Conclusions

In the near future, many measurements will be needed in the submillimeter region. If masers are to be developed in this range, considerable spectroscopic data on solids and gases will be required. Data are also needed on dielectric constants and loss tangents of a wide variety of materials. The latter measurement tasks may sound plebian, but they are no less necessary, and they require an operating submillimeter system. For any such usage, it is felt that the measurement system described here has many advantages, including economy and versatility. The principle has been proven and the practicality demonstrated.

Acknowledgment

The author would like to express his appreciation to Dr. M. Cohn for his proposal of the work reported herein and for his many helpful discussions on the course of its progress. Thanks also go to J. O. Wedel for his circuitry design work and to J. D. Rodgers for his assistance in laboratory measurements.

New Technique for Microwave Radiometry*

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Summary—An interference modulation technique for microwave radiometry is described. Use of the technique is considered for the development of a radiometer for tuning over a wide range of frequencies, a radiometric device for determining the absolute sensitivity of detectors over a wide range of frequencies, and a means for determining the power level of coherent sources as a function of frequency. A device using this technique does not require the use of a coherent source, and the technique is applicable to wavelengths well into the low millimeter region.

A tunable radiometer using interference modulation has been operated as a measurements system within the 50–90 Gc region. Successful performance required the use of a sensitive detector which consisted of a barretter operated in an evacuated atmosphere. A noise source having known temperature was used as the source of microwaves for determining the sensitivity of detectors as a function of frequency. It has been found that the sensitivity of barretters is greatly improved by a reduction in air pressure and that, when evacuated, a commercially available barretter will provide a sensitivity of approximately 6×10^{-11} watts for an audio bandwidth of one cycle per second. The technique discussed provides the opportunity for developing a calibrated power meter for millimicrowatt levels from coherent sources.

Introduction

A CONTINUOUSLY tunable measurements device based on an interference technique has been operated throughout the 50–90 Gc region. The instrument, which does not require the use of a coherent source, is useful for making transmission measurements as a function of wavelength; and detector sensitivity measurements can be made without the use of a power meter, per se. In fact, the technique used should also serve for measuring the output power of weak coherent sources.

Determination of detector sensitivity vs frequency throughout the low millimeter region by other techniques is a formidable task because of the low-power levels (harmonics) available, and the difficulty of obtaining harmonics throughout all of the lower millimeter region. Transmission measurements have not been possible with superheterodynes in much of the lower millimeter region because suitable coherent sources (and sensitive fast detectors) are unavailable. Use of direct-detection systems is limited by the need for a continuously tunable, band-pass filter for microwaves. The effect of a tunable filter is accomplished in the present system through the use of an interference technique and filtering in the audio range.

The heart of the new instrument is, in essence, a continuously tunable, electromechanical, band-pass filter. It has been operated throughout the 50–90 Gc region where 4-mm components could be used, but the technique used should be useful throughout the low millimeter-wave region. Operation is based on the fact that there is a one-to-one correspondence between mechanical speed and the frequency of Doppler components. Power from a microwave source is divided between two paths by a waveguide T and is recombined with a T before reaching the detector. Doppler components are produced by varying one of the paths with a trombone-shaped section which is moved back and forth by the drive rod at a nearly constant speed. Thus, an interference modulation frequency is produced which is

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uniquely related to the wavelength within the waveguide. The audio frequency modulation spectrum at the detector for a wide-band microwave source will be continuous; thus, a tuned audio amplifier (or other frequency analyzer) in conjunction with the varying path length will serve as a band-pass filter.

Theory of Interference Modulation

Principles of Operation

The fact that there is a one-to-one correspondence between mechanical speed and the frequency of Doppler components provides an interesting new concept for microwave instrumentation. Assume that the difference in path length between Paths 1 and 2 of Fig. 1 is varied at a constant speed s; then an interference modulation frequency f_a would be produced which is related to the wavelength within the waveguide λ_g by the equation

$$f_a = s/\lambda_g. \tag{1}$$

Thus, the tuned audio amplifier in conjunction with the varying path length will serve as a microwave band-pass filter because the amplifier output will depend only on those frequency components corresponding to λ_g of (1).

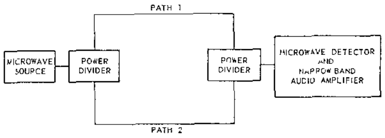

Fig. 1—Equivalent circuit of interference modulation device.

In order to visualize amplifier waveform, assume that the difference in lengths of Paths 1 and 2 is varied at a nearly constant rate by alternately lengthening and shortening one of the paths. With this arrangement the Doppler frequencies will be unchanged except during the short time required for reversing the motion. It is shown in the Appendix that detector power from a noise source depends on the difference in lengths of Paths 1 and 2 and that the maximum power occurs when the paths are of equal length. When operated at low-power levels (square-law region), detector output voltage V_d is related to incident microwave power P_d by

$$V_d = RP_d. \tag{2}$$

For a given detector the responsitivity R is approximated by a constant. Thus, the voltage waveform out of a tuned amplifier following the detector will contain pulses occurring periodically, and these pulses will have a maximum each time the path length difference is zero.

