A FAST AND SENSITIVE SUBMILLIMETER WAVEGUIDE POWER METER

Neal Erickson
Department of Physics and Astronomy
Lederle 619, Univ. of Mass.
Amherst, MA 01003

Abstract

A sensitive, fast and stable power meter has been developed which responds over a frequency range extending from 75 GHz up to the THz range. The method of measurement is essentially calorimetric, but with feedback circuitry added to speed up the response. The measurement time constant is 0.5 sec for power levels from 2-100 mW, but increases to 9 sec below 50 μW. The drift level is 2-5 μW in a typical measurement, while accuracy is better than 3%.

Introduction

Measurement of power at frequencies in the mm and submillimeter wave range has always been problematic. Power sensors suffer from a number of problems, all of them worsening as frequency increases. These include poor sensitivity, slow response time, poor impedance match and frequency dependent calibration. At this time commercial sensors are available which largely eliminate these problems up to 110 GHz, but at higher frequencies there are no fully satisfactory measurement options. This leads to a situation where it is not known within a factor of two or more the power required to operate SIS or HEB mixers, and where errors in the measurement of multiplier efficiencies can be comparably large.

In fact, it has always been possible to make accurate measurements at any frequency through the use of calorimetric power measurement [1,2,3]. The problem with these instruments is that they are very slow in response, with sufficient zero drift that accuracy suffers severely below a power of 50 μW. The accousto-optic sensor [4] can offer fast response and reasonable accuracy, but requires careful matching to an optical beam and requires correction for input reflection and partial absorption. Its drift level is also fairly large, and it requires a chopped signal. Other very sensitive detectors such as cooled bolometers are also fast, but are nearly impossible to calibrate with any accuracy.

The goal of this work was to develop a power sensor which could be used like a conventional waveguide power sensor at any frequency in the ~100GHz to THz range, with fast enough response to permit its use in tuning a source. The sensor had to have minimal frequency variation to its response and a sufficiently good input match that the error due its VSWR would be insignificant even in the measurement of a source with a VSWR of 4:1. In addition it had to be sensitive enough to reliably measure a power level

of a few microwatts, since this is a typical power level for submillimeter sources. This sensor incorporates a number of features used in the calorimeter of refs [1,2], but uses a construction much more optimized for response speed.

Fundamentals of calorimetry

The only method of power measurement that is frequency insensitive is calorimetry, that is, simply measuring the temperature rise of a well-matched wideband absorber in response to applied power. In principle, this is a very simple technique and is easily calibrated because the same absorber can be heated using a known amount of power applied through other means. It is important in this calibration that the reference heat produce the same response as the microwave power. In practice, the difficulty is in making a sensor with a fast response time and uniform temperature, while maintaining acceptable sensitivity. 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 common mode temperature variations, but this requires extreme care in their fabrication.

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, but 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.

The accuracy of power measured in this way is quite good if the input loss is small. In the case where the input loss is negligible, it is obvious that all of the power must appear as heat in the load, and that this heat can easily be calibrated by heating the load with a known power. The problem when the loss is high is that it is not easy to determine exactly how much of the input power dissipated in the waveguide loss is also heating the load. However, this factor is calculable assuming the waveguide loss is uniform over its length. 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%.

