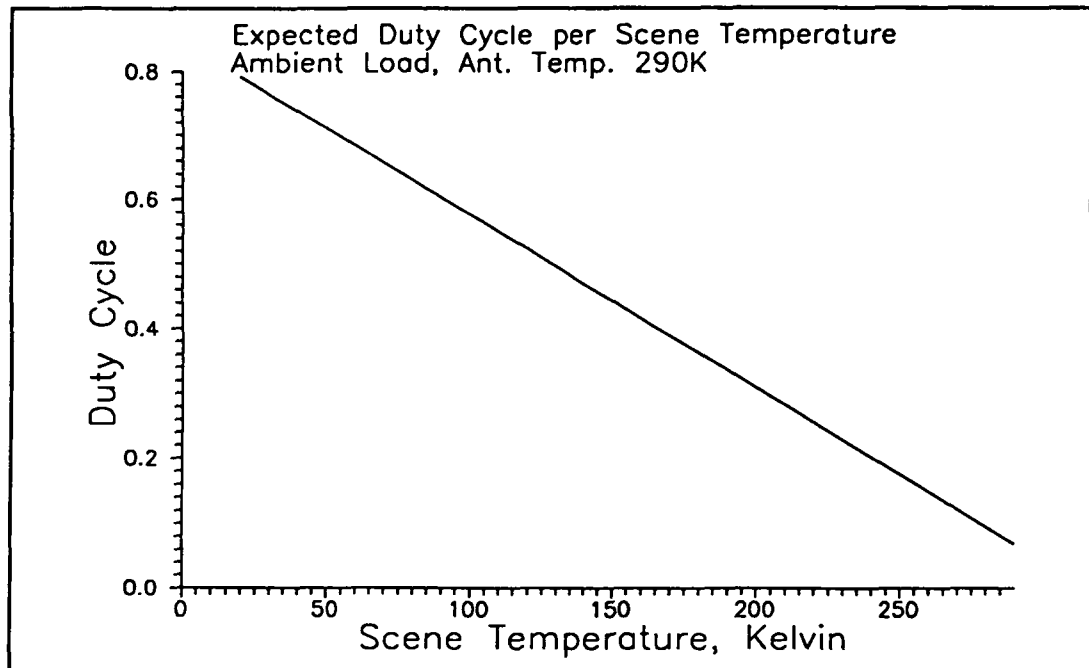


Figure 5.2 Calculated Radiometer Transfer Function.

The transfer function for the hot reference is similar, but shifted slightly upward on the duty cycle axis.

6.0 AF Section Analysis

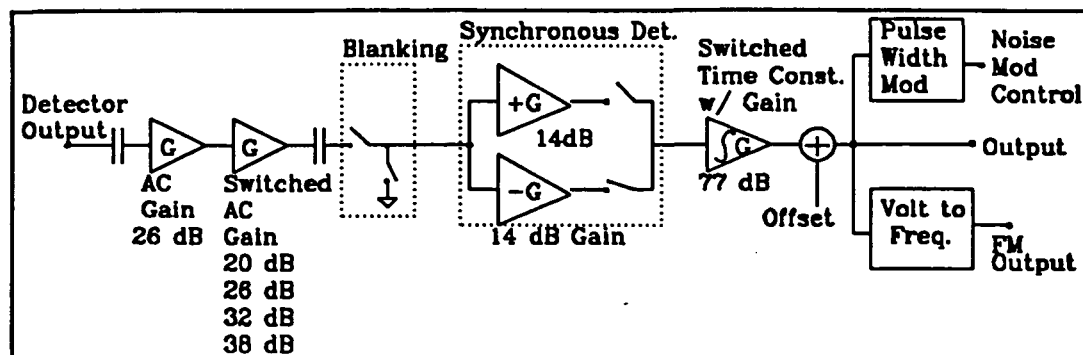


Figure 6.1 AF Diagram.

The audio frequency section is shown in figure 6.1. The gain values shown are the final determined values (table 12.9). The output of the IF detector is AC coupled to the first op-amp stage. The coupling capacitor removes the negative dc bias from the detector. The output of the first op-amp is amplified by a switchable gain stage. The gain is switched in conjunction with the integrator time selection to maintain loop stability with a broad range of integrator values. The first and second op-amps are the AC amplifiers. The switched-gain stage is placed in the AC section because the change in gain will vary the dc offset out of the op-amp. The capacitor after the AC stages removes the DC offset.

Following the variable gain stage is a blanking switch. The signal is blanked during the latching circulator switching time. When blanked, the signal path is disconnected and the input to the synchronous detector is

grounded. The synchronous detector separates the time-multiplexed signal of antenna scene + injected noise and reference load. The synchronous detector switches with the latching circulator at a 2-KHz frequency. The output of the synchronous detector will be a DC offset in the unbalanced case, near zero in the balanced case. The synchronous detector stage has a small gain to satisfy the particular op-amp requirements for stability. The op-amps used until the end of the synchronous detector have high gain-bandwidth products and high slew rates. In the configurations used, the op-amp cutoff frequency is at least ten times the synchronous detector switching frequency. This amplifier bandwidth is needed to maintain square-wave reproduction. If the amplifier bandwidth is lower, near the synchronous detector switching frequency, then the sensitivity factor of "2" in eq (4.8) will be increased up to a factor of 2.22 [16]. This is undesirable because the sensitivity is decreased.

Following the synchronous detector is the switched time-constant integrator. The integrator incorporates the high DC gain. The high gain is acceptable here because a large bandwidth is not required. The output of the integrator is summed with an offset value to produce a control voltage that ranges from zero to +10 volts.

The control voltage is the output signal of the polarization channel. The control voltage also sets the duty cycle to the noise PIN switch and sets the voltage to frequency (v/f) rate. The v/f circuit is best for transmitting the output signal under external noisy conditions. At the far end of the cable a frequency-to-voltage converter (f/v) changes it back to a DC signal. However, the f/v circuit inherently has a temperature dependance and cannot be used for calibration.

6.1 Time Constant Calculation

Because the AF section completes a closed-loop system, the integrator RC time constant is not the closed-loop time constant [10]. The integrator RC time constant is the open loop time constant, which only sets the noise bandwidth in the AF section.

Starting from control theory, the simple closed loop response is calculated. The control block diagram is shown in figure 6.2:

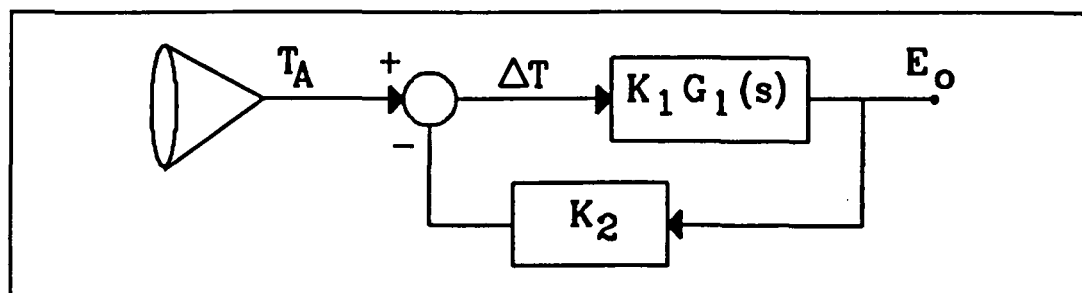


Figure 6.2 Feedback Control Loop.

The closed loop transfer function is:

$$\frac{E_o}{T_A} = \frac{K_1 G_1(s)}{1 + K_1 K_2 G_1(s)} \quad (6.1)$$

This transfer equation can be rearranged to a product of two terms:

$$\frac{E_o}{T_A} = \frac{K_1}{1 + K_1 K_2} \cdot \frac{1}{1 + x} \quad (6.2)$$

The unknown "x" term is solved for, resulting in:

$$x = \frac{\frac{1}{G_1(s)} - 1}{1 + K_1 K_2} \quad (6.3)$$

Then the $G_1(s)$ term is determined. For an ideal op-amp integrator shown in figure 6.3, the integrator transfer

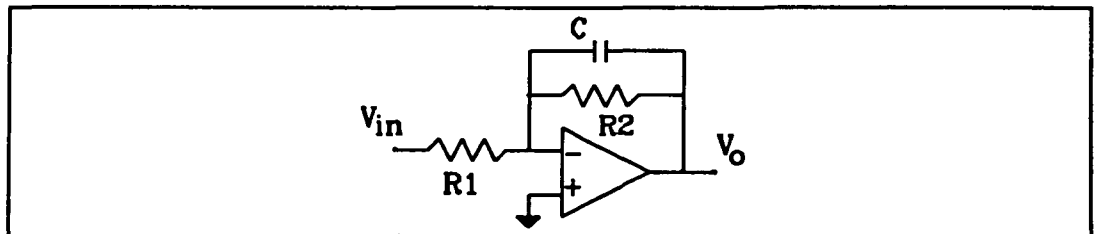


Figure 6.3 Op-Amp Integrator Circuit.

function is:

$$\frac{V_o}{V_{in}} = G(s) = \frac{R_2}{R_1 + R_1 R_2 s C} = \frac{R_2}{R_1} \cdot \frac{1}{1 + R_2 s C} \quad (6.4)$$

the R_2/R_1 gain term can be dropped off the $G_1(s)$ and included in the K_1 gain term. Substituting the simplified $G_1(s)$ results in:

$$x = \frac{R_2 Cs}{1 + K_1 K_2} \quad (6.5)$$

Substituting the "x" term in the closed-loop transfer equation:

$$\frac{E_o}{T_A} = \frac{K_1}{1 + K_1 K_2} \frac{1}{1 + \frac{R_2 Cs}{1 + K_1 K_2}} \quad (6.6)$$

The first product term is the closed-loop gain and the second term is the closed-loop pole response. It can be seen that the closed-loop time constant is the open-loop time constant, $R_2 C$, divided by the factor $(1+K_1 K_2)$.

In selecting loop gain constants, it is desirable that the loop error, ΔT , does not exceed the radiometer sensitivity [13]. The loop error expression is calculated by:

$$DE_{WAT} = \frac{E_o}{K_1 G_1(s)} \quad (6.7)$$

At steady state, the loop error is:

$$\Delta T_{ss} = \frac{E_o}{K_1} \quad (6.8)$$

K_1 will be chosen such that the maximum error will be equal to the sensitivity. The maximum error occurs when $E_o = 5$ Volts.

The feedback gain, K_2 , is determined from previous calculations in the RF section. The units for K_2 are Kelvin/Volt. From the duty cycle calculations, it was determined that the duty cycle would change by 0.75 from a 20 K scene to a 300 K scene. The duty cycle change of 0.75 corresponds to a 7.5-volt change to the noise duty cycle controlling circuit. The amount of injected noise for a 20 K scene is 182.93 K. The amount of injected noise for a 300 K scene is 9.13K. Therefore, the feedback gain constant is:

$$K_2 = \frac{182.9K - 9.134K}{0.75 \text{ duty}} \frac{1 \text{ duty}}{10 \text{ V}} = 23.17 \text{ K/V} \quad (6.9)$$

The open-loop AC*DC gain is determined from K_1 . K_1 is the change in output voltage per 1 K change in antenna scene. The combined RF and IF gain is approximately 54 dB, and the detector conversion is 800 mV/mW.

$$G_{AC} G_{DC} = \frac{K_1}{(kB10^{5.4}) \frac{800 \text{ mV}}{\text{mW}}} \quad (6.10)$$

The closed-loop time constant is determined from the system noise, T_{sys} , of 1982 K.

$$\tau = \frac{4 (T_{Ref} + T_{sys})^2}{W (\Delta T)^2} \quad (6.11)$$

Either the desired sensitivities or closed-loop time constants can be chosen first before other values are computed. It is desirable to have good sensitivity for ground measurements (< 0.5 K) and short time constants for flights (0.1 sec.). The computed constants are shown in table 6.1.

The open-loop time constant is determined by the single-pole op-amp integrator. The desired time constant is the effective integration time, which is determined from the noise bandwidth of the op-amp filter. The noise bandwidth for a low pass filter is defined by eq. (6.12) where $G(f)$ is the power transfer function.

Table 6.1 Computed Loop Constants.

Loop Constants				
Sensi- tivity	Closed- Loop Time Constant	K1	AC*DC Gain	Open-Loop Time Constant
2.0K	0.0176	2.5	3.005E6	1.037
1.0K	0.0702	5	6.010E6	8.203
0.5K	0.2809	10	1.202E7	65.37
0.2K	1.756	25	3.005E7	1019.

$$B_N = \frac{1}{G(0)} \int_0^{\infty} G(f) df \quad (6.12)$$

The power transfer function of the op-amp integrator is:

$$G(f) = \frac{1}{1 + (f R_2 C)^2} \quad (6.13)$$

Using this transfer function, the noise bandwidth is:

$$B_N = \frac{1}{4 R_2 C} \quad (6.14)$$

The effective integration time is then [11]:

$$\tau_I = \frac{1}{2 B_N} = 2 R_2 C \quad (6.15)$$

The AF gain breakdown and integrator components are shown in table 6.2.

Table 6.2 Computed Loop Gain and RC Values.

Gain Budget and Integrator Values				
AC * DC Gain	Total AC Gain	Total DC Gain	$R_2 * C$ Product	C, for $R_2=10M$
3.005E6	2,500.	1202.	0.518	0.051 uF
6.010E6	5,000.	1202.	4.10	0.41 uF
1.202E7	10,000.	1202.	32.7	3.27 uF
3.005E7	25,000.	1202.	509.	50.9 uF

The first-stage AC gain was selected to be 100. The minimum switched AC gain is 5 and the synchronous detector gain is 5. So the minimum total AC gain is 2500. The DC gain includes the gain in the integrator, R_2/R_1 .

The noise added to the signal by the AF section is also analyzed. The signal out of the detector is small, so the first amplifier should add low noise and have a high gain. From previous calculations. the detector output is expected to be 930 uV when balancing to the ambient load. The first-stage op-amp has a noise voltage rating of 6 nV/ $\sqrt{\text{Hz}}$ at 1 Hz. The noise voltage rating decreases with increasing frequency, so this is considered worst case. The lowest effective integration time is 1.037 sec for the 2 K sensitivity. This corresponds to a noise

bandwidth of 0.482 Hz. The largest added noise voltage is then approximately 4 nV, very much below the input level.

7.0 Temperature Control

The noise injection radiometer has good absolute accuracy when the temperature of the interior components are the same temperature as the reference load. The temperature of the components also need to be stable over time. Good temperature control and thermal isolation of the enclosure is needed to attain these requirements.

In radiometer design, it is easier to maintain a constant enclosure temperature using heat rather than cooling. Heating circuits require less volume and are more precisely controllable than refrigeration. With heating control, the enclosure temperature is maintained at a higher temperature than the expected exterior air temperature. The chosen interior temperature is 40°C.