For the special case of the narrow microwave pass band B considered in the Appendix, amplifier output would be independent of audio bandwidth provided that all Doppler frequencies are within the amplifier pass band. Conversely, the output of a tuned amplifier is independent of microwave bandwidth if the audio amplifier pass band is contained within the complete Doppler spectrum. Assume that 1) the frequency limits of the amplifier pass band correspond exactly to the two Doppler frequencies for the edges of the microwave band $(F-B/2, F+B/2)$, and 2) that the detector responsivity R is constant over the frequency interval B. Under these conditions, and for the realistic condition of a microwave source spectrum wider than B, (2) and (17) can be used to express amplifier waveform for the hypothetical case of a constant change in the path length difference. Thus, if time zero is defined so that it exists each time the path difference is zero, the voltage out of the tuned amplifier with center frequency f_a and unity gain may be expressed as

$$v = \frac{1}{2} RkTB \frac{\sin\left[\left(\frac{\omega_a t}{F}\right)\frac{B}{2}\right]}{\left[\left(\frac{\omega_a t}{F}\right)\frac{B}{2}\right]} \cos \omega_a t = g(t) \cos \omega_a t. \tag{3}$$

The radian frequency of the output wave is ω_a, and it is equal to the center frequency of the tuned amplifier. Notice that the argument of the sine term of (3) differs from that of the cosine term by the fraction $B/2F$. The ratio B/F is the system resolution factor, and it might typically be of the order of 0.1 or less. Thus, the effect of $g(t)$ is to slowly modulate the envelope of the cosine output wave.

In practice, the difference in lengths of Paths 1 and 2 is changed at a nearly constant rate by alternately lengthening and shortening one of the paths. The path length change is such that equal path lengths occur approximately midway between the path length extremes. In this arrangement, pulses described by $g(t)$ occur periodically and reach a maximum each time the difference between path lengths is zero. The pulses described by $g(t)$ exist throughout the sweep period, but with reduced amplitude. Thus, if (3) is to be valid, the sweep period must be sufficiently long so that effects of sweep-to-sweep interference are negligible. Near each envelope peak where values of t are sufficiently small (by definition) that

$$\sin\left[\left(\frac{\omega_a t}{F}\right)\frac{B}{2}\right] \text{ can be approximated by } \left[\left(\frac{\omega_a t}{F}\right)\frac{B}{2}\right],$$

amplifier output voltage may be expressed as

$$v = R\tfrac{1}{2}kTB \cos \omega_a t. \tag{4}$$

Notice that (4) indicates that the magnitude of the peaks in amplifier waveform will be directly proportional to the power (kTB) contained in the microwave band B; recall that for a wide-band noise source, B is determined by the bandwidth of the tuned audio amplifier. Also recall that the center of the band to which the system responds is determined by the rate of change in path length and/or amplifier center frequency. Thus, the band of microwaves to which the system is tuned can be controlled solely by the characteristics of the audio amplifier.

Since the magnitude of the peaks in the amplifier output caused by a coherent source will also be proportional to source power, comparison of peak output with that caused by a calibrated noise source provides a convenient method for determining the power level of a weak coherent source. In principle, power measuring devices can be calibrated as a function of frequency with a modulated noise source and a tunable microwave band-pass filter. Such measurements are not currently possible in the low millimeter region because band-pass filters with variable but known center frequency and effective bandwidth are not available.

System Resolution and Useful Source Power Available

As stated above, the magnitude of the peaks in the amplifier waveform is directly proportional to the power contained in a microwave band of width B which is determined by the bandwidth of the tuned audio amplifier. Let f_1 and f_2 be the lower and upper frequency limits of the audio pass band, respectively. Then the lower and upper microwave frequency limits to which the amplifier responds, F_1 and F_2, may be expresses as

$$\left[F_2\left(\frac{\lambda_2}{\lambda g_2}\right) - F_1\left(\frac{\lambda_1}{\lambda g_1}\right)\right]\frac{s}{c} = f_2 - f_1,$$

where

λ_1 and λ_2 are free space wavelengths for the frequencies F_1 and F_2,
λg_1 and λg_2 are guide wavelengths for the frequencies F_1 and F_2,
s is the rate of change in path length and
c is the free space propagation velocity.

Let F represent the center of the microwave spectrum, for which the Doppler frequency equals amplifier center frequency f_a, and let B and B_a represent the microwave and audio bandwidths, respectively. Since the pass bands of interest are narrow, it is reasonable to assume that $\lambda_1/\lambda g_1$ equals $\lambda_2/\lambda g_2$. With this assumption, it is seen that the relation

$$\frac{B}{F} = \frac{B_a}{f_a} \tag{5}$$

is a good approximation for narrow pass bands. In other words, the percentage microwave bandwidth resolved by the system is equal to the percentage audio bandwidth.

For a constant rate of change of path length and a coherent source, detector power [see (16)] modulated so that it passes an amplifier tuned to the Doppler frequency ω_a may be expressed $(P_s/2)\cos\omega_a t$. P_s is the power level of the unmodulated coherent source. From (4) it may be seen that detector power from an incoherent source during the interval of maximum output is $(kTB/2)\cos\omega_a t$. Noise power from a matched microwave load can be determined from the approximation[1]

$$P = kTB, \tag{6}$$

where k is Boltzmann's constant (1.38×10^{-23} joules per degree Kelvin), T is the temperature of the load in degrees Kelvin, and P is the power in watts available in the bandwidth B in cycles per second. Thus, equivalent source power for the case of an incoherent source is equal to the total noise power, kTB, contained within the resolution of the system.

