Correspondence

Three Microwave Calibration Services Extended: Noise Sources, Attenuation, Reflection Coefficient

SUMMARY

The Radio Standards Laboratory of the NBS Institute for Basic Standards, Boulder, Colo., has announced three changes in the microwave calibration services it offers: 1) Calibration of noise sources has been extended to WR62 waveguides. 2) Calibration of coaxial attenuators and couplers has been pushed beyond the former ceiling of 12 GHz up to 18 GHz. 3) Measurements of reflection coefficient magnitude are now also available for WR137 waveguides.

The following three changes in microwave calibration services were recently announced by the Radio Standards Laboratory, Boulder, Colo. The RSL is part of the Institute for Basic Standards of the National Bureau of Standards (U. S. Department of Commerce).

MICROWAVE NOISE SOURCES

The Radio Standards Laboratory announces that the measurement of effective noise temperature of waveguide noise sources is now available in WR62 waveguides (frequency range 12.4 to 18.0 GHz). Suggested calibration frequencies are at 13.5, 15.0, and 17 GHz. For noise sources with an effective noise temperature of approximately 11 000 °K and a reflection coefficient magnitude of 0.09, or smaller, the calibration will be reported to an uncertainty of ±250 °K. The calibration system can measure noise sources with effective noise temperatures in the range of 1000 to 300 000 °K.

The noise source must be fitted with an output connector that is compatible with the UG 419/U waveguide flange connector. Direct current for operation of the gas discharge tube should be selected and specified within the range of 1 to 300 mA. With the noise source in an operating condition, the source must have an input reflection coefficient magnitude no greater than 0.09 (i.e., approximately VSWR<1.2).

COAXIAL ATTENUATION SERVICES EXTENDED TO 18 GHz

Radio Standards Laboratory calibration services for the measurement of attenuation of coaxial attenuators and couplers have been extended continuously in frequency to 18 GHz from the former range of 200 MHz to 12 GHz. Measurements are made over an attenuation range of 60 dB, with an uncertainty in measurement not exceeding 0.2 dB/10 dB for attenuators and couplers having a VSWR of 1.3 or less.

Attenuators and couplers submitted for calibration should be equipped with Type-N connectors complying with the MIL C 39012 specification, or with the new precision 7-mm connectors. The critical mating dimensions used by NBS are as shown in Fig. 1.

Equipment fitted with connectors designed for 0.141-inch (3.5-mm) coaxial lines will be measured with an uncertainty determined by the stability and VSWR of each unit.

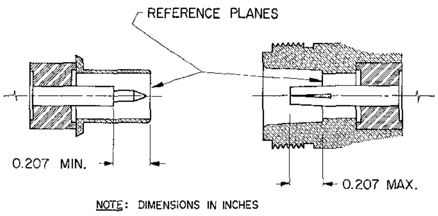

Fig. 1. Critical connector dimensions for coaxial attenuation calibrations.

WAVEGUIDE REFLECTORS CALIBRATED IN WR137 WAVEGUIDES

Measurements of the reflection coefficient magnitude of waveguide reflectors (mismatches) in WR137 waveguides (5.85–8.20 GHz) has been announced as a new calibration service by the Radio Standards Laboratory. A calibration service in two other waveguide sizes, WR90 and WR62, is also available.

The calibration can be performed at any frequency within the range of 5.85 to 8.20 GHz. Some degree of economy to the customer results if calibrations are requested at the selected frequencies of 6.45, 7.00, and 7.40 GHz. Measurements can be made over a range of 0.024 to 0.2, with an uncertainty of the reflection coefficient magnitude expressed as $\pm(0.0002+0.002\,|\Gamma|)$, where $|\Gamma|$ is the numerical value of the measured magnitude.

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Determination of Equivalent Network Parameters for a Lossless Two-Port Junction

We wish to point out a graphical method for determining the parameters of a lossless two-port network, which we have found useful when the number of available data points is small. This situation may arise when the output line is not easily terminated by an adjustable short circuit, which is the case, for example, for a radial waveguide.

In the usual way,[1] measurements of the admittance are made in the input line for short circuits at various locations in the output line. If losses in the two-port and the load are neglected, the output susceptance B_o at an arbitrary output reference plane T_o is related to the measured input susceptance B_i by

$$B_o = -B_{22} - B_{12}^2(B_i - B_{11})^{-1} \tag{1}$$

where B_{11}, B_{22}, and B_{12} are the network parameters in an admittance representation ($Y_{11}=jB_{11}$ etc.). A plot of B_o, which can be calculated using transmission line theory, against $(B_i-B_{11})^{-1}$ should, therefore, be a straight line with slope $-B_{12}^2$ and intercept $-B_{22}$. An initial value for B_{11} is easily obtained if the output reference plane can be chosen to be at one of the short circuits, so that B_o is infinite and B_{11} equals B_i for that short circuit. Otherwise, a value which gives a straight line plot must be found by trial.