Calorimeter sensor

The geometry of the sensor used in this work is shown in figure 1. The input was chosen to be WR10 waveguide so that cross calibration could be done against commercial 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. 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 simple slab with no critical shape. The waveguide cutoff angle was chosen as 10° to give a VSWR better than 1.2:1 for the best available load material in the WR10 band. A variety of materials have been tested as loads including several ferrite loaded resins and other lossy materials. The best found so far in having a good match with very low mass is a sandwich of 2 silicon wafers. The inside layer is a 100 $\Omega/\square$ metal film deposited on a moderate resistivity silicon wafer with a thickness of 0.2 mm. The outer layer is a highly doped silicon wafer with a thickness of 0.3 mm. The metal film provides a well matched load while the very lossy silicon top layer absorbs all of the energy that leaks through the film. All of the edges of the inside 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. Silicon is an excellent substrate material because it is a good heat conductor, is easily coated with metal films, and is readily available in various dopings.
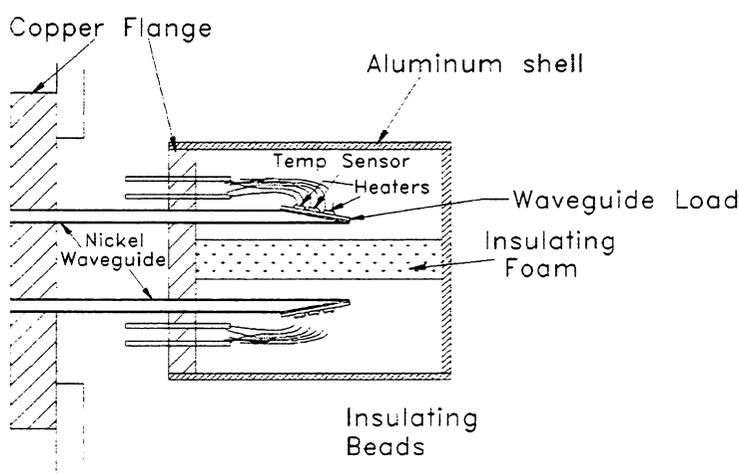
Figure 1. Diagram of the power sensor head, showing the waveguide loads, heaters and temperature sensors.

The waveguide is electroformed nickel with a gold plated interior. Nickel is a fairly poor thermal conductor, and is readily electroplated with good mechanical properties. 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 compared to that of the load. A numerical simulation shows that the effective heat capacity of the load includes a contribution of one third of the heat capacity of the full waveguide length. A practical minimum waveguide thickness is ~60 μm and a reasonable length is 10-15 mm. For this sensor, the waveguide adds about 10% to the effective heat capacity of the load. The time constant of this type of load is found to be 9 sec for a responsivity of ~500K/W. While this is faster than typical calorimeters, it is still too slow for a primary power sensor.

Stainless steel would be better for use in the waveguides since its thermal conductivity is 20% that of Ni, but it is only available in 0.25 mm thickness, and must be thinned to be useful. Thinning this small waveguide in a reproducible manner is possible but tedious. Stainless steel waveguide also must be plated on the interior to reduce the loss, and this process is somewhat variable in its results.

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 have been found to be stable to <0.2 mK, with noise determined largely by the readout electronics. The only problem with these sensors is that the smallest standard package is somewhat large, and adds 20% to the mass of the load.

Drift due to external temperature fluctuations is a serious problem with such a small responsivity. To minimize this problem the sensor uses a second stage of thermal isolation by connecting the waveguides together at their midpoint with a copper plate. Any differential heat leaking in along the waveguides is short circuited at this point, and the mass of the midplate leads to a very slow propagation of this (now common mode) heat to the sensors. The sensors are isolated from any 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. To eliminate this effect the sensors must be thermally isolated from each other by an insulator, which completely blocks any convection within 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/τ, 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 20 works well, giving a time constant of 0.45 sec, and faster response is probably possible. However, this gain is only practical with the present circuit for measurements above 2 mW. 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 2. The feedback circuit is somewhat complicated because the error voltage from a linear temperature sensor applied to a heater resistor produces a $V^2 \propto \Delta T^2$ response. Such a nonlinear loop response is difficult to compensate. To simplify the loop response and to provide a linear readout, an analog square root module is included within the loop. These modules are now so accurate that their response has no measurable affect on the overall linearity of response. The electronics module uses a bridge circuit to measure the temperature imbalance of the two loads, with AC excitation of the bridge to minimize drift in the following amplifier. The AC imbalance signal is amplified and the synchronously detected to minimize noise. Following DC amplification, the square root is taken and the output is then offset from the full scale error voltage. This offset error is then fed back to the active sensor to heat it in response to its difference in input power from full scale. The reference sensor is always heated with a power equivalent to full scale input.
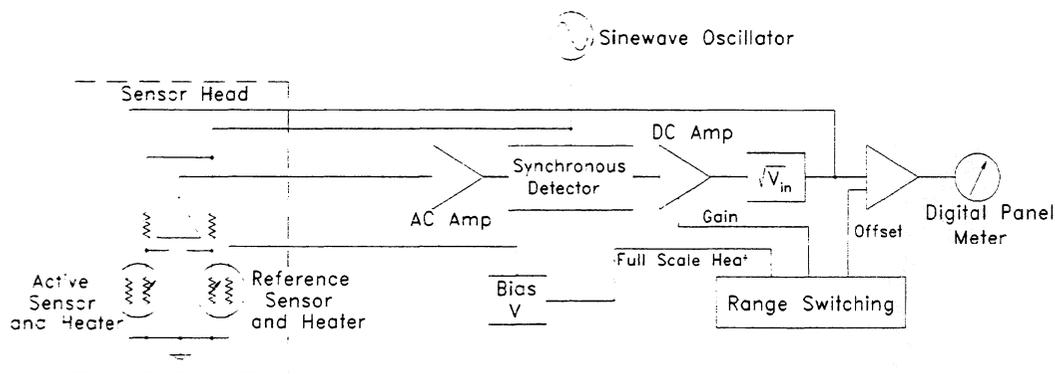