This paper describes a system which uses an audio amplifier having a center frequency of 28.5 cps and a bandwidth of 4 cps. Assume that the change in path length and audio amplifier center frequency is such that the system responds to a band centered at 70 Gc; then from (5) we find that the system responds to a microwave band B as follows:

$$B = \frac{B_a}{f_a}F = 9.8 \text{ Gc}.$$

Under these conditions and with the noise source employed (14,500°K), the equivalent power kTB is 2×10^{-9} watts. In comparison with power levels for coherent sources, the equivalent power is small but useful if a sensitive detection system is employed.

Experimental Results

Modulation Device

Fig. 2 illustrates the microwave circuit of the modulation device as first constructed and operated. The moving short circuit was driven back and forth in a waveguide at a nearly constant speed except for a short reversal time. The period of the mechanical motion was two seconds, *i.e.*, pulses occurred at the rate of one per second. With this drive mechanism, tuning to a desired microwave frequency of observation was accomplished by adjusting the length of the travel of the short without changing the period. Length of travel was continuously adjustable over the range of ½ inch to 6 inches; the corresponding velocities with the short circuit mounted in RG98/U waveguide provide 30-cycle Doppler modulation for wavelengths as long as six mm and extending throughout the lower millimeter region. Thus, a noise source, a 30-cycle amplifier, a waveguide system similar to that illustrated in Fig. 2, in conjunction with appropriate detectors and noise sources could, in principle, be used to make measurements over all of the lower millimeter region.

[1] This approximation is usually valid for the millimeter region; it causes only a small error in calculations for the submillimeter region providing low temperatures are not involved, and it is not usually valid for infrared. For example, see G. R. Nicoll, "The measurement of thermal and similar radiations at millimeter wavelengths," *Proc. IEE*, vol. 104, pp. 519–527; September, 1957.

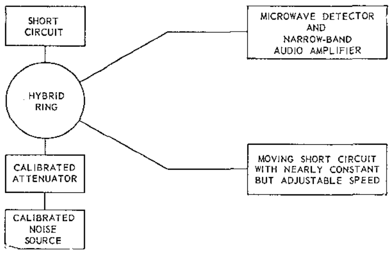

Fig. 2—Block diagram of interference modulation device as first constructed.

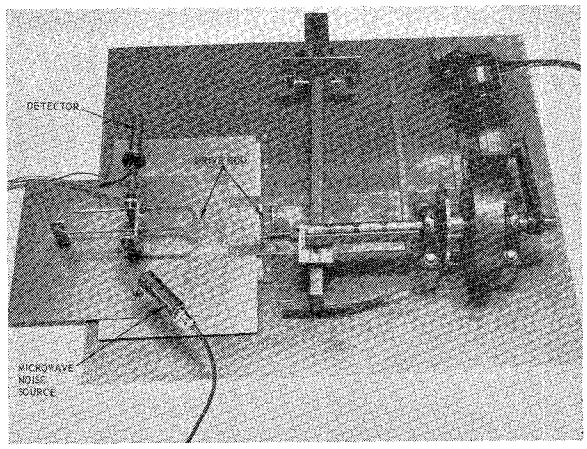

Fig. 3—Trombone version of interference device.

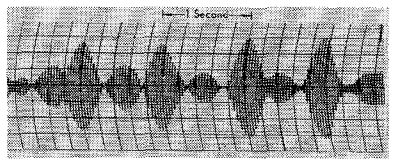

Fig. 4—Amplifier output vs time.

The system as outlined above was operated with a commercially available four-mm noise source[2] and a PRD-634 barretter mounted in an evacuated PRD-632 holder. System performance was poor outside the design region of the hybrid ring, and a somewhat wider bandwidth was later obtained by replacing the hybrid ring with a magic T.

Fig. 3 shows a new system which has been operated over the 50–90 Gc region with the four-mm noise source and evacuated barretter. A trombone-type device provides the variable path length required for the circuit illustrated in Fig. 1. Power is divided by waveguide T's, and path length is changed with a sliding trombone made of circular waveguide.

The power dividers are E-plane T's in which the output branches near the junction are half the height of the input branch. The output branches are gradually tapered up to the size of RG 98/U guide. The movable section of the trombone is constructed of round tubing with an inside dimension of 0.189 inch. Tubing was used because of the availability of slide fit telescoping brass tubing and the relative ease of fabricating a bend. Operation of the system could be limited by the presence of higher order modes that might be generated within the bend. For this reason, a rectangular trombone was constructed. The movable section was fabricated of RG 98/U waveguide and it slides over rectangular waveguide fabricated of 0.003-inch brass stock. No discernible differences were observed between the system performance obtained with the circular and rectangular trombone sections.