7.1 Control Circuit

The type of heating control chosen is a proportional controller. There are six enclosure heating circuits mounted in the radiometer, with each element controlled

separately. The block diagram of each controlling circuit is shown in figure 7.1.

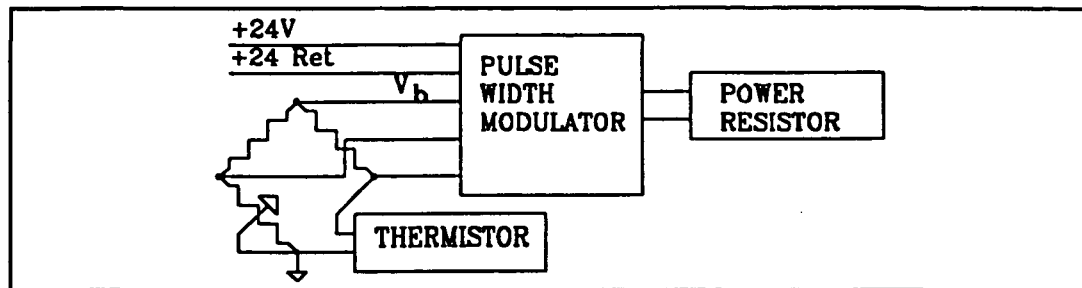


Figure 7.1 Bridge Sensor and Heating Circuit.

The thermistor is the heat-sensing element and the power resistor is the heat-delivering element. The thermistor completes a resistive bridge powered by a stable voltage V_b . The desired temperature of the thermistor is set by adjusting one of the bridge resistances. The difference between the midpoints of each side of the bridge is the error voltage. The error voltage controls the duty cycle of the pulse-width modulator (PWM). The resulting transfer function of the circuit is shown in figure 7.2. The X-Axis temperature in figure 7.2 is that at the thermistor. The limits on the duty cycle axis show that the heating element can be full off, 0 duty cycle, but not quite full on. This is a characteristic of the controller microcircuit. The transfer function shows that the duty cycle will vary linearly when the thermistor temperature is within $\pm 0.6^\circ\text{C}$ of the median temperature. When the duty cycle of a heating element is greater than 0 and less than 0.9, the heating unit is

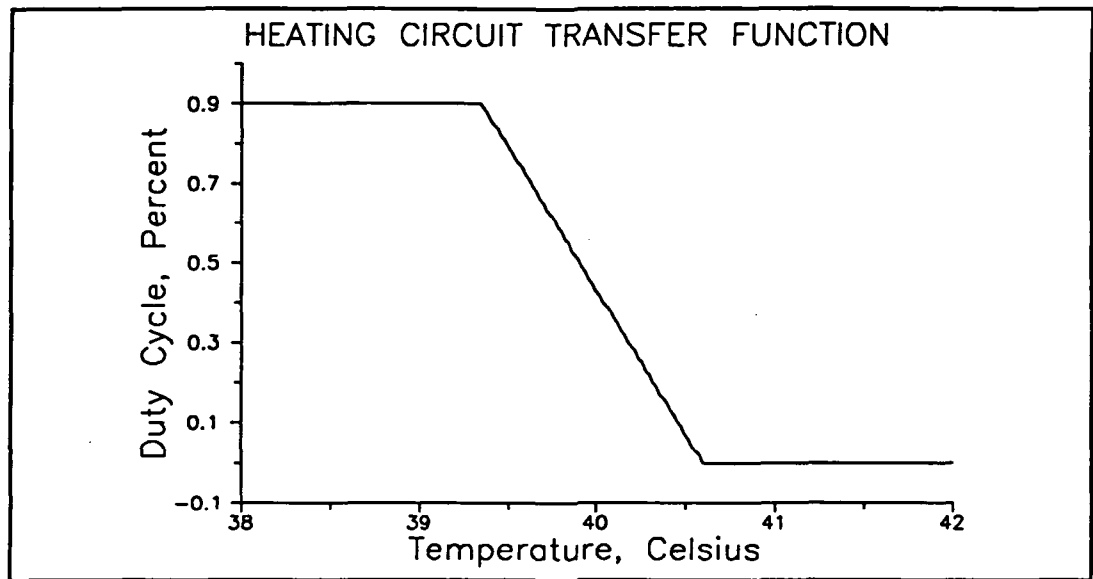


Figure 7.2 Heating Circuit Transfer Function.

"tracking." If the units are not tracking, then the temperature of the enclosure is uncontrolled.

The final temperature that each heating circuit attains is dependant upon its thermal load. The duty cycle is proportional to the amount of heat supplied. If the external environment is cold, much heat will need to be supplied and the thermistor temperature will stabilize below the median. If the external environment is warm, less heat is needed and the thermistor temperature will stabilize above the median. This is the way a proportional controller behaves.

The temperature control of the enclosure is a separate consideration than the control of each heating circuit.

The transfer function in figure 7.2 is only for each heating unit, not the whole enclosure.

7.2 Control Loop

The closed-loop heating diagram is shown in figure 7.3.

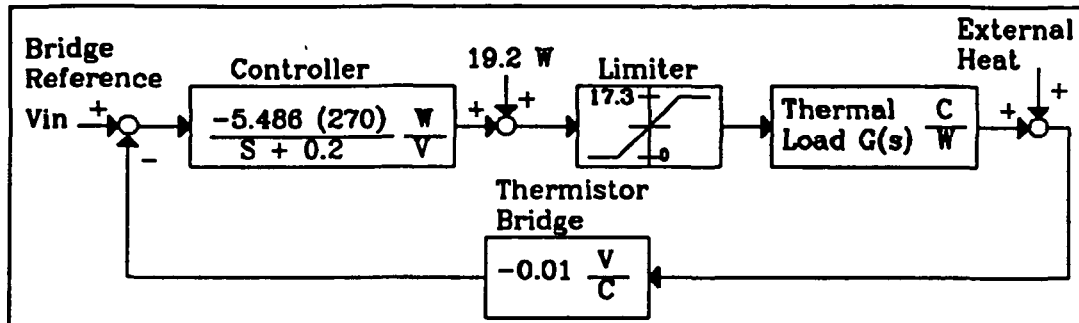


Figure 7.3 Heating Closed- Loop Diagram.

The closed loop is difficult to model and analyze. The thermal load characteristics change with temperature. The added heat depends upon the convection coefficient, which is difficult to determine. The limiter in the loop changes the problem from linear controls to a nonlinear control problem [14]. In general, nonlinear controls are much more difficult to analyze. One method, the graphical Phase Plane method, could be used to determine stability if the closed loop is second order or less [14].

Because of the modeling complexities, the stability of the heating loop was evaluated by experimentation. One of the problems encountered is accurate control at each

heat source vs. accurate control of the box interior temperature.

If each temperature sensor is placed very close to its heat source, accurate control of each heat source is possible due to the shorter time constant of the heat flow. In terms of the diagram of figure 7.3, the controller gain term (270) could be increased while maintaining stability. However, the time constant does have a minimum limit. The minimum limit is determined by the heater core (probably ceramic) and the thermistor (epoxy coated). In experimentation it was noticed that the thermistor time constant is noticeably greater than that of a thermocouple temperature probe. However, the controlled temperature is only at certain points on the enclosure. The larger the thermal load, the larger the gradients between heat sources and the enclosure interior.

On the other hand, if the heat sensor is placed far away from the heat source, control of each heat source will be poorer because the gain of the bridge error will have to be lowered to maintain stability with a long thermal time constant. However, the enclosure interior temperature will be better regulated with respect to change in thermal loads. The improved interior regulation makes

use of the small thermal gradients between source and sensor. The temperature of the source will be slightly greater than that of the sensor by an amount that depends on the thermal load. The greater the thermal load, the higher the difference, but yet the more constant the average temperature. Therefore, the enclosure interior temperature is expected to be closer to the sensor temperature.

The layout chosen for heating three of the enclosure's six sides is depicted in figure 7.4.

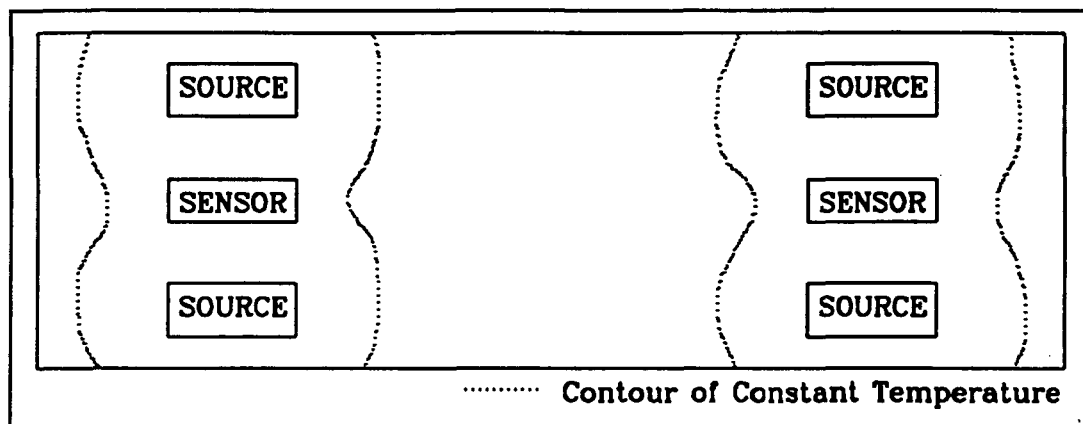


Figure 7.4 Panel Layout for Heating.

Two heat sources surround and are controlled by one heat sensor. The six heating units are separated enough so that they do not directly interfere with each other. For the four units mounted on the enclosure's two sides, the separation between heat source and sensor is two inches. For the two units mounted on the enclosure lid, which has a larger area, the separation is three inches. Stability

of the control loop was examined at one sensor as the box heated up. The control loop is underdamped and has a long settling time, in the order of a few minutes. However, the settling time improves as the entire box warms up. The long settling time of the control loop is acceptable because sudden changes in thermal loads are not expected.

A model of a single enclosure panel was developed to test the above judgment on heat source layout. Two-dimensional heat flow along a panel was modeled using finite-difference approximations. In this method, the steady state temperature at a point is the average of its surrounding temperatures, minus the amount of heat loss [12]. The temperature of each pair of resistors was adjusted so that the thermistor temperature was at the temperature determined by a heating transfer function. The heat loss at each grid point is the loss through $7/8$ inch of material with thermal conductivity 0.2 W/(m K) . The magnitude of the thermal gradients along the panel is determined by the ratio of the panel thermal conductivity to the insulation conductivity. To show the gradients better, the thermal conductivity of the metal panel was exaggerated down to 10 W/(m K) .

Two layouts were examined. The first had resistor pairs 2 inches apart from the sensor, and a heating transfer function ranging from 39.5°C to 40.5°C. The second layout consisted of the resistor pairs 0.5 inches from the sensor and a heating transfer function with a tighter range, from 39.6°C to 40.4°C. The temperature contours for an external temperature of 0°C are shown in figures 7.5 and 7.6.

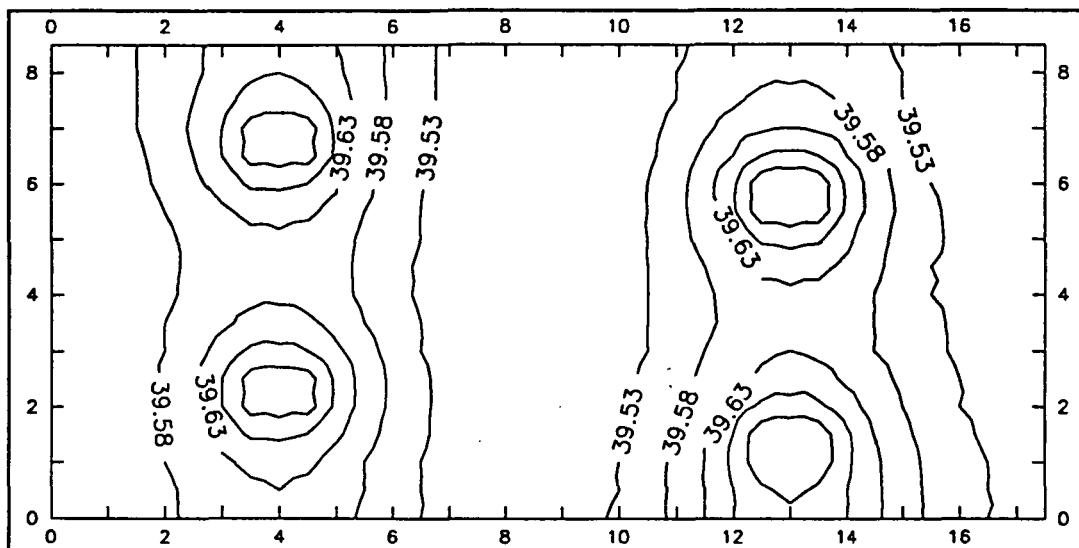


Figure 7.5 Steady State Temperature Gradients, Case 1.

Table 7.1 shows computed values from the model computations. The values show that variation of temperature over the whole panel is greater for the case 2 panel layout. The difference in temperature within the whole enclosure is expected to be greater than the difference on the panels. Therefore, the case 1 layout is better for uniform temperature. In addition, the

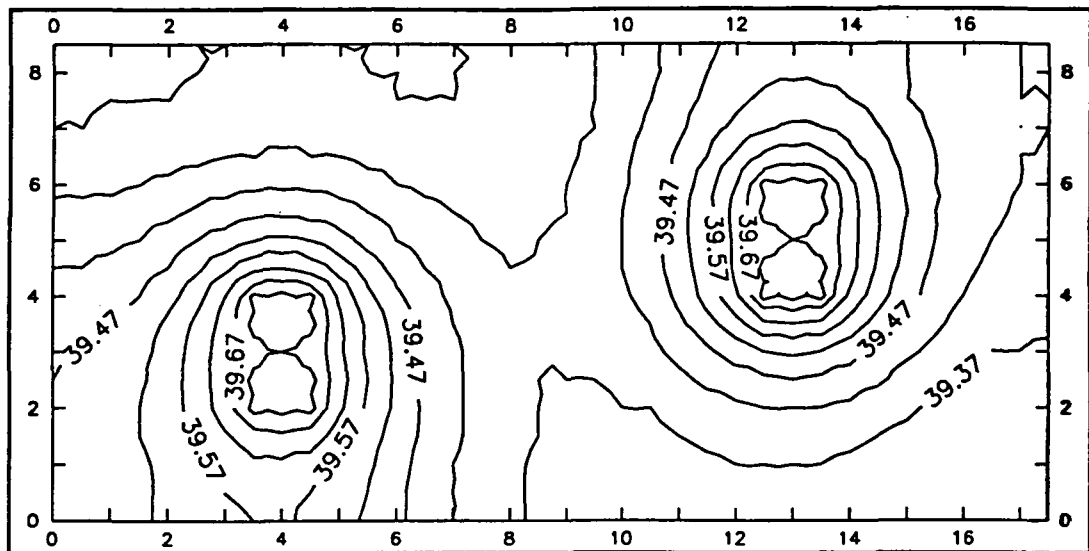


Figure 7.6 Steady State Temperature Gradients, Case 2.

