A Fast, Very Sensitive Calorimetric Power Meter for Millimeter to Submillimeter Wavelengths

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Abstract

A calorimetric power meter is now commercially available having sensitivity and stability sufficient to measure a power level well below 1 μW. The Erickson model PM1B sensor has an input in WR10 waveguide, with a very well matched termination. The frequency response is believed to be flat throughout the submillimeter wave region. The response time of the sensor is 7 sec, but this time can be reduced through the use of thermal feedback, at the price of higher noise. An effective time constant of 1 sec is achieved with 0.15 μW rms noise, and 0.25 sec with 1 μW noise. The maximum input power is 200 mW, making the instrument useful with a wide variety of sources. The sensor head is quite small, yet well insulated from its environment.

Introduction

Calorimeters have long been used to measure power over extremely wide frequency bands. The method is very simple, just measuring the temperature rise of a well-matched wideband absorber in response to applied power. In principle, this technique is easily calibrated because the same absorber can be heated using a known amount of power applied through other means. In practice, the difficulty is in making a sensor with a fast response time and uniform load temperature, while maintaining acceptable sensitivity. It is important in this calibration that the reference heat produce the same response as the microwave power. In addition, drift due to ambient temperature variations is very difficult to control. The usual technique to reduce drift is to use a matched pair of sensors having nearly identical thermal environments, and to use one as the measurement sensor and the other as a reference. So long as the sensors are identical, they show no response to ambient temperature variations, but this requires extreme care in their fabrication.

In the simplest situation, the response time, τ, of the sensor is determined by the heat capacity of the sensor load, C, and the thermal resistance to thermal ground, ρ:

$$\tau = C \times \rho$$

while the temperature rise ΔT is given by:

$$\Delta T = \rho \times P, \text{ where P is the absorbed power.}$$

Since a large ΔT response is desired, ρ must be made fairly large, and the only way a fast response can then be achieved is by making the heat capacity (and thus the mass) of the sensor as small as possible. However, a practical sensor can not be made arbitrarily small. A sensor must be large enough to be an efficient absorber with no frequency resonances. In practice this means that a slowly tapered load is required, and the load material must be thick enough that no power can pass through it. In addition the load must include a temperature sensor and a heater resistor for calibration.

This paper describes an improved calorimeter which has been designed to measure power levels below 1 μW with an acceptable time constant and with accuracy and reproducibility of ~2-3%. It is similar in many ways to the unit in [1] (which in turn uses concepts from earlier work [2,3,4]), but the improvements lead to nearly an order of magnitude increase in sensitivity, with faster response. This power meter is now available commercially as the Erickson Instruments PM1B.

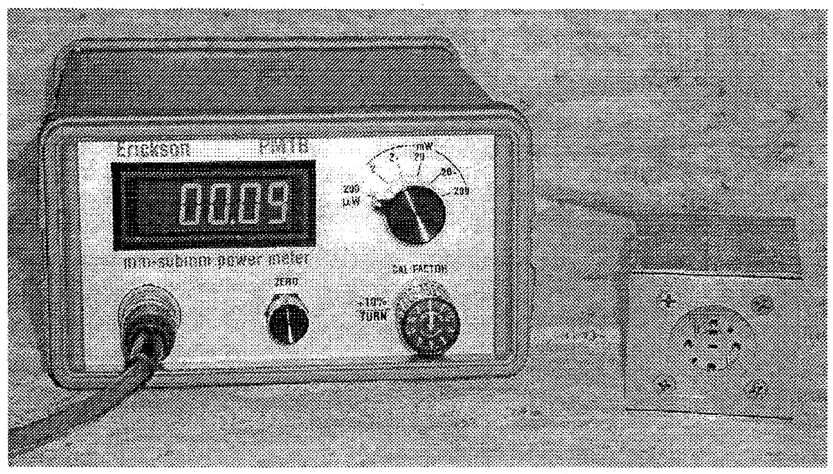

Figure 1. Complete PM1B. The sensor head is 5 cm square.

Calorimeter sensor construction

The geometry of the sensor used in this work is shown in figure 2. The input was chosen to be WR10 waveguide so that cross calibration could be done against other waveguide sensors, while being small enough to keep the heat capacity low. It is quite practical to couple from this waveguide into any smaller size using linear tapers. While many submillimeter sources are optically coupled, nearly all use waveguide internally, and are easily coupled back into waveguide using just a feedhorn. An unconventional waveguide termination geometry was chosen which places the load entirely on the outside of the waveguide, where it is easy to attach the thermometer and heater. The load itself is a rectangular slab of silicon with a thickness of 0.4 mm, having a 100Ω/ metal film on the inside. This load has very low mass and high thermal conductivity. The waveguide is cutoff at an angle of 10° to produce a VSWR better than 1.15:1 within the WR10 band. The expected VSWR at much higher frequencies (at least 2 THz) is comparable. All of the edges of the wafer are coated with conductive epoxy to eliminate loss of power, and the absorber is carefully sealed to the waveguide without any gaps where power can be lost.