Fig. 4 is a recording of output voltage vs time of the audio amplifier which had a center frequency of 28.5 cps and a 4-cps bandwidth. The trombone modulator, the Roger White noise source, and an evacuated barretter were used to obtain Fig. 4. The waves under the envelopes with large peaks occurring at one-second intervals are produced by the interference phenomenon; the waves under the smaller (and unevenly spaced) envelopes are produced by amplifier and detector noise that passes the tuned amplifier. As expected, a change in temperature of the noise source changes the average level of the periodically occurring envelopes but does not modify the average for the envelopes produced by amplifier and detector noise. Eq. (3) describes the waves produced by the interference phenomenon; by definition t equals zero in (3) each time a large peak is produced by constructive interference.

System resolution can be determined from Fig. 4 without explicit knowledge of amplifier bandwidth or center frequency. There are about 14 cycles under the major envelopes of Fig. 4; thus, (3) indicates that $2F/B$ is 14, or that the ratio of center microwave frequency to microwave bandwidth is approximately seven. Parenthetically, note that the indication of an average value meter depends only on envelope peak amplitude because envelope width is controlled only by audio amplifier bandwidth. Since $2F/B$ is equal to $2f_a/B_a$ (5), it can be seen that envelope width is $2/B_a$ seconds.

Fig. 5 includes data measured with the trombone system, the Roger White noise source, the evacuated barretter, and the audio amplifier described above. The output of the amplifier was observed on a meter whose deflection is proportional to the average value of audio amplifier output voltage. Since the pulse lengths are independent of microwave frequency, average voltage output depends on microwave power contained in the bandwidth corresponding to the system resolution, 14 per cent for the operating conditions described above. Care should be taken in interpreting Fig. 5 because 1) output voltage is proportional to input microwave power (for example, a change of 10 db in amplifier output corresponds to a change of 5 db in RF input), and 2) a constant percentage system resolution means that resolution expressed in gigacycles (and thus available noise power) is directly proportional to center microwave operating frequency.

[2] Noise source Model GNW-V18, for which the noise temperature was given as 14,500°K ± 1800°K in private communications with R. White, Roger White Electron Devices, Inc.

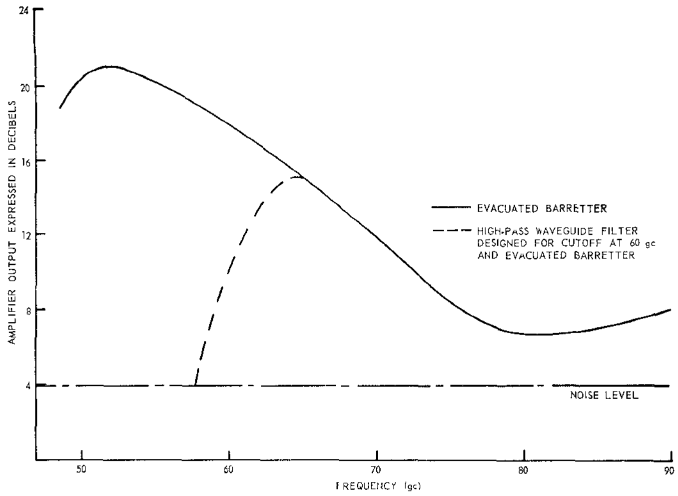

Fig. 5—Response of system to transmission through high-pass filter.

The solid curve of Fig. 5 indicates signal-to-noise ratio at the amplifier output vs frequency with a PRD-634 barretter mounted in a fixed tuned, evacuated holder (PRD-632). This curve provides data required to determine[3] relative detector sensitivity vs frequency. Similar detector measurements have also been made with a tunable holder developed at the Georgia Institute of Technology, Atlanta. If tuned for maximum response at each frequency, sensitivities can be made comparable to that at 55 Gc (Fig. 5) for frequencies up to approximately 75 Gc. This type of measurement would be difficult to accomplish by conventional techniques.

The dashed curve of Fig. 5 was measured by inserting a waveguide designed[4] for cutoff at 60 Gc in the setup used to obtain the solid curve; thus, transmission loss introduced by the waveguide can be determined by taking the difference between the two curves. Because the detector operates in the square-law region, waveguide insertion loss expressed in decibels is one half the difference between the two curves. Within a first approximation, the waveguide cutoff frequency is that frequency for which the difference in the two curves is 6 db, and system resolution is two times the difference between the 6-db frequency and the frequency for which insertion loss is essentially zero. System resolution obtained in this manner indicates a bandwidth of about 9 Gc, which is in approximate agreement with that predicted by (5).

In conclusion, the experimental system shown in Fig. 3 was constructed in accordance with the diagram of Fig. 1. Fig. 5 illustrates average amplifier output obtained with the system as a function of trombone speed—alternatively, average output voltage as a function of center frequency for a fixed percentage bandwidth. The solid curve contains data on detector sensitivity vs frequency; the dashed curve was obtained with a high-pass waveguide filter inserted at the output of the noise tube. Thus, the difference between the solid and dashed curves provides a direct measurement of microwave resolution without the use of a coherent source. Percentage microwave resolution determined by this method agrees with the resolution calculated from Fig. 4 (same as percentage bandwidth of the audio amplifier).