Table 7.1 Modeled Heat Flow Comparison.

Single Panel Heat Flow Comparison for Two Layouts				
	External Temperature 0°C		External Temperature 30°C	
	Case 1	Case 2	Case 1	Case 2
Power Dissipated	34.528 Watts	34.419 Watts	8.954 Watts	8.861 Watts
Maximum - Minimum Temperature	0.30°C	0.43°C	0.08°C	0.12°C
Average Temperature	39.57°C	39.44 C	40.26°C	40.16°C

panel of case 1 will reach steady state quicker than the panel of case 2.

7.3 Range of Thermal Stability

It is desirable to know the range of external temperatures at which the radiometer can be operated. The range of operation is determined by the thermal conductivity of the enclosure and the power dissipated within the enclosure.

The minimum power dissipated in the enclosure is determined by the power consumed by the circuits. A DC ammeter with 0.1-amp resolution was used to measure the currents flowing into the radiometer under normal operation. Table 7.2 shows the computations for minimum power consumed.

Table 7.2 Minimum Power Dissipated as Heat.

Minimum Power Dissipated into Radiometer	
+18 V, 0.3 Amp	5.4 Watts
-18 V, 0.2 Amp	3.6 Watts
+10 V, 1.7 Amp	17 Watts
+28 V, 0.1 Amp	2.8 Watts
Internal Fan	12 Watts
Hot Loads, 2* 8.22W* 0.1 Duty	1.65 Watts
Total Minimum Power	42.5 Watts

The maximum power is the power consumed by the circuits added to the maximum power dissipated by the heating system. The maximum additional power is computed in table 7.3.

Table 7.3 Available Heating Power.

Maximum Additional Available Power for Heating	
Each Heater, 19.2W *0.9 Duty	17.3 Watts
6 Heaters Inside Enclosure	x 6
Total Maximum Additional Power	103.7 Watts

The thermal conductivity of the enclosure is approximated by using observations. When the radiometer is operating at room temperature, the minimum dissipated power is a little too much for the heating system to maintain tracking. When the external fan is turned on, the duty cycle of the heating system goes from zero to approximately 20%. Therefore, it is estimated that with no external fan, the heating system will track up to 20°C as the external temperature. When operating at room temperature with the external fan turned on, the heating system operates at 20% duty cycle. From these approximations, the full range can be determined.

For determining the thermal conductivity with no external fan, the minimum power maintains a difference of 20°C.

$$k = \frac{42.5W \cdot 0.0222m}{20K \cdot 0.731m^2} = 0.065 \frac{W}{m \cdot K} \quad (7.1)$$

This effective thermal conductivity is an acceptable value because the conductivity of the foam is approximately 0.035. The wiring and waveguide act as good heat conductors and will raise the value of the conductivity. Using this value of conductivity, the lower operating range can be determined.

$$\Delta T = \frac{145W \cdot 0.022m}{0.065 \cdot 0.731m^2} = 67 \text{ K} \quad (7.2)$$

Therefore, the lower limit of operation is -23°C .

For determining the effective thermal conductivity with the external fan in operation, 20% added power maintains a difference of 17°C . The effective thermal conductivity is calculated in equation 7.3.

$$k = \frac{(42.5 + 0.2(104))W \cdot 0.0222m}{17K \cdot 0.731m^2} = 0.113 \frac{W}{m \cdot K} \quad (7.3)$$

Then the upper operating range is determined,

$$\Delta T = \frac{42.5W \cdot 0.0222m}{0.113 \cdot 0.731m^2} = 11.4 \text{ K} \quad (7.4)$$

as is the lower operating range.

$$\Delta T = \frac{145W \cdot 0.0222m}{0.113 \cdot 0.731m^2} = 39.0 \text{ K} \quad (7.5)$$

From these results, the estimated operating range is shown in table 7.4.

Table 7.4 Estimated External Temperature Operating Range.

Estimated External Temperature Operating Range	
Without External Fan	-27°C to +20°C
With External Fan Operating	+1°C to 28.6°C

7.4 Hot Reference Load

A diagram of the hot reference load is shown in figure 7.7.

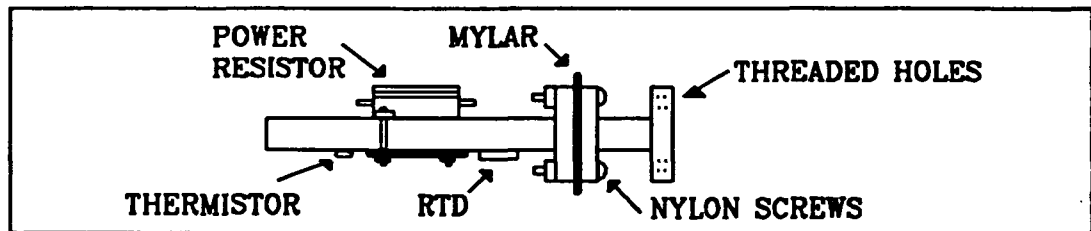


Figure 7.7 Hot Reference Load.

The components are mounted on a waveguide matched load. The same temperature control circuit used for heating the radiometer box is used for the hot loads (figure 7.1). The gain for the hot load control is reduced to 200. The lower gain gives the heating transfer function a slightly greater temperature span than shown in figure 7.2. However, the hot load's thermal load will not vary much because of the temperature-stabilized enclosure. Therefore, after warmup, the power delivered will be steady, resulting in a very steady hot load temperature.

It is desirable to keep the temperature of the hot load isolated from the rest of the system. The three elements of heat transfer are considered: convection, radiation, and conduction. To prevent exterior convection transfer, the waveguide load is enclosed in liquid-formed foam. The foam was trimmed after solidifying and wrapped with heavy aluminum tape. The aluminum shielding helps minimize transfer through radiation. A sheet of mylar and nylon screws are used to help prevent thermal conduction. The mylar has a 0.5-mm-diameter pinhole in the center to equalize interior air pressures. The mylar also helps to minimize transfer through interior convection.

Figure 7.8 shows the external physical connection to the hot load.

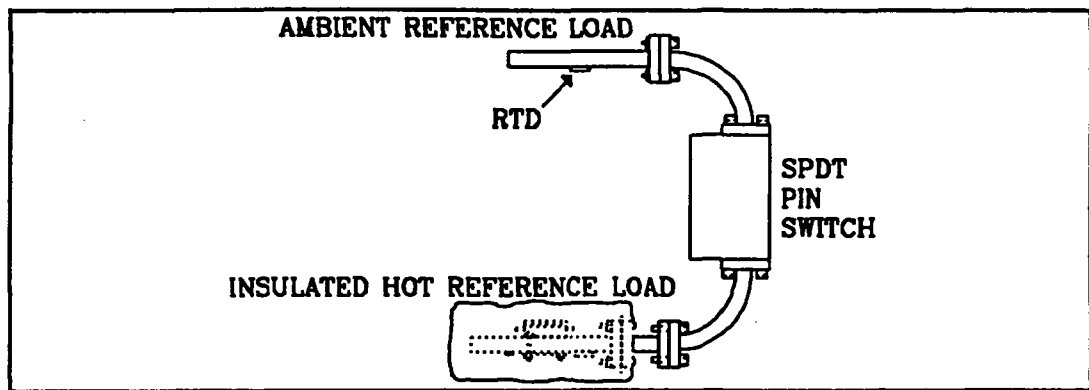


Figure 7.8 Reference Loads Assembly.

The air circulating around the connecting waveguide removes heat conducting out of the hot load. However, some heat may flow past the hot load bend. Therefore the SPDT PIN switch and its connecting waveguide may be at a higher temperature than the ambient air.

8.0 Temperature Sensing

The ideal case of stable and uniform temperature control will not be obtained, so selected temperatures are monitored to make corrections in the data. The temperature monitoring is separate from the temperature control. The block diagram of a single temperature-sensing circuit is shown in figure 8.1.

The temperature-sensing circuits make use of a resistive temperature device (RTD), which has better linearity than thermistors. The RTD driver circuit is effectively a

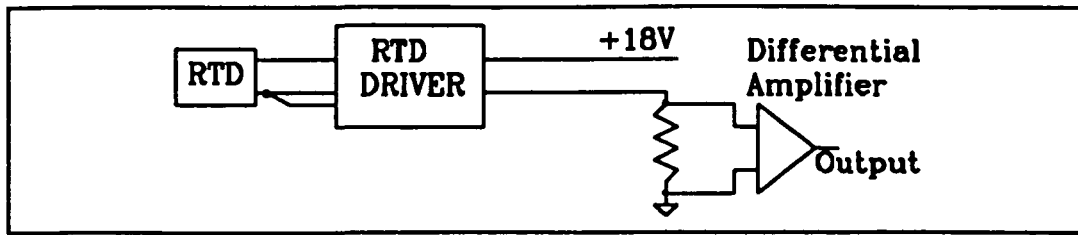


Figure 8.1 Temperature- Sensing Circuit.

resistance to current converter. The change in resistance of the RTD with temperature linearly corresponds to the change in current taken from the source voltage. A resistance in series with the supply voltage return generates a voltage proportional to the current. The instrumentation amplifier detects the voltage across the resistor. One advantage of using a current signal over a length of wire is that a change in line resistance does not change the signal. Switching cable length will not change the calibration.

The placement and range of the RTD's are shown in table 8.1.

The placement of the interior sensors was chosen by experimental results in the reference by Skou [13]. In the reference, interior temperature changes were forced with a heat blower. Output fluctuations in a NIR radiometer were found to be closely correlated with the reference load temperature and almost perfectly correlated with the noise source temperature. The external sensors are placed at the locations of

Table 8.1 RTD Placement.

RTD LOCATIONS AND TEMPERATURE RANGE			
RTD Num.	Location	Temperature Span, °C	Measurement Resolution, °C
1	V Ambient Reference Load	25 to 80	0.017
2	V Noise Source	25 to 80	0.017
3	V Hot Reference Load	25 to 80	0.017
4	Antenna	-50 to 50	0.03
5	H Ambient Reference Load	25 to 80	0.017
6	H Noise Source	25 to 80	0.017
7	H Hot Reference Load	25 to 80	0.017
8	Orthomode Transducer	-50 to 50	0.03

significant losses.

The temperature span of the interior RTD's are set the same for user convenience. The noise source and ambient reference RTD's are expected to be near 40°C. The hot load references are expected to be near 70°C. The measurement range and voltage offset was selected so that 40°C corresponds to zero volts output. Therefore, if the readings on the temperature LED display stabilize between -100 and +100, then the enclosure is stabilized between 38°C and 42°C.

The measurement resolution may appear to be slightly worse than expected for a 12-bit system. This is because

the listed temperature extremes generate a 16-volt difference, whereas the A/D handles a 20-volt difference. The missing 4-volt range, 820 bits, are not utilized.

9.0 Data Acquisition

The data storage medium is a rack-mount 80286-based computer with a analog-to-digital (A/D) sampling board. The sampling board has a maximum of 8 inputs of which 5 are used. The input channel assignments are listed in table 9.1.

Table 9.1 A/D Channel Assignments.

A/D CHANNEL ASSIGNMENT	
Channel	Usage
0	V Polarization
1	H Polarization
2	Optocoupled V Polarization
3	Optocoupled H Polarization
4	Multiplexed Temperature and Mode

The multiplexed data order is further explained in table 9.2. A marker is set into the multiplexed data so the data can always be properly aligned. The marker should always be 0 counts at the computer because it is a -15-volt input to an A/D board that accepts -10 volt as the lowest input.

Table 9.2 Multiplexed Data Order.

Multiplexed Data Order	
Number	Data
1	RTD 1
2	RTD 2
3	RTD 3
4	RTD 4
5	RTD 5
6	RTD 6
7	RTD 7
8	RTD 8
9	Mode Information
10	Marker, -10V, 0 Counts

The mode information is a discrete voltage, which indicates the radiometer reference load and integration time selected. The discrete levels are explained in table 9.3.

Table 9.3 Mode Information Levels.

MODE REFERENCE VOLTAGES			
Reference Load	Integration Time	Mode Voltage	Corresponding Count Range
Ambient	0.025 s	-1	1740 > C < 1944
Ambient	0.10 s	-2	1535 > C < 1739
Ambient	0.40 s	-3	1330 > C < 1534
Ambient	1.6 s	-4	1125 > C < 1329
Hot	0.025 s	+1	2150 > C < 2354
Hot	0.10 s	+2	2355 > C < 2560
Hot	0.40 s	+3	2561 > C < 2764
Hot	1.6 s	+4	2764 > C < 2969

10.0 System Noise Considerations

The accuracy and stability of the radiometer design theory is worthless if noise corrupts the system. The system consideration is designing the connections of the sub-assemblies in such a way that they work well together, and added noise is minimized.

The sub-assemblies are:

1. Radiometer Box.
2. Power/Control Box.
3. Computer.

Noise added to the radiometer outputs can be caused by 3 sources:

1. External fields picked up by the input lines.
2. Ground Loops.
3. Internally Generated Noise.

The design of the radiometer accounts for these factors to minimize the added noise.

10.1 External Fields

The only input signals to the radiometer are the power lines. Noise that these lines pick up could pass into the radiometer and contaminate the sensitive receiver circuits. The power wires are in a single bundle and thus present minimal loop area to any external magnetic fields. Therefore, noise picked up by magnetic induction will be minimal. Filters are placed at the input to the radiometer to attenuate high frequency noise picked up by external electric fields. A diagram of the filters is shown in figure 10.1, and photographs in figure 10.2.