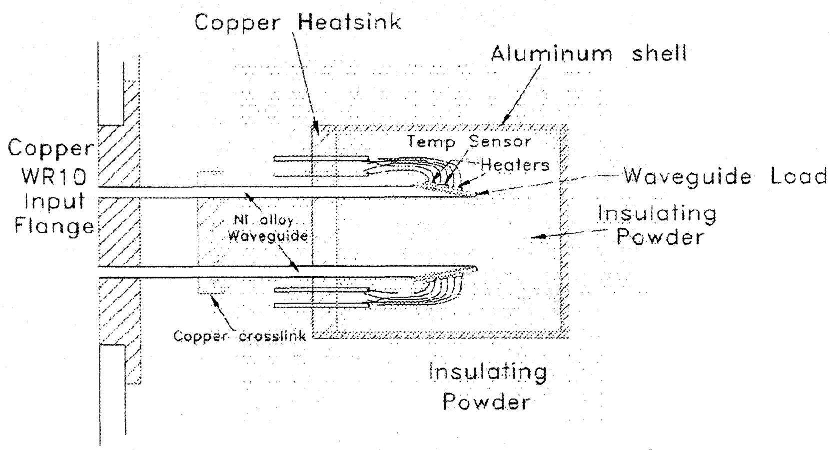

Figure 2. Diagram of the power sensor head, showing the waveguide loads, heaters and temperature sensors.

The waveguide is a low thermal conductivity Ni alloy having a thickness of 0.07 mm and a gold plated interior. The waveguide must be very thin so that it has sufficient thermal resistance in a short length, and so that its heat capacity is small. In this sensor the waveguide contributes ~1/3 of the total heat capacity of the load. The time constant of this sensor is 7 sec with a responsivity of ~180K/W. While this is faster than typical calorimeters, even faster response may be achieved as described in the next section.

The temperature sensors used must be very linear if the sensor is to work over a wide power range, and they also must have very low drift and noise. Thin film Pt resistance sensors (RTD's) have been found to be stable to <0.01 mK, with noise determined by the readout electronics. At an RTD bias of 0.5 V, the noise of the preamplifier is equivalent to ~.02 μW rms, and is less than typical drift levels. The bias resistors can add a very large amount of noise and the exact manufacturing process used seems to make a difference. Thin metal film resistors are best but some of these are still noisy, while the best add no detectable noise. Noise becomes quite important with the feedback circuit described later.

Drift in thermal sensors comes from two sources. In analogy with differential amplifiers, inputs can be considered common mode if a thermal influence is applied to both temperature sensors, and differential if it applies to one only. Common mode drift, as is caused by ambient temperature fluctuations, occurs because the sensors do not have matched response. The sensors show a linearity error of up to 3×10^{-3} which is equivalent to a common mode drift of 15 μW/K. The match of the sensors is trimmed with external resistors to bring this drift to a level ~2 μW/K (ambient), which leads to much improved stability for the sensor.

Differential drift due to external temperature fluctuations is a more difficult problem. The sensor response of 0.22 mK/μW means that differential influences must be very highly attenuated to measure power below 1 μW. The sensor's differential response is greatly reduced by connecting the waveguides together with copper plates at three points.

Any differential heat leaking in along the waveguides is short circuited at these points, and the mass of the last plate leads to a very slow propagation of this (mostly common mode) heat to the sensors. In the earlier sensors, the separation between waveguides was 10 mm, which has now been reduced to 5 mm. This helps considerably to reduce differential heat input. The total waveguide length to the load is 24 mm. Most of the length is used for thermal isolation and only 3 mm is used for power measurement. The sensors are isolated from differential heating from other directions by enclosing them in a metal shell, so that any external influence is nearly entirely common mode.

One subtle source of drift is heat generated in the temperature sensors due to the current used to measure their resistance. Convection currents from one load to the other cause drift which depends on the orientation of the sensors. This problem is potentially more serious as the sensors are moved closer together to reduce differential drifts. To eliminate this effect the sensors are thermally isolated from each other by filling the shell with a silica powder insulator, which almost completely blocks convection, and also has extremely low conduction. This same insulator fills the sensor housing surrounding the metal shell.