Detector Sensitivity Measurements

The interference device provides a convenient instrument for measuring the absolute sensitivity of detectors operating in the square-law region. This has previously been a formidable task because sensitive power standards and coherent sources to cover a wide spectrum are expensive and sometimes difficult to obtain.

In certain cases, detector sensitivity may be too poor to measure by the direct method that will be described below. In that case, it is possible to measure the output power level from a harmonic generator with the interference device, and determine sensitivity by usual techniques. This in itself provides some extension in the state-of-the-art because the interference device will serve as a highly sensitive power meter.

For this paper, minimum detectable change in power $\Delta P_{\min}$ is defined as that change in signal power for which the rms value of signal voltage out of the tuned amplifier is equal to the rms value of output noise. From (4) the rms voltage out of a unity gain amplifier near a peak in envelope from the interference modulation device is $(R/2\sqrt{2})kTB$. For the purpose of using the interference device to determine detector sensitivity applicable to other systems, let S/N represent the ratio of rms value of signal voltage to rms value of noise voltage V_N. Then we have

$$\frac{R}{2\sqrt{2}}kTB = \left(\frac{S}{N}\right)V_N. \tag{7}$$

Calculations for $\Delta P_{\min}$ made below will be applicable to square-wave modulation, whether it be random noise square-wave modulated at the rate f_a, or coherent power that is square-wave modulated at the rate f_a. The rms voltage out of a unity gain amplifier tuned to the fundamental frequency is $(\sqrt{2}/\pi)v_{\text{peak}} = (\sqrt{2}/\pi)R\Delta P$, where v_{peak} is the magnitude of the square-voltage pulse out of the detector. $\Delta P_{\min}$ for square-wave modulation can be expressed in terms of rms noise voltage as

$$\frac{\sqrt{2}}{\pi}R\Delta P_{\min} = V_N. \tag{8}$$

Substituting the value of V_N from (7) into (8) gives

$$\Delta P_{\min} = \frac{\pi}{4}kTB\left(\frac{N}{S}\right). \tag{9}$$

[3] For example, minimum detectable temperature at 50 Gc is 6 db smaller than at 75 Gc. Thus, since percentage resolution is constant, minimum detectable power at 50 Gc is approximately 8 db smaller than at 75 Gc.

[4] A one-inch length of 0.099- by 0.074-inch guide was used to obtain a high-pass filter with a cutoff at 60 Gc. One-inch linear H-plane tapers are used as transitions to RG 98/U dimensions.

Fig. 6 shows a recording of system output consisting of several sweeps of the type shown in Fig. 4. Parameters pertinent to the system sensitivity are as follows:

$$T = 14{,}500°\text{K} \qquad f_a = 28.5 \text{ cps}$$
$$F = 55 \text{ Gc} \qquad B_a = 4 \text{ cps.}$$

From (5) and (6), we find that B is 7.7 Gc and kTB is 1.5×10^{-9} watts.

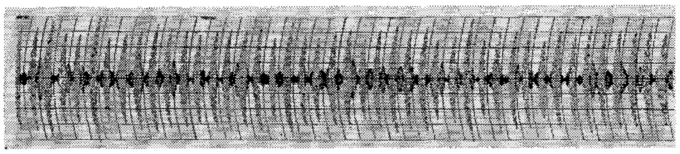

Fig. 6—Amplifier output vs time (one second per division).

Measurements of the positive peaks of Fig. 6 indicate an average height of 3.1 divisions with an rms fluctuation of 0.37 division. The rms signal corresponding to an average height of 3.1 divisions is 3.2 divisions. Thus S/N is 5.9. Then from (9),

$$\Delta P_{\min} = \frac{\pi}{4} 1.5 \times 10^{-9}\left(\frac{1}{5.9}\right) = 2 \times 10^{-10}W.$$

For a detector operating in the square-law region, minimum detectable power $\Delta P_{\min}$ varies as

$$\Delta P_{\min} = K\sqrt{B_a},$$

where K is the minimum detectable change in power referred to a one-cps audio bandwidth and B_a is the audio bandwidth. Since the audio bandwidth used is 4 cps, minimum detectable change in power referred to a one-cps audio bandwidth is approximately 10^{-10} watt. The measurement described above includes ohmic losses in the waveguide and losses due to mismatches at the various junctions. An estimate based on limited measurements of the average loss in the waveguide over the pass band of 2–3 db gives a minimum detectable power referred to a one-cps audio bandwidth for the evacuated barretter of 6×10^{-11} watts.

Detectors

Evacuated PRD-632 barretters are used because they are more sensitive than individual crystal diodes (IN53, IN2792, and MA441) available to us; in fact, crystal diodes tested are too insensitive to detect the signal from the interference device. Sensitivity measurements were made at 70 Gc with a harmonic generator. The diode units, when tuned for optimum performance, are 10–18 db (minimum detectable power) less sensitive than evacuated barretters. Furthermore, diodes and their holders are more frequency sensitive than is desirable for this application. For the comparative measurements, the barretter was mounted in a tunable holder developed at Georgia Tech.

The input circuit for the 30-cps amplifier uses a Triad model G10 transformer. For these measurements, no dc bias was used with the diodes, and the diodes were connected so that their load impedance was that of the transformer primary. DC bias did not improve sensitivity for the crystal diodes because 1/f noise is excessive at 30 cps.