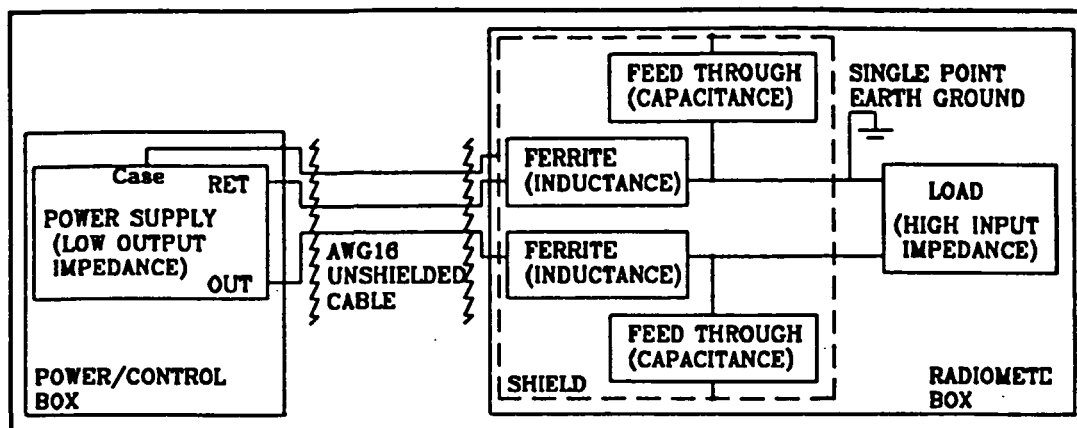


Figure 10.1 RFI Filter.



Figure 10.2 Photos, Filter Interior Left and Installed Right.

The filter rejects the higher frequencies by appearing as an impedance mismatch between the power supply and load. At high frequencies, the toroid appears as a large impedance and the capacitor as a low impedance. Therefore, at high frequencies, the toroid mismatches the supply's low output impedance and the capacitor

mismatches the load's higher impedance. Shielding is placed around the toroid ferrite-feedthrough capacitor combination so fields are not radiated inside the radiometer box.

10.2 Ground Loops

Ground loops occur when return power or a return signal has multiple paths to take. One effect of ground loops is that multiple paths present larger loop areas for magnetic induction pickup. Another problem is that the actual potential at the ground point is not as expected. Figure 10.3 shows the simplified grounding scheme for the radiometer. The power return lines are only connected at one point, which defines the common ground.

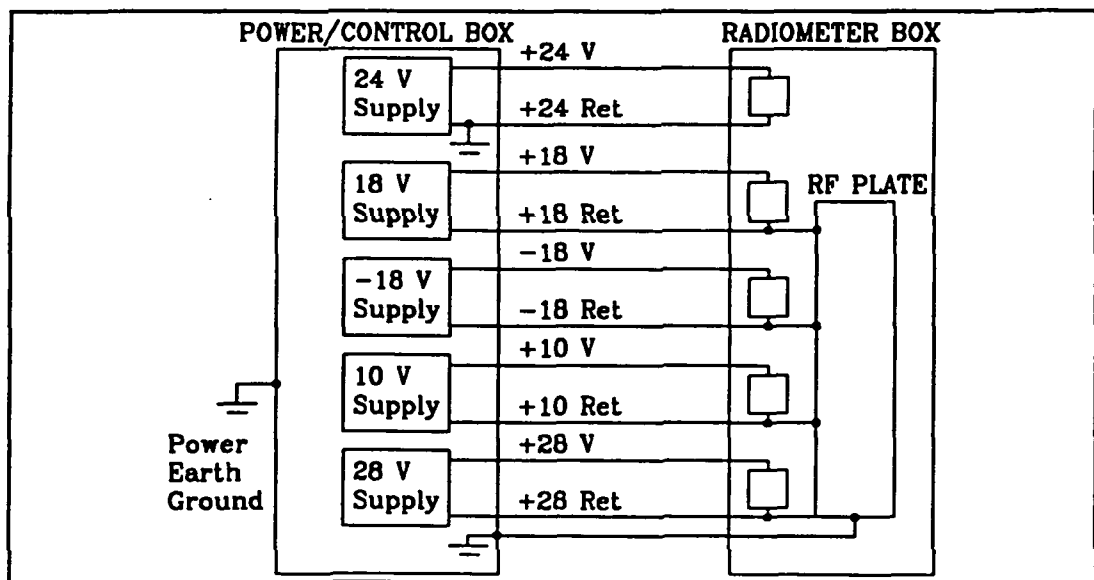


Figure 10.3 Simplified Grounding Scheme.

In addition, the ground potential in a supply return line will be higher at the radiometer than at the power/control box because of the $I \cdot R$ drop on the line. If another line between the boxes is grounded on both sides, D.C. current will flow on that path. The added path for return current increases the power line loop area, picking up magnetic induction noise.

Therefore, signal grounding between the boxes is also important. The polarization channel signals that travel from the radiometer to the power/control box, and then to the computer, have only one reference ground- at the radiometer box.

The frequency polarization signal requires a ground at the point where it is converted back to a DC level. An optoisolator is used to isolate the grounds as shown in

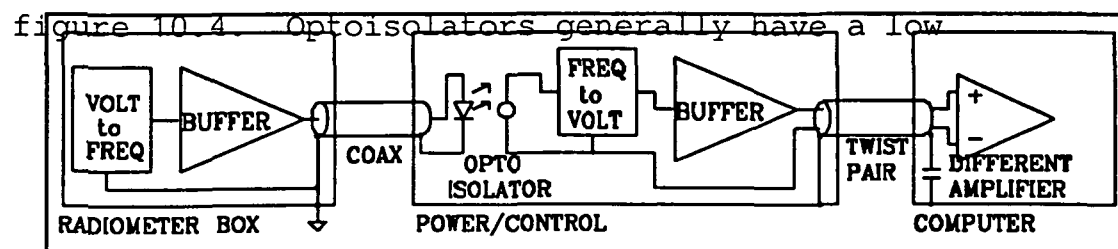


Figure 10.4 F/V Signal Isolation.

bandwidth, but recent advances enabled us to use one of 1 MHz bandwidth.

The RTD temperature signals obtain ground isolation in that each signal line is both a signal and a power line. The signal is a loop current that is not grounded inside the radiometer. This was previously shown in figure 8.1.

10.3 Internally Generated Noise

The audio frequency section of the radiometer is susceptible to picking up noise from one of the ma^X switching elements inside the radiometer. The noise can be picked up by electromagnetic waves and through power lines. Switching circuits create glitches on power supply lines. Switching is used for temperature control, noise diode modulation, synchronous detection, and polarization signal output.

The heating control power supply is dedicated for that function and does not connect to the ground used in the AF circuits. Therefore, its switching does not directly interfere with other supplies. However, it does switch large currents through the wires to the heating resistors, which could cause electromagnetic interference problems (EMI,RFI). Therefore, lengths of wire carrying the switched heating current that are longer than 4 inches are wired in twisted shielded pairs. The shielding is connected to earth grounds at both ends.

The noise diode modulator uses a pulse-width modulation (PWM) chip that puts undesired switching glitches on the power supply line. The glitches were reduced by slowing down the switching transitions and reducing the voltage span. The PIN diode driver has only one threshold value, so a slightly slower rise and fall time of its trigger does not degrade pulse width accuracy. The output driver of the PWM circuit is shown in figure 10.5.

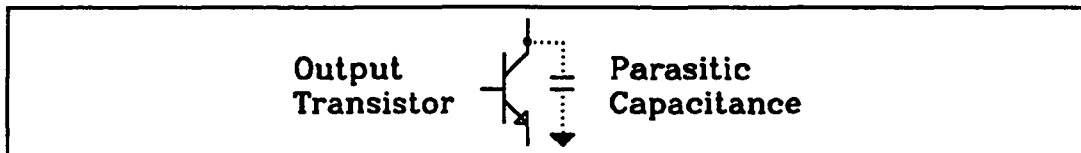


Figure 10.5 PWM Output Driver.

The parasitic capacitance is due to the high-current-handling transistor in the PWM, suited for switching power supplies. The best use of this device for high-speed switching is shown in figure 10.6.

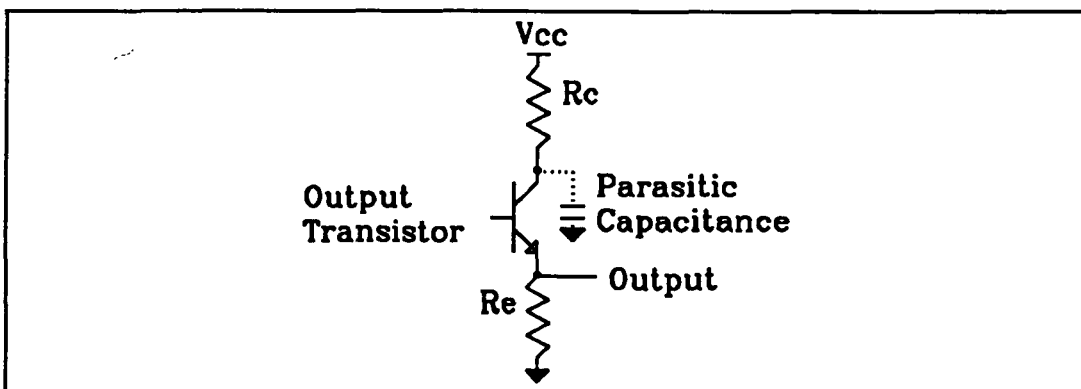


Figure 10.6 Implementation of PWM Output Circuit.

In this configuration, emitter resistance, R_e , limits the current drain from the parasitic capacitance, slowing down the turn on time. The collector resistance, R_c , limits the capacitance charge time, slowing down the turn-off time. The selected resistor combination sets the voltage swing at the emitter to 1 volt above and below the PIN driver threshold.

The synchronous detection circuit puts noise in the system in two ways. The first is by power line glitches (changes in voltage level) caused by the energy drawn to switch the latching ferrite circulators. The second is by radiation from chopper circuit control lines in proximity to the AF amplifiers. The ferrite switching glitches were reduced by the circuit shown in figure 10.7, placed on the connector to the latching circulator.

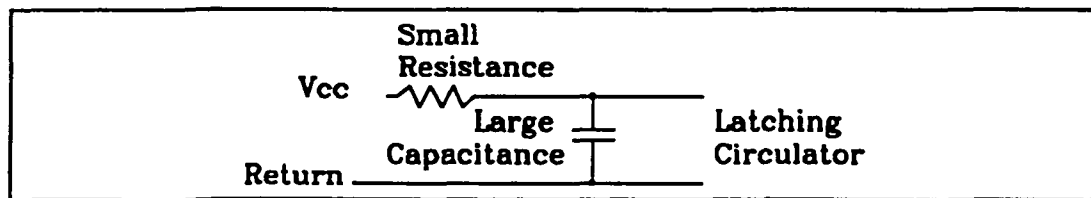


Figure 10.7 Latching Circulator Power Line Filter.

When the circulator switches, it draws the majority of its current from the large capacitor shown in figure 10.7. The capacitor recharges through the small resistor between current draws. This keeps the power supply line, which is shared with amplifiers, clear of big glitches. To minimize radiation by the synchronous detector control

lines, the wires are attached on the back of the circuit card with a grounded, shielded covering.

The voltage-to-frequency (v/f) circuit also has the potential for causing noise on the supply line. The v/f converter chip itself does not glitch the supply line much. However, the buffer that follows would normally glitch the supply line because it drives a low impedance optoisolator. The solution for reducing the wise is shown in figure 10.8.

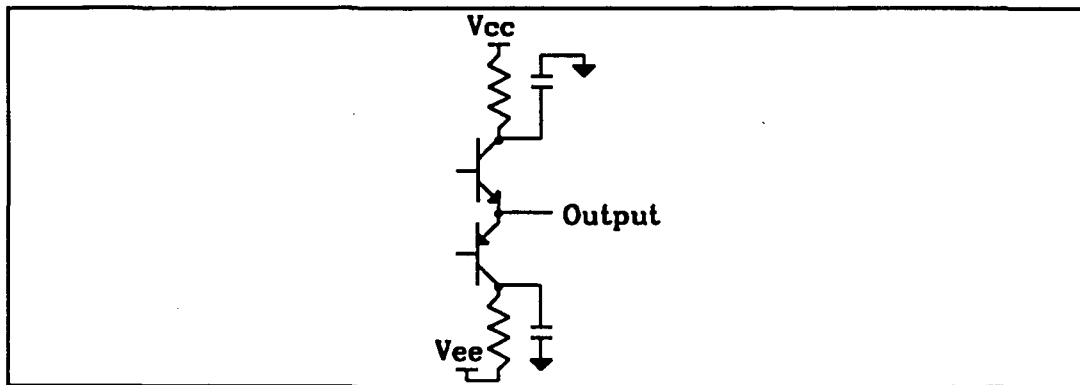


Figure 10.8 V/F Output Driver Circuit.

The resistors to the push-pull transistor configuration cause a drop in voltage to the buffer circuit when switching, but only a small glitch to the supply line. The voltage drop to the driver is acceptable because the output swing is 5 volts, much smaller than the supply voltage.

11.0 Radiometer Housing

The housing for the radiometer is important from structural and thermal viewpoints. The container must withstand transport and use, yet be accessible for repair. In addition, the thermal heat transport between the interior and exterior must be slow to facilitate uniform temperature control in the interior.

11.1 Exterior Radiometer Box

The radiometer housing consists of an interior and exterior box.

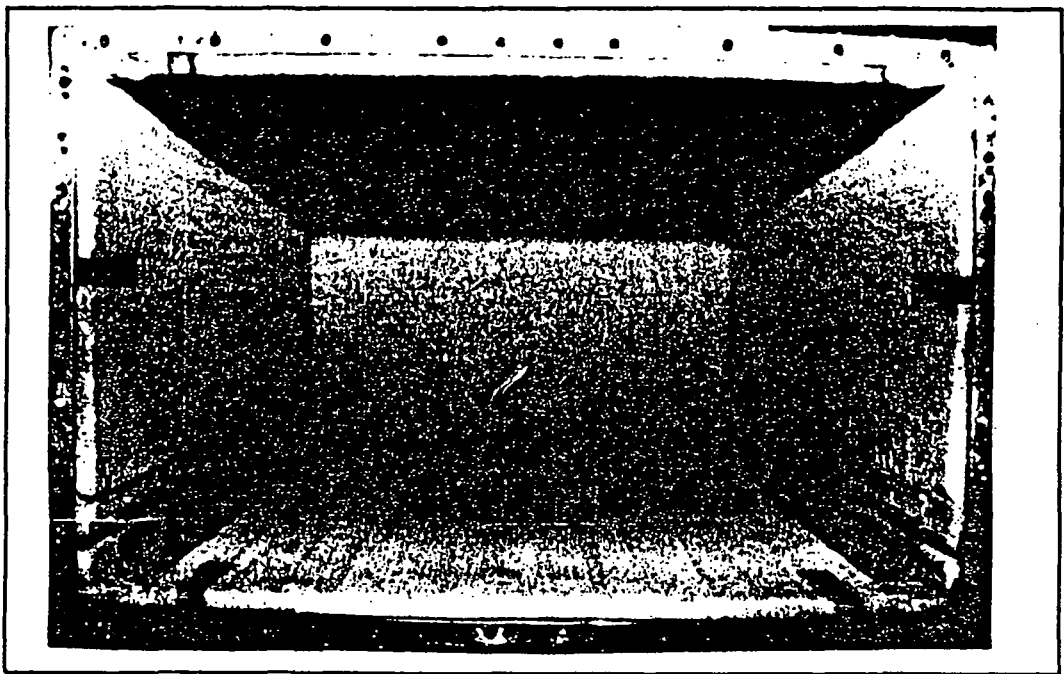


Figure 11.1 Photo, Exterior Radiometer Box.