Feedback circuit for faster response

Improvement in the speed of response can be obtained electronically with the readout circuit. While in principle the readout response can be tuned to give an apparent increase in sensor speed, in practice this results in overshoot of the measured power, and very unpredictable responses while tuning. An alternative method is to use the readout circuit to maintain a nearly constant temperature for the load. This can be done by biasing the temperature of the loads (both sensor and reference) to a point somewhat above that which would be produced by the highest power that is to be measured. As RF power is applied, the heater power is reduced to maintain a constant temperature. Rather than measuring temperature rise, one then measures the change in heater power. If the feedback circuit maintaining the temperature has a gain, G, then the change in temperature for a given RF power is reduced by a factor of G, and the response time is decreased by the same factor. There is no problem in this case with spurious behavior in the indicated power, no matter how the RF power is varied, and arbitrarily fast response may in principle be obtained. However there are practical limits to speed because the sensor noise, for frequencies $> 1/\tau$, is increased by a factor of G, and also because the response can not be faster than the thermal equilibrium time within the load element (due to the thermal resistance between the heater and the thermometer). For the load used in these sensors, a gain of 30 is about as high as is practical, giving a predicted time constant of 0.25 sec. At this gain there is an overshoot in response to a power impulse of 1-2%. This gain leads to an rms noise level of $\sim 1\ \mu W$ and is only practical on the higher scales. At lower power levels the gain is reduced for lowest noise and highest stability.

A block diagram of the readout electronics is shown in Fig 3. The sensor uses a bridge circuit to measure the temperature imbalance of the two waveguide loads, with AC excitation of the bridge to minimize drift in the following amplifier. The AC imbalance signal is amplified and then synchronously detected to minimize noise. The reference sensor is always heated with a power 5-10% above full scale input. A feedback circuit amplifies this error and heats the active sensor to approximately the same temperature as the reference. The feedback circuit is somewhat complicated because the error voltage from a linear temperature sensor applied to a heater resistor produces a feedback proportional to the error squared. Such a nonlinear loop response is difficult to compensate. To simplify the loop response and to provide a linear readout, an analog square root IC is included within the loop. These IC's (at the highest level of performance) are sufficiently accurate that their response has no measurable affect on the overall linearity of response. The feedback voltage (prior to the square root) is linear with input power, being largest with no input and decreasing to nearly zero at full scale input. After subtracting an offset this voltage is used for power indication.

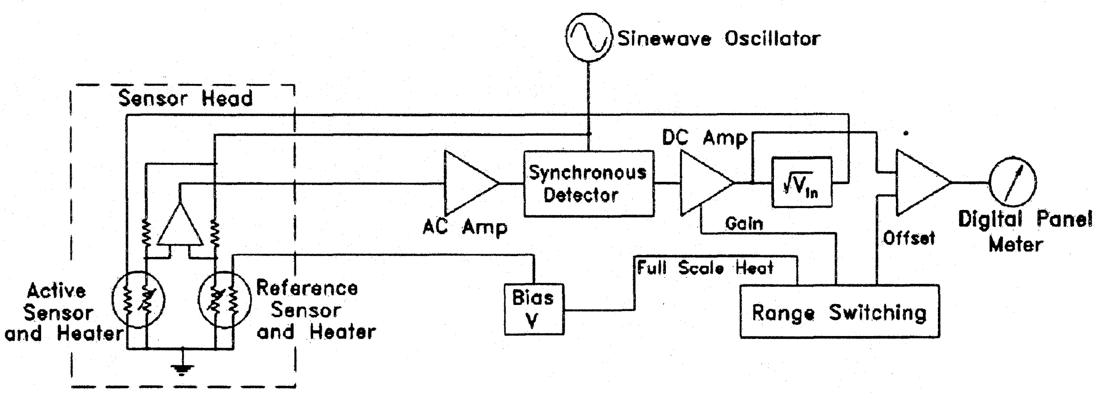

Figure 3. Block diagram of the readout electronics with feedback to increase the response speed.

The stability of this type of sensor depends critically on maintaining a constant heat path to thermal ground for both waveguide loads. Since the bias temperature is greater than full scale power, any variation in the thermal path will cause a significant zero drift. For this reason, it is necessary to change the bias heat when changing measurement scales so that the drift will be only a fraction of each full scale reading. Changing scales requires changing several gains and offsets in the electronics.

Performance

The meter has four ranges, 200 mW, 20 mW, 2.0 mW and 200 μW FS, using a $4^1/_2$ digit panel meter. The maximum power level of 200 mW is limited mainly by concerns for the sensor stability if operated continuously at higher temperatures, as is necessary due to the bias heat used. At the bias point, the sensor temperature is ~60 C. Table 1 shows the properties of the meter on the four ranges. Measurements typically require three time constants for a response to within 95% of the true input power, but the true sensor response is not a simple exponential due to multiple heat paths to ground with various intermediate masses. The minimum readout noise is ±1 in the last digit of the meter, and this must be added to the rms noise of the electronics from the last column.