In order to check the accuracy of the network parameters obtained in this way, one may rewrite (1) as

$$B_i = B_{11} - B_{12}^2(B_o + B_{22})^{-1} \tag{2}$$

and plot B_i against $(B_o+B_{22})^{-1}$, using the value of B_{22} from the intercept on the previous graph. A straight line should again be obtained, with slope also $-B_{12}^2$ and with intercept B_{11}. Small changes can now be made to B_{11} and B_{22}, and the data replotted until both graphs yield good straight lines with identical slopes and intercepts consistent with the values chosen for B_{11} and B_{22}.

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[1] L. B. Felsen, *Handbook of Microwave Measurements*, vol. 1, Brooklyn, N. Y.: Polytechnic Press, 1963, p. 227.

On High-Frequency Stability of Tunnel Diode Amplifiers

It is well recognized that in a shunt tuned tunnel diode amplifier, where the tunnel diode is parallel resonated by an inductance at band center, stability at high frequencies requires that diode series resonant frequency be greater than its cutoff frequency. In terms of diode equivalent circuit elements, this implies that, for stability,

$$R_s > \frac{L}{R_N C_j} \tag{1}$$

where R_s is series resistance, L is series inductance, C_j is junction capacitance, and R_N is negative resistance. This condition implies, and is implied by the condition that the diode be short-circuit stable.

In a series tuned amplifier,[1] where induc-

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[1] H. Plutchok, "Octave-bandwidth tunnel diode amplifier design," *Proc. National Electronics Conf.*, vol. 21, 1965, pp. 119–124.

tance is added in series with the diode to produce resonance near band center, the diode is effectively decoupled from the external circuitry at high frequencies. Hence, the stability condition must be that the diode be open-circuit stable. This implies that

$$C_p\left(\frac{L}{R_N C_j} - R_s\right) < \frac{R_s R_D C_j}{R_D - R_s} \qquad (2)$$

where C_p is package capacitance, a condition which is more easily met than is (1). In fact it is easily seen that (1) guarantees (2), but not vice versa. Physically this is understood by observing that resonance with C_p must occur above diode series resonance.

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X To K Band Broadband Varactor Frequency Doubler

This correspondence describes a broadband X to K band varactor frequency doubler. Instantaneous bandwidths of 12 percent at 20 percent efficiency or 19 percent at 2 percent efficiency were obtained. Typical bandwidths of varactor multipliers at these frequencies had been reported at about one percent. Broadband varactor frequency multipliers with C-band or lower output frequencies, designed in coaxial or strip transmission line circuits, had been reported by C. L. Cuccia [1] and by R. J. Wenzel [2].

Our X to K band doubler is designed in the waveguide circuit, as shown in Fig. 1.

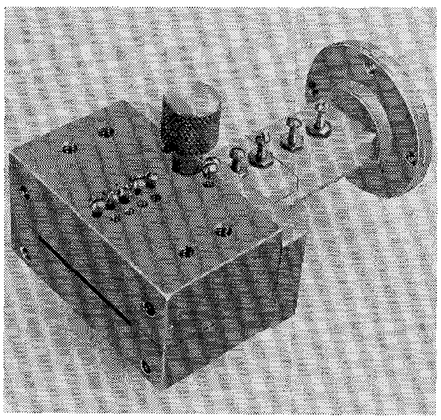


Fig. 1. Doubler with tuning screws in waveguide section.

The input circuit consists of a reduced-height WR90 waveguide section with several tuning screws preceding the Sylvania D5047B varactor. The floating-bias arrangement was used—that is, one end of the varactor open-circuited. The output waveguide WR42 contains a three-quarter wavelength section stepped-impedance transformer (binomial characteristic impedance distribution) with trim-tuning screws. A variable series capacitance tuner is located at the open end of the varactor. The input to the doubler was provided by an X band frequency sweeper followed by a TWT amplifier. The power inputs were in the 15 to 160 mW range with a ± 0.5 dB maximum ripple in the doubler input band. The doubler was tuned during the tests to various frequencies, and different input-power levels were used. The results in terms of bandwidth vs. efficiency are shown in Fig. 2, where the output frequency bands and input power levels are indicated. A typical output power vs. frequency curve is shown in Fig. 3 (10 mW output, 16 to 18 Gc/s band). The skirts of the doubler frequency band fall off sharply at the edges and no off-band responses have been observed. The ripple in the pass band seen in Fig. 3 was due in part to slight variations of input-power level. The doubler output bandwidth was sensitive to input-power levels, but could be corrected by slight retuning of the input screws.