The exterior container shown in figure 11.1 is a five-sided box with plastic female guide rails leading to the opening. Approximately 7/8-inch-thick styrofoam is adhered to each wall. At the opening there are guide

blocks on each wall to assist placement of the interior box into the exterior box. The guide blocks prevent the styrofoam near the opening from being crushed upon insertion of the inner into the outer box. The blocks are made from phenoly, which is a composite material of relatively low thermal coefficient.

11.2 Interior Radiometer Box

The interior container is a fully enclosed box. The box is shown in figure 11.2 with the front and top removed.

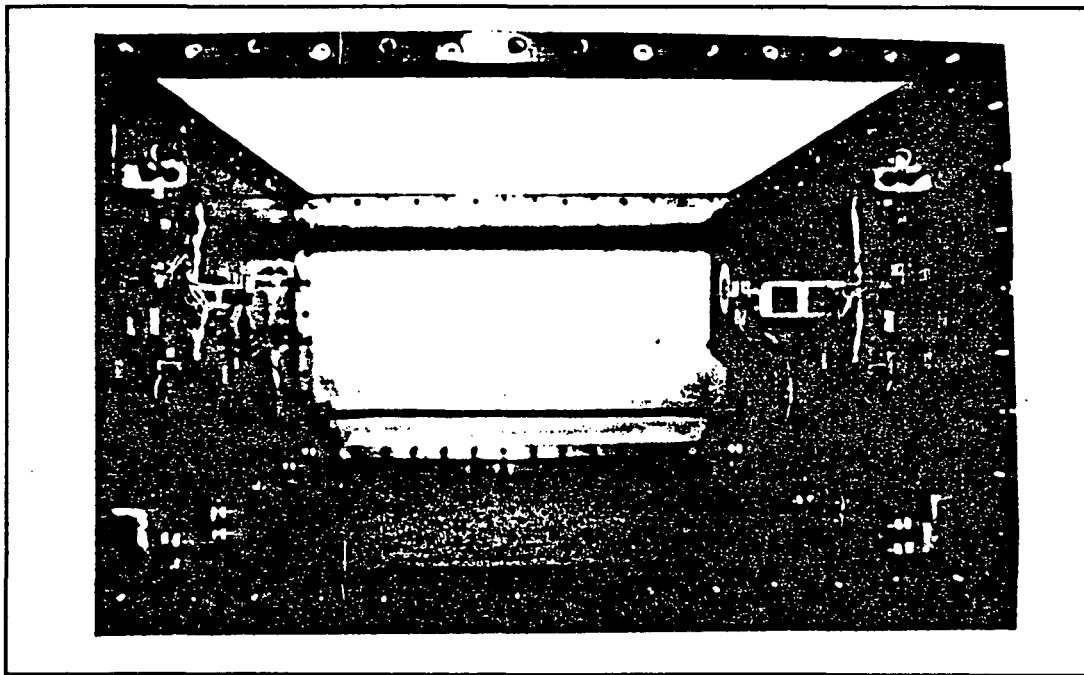


Figure 11.2 Photo, Interior Radiometer Box.

The front panel of the interior box shown in figure 11.3 incorporates the waveguide (H and V ports) and cable connections (P1, P2, and J3).

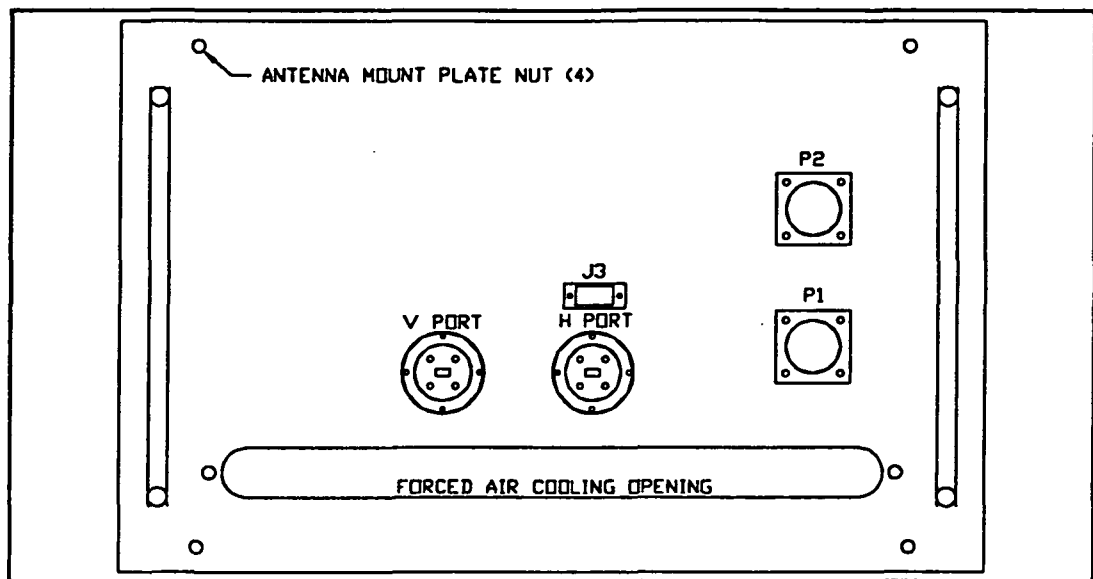


Figure 11.3 Radiometer Face Plate.

Therefore the front panel serves as the face of the exterior box as well. The front panel connects to the interior frame with 1-inch-thick phenoly. The front panel and interior frame are insulated from each other by the phenoly. The phenoly bolts to the frame, and the face plate separately bolts to the phenoly. One inch styrofoam adhered to the inside of the front face. Near the bottom of the front face there is a long oval cutout of 1-inch in height. This opening allows external air to be circulated underneath the interior base plate. The external air does not mix with the internal air; they are separated by a sheet of aluminum. The external air circulation increase the effective thermal conductivity of the enclosure.

The barrier between the external and internal air is the radiometer base plate. The circuit card holder, power supply cards, and RF plate supports are mounted to the base plate, which slides into the radiometer from the front. This 1/16-inch aluminum plate is supported at its sides along the full length.

The microwave components are mounted on a plate 1 inch above the base plate. This plate is supported by 1/2-inch-thick phenoly, which insulates it from the base plate. The phenoly supports are bolted to the base plate, but at the sides where the base plate itself is supported. Therefore there is no flexing of the base plate caused by the weight of the microwave components. The 1-inch height of the microwave component plate allows air to circulate between it and base plate. This microwave component plate is 3/32-inch-thick aluminum and spans 15 inches between supports. This thicker plate minimizes flexing during vibrations. Minimal flexing is required to limit shearing forces in the waveguide leading to the front plate.

The internal waveguide is semi-insulated from the bulkhead feedthrough by a sheet of mylar. Figure 11.4 shows the assembly with nylon screws.

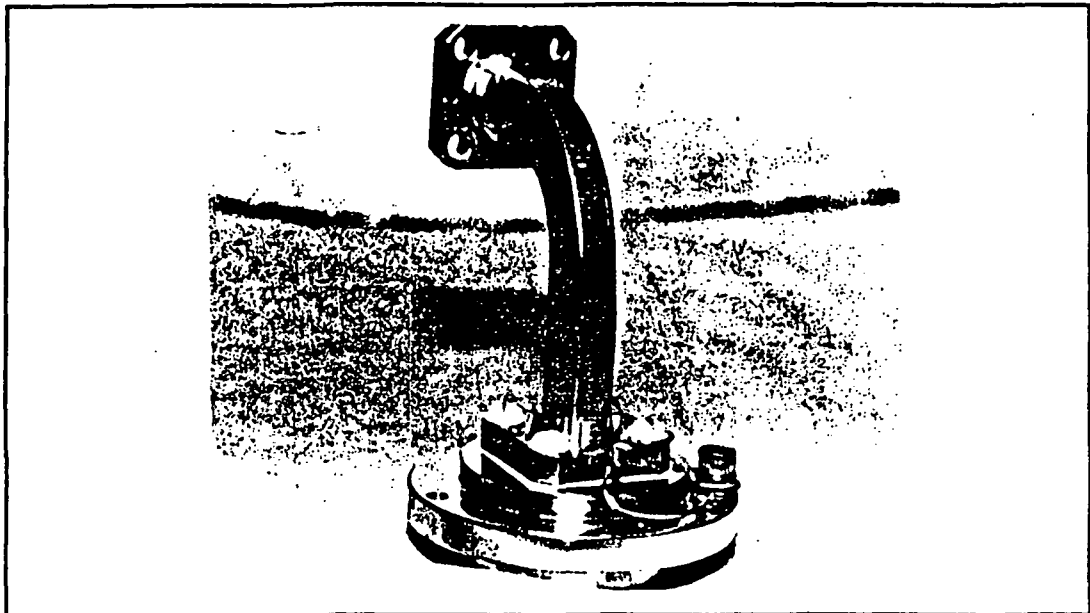


Figure 11.4 Photo, Bulkhead Waveguide Assembly.

Waveguide grounding continuity is maintained across the mylar by a thin two-inch long wire. The nylon screws were later replaced by custom-made steel screws. Steel screws were needed because of the shearing forces between the front plate and waveguide assembly during vibration. The modified steel screws are shown in figure 11.5.

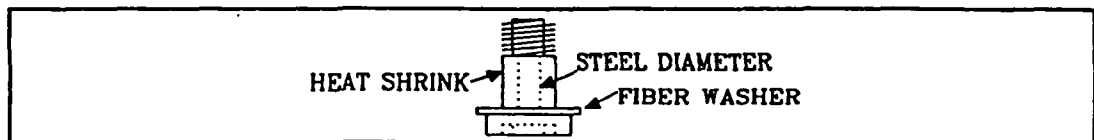


Figure 11.5 Insulated Screw.

The top half of the screw is filed down to a narrow cylinder. Heat shrink is placed on the reduced portion to bring the diameter back up to the size of the threads. Heat shrink is hollow, cylindrical plastic which reduces in diameter when hot air is blown on it. The cap of the bolt is separated from the waveguide by a fiber washer.

The two male slides are mounted on the interior box sides. The sides, in turn, are mounted to the interior box bottom frame by long steel hinges. The slides are mounted just above the hinges. The mating of a slide and plastic guide rail is shown in figure 11.6.

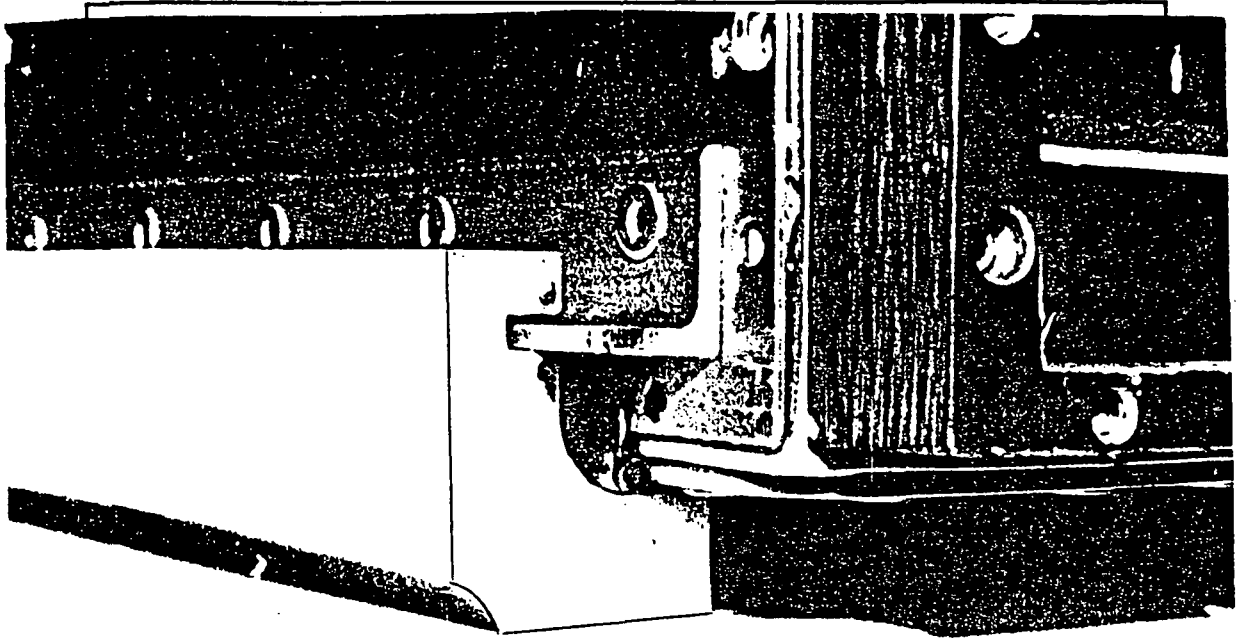


Figure 11.6 Photo, Mating of Guide.

The slide/guides are satisfactory to support the inner box weight. When the radiometer is in the upright position, the plastic guide also supports weight at the corner edge of the inner box. This can be seen in figure 11.6.

The sides and top of the inner box are fastened with screws. Receptacles for the screws are, plate nuts riveted into the frame. Plate nuts are also riveted to

the interior base plate to hold down the circuit card holder and power supplies. The plate nuts allow those items to be removed easily.

The circuit card holder is shown in figure 11.7.