Scale	Gain	1/e time const.	95% response	rms noise
200 μW	none	7 sec	20 sec	~.02 μW
2.0 mW	5	0.9 sec	4 sec	~.15 μW
20 mW	30	0.25 sec	0.75 sec	~1 μW
200 mW	30	0.25 sec	0.75 sec	~0

Table 1

The 200 μW scale operates open loop for best stability, and has a drift level below 0.1 μW during a typical measurement interval, if the sensor is not thermally disturbed. However, most measurements introduce a differential thermal disturbance, and the drift becomes ~0.5-1.0 μW typically. This can be reduced if the sensor remains connected to the source for a time so that the transient dies out. Wrapping the sensor head in insulating foam significantly reduces drift. In one measurement, a power level of 0.2 μW at 1.5 THz was measured with confidence in a set-up where the source could be turned on and off without disturbing the sensor.

The large change in bias heat when changing scales leads to the only obvious difference in use between this meter and a more conventional power sensor. Upon changing scales, there is a large thermal transient that must die away before the sensor is stable once more. This stabilization time depends on the inherent time constant of the sensor element, and when switching scales to a lower power, at least five minutes are required for stabilization. This problem would be less if the two loads were perfectly matched so that they both settle at the same rate. When switching scales upward in power the settling time is quite short.

For some purposes, as when measuring the power from cryogenic components, it is useful to place the sensor in a vacuum to avoid the loss and calibration errors introduced by a vacuum window. This sensor works in vacuum (if the sensor case is provided with vent holes) with some minor changes in its properties. The open loop sensitivity increases by 30% (for the 200 μW scale), but on the higher ranges this error is reduced by feedback. Thus the sensitivity increase on the 2 mW scale is only 5%, and on the 20 and 200 mW scales only 1%. The exact changes can be measured in real time with the internal calibration. There is a change in the zero by ~70 μW, but otherwise operation is normal.

Accuracy

It is extremely hard to check the accuracy of a sensor such as this given the lack of power standards. In the case where the input loss is negligible, all of the power must appear as heat in the load, and this heat can be calibrated by heating the load with a known power. There is some uncertainty in this case because the load is not isothermal. The temperature gradient over the full length of the silicon load is ~15% of that within the waveguide. This can be largely resolved by placing the calibration heater in the same area where most of the heat is expected to be dissipated. At lower frequencies the loss is primarily near the extreme end of the load (where the waveguide height becomes very low), and so the calibration heater is placed at this point. At very high frequencies the mode distribution may change this somewhat, but still the power tends to be dissipated in the same region. The uncertainty in the exact temperature distribution contributes an error of ~2% in power.

Calibration of waveguide loss is fairly simple in principle. The effective waveguide loss length is the full distance from the input to the last copper plate, plus half the distance from this point to the load. Assuming the waveguide loss is uniform over its length, and that the loss for all modes is not very large, this loss may be measured by inserting a piece of identical waveguide in front of the sensor and measuring the change in power. Standard coin silver waveguide has only slightly less loss than the plated waveguide used in the sensor and so a somewhat longer piece (~3 cm) may be used to measure this correction. In practice, the inaccuracy in the use of a calorimeter of this type is dominated by the much more uncertain loss in the tapers used to transition into WR10 waveguide. Within the 75-110 GHz band the input loss is small enough to introduce an uncertainty of only 1-2%. At 300-500 GHz the loss of a taper is ~1 dB, so the uncertainty is still acceptable. In direct comparisons of all meters with each other, relative errors never exceed 3%, in the test frequency range of 100-150 GHz.

Conclusion

A new ultra-broadband power meter with accuracy of ~3% has very high sensitivity, and a moderately fast response time. The sensitivity of under 1 μW is the highest ever achieved with a well calibrated thermal sensor. The unit is quite stable under typical ambient temperature variations, and is small enough to be used like a conventional microwave power sensor. The input is in WR10 waveguide but signals at any frequency above 70 GHz may be measured with appropriate tapers. Accuracy is expected to be good to >2 THz, and measurements may be made even in the visible.

References

[1] N. Erickson, "A Fast and Sensitive Submillimeter Waveguide Power Meter," *Tenth International Symposium on Space Terahertz Technology,* pp. 501-507, 1999.

[2] B. Volwinkle, "Broad-band Calorimeter for Precision Measurement of Millimeter and Submillimeter-Wave Power," *IEEE Trans. on Instrumentation and Measurement,* Vol 29, pp. 183-189, 1980.

[3] B. Vowinkel and H.P. Roser, "Precision Measurement of Power at Millimeter and Sub-Millimeter Wavelengths Using a Waveguide Calorimeter," *Int'l J. of Infrared and Millimeter Waves*, Vol 3, pp 471-487, 1982.

[4] A.C. Macpherson and D.M. Kerns, "A Microwave Microcalorimeter," *Rev. Sci. Instr.*, vol. 26, pp. 27-33, 1955.