Figure 2. Block diagram of the readout electronics with feedback to increase the response speed.

The feedback loop requires that there be a heater resistor on the load, and in addition there needs to be a second heater resistor for calibration purposes. These resistors add mass to the sensor, and their functions could all be performed by metal films deposited directly on the silicon load, but this would require an investment in masks and lithography.

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 somewhat 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 actual performance achieved in the prototype meter built using this technique is not optimized, but the meter still is an extremely useful instrument. This sensor used waveguides with relatively poor thermal match, having 10% difference in responsivity, and thus significant common mode response. The maximum power level that can be measured is 100 mW, 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 70 C. Other scales provided are 20 mW, 2.0 mW and 200 μW FS, using a $3^1/_2$ digit panel meter. On the 100 mW scale, the drift is negligible, being <0.2 mW over several months. On the 20 mW scale, the drift is ~0.1 mW over the same time, with much smaller short term drift. For both of these scales the gain is 20 so the time constant is 0.45 sec. For the 2.0 mW scale the gain is reduced to 5 giving a time constant of 1.8 sec. This scale is useful down to a power level of 50 μW with good stability, with a typical drift and noise level of 5-10 μW. The 200 μW scale operates open loop for best stability, and has a typical drift level <2 μW during a typical measurement interval, if the sensor is not thermally disturbed. However, most measurements introduce a thermal disturbance due to the common mode response, and the drift becomes ~5 μW typically. This can be reduced if the sensor remains connected to the source for a time so that the transient dies out.

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 time constants (45 sec) are required for stabilization. This problem would be less if the two loads were closely matched so that they both settle at the same rate. When switching scales upward in power the settling time is quite short.

The potential for much higher stability is apparent from another sensor (without feedback), which was made using stainless steel waveguides that were electropolished

down to a thickness of 80 μm and matched in their thermal conductivity. They also have a longer length and significantly lower thermal conductivity. However, the overall time constant is only twice as large, in part because the waveguide conductivity is only a part of the total heat loss of the sensor. In this sensor the typical drift level is <0.5μW over a period of 5 minutes without perturbation, and <1 μW during typical measurements. If such low drift were present in a sensor with fast response, then the instrument would be extremely useful for most submillimeter work. These refinements are planned for future versions of this instrument.

Conclusions

A very wideband power meter has been built which is fast and stable. The useful power range is 10 μW-100 mW with a measurement time constant of 0.5-9 sec depending on power. Construction is somewhat complex, but could be simplified in production. This meter already is useful for a wide range of measurements on LO sources, and with some refinements could fill an even larger range of needs.

References

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

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

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

[4] Thomas Keating Ltd, Billingshurst, W. Sussex, England.