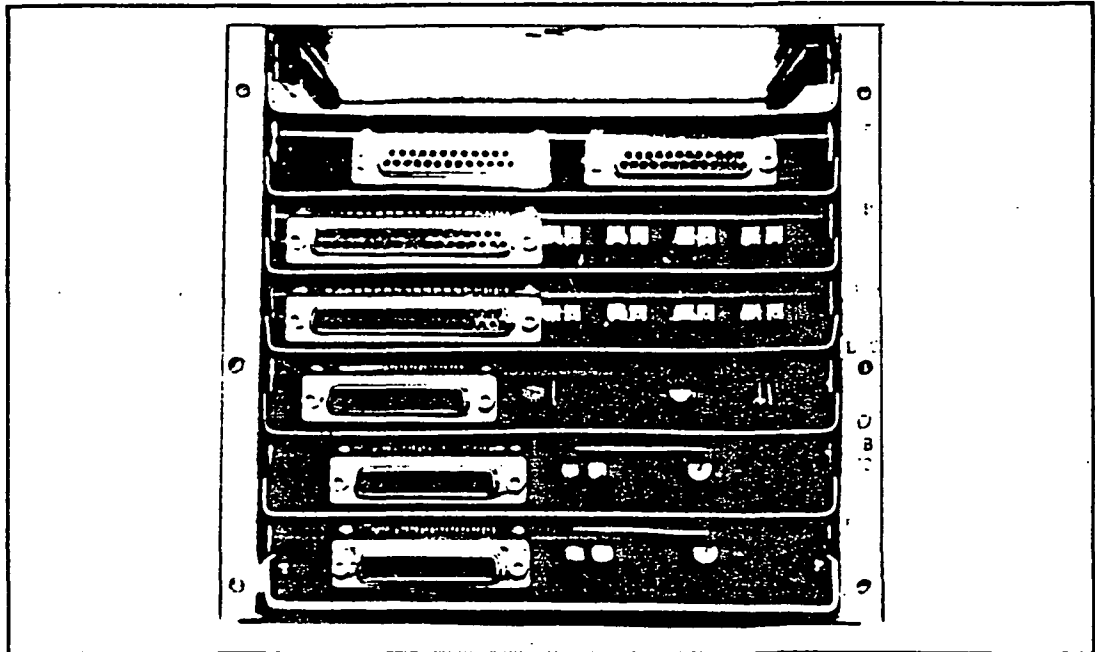


Figure 11.7 Photo, Circuit Card Holder.

Plastic edge guides hold the cards firmly in place. Sheet aluminum is placed between each card for RFI shielding. The edges of the sheets are bent to hold them firmly in place between the card edge guides. The sheets slide out for access to the bottom of the card holder enclosure. A lid with cutouts for the connectors is screwed into plate nuts on the top edge of the circuit card holder.

12.0 Calibration

12.1 Calibration of RTD's

The first step in calibrating the radiometer is calibrating the temperature sensors. Calculations of brightness temperature will be adjusted according to the temperature change of key microwave components. The temperature sensors are platinum-resistive temperature devices (RTD's). RTD's are used because they have a more linear response than thermistors.

The six RTD's used in the interior radiometer were calibrated in warm and hot environments. The two external RTD's were calibrated in warm, cold, and freezing environments. For the warm measurements, all eight RTD's were heat-sinked in a circular manner on a small sheet of aluminum. A RTD temperature probe with 0.01°C display resolution was affixed in the circle center. The assembly was placed in a small cardboard box with an opening to allow heat input, as shown in figure 12.1.

Figure 12.1 shows an earlier setup using an uncalibrated thermocouple probe at the center of the plate. A small heat blower was used to inject heat into the covered box. The data system was set up to record the RTD data. Temperature data was taken visually with the temperature

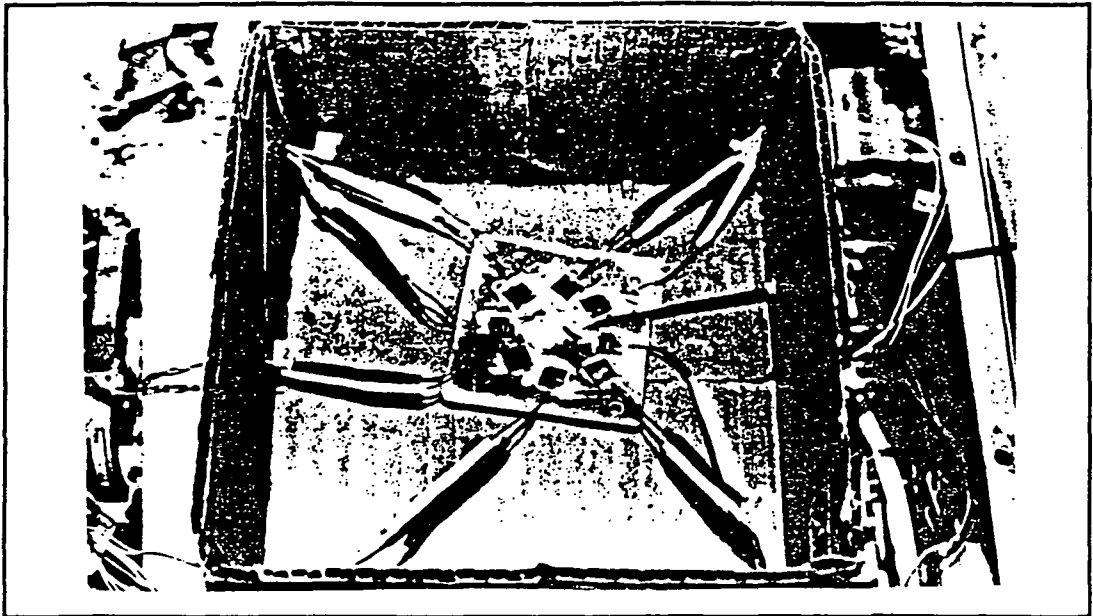


Figure 12.1 Photo, RTD Calibration Layout.

probe display. The procedure was as follows:

- Inject a small amount of heat into the box.
- Wait approximately 5 seconds for the internal metal components to reach their temperature apex. The temperature apex is the transition point between warming and cooling. The metal components will have a more uniform temperature at the apex.
- Record data.
- Repeat at a higher temperature.

The external RTD's were removed from the plate for measurements over 50°C. The six internal RTD's were removed from the plate for the cold and freezing measurements. A lab refrigerator and a freezer were used for those measurements.

The commercial software Grapher was used to perform the RTD equation fitting. The function of the fit is degrees Celsius, and the input variable is the computer count (0 to 4095). All interior RTD's were linearly fit after checking that only a small reduction in error was obtained with higher-order polynomials. The exterior RTD's were fit with second-order polynomials because they would be used over a wider temperature span. The resulting coefficients are presented in table 12.1.

Table 12.1 RTD Calibration Coefficients.

RTD Curve Fitting					
RTD Num	1st-Order Coeff.	2nd-Order Coeff.	Offset	Num. Data Points	Sum of Sq. Error
1	0.0207708		-2.14048	10	0.15
2	0.0208032		-2.36763	10	0.067
3	0.0205284		-2.21926	10	0.12
4	0.0288891	5.11548E-7	-74.6675	9	0.12
5	0.0207006		-2.06336	10	0.083
6	0.0206893		-1.80459	10	0.17
7	0.0205621		-2.06035	10	0.077
8	0.0288949	4.66599E-7	-75.137	9	0.13

12.2 Calibration of Microwave Component Losses

The second step in calibration is determining the actual losses of the microwave components. In the RF analysis

section, losses used in computations were the manufacturer's maximum specification. For actual measurements, accurate losses should be used in brightness temperature computations.

Loss and reflection coefficients were measured with a scalar network analyzer. The test setup is shown in

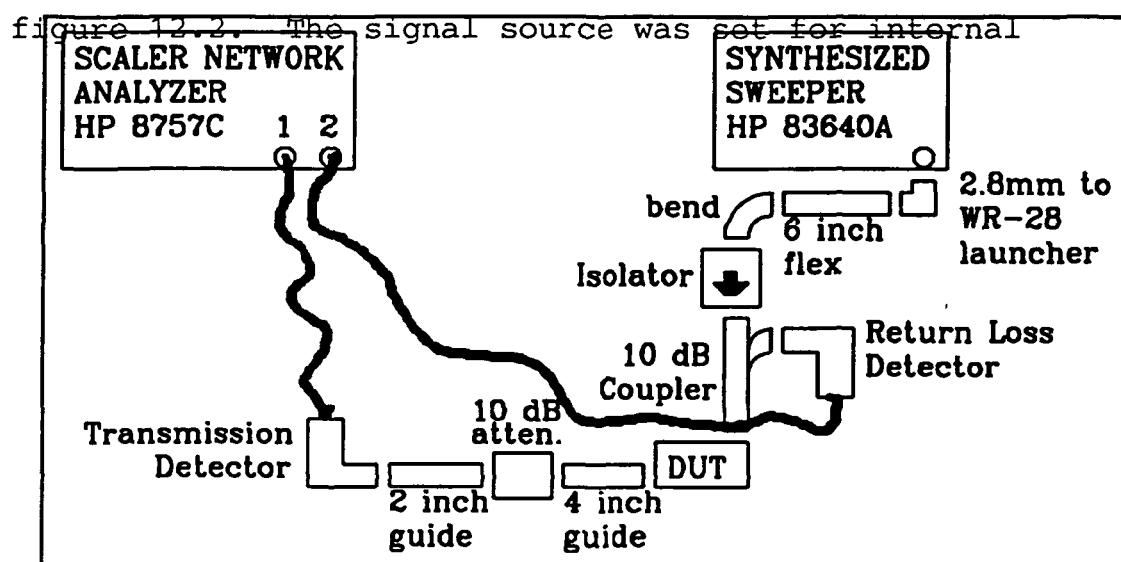


Figure 12.2 Scalar Network Analyzer Setup.

leveling and a sweep from 36 to 38 GHz in AC modulation mode. The source power was set so the power received at the detectors was in the square-law region, -50 dBm to -15 dBm. The source power was set at -10 dBm so the maximum power each detector received was -20 dBm.

The network analyzer setup was calibrated before measurements were performed. To calibrate the transmission detector, the device under test (DUT) was

removed and the power received in channel 1 over the desired frequency sweep was saved. To calibrate the return loss detector, a waveguide short was placed at the DUT end of the coupler and the detected power in channel 2 was saved.

For both detector calibrations, standing waves in the setup reduce measurement accuracies. The source and detectors must be well matched to reduce standing waves. The isolator helps to match the source, and 10-dB attenuation helps to match each detector. However, the increased number of connections increases the standing waves somewhat from the expected improved value. Standing waves are visible in the measurements as periodic changes in power, and they decrease the measurement accuracy. The measurement results in tables 12.2 and 12.13 are a visual average over the 1-GHz bandwidth of interest.

12.3 Spectrum Usage

The V channel oscillator was too unstable for both oscillators to be set to the same frequency. The V channel oscillator changed as much as 40 MHz at room temperature. As the box heated up, the drift apparently increased and caused interference between channels.

Table 12.2 Measured Insertion and Return Losses, H Polarization.

H Component Loss/Reflection Measurements		
ITEM	Loss	Return Loss
Antenna Connection	0.25 dB*	-21 dB
OMT H Port Insertion	0.14 dB	-17 dB
H Isolator Insertion	0.25 dB	-27 dB
H Isolator Isolation	28 dB	-22 dB
H Directional Coupler Insertion	0.25 dB	-18 dB
H Directional Coupler Isolation	20.5 dB	-38 dB
H SPDT PIN Sw., Amb. Ref. port, Insertion	1.17 dB	-18 dB
H SPDT PIN Sw., Amb. Ref. port, Isolation	29 dB	-7 dB
H SPDT PIN Sw., Hot Ref. port, Insertion	2.67 dB	-11 dB
H SPDT PIN Sw., Hot Ref. port, Isolation	27 dB	-1.11 dB
H Latch Circ., Antenna port, Insertion	0.30 dB	-25 dB
H Latch Circ., Antenna port, Isolation	28 dB	-26 dB
H Latch Circ., Ref. port, Insertion	0.36 dB	-30 dB
H Latch Circ., Ref. port, Isolation	26 dB	-31 dB
H Bulkhead to mixer input, Insertion	1.0 dB	

Determined from Noise Figure Measurements.

Therefore, the two oscillators were separated by 800 MHz. The resulting spectrum is shown in figure 12.3.

12.4 Noise Figure Calibration

Three noise figure measurements are desired.

- 1) Noise figure of each mixer/preamp.

Table 12.3 Measured Insertion and Return Losses, V Polarization.

V Component Loss/Reflection Measurements		
ITEM	Loss	Return Loss
V Isolator + 1.4-inch guide, Insertion	0.36 dB	-25 dB
V Isolator + 1.4-inch guide, Isolation	22 dB	-25 dB
V Directional Coupler Insertion	0.28	-19 dB
V Directional Coupler Isolation	20.4 dB	-37 dB
V SPDT PIN Sw., Amb. Ref.	2.36 dB	-14 dB
V SPDT PIN Sw., Amb. Ref. port, Isolation	26 dB	-20.5 dB
V SPDT PIN Sw., Hot Ref. port, Insertion	1.17 dB	-15 dB
V SPDT PIN Sw., Hot Ref. port, Isolation	28 dB	-6 dB
V Latch Circ., Antenna port, Insertion	0.28 dB	-27 dB
V Latch Circ., Antenna port, Isolation	20.8 dB	-28 dB
V Latch Circ., Ref. port, Insertion	0.33 dB	-25 dB
V Latch Circ., Ref. port, Isolation	13.5 dB	-25 dB
V Bulkhead to Mixer input, Insertion	1.0 dB	

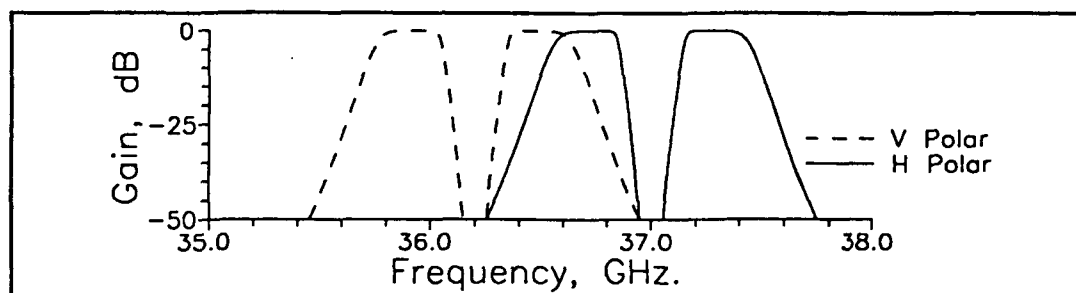


Figure 12.3 Receiver Spectrum Usage.

2) Noise figure of the radiometer from the waveguide bulkhead.

3) Noise figure of the radiometer with antenna installed.

The basic equation for determining the component or system noise figure uses the Y factor. The Y factor is the measured power ratio at the system output for two known temperatures presented at the input. The ratio can also be measured as a voltage ratio after a square-law detector. The Y-factor equation is:

$$Y_{factor} = \frac{P_H}{P_C} = \frac{T_h + T_{sys}}{T_c + T_{sys}} \quad (12.1)$$

The terms in equation 12.1 can be rearranged to form an equation for system noise temperature.

$$T_{sys} = \frac{T_h - YT_c}{Y - 1} \quad (12.2)$$

The noise figure, F, in dB, is calculated from the noise temperature as:

$$F_N = 10 \text{Log} \left(\frac{T_{sys}}{290} + 1 \right) \quad (12.3)$$

The test configuration for the mixer noise figure measurement is shown in figure 12.4.