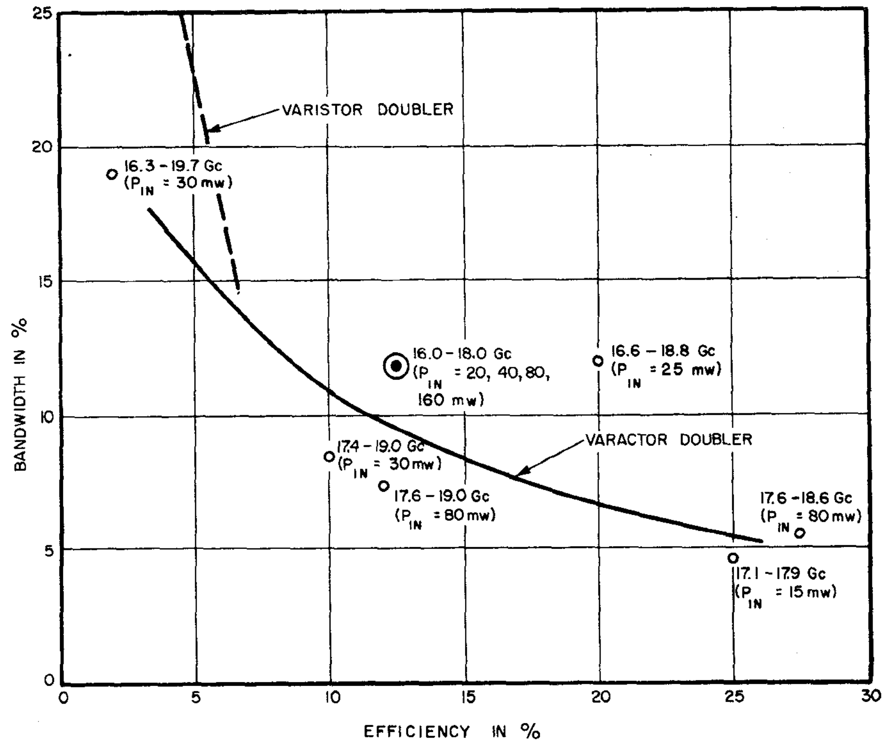
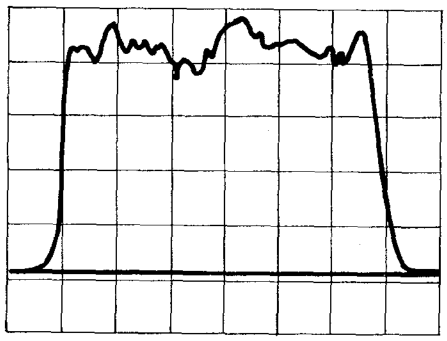


Fig. 2. Bandwidth vs. efficiency.



OUTPUT FREQUENCIES 16.08 – 18.08 Gc
INPUT POWER (AVG) – 80 mw
OUTPUT POWER (AVG) – 10 mw
BANDWIDTH = 11.7 PERCENT
EFFICIENCY = 12.5 PERCENT

Fig. 3. Output power vs. frequency characteristics.

It should be pointed out that all measurements were performed at the input frequency sweep rate between 1 and 100 c/s. Varying the sweep rate normally caused a reduction in bandwidth and/or ripple in output-power level. This was determined to be due to the floating bias arrangement and could be corrected by introducing a constant dc bias voltage across the varactor.

The data in Fig. 2 indicate that a considerable advantage in efficiency can be obtained by replacing the variable resistance element by a varactor in an X to K band frequency doubler when bandwidths of 12 percent or below are required. Theoretical analysis of a simplified equivalent circuit of a doubler indicates that 25 to 30 percent bandwidths with present-day varactors should be possible at these frequencies but may require a different varactor imbedding circuit.

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References

[1] C. L. Cuccia, "Broadband multiplier chains with interdigital filters," *Microwaves*, vol. 3, pp. 22–25, June 1964.
[2] R. J. Wenzel, "Wideband varactor harmonic multipliers," 1965 G-MTT Symposium Program and Digest, Clearwater, Fla.