Improvements of approximately 14 db in minimum detectable power have been obtained with PRD-634 barretters mounted in PRD-632 holders which have been evacuated to a pressure of 50 μ or less; this is a somewhat greater improvement than has been reported[5] for an X-band barretter. Minimum detectable power for evacuated barretters has been as low as 6×10^{-11} watts (Fig. 6), but the barretter presently in use is about one-db less sensitive. Results obtained with the currently used barretter and the 30-cps amplifier, which has a noise figure of about 2 db, are given below. The barretter is biased with direct current to an operating resistance of about 200 ohms; when evacuated, biased current is reduced from the nominal four ma to one ma to prevent burn-out and to maintain resistance at 200 ohms. Noise power contributed to amplifier output by the barretter is 20 db for atmospheric pressure and is reduced to seven db when evacuated. The above data indicate an improvement in signal-to-noise ratio of 28 db, and 15 db of this improvement results from increased responsivity caused by evacuation.

Power Measurements

As previously noted, the magnitude of the peaks in the amplifier waveform are directly proportional to the noise source power contained in a microwave band determined by the bandwidth of the audio amplifier. Since the magnitude of the peaks in the amplifier output caused by a coherent source is also proportional to source power, comparison of peak signal with that caused by a calibrated noise source should provide a convenient method of determining the power level of a weak coherent source. Power meters are sometimes required which are far more sensitive than the calorimetric type[6] currently used. An example of a severe power level problem is that of determining the power available from crystal harmonic generators.

Video detectors are more sensitive than water calorimeters, but normally they are not used for absolute power measurements because of calibration difficulties. The power-measuring technique described herein uses the sensitivity available from video detectors, and calibration is obtained by comparison with a noise source of known temperature. It is possible to specify available power from knowledge of source temperature because the interference device serves as a band-pass filter for

[5] M. W. Long and W. K. Rivers, Jr., "Submillimeter wave radiometry," Proc. IRE, vol. 49, pp. 1024–1027; June, 1961.

[6] See, for example, J. B. Thaxter and J. McGowan, III, "100–200 kMc water calorimeter," *Rev. Sci. Instr.*, vol. 32, p. 605; May, 1961.

microwaves. Thus, the power level of a coherent source can be determined by comparing the relative levels of the peaks caused by the coherent source (many peaks will occur periodically within a single sweep for this mode of operation) with those caused by the noise source.

It should be possible to develop a simple, highly sensitive power meter based on interference modulation. Variability of detector characteristics could be calibrated out if a noise source were made an integral part of the instrument. However, care must be taken in accounting for reflection losses because the average loss for a band of frequencies may differ from that for a single frequency. The peaks of Fig. 6 were produced by an input power of 1.5×10^{-9} watts. With a power level of one mμw, the "probable" error[7] in the peaks due to noise would be about $\frac{1}{2}$ db.

Effects of Higher Order Waveguide Modes

Thus far the analysis has been made as if the noise source radiates only the dominant mode (TE_{01}), and with the assumption that the radiation temperature is independent of frequency. Errors can be introduced in the system as a result of the noise source radiating higher order modes or by higher order modes being generated from the dominant mode by waveguide discontinuities. In principle, each mode generated by the source can have a radiation temperature which approaches source temperature. However, since gaseous noise sources are designed to maximize radiation temperature for the dominant mode, it would seem likely that the temperatures for higher modes are substantially less than that of the dominant mode. The temperatures for higher order modes generated by waveguide discontinuities are less than the temperature of the dominant mode because the higher order mode energy is obtained by conversion from the dominant mode.

Fig. 7 shows guide wavelength for RG98/U (inside dimensions of 0.148×0.074 inch) vs frequency for several modes. For each mode there exists a frequency for which guide wavelength is equal to guide wavelength for the dominant mode at a lower frequency; thus, there are Doppler frequencies associated with higher modes which are produced by path length changes that are equal to those for the dominant mode. Eq. (5) is invalid near cutoff because λ/λ_g changes rapidly within the bandwidth B. The rapid change in λ/λ_g near cutoff results in a smaller microwave bandwidth for a given audio bandwidth. Higher order contributions are also reduced because waveguide attenuation is large for modes near the cutoff frequency.

[7] For the peaks of Fig. 6, the average is 3.1 units and the standard deviation (independent of microwave power) is 0.37 unit. Thus probable error, defined as 0.67 times standard deviation for a Gaussian distribution, is 0.25 unit. Since meter deflection is proportional to microwave power, a one-mμw power level would produce peaks with an average level of 2.0 units; thus, the probable error would be about 1/2 db.

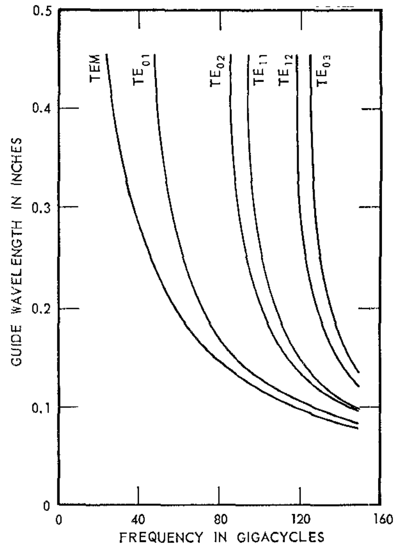

Fig. 7—Mode chart for RG98/U.