The 6-dB attenuator and isolator are used to improve the match for the noise source. The loss of the isolator and

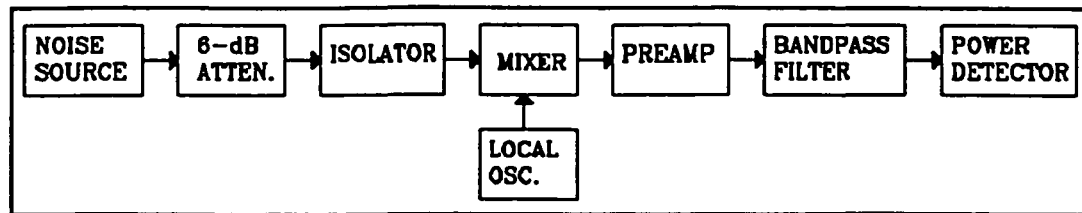


Figure 12.4 Mixer Noise Figure Measurement Setup.

attenuator were measured using the scalar analyzer. The hot temperature is calculated by:

$$T_h = 290 \left(10^{\frac{(ENR - Attn)}{10}} + 1 \right) \quad (12.4)$$

The cold temperature was room temperature and was produced by removing power to the noise source. The unpowered noise source produced the same thermal power as a substituted matched load.

The average ENR of the noise source is calculated in table 12.4 using the manufacturer's calibrations.

Table 12.4 Calculation of Average ENR for Source #2917.

Average ENR Determination	
Frequency	Manufacturer's Calibration
36.5 GHz	22.76 dB
37.0 GHz	23.29 dB
37.5 GHz	23.47 dB
1 GHz Bandwidth Avg.	23.17 dB

The combined attenuation of the isolator and attenuator was measured using the scalar network analyzer as 7.30 dB. Therefore, the hot temperature is:

$$T_h = 290 \left(10^{\frac{(22.17-7.30)}{10}} + 1 \right) = 11851.4 \text{ } ^\circ\text{K} \quad (12.5)$$

With the temperatures known, the noise figure is computed and shown in Table 12.5. Mixer Noise Measurement polarization channel

Mixer/Preamp Noise Measurement				
	Relative Power	Y Factor	Noise Temp	Noise Figure
H Pol	11.79 dB	15.10	494.97°K	4.324 dB

is measured using the noise source. The V-channel oscillator is currently offset from 37.0 GHz, and the noise source is not calibrated at its frequency range. The noise figure of each mixer is dependant on the

oscillator frequency, so measurements made at 37.0 GHz for the V mixer cannot be used.

The noise figure at the bulkhead of the radiometer was then measured. The setup is shown in figure 12.5.

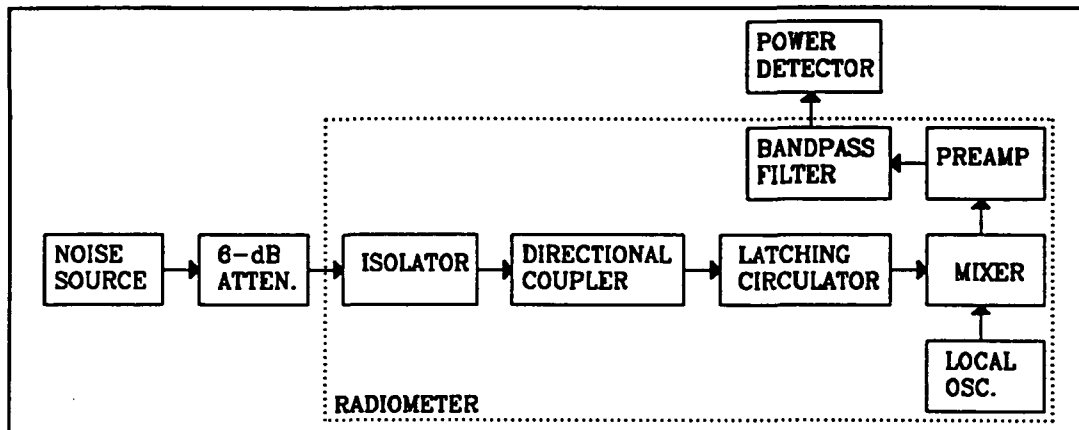


Figure 12.5 Noise Figure Measurement from Bulkhead, Setup.

The 6-dB attenuator was measured as 6.6 dB. The hot temperature reference temperature was calculated to be 13454.3°K. The resulting radiometer noise temperature is

Table 12.6 Noise Figure Measurements from Bulkhead.

Radiometer Noise Measurement from Bulkhead				
	Relative Power	Y Factor	Noise Temp	Noise Figure
H Pol	11.50 dB	14.125	703.3 °K	5.347 dB

appears correct because it is approximately 1 dB greater than the noise figure of the mixer alone. This is the same as the measured 1.0-dB loss between the bulkhead and mixer input (table 12.2). Losses before the preamp add directly to the noise figure.

The noise figure of the complete system was then determined with the antenna installed. The measurement setup is shown in figure 12.6.

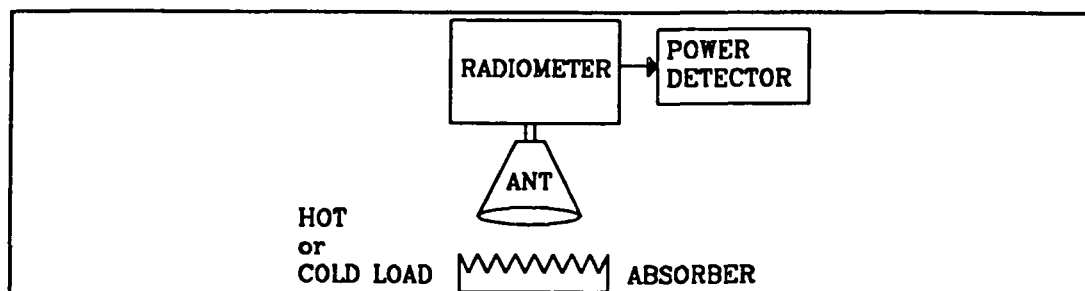


Figure 12.6 System Noise Temperature Measurement, Setup.

The hot load was microwave absorber at room temperature, and the cold load was absorber soaked in liquid nitrogen, LN₂. LN₂ has a boiling point of 77 K. The actual loads are shown in figure 12.7. The cold load pot was later wrapped in two inches of foam to greatly reduce the time

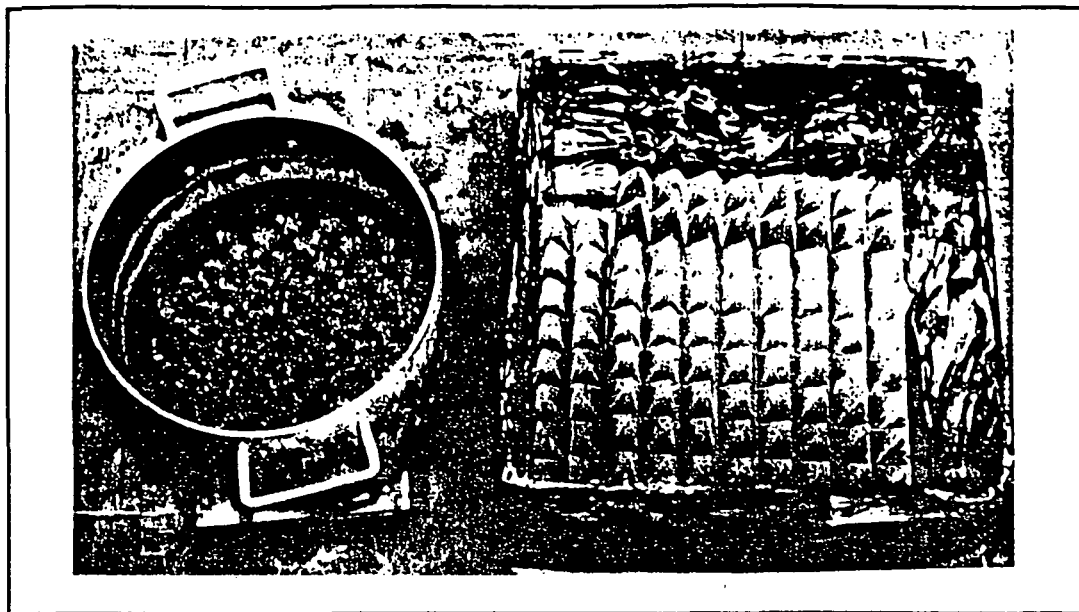


Figure 12.7 Photo, Cold Load in Pot, Warm Load in Box.

in which LN_2 boiled away. The noise figures are shown in table 12.7

Table 12.7 System Noise Temperature Measurements.

Complete Radiometer Noise Temperature Measurement				
	Relative Power	Y Factor	T_{sys}	Noise Figure
H Pol	0.95 dB	1.244	814.5 °K	5.808 dB
V Pol	0.79 dB	1.199	1015. °K	6.535 dB

The antenna loss can be determined from the increase in the H polarization noise figure between tables 12.6 and 12.7. The combined loss of OMT, waveguide, and antenna is the noise figure difference, 0.461 dB. The OMT H loss was measured to be 0.15 dB (table 12.2), and two inches

of waveguide are 0.053-dB Loss. Thus the antenna loss is 0.25 dB.

The noise figure for the V-polarization mixer can be determined from the V-system noise figure. It is assumed that the OMT and antenna present equal losses for each polarization. The loss between the bulkhead and mixer input was measured to be the same for each polarization. The estimated V-channel noise figures are shown in table 12.8

Table 12.8 V Channel Estimated Noise Figures.

V-Polarization Noise Figures Determined from T_{sys}		
Location	Noise Figure	Noise Temperature
From Bulkhead	6.074 dB	884.3 °K
Of Mixer	5.051 dB	637.9 °K

The V-polarization mixer is capable of a better noise figure with a 37-GHz oscillator. The mixer noise figure at that frequency was measured as 4.314 dB, slightly better than the H-polarization mixer. However, the mixer would need to be retuned to achieve that noise figure at 37 GHz.

12.5 Recomputation of Loop Time Constants

With component losses and the system noise temperature better known, the loop values estimated in tables 6.1 and 6.2 were recomputed. The H-channel system noise temperature was measured as 815 K for the truck-mounted antenna connection. The system noise temperature for aircraft installment was estimated as 850 K for another half-foot of waveguide to the antenna. The calculations were started with desired closed loop integration times of 0.025, 0.10, 0.40 and 1.6 seconds. The feedback gain constant, K_2 , was re-estimated as 28.9 K/V. The recomputed values are shown in tables 12.9 and 12.10.

Table 12.9 Recomputed Feedback Loop Constants.

Recomputed Loop Constants				
Closed-Loop Integration Time	Sensitivity for $T_{sys}=850$ K	K_1	AC*DC Gain	Open-Loop Time Constant
0.025 sec	0.82 K	6.1	6.874E6	4.43 s
0.10 sec	0.41 K	12.2	13.75E6	35.36 s
0.4 sec	0.21 K	24.3	27.38E6	281.3 s
1.6 sec	0.10 K	49.0	55.22E6	2267. s

Table 12.10 Recomputed Loop Gain and RC Values.

Recomputed Gain Budget and Integrator Values				
AC*DC Gain	Total AC Gain	Total DC Gain	$R_2 * C$ Product	C, for $R_2 = 10 \text{ M}$
6.874E6	1000	6874	2.21	0.22 μF
13.75E6	2000	6874	17.68	1.76 μF
27.38E6	4000	6874	140.65	14.0 μF
55.22E6	8000	6874	1133.5	113 μF

12.6 Radiometer Sensitivity Calibration

The sensitivity of the radiometer is the smallest increment in scene temperature that would produce a discernable change in the output. The discernable amount of change at the output is defined by convention as "A dc-output voltage equal to the effective value of the output fluctuations due to the system noise" [16 p.931]. In simpler terms, the sensitivity is defined as " ΔT is the root mean square temperature fluctuation of the output meter" [3 p.275]. Root mean square is the same as the standard deviation [11].

The sensitivity of the radiometer was measured by taking many samples of the output for a constant temperature scene. One thousand samples were taken over 10 seconds. Both warm and cold temperature scene were used to

determine the relation between computer counts and radiometric temperature. The relation was used to convert the output statistics from terms of computer counts to units of degrees Kelvin.

Table 12.11 Mean Radiometer Counts For Constant Loads.

Mean Counts For Constant Scene			
Scene Temp.	Integ. Time	V Pol Mean, Counts	H Pol Mean, Counts
281 K	0.025 s	3716.18	3744.10
77 K	0.025 s	1137.27	1160.58
281 K	0.10 s	3748.59	3755.35
77 K	0.10 s	1149.70	1152.28
281 K	0.4 sec	3766.17	3767.93
77 K	0.4 sec	1183.91	1172.17
281 K	1.6 sec	3770.54	3775.69
77 K	1.6 sec	1225.27	1195.99

The mean counts in table 12.11 were used to determine the relation between temperature difference in Kelvin and computer counts. The resulting measured radiometer sensitivities are shown in table 12.12.

Table 12.12 Measured Radiometer Sensitivity.

Measured Radiometer Sensitivity for Warm Scene					
Integration Time	Pol	Standard deviation, Counts	K/Count Relation	Measured Sensitivity, °K	Calculated Sensitivity, °K
0.025s	V	9.23	0.07900	0.730	0.94
0.025s	H	7.66	0.07897	0.605	0.80
0.10 s	V	6.04	0.07849	0.474	0.47
0.10 s	H	5.21	0.07837	0.408	0.40
0.40 s	V	4.25	0.07900	0.3355	0.23
0.40 s	H	3.99	0.07859	0.3135	0.20
1.6 s	V	4.014	0.08015	0.320	0.12
1.6 s	H	3.56	0.07910	0.2815	0.10

12.7 Radiometer Response Calibration

The radiometer response to scene brightness temperature was calculated using the mean values from table 12.11. The radiometer response was expected to be linear to changes of scene brightness temperature. The best way to calibrate the system was with hot and cold loads placed in front of the antenna. A good estimate can also be obtained by using a constant scene temperature and changing the radiometer reference between the hot and warm loads.

For each integration time of the radiometer, four sets of samples were recorded. The four sets consisted of:

- warm scene with a warm and hot reference.
- cold scene with a warm and hot reference.

From the mean of these measurements (table 12.11), four slopes of the radiometer response were calculated. If the response is linear, the four slopes should fairly well match.

Using the measured component losses, calibrated temperatures, and known scene temperature, the effective scene temperature into the directional coupler and the matching temperature at the output of the directional coupler can be computed as shown in sections 5.2, 5.3, and 5.4. The required amount of injected noise is then calculated and calibrated with the polarization output signal.