The condition of simultaneously having the lengths of Paths 1 and 2 in Fig. 1 equal electrically for all modes is satisfied only if the paths are physically identical. For example, if the trombone (part of Path 1) is constructed of circular guide and Path 2 is made entirely of rectangular guide, peaks for higher modes will be displaced from the strong, dominant mode peak. It is conceivable that such a displacement would be useful for affecting a significant minimization of higher mode contributions. Two trombones have been used with the system described herein, one was constructed from RG98/U and the other constructed from circular waveguide. No differences have been observed between the results obtained from the two trombones; thus, the contributions due to higher order modes are not prohibitive for this system.

Probably the greatest minimization of the effects of higher order modes is due to detector design. The barretter and other waveguide mounted detectors are designed for optimum performance for the TE_{01} mode. It is expected that such detectors are quite insensitive to higher order modes.

Conclusions

An interference modulation technique for microwave radiometry is described. Operation is based on the fact that there is a one-to-one correspondence between mechanical speed and frequency of Doppler components. One of the attractive features of the technique is that it is applicable to wavelengths throughout much of the low millimeter region where coherent sources are virtually nonexistent. The complete system consists of simple microwave components, a detector, and a tuned audio amplifier.

A radiometer using interference modulation has been operated as a measurements system within the 50–90 Gc region. Microwave performance of this instrument is consistent with theoretical predictions. A commercially available noise source having a known temperature was used as the source of microwaves for determining the sensitivity of detectors as a function of frequency. It has been found that the sensitivity of a commercially available barretter evacuated so that the air pressure is 50 μ or less is approximately 6×10^{-11} watts for an audio bandwidth of one cycle per second. From the detector investigation it has been learned that signal-to-noise ratio at the detector output is improved by 28 db by evacuating the barretter, and that minimum detectable power for evacuated barretters is 10–18 db smaller than for available crystal diodes.

The interference modulation techinique provides the opportunity for developing a highly sensitive power meter for coherent sources. However, there are numerous details to be solved before a simple, highly sensitive power meter can be made commercially available. Present indications are that it should be possible to develop a calibrated power meter for millimicrowatt levels from coherent sources.

Appendix

Microwave Power at Detector vs Path Length

Assume that power from the microwave source of Fig. 1 is divided equally between Paths 1 and 2, and is recombined before reaching the detector. Let the difference in lengths of Paths 1 and 2 be designated δ. Then for any single microwave frequency, power reaching the detector, P_d, is expressed in terms of source power, P_s, is

$$P_d = \frac{P_s}{2}\left[1 + \cos\frac{2\pi\delta}{\lambda_g}\right], \tag{10}$$

where λ_g is the guide wavelength corresponding to the microwave frequency f. For an air-filled waveguide, guide wavelength and frequency are related by the well-known equation

$$\lambda_g = \frac{c/f}{\sqrt{1 - \left(\frac{c/f}{\lambda_c}\right)^2}}, \tag{11}$$

where c is free space propagation velocity and λ_c is guide cutoff wavelength. Notice that the detector power is equal to the total source power when the path difference is zero, and that it changes periodically with path length.

If the microwave source is a noise tube filtered so that only frequencies within a bandwidth B about the center frequency F are passed, power incident upon the detector can then be expressed as

$$P_d = \frac{k}{2}\int_{F-B/2}^{F+B/2} T(f)\left[1 + \cos\frac{2\pi\delta}{\lambda_g}\right] df, \tag{12}$$

where k is Boltzmann's constant and $T(f)$ is effective noise temperature which may depend on frequency. Integration of (12) is not simple because of the dependence of λ_g on f. It may be seen, however, that for the special case in which path length δ is zero, detector power is equal to total source power.

Guide wavelength λ_g can be expressed in terms of free space propagation velocity and frequency as

$$\lambda_g = \left(\frac{\lambda_g}{\lambda}\right) c/f.$$

The factor λ_g/λ is always greater than unity in waveguide, and is less than 1.7 within the recommended bands for the various standard waveguides. Since pass bands of interest are narrow (B/f small), it is reasonable to assume that λ_g/λ and $T(f)$ are constant over B. Then (12) becomes

$$P_d = \frac{kT}{2}\int_{F-B/2}^{F+B/2}\left[1 + \cos\frac{2\pi\delta}{c}\left(\frac{\lambda}{\lambda_g}\right)f\right] df. \tag{13}$$

The result of integrating (13) is

$$P_d = \frac{1}{2}kT\left\{B + \frac{c}{2\pi\delta}\left(\frac{\lambda_g}{\lambda}\right)\left[\sin\frac{2\pi\delta}{c}\left(\frac{\lambda}{\lambda_g}\right)(F + B/2) - \sin\frac{2\pi\delta}{c}\left(\frac{\lambda}{\lambda_g}\right)(F - B/2)\right]\right\}. \tag{14}$$