These calculations were run and computations of change in temperature per change in polarization output signal were computed. Tables 12.13 and 12.14 show the results.

The change in temperature is the difference between the scene temperature at the input to the directional coupler and the temperature to be matched at the output of the directional coupler.

Table 12.13 H-Channel Radiometer Transfer Slopes.

Change in Temperature/Change in H-Polarization Signal				
Inte- gration Time, Seconds	Warm Scene	Warm Ref.	Cold Scene	Hot Ref.
	Warm Ref vs Hot Ref.	Warm Scene vs Hot Scene.	Warm Ref vs Hot Ref.	Warm Scene vs Hot Scene
0.025	-0.06396	-0.06542	-0.06245	-0.06526
0.10	-0.06442	-0.06493	-0.06287	-0.06476
0.40	-0.06531	-0.06511	-0.06116	-0.06565
1.6	-0.06562	-0.06552	-0.06296	-0.06523

Table 12.14 V-Channel Radiometer Transfer Slopes.

Change in Temperature/Change in V-Polarization Signal				
Integra- tion Time, Seconds	Warm Scene	Warm Ref.	Cold Scene	Hot Ref.
	Warm Ref vs Hot Ref.	Warm Scene vs Hot Scene	Warm Ref vs Hot Ref.	Warm Scene vs Hot Scene
0.025	-0.06465	-0.06435	-0.06320	-0.06419
0.10	-0.06513	-0.06388	-0.06365	-0.06372
0.40	-0.06612	-0.06405	-0.06188	-0.06359
1.6	-0.06638	-0.06445	-0.06371	-0.06417

The temperatures of the losses used in the tables 12.13 and 12.14 computations are listed in table 12.15. The difference in the SPDT PIN switch temperatures shown in table 12.15 is because V-channel PIN switch is physically

Table 12.15 Temperatures used in Transfer Slope
Computations.

Temperatures of Losses Used for Radiometer Response		
Loss Items	H Channel	V Channel
Antenna	Ant RTD	Ant RTD
OMT and Exterior Waveguide	OMT RTD	OMT RTD
Isolator and Waveguide to Directional Coupler	Noise Source RTD	Noise Source RTD
Directional Coupler and Waveguide to Circulator	Noise Source RTD	Noise Source RTD
Latching Circulator	Ambient Ref. RTD	Ambient Ref. RTD
SPDT Ref. Switch	$(T_{hot} - T_{amb}) / 2.5 + T_{amb}$	$(T_{hot} - T_{amb}) / 1.5 + T_{amb}$

located further from the circulating air fan. Therefore more of the heat from the V-channel hot reference load will be conducted through the waveguide to the PIN switch. The temperature assignments in table 12.15 are good approximations.

During radiometer use, the amount of injected noise per polarization output signal should also be computed. Changes in power delivered by the noise source can then be taken into account. The noise source output power is expected to vary with the device temperature. The ENR rating is measured at room temperature. The manufacturer estimates that the noise power for our bandwidth will

change 0.005 dB per degree Celsius. The maximum specification on the sources is 0.01 dB per degree Celsius.

A quick look at the noise source temperature dependence was performed when noise source #2917 was set up for mixer noise figure measurements (figure 12.4). Using a heat blower to vary the case temperature over a range of 30°C, the power out of the preamp varied by 0.40 dB. However, it is difficult to determine how much of the change was due to the noise source and how much, to drift in other components (power meter). Therefore the manufacturer's suggested variation should be used in computations.

13.0 Conclusion

The radiometer was shown to have a measured sensitivity of 0.5 K for aircraft use, and down to 0.3 K sensitivity for truck-mount use. The radiometer was also shown to have good linearity throughout its range of detection. The range of detection is ranges from a low of 20 K to the temperature of the selected reference load. The noise injection principle and thermally controlled enclosure enable the radiometer to achieve good absolute stability. Temperature sensors have been calibrated so

small enclosure temperature changes can be accounted for. Problems with one microwave oscillator prevented each of the two polarization channels to detect the same bandwidth. However, the radiometric physics are, for all practical purposes, the same for the similar wavelengths. The operation of the radiometer, power/control box, and data acquisition system were closely monitored and the system is judged to be reliable and of scientific value.

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Appendix

A.1 Radiometer Equipment Setup

1. Assemble the following components:
 - a. Radiometer
 - b. Power/Control Box
 - c. Radiometer Cable (2 circular connectors at each end)
 - d. Computer Cable (Circular connector to 37 pin 'D' connector)
 - e. Rack mount computer
 - f. Computer video display
 - g. Computer and radiometer power cords
2. Connect the radiometer cable between the radiometer and the power/control box. Connect the computer cable between the power/control box and the computer. Connect the video monitor to the computer. Install power cables to the computer and power/control box
3. Turn on the power/control box. The radiometer will slowly begin warming.
4. Monitor the radiometer temperature using the temperature LED display and its rotary select switch.

The radiometer is ready to use when the display stabilizes between approximately +100 and -100 for selected temperatures 1,2,5 and 6. The approximate time for the warmup is one hour.

5. Calibrate system with warm and cold loads in front of the antenna.

A.2 Recording Radiometric Measurements

1. Turn on the computer.
2. Change to the program directory (type "cd program
<enter>").
3. Run the data acquisition program (type "37das
<enter>").
4. Select desired data recording parameters using the 'F' keys.
5. Hit 'F9' key. Continue depressing <enter> to begin data acquisition.
6. Observe Displays during use to check proper data storage.

A.3 Additional Photographs of System

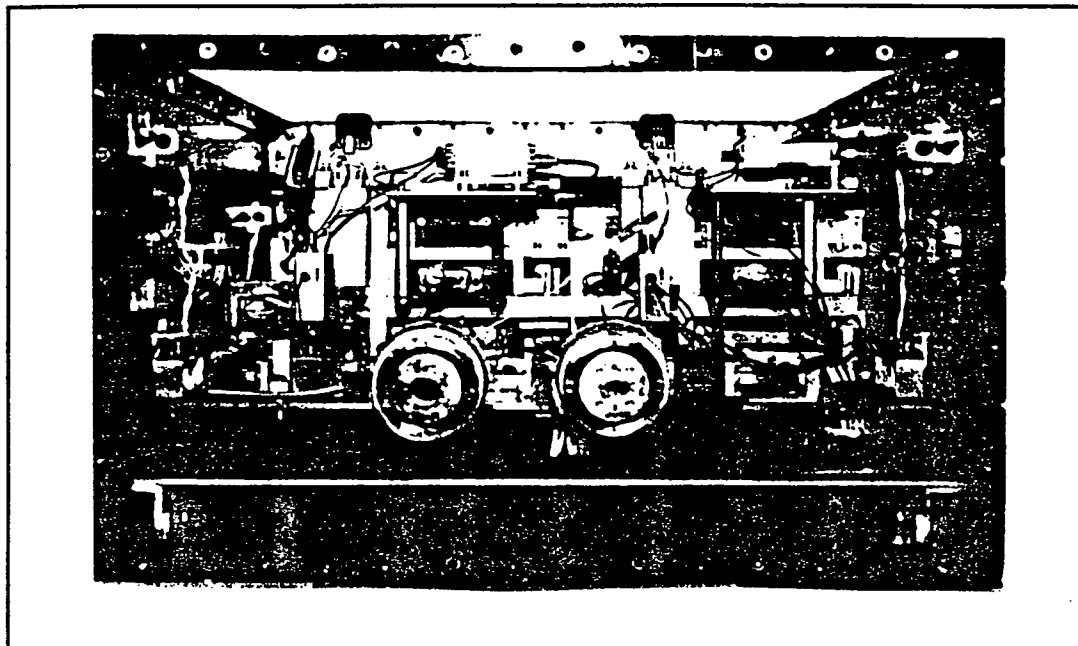


Figure A.1 Photo, Front View into Radiometer.

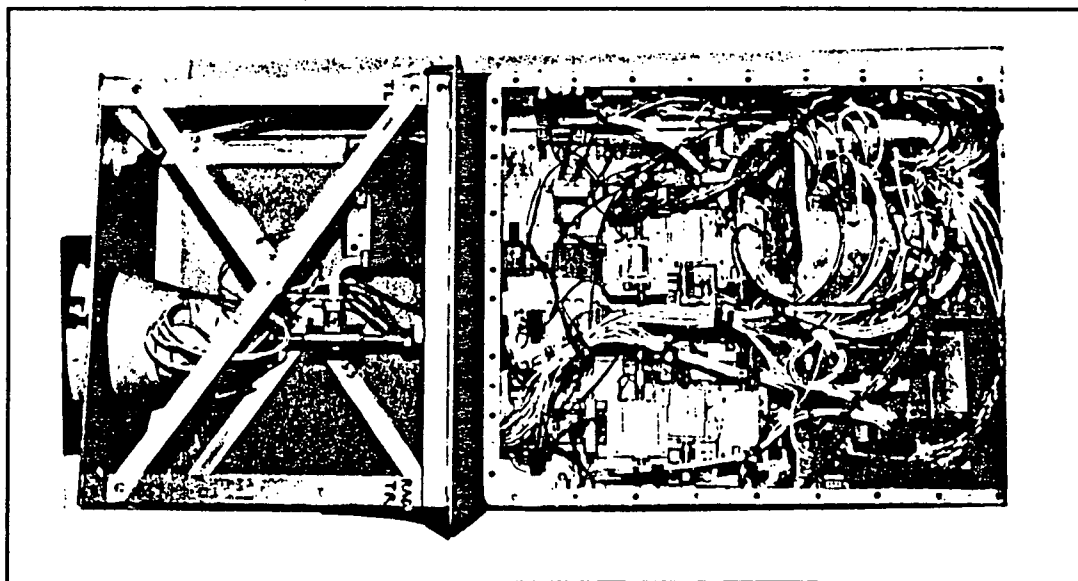


Figure A.2 Photo, Top View Into Radiometer with Antenna.

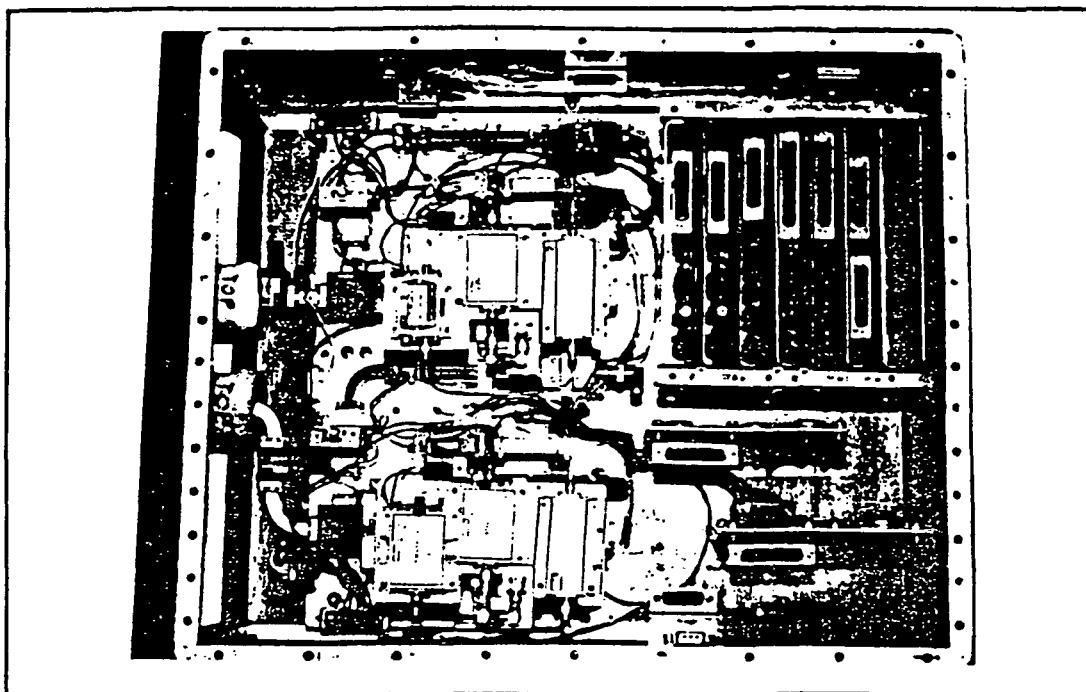


Figure A.3 Photo, Top View of Radiometer, No Wiring.

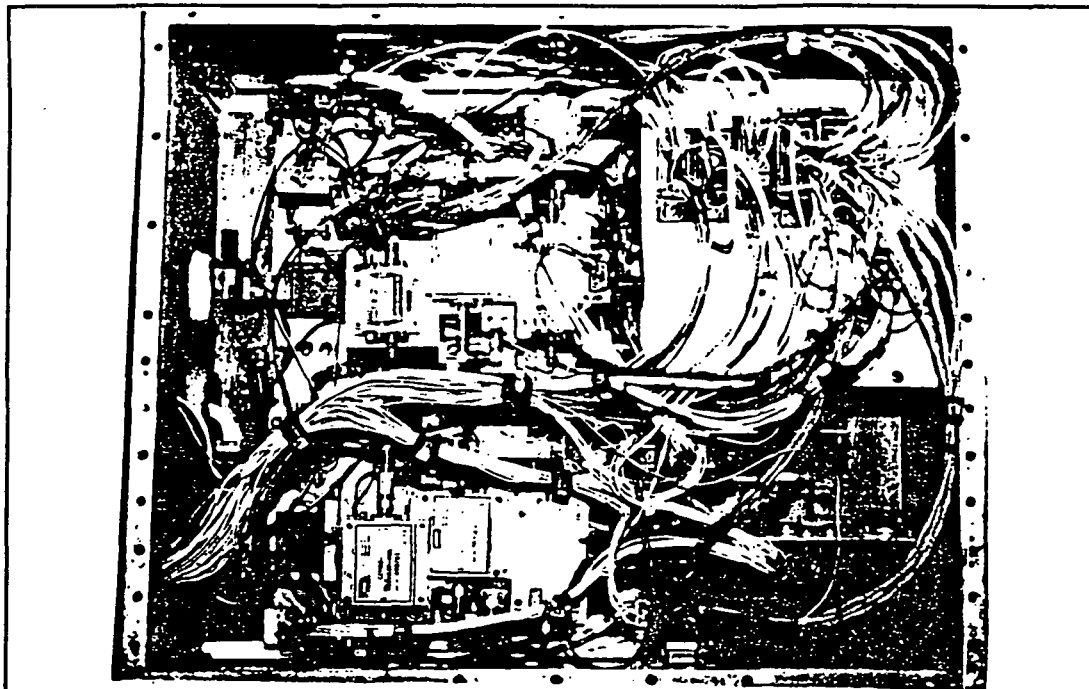


Figure A.4 Photo, Top View of Radiometer, Wiring Installed.

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