Use of the identity

$$\sin(\alpha + \beta) - \sin(\alpha - \beta) = 2\cos\alpha\sin\beta$$

in (14) gives

$$P_d = \frac{1}{2}kTB\left\{1 + \frac{\sin\frac{2\pi\delta}{c}\left(\frac{\lambda}{\lambda_g}\right)\frac{B}{2}}{\frac{2\pi\delta}{c}\left(\frac{\lambda}{\lambda_g}\right)\frac{B}{2}} \cdot \cos\frac{2\pi\delta}{c}\left(\frac{\lambda}{\lambda_g}\right)F\right\}. \tag{15}$$

Notice that (15) indicates that P_d is equal to the total source power kTB only if the path length difference δ is zero. From (10), for a coherent source, it was seen that detector power is equal to total source power if δ is zero, but it also reaches this value periodically with changes in δ. For both types of microwave sources, detector power consists of a constant term which is equal to one half the source power plus a term which depends on δ. For the incoherent source and the limit of large path differences, (15) indicates that detector power is equal to the constant term only, *i.e.*, one half the total source power.

Assume that the difference in paths δ is varied at a constant speed s. Let time t be defined as zero when δ is zero, *i.e.*, let δ be expressed as $\delta = st$. Then (10) becomes

$$P_d = \frac{P_s}{2}\left[1 + \cos\frac{2\pi st}{\lambda_g}\right] = \frac{P_s}{2}\left[1 + \cos\frac{\omega_a t}{\lambda_g}\right]. \tag{16}$$

The interference modulation rate for a band of frequencies which is produced by a constant change in δ may be obtained from (15). It can be seen that the term $1/c(\lambda/\lambda_g)F$ is equal to λ_g^*, where λ_g^* is the guide wavelength corresponding to the microwave center frequency F. Thus, for uniform change in δ, the microwave power incident upon the detector may be expressed as

$$P_d = \frac{1}{2} kTB \left\{ 1 + \frac{\sin\left[\left(\frac{\omega_a t}{F}\right)\frac{B}{2}\right]}{\left[\left(\frac{\omega_a t}{F}\right)\frac{B}{2}\right]} \cos \omega_a t \right\}. \qquad (17)$$

Notice that the magnitude of the periodic peaks caused by a coherent source of power P_s [see (16)] are the same as the peak for an incoherent source [$t=0$ in (17)] for which kTB is equal to P_s.

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Superheterodyne Radiometers for Use at 70 Gc and 140 Gc*

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Summary—In this paper four different millimeter wave equipments, which have been made for plasma diagnostic work, are described. They are:

1) A straightforward 70-Gc superheterodyne radiometer with an over-all noise factor of 13 db;
2) An early 140-Gc radiometer, with second harmonic mixing, which has an over-all noise factor of about 25 db;
3) A later and more sensitive 140-Gc radiometer which contains a fundamental local oscillator, VX 3352 mixer crystals and a 408-Mc IF amplifier commencing with an Adler tube;
4) A very simple 140-Gc transmission measuring equipment containing a 1-watt source and a crystal video receiver which has a tangential sensitivity of −42 dbm.

The last part of this paper discusses the minimum temperature changes which can be detected, at short millimeter wavelengths, with various types of superheterodyne radiometers, the Golay cell, the barretter, the crystal video radiometer, the 1.5°K carbon bolometer and the 1.5°K InSb photoconductive detector. The performances expected from straight traveling-wave tube radiometers and traveling-wave masers at short millimeter wavelengths are also considered.

The Appendices are devoted to mixer crystal performance in the millimeter and submillimeter regions, a theory of second harmonic mixing and the voltage sensitivity of a forward biased detector crystal.

Introduction

A TOROIDAL shaped fusion machine called ZETA was built at the Atomic Energy Research Establishment (AERE), Harwell a few years ago [1]. The pulses applied to this machine last for a few milliseconds and the electron densities produced in the plasma are in the region of 10^{14} per cm^3. The plasma frequency therefore lies in the gap between the microwave and infrared regions. In 1958, there was some doubt about the value of the temperature being reached inside ZETA and an urgent need arose for sensitive radiometers suitable for making temperature measurements on ZETA in the neighborhood of the plasma frequency.

The Royal Radar Establishment (RRE) was asked to tackle this problem from both the microwave and infrared sides of the gap. This paper describes the extension of superheterodyne radiometer techniques into this part of the spectrum. The outstanding progress made by the RRE infrared team has already been described elsewhere [2]–[5]. Considerable overlapping of radio and infrared techniques has now occurred.

In addition to needing sensitive radiometers at short millimeter wavelengths, a need also arose at AERE for equipments suitable for making transmission and phase shift measurements through plasmas at various wavelengths throughout the millimeter region. One equipment falling into this category is also described in this paper.

Theoretical Performance of Mixer Crystals in the Millimeter and Submillimeter Regions

At short millimeter wavelengths, the performance of a superheterodyne radiometer is mainly determined by the mixer crystals, so four years ago an attempt was made to work out the over-all noise factors which it should be possible to obtain with silicon, germanium and gallium arsenide mixer crystals, over the wavelength

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