IEEE TRANSACTIONS ON INSTRUMENTATION AND MEASUREMENT, VOL. IM-35, NO. 4, DECEMBER 1986

1.25-MHz Attenuation Measurement System

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Abstract—A system has been developed to make highly accurate measurements of nominally 6-dB increments of attenuation at 1.25 MHz. Initial experiments indicate a typical systematic error of 5 μB (1 μB = 0.00001 dB) with a resolution of 1 μB. A special linearity measurement system (LMS) using NBS-constructed linear tuned hybrids and power detectors has been used to determine the nonlinearity of a tuned 1.25-MHz power detector. This detector utilizes a single thermistor bead design with thermal isolation to obtain nearly linear tracking of a 4:1 change in input power. The nonlinear correction for this change, determined by the LMS, is on the order of 13 μB for the detector presently in use. This calibrated detector and another of similar design are used in the attenuation measurement system (AMS) to make power ratio measurements to determine the change in attenuation of the device under test (DUT). It is anticipated that changes of approximately 6 dB can be measured with initial insertion loss of up to 100 dB with an accuracy of 0.001 dB. A special design consideration will be required for units used for calibration in order to keep mismatch errors from significantly degrading the accuracies estimated above.

I. Introduction

THERE are several variable RF attenuators and voltage doublers that operate specifically at 1.25 MHz. These devices require calibration [1]. Several methods are available. The majority of these methods utilize reference standards of various types, but a few are self-calibrating. The method we have chosen, for the best possible accuracy, is a self-calibrating attenuation measurement system which measures changes in insertion loss of a 6-dB step.

Two power detectors are used to determine these changes in insertion loss. The ratio of these detector responses rather than the absolute power at the detectors is the parameter of interest. This ratio measures the changes in power readings which occur with the change of insertion loss of 6 dB. For this type of measurement, a high degree of linearity in the response of one of the detectors is required over the range of interest. For the desired accuracy the response of this detector must also be stable, and the precise degree of its nonlinearity must be known.

The first step in finding a suitable detector was to determine what commercial detectors might be available that would satisfy our needs. Dual thermistor bead designs [2] are not sufficiently linear due to bead variations within the individual units [3]. Barretters are too easily burnt out and their availability is poor. Other more exotic designs are too hard to obtain and most are incompatible with the electronics measurement system. A problem common to virtually all commercial detectors when operated at 1.25 MHz is the insufficient filtering of the RF signal in the power meter sense-and-bias leads. Simple capacitive filters do not adequately remove the unwanted signal. This adversely affects measurement accuracy. As no commercial detector can meet the imposed needs of the measurement system, a new detector design was found to be essential.

We have developed a new design which has overcome the problems in the commercial detectors. A single thermistor bead is used for increased linearity. Several tuned LC filter sections have been placed in the power meter bias-and-sense leads; these filters attenuate the unwanted RF signal level by at least 120 dB. Circuit sections have been placed in separate compartments constructed of double-sided copper-clad fiberglass to insure isolation. Finally, the design is simple and easily reproducible, and utilizes components that are readily obtainable.

This new detector design, incorporated in the attenuation measurement system (AMS), provides a measurement of nominally 6-dB changes in insertion loss with an accuracy of approximately 5 μB. The system operates exclusively at 1.25 MHz. The detectors have an input power range of approximately 1–10 mW. The AMS can measure the 6-dB change (plus an additional insertion loss of <20 dB) with a resolution of 1 μB, a random error of less than 1.0 μB, and a systematic error of 5.0 μB. Our data support a 1-μB resolution.

II. System Description and Measurement Procedures

Fig. 1 illustrates the experimental arrangement for measuring the change in attenuation of the device under test (DUT). The signal generator and the bandpass (BP) and low-pass (LP) filters deliver a clean 1.25-MHz signal to the dual detector system. The DUT is any device with a switchable change of approximately 6 dB in insertion loss (or a voltage doubling or halving). A measurement is made by determining the ratio of the response of detector $P3$ to the response of detector $P0$ when the DUT is in its low-loss setting and then determining the same ratio when the DUT is in its high-loss setting. The difference between the two ratios is the change in insertion loss. A constant power level applied to $P0$ is maintained by adding 6 dB of attenuation to the attenuator preceding the reference junction when 6 dB of insertion loss is removed from the DUT. (The amplifier is used to compensate for

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Manuscript received March 25, 1986. This work was supported by the Calibration Coordination Group of the three Armed Services of Department of Defense under Contract CCG-156.

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IEEE Log Number 8610868.

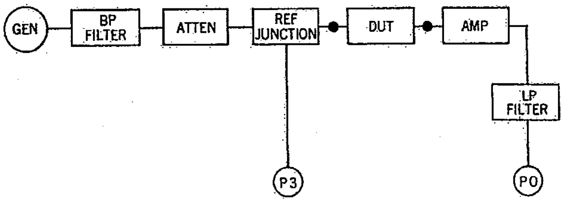

Fig. 1. Block diagram of system to measure attenuation change of DUT with two power meters.

up to approximately 100 dB of additional insertion loss in the DUT). The heart of the measurement system is the detector $P3$. During a measurement $P3$ experiences a 6-dB change in input power. The response of $P3$ is very nearly linear with a known nonlinearity function over this change in input power. Linearity in the response of detector $P0$ is not necessary as the power input to $P0$ is held constant.

The measured attenuation A is defined by

$$A = 10 \log (P3/P0).$$

The change in attenuation ΔA in a system with a nominal impedance of 50 Ω is

$$\Delta A = A_h - A_l$$

where A_h is the value of A at the high attenuation setting of the DUT and A_l is the corresponding value at the low setting. In this case, each value of A is the average of ten measurements of $P3$ and $P0$. $P3$ and $P0$ are measured simultaneously to decrease the sensitivity of the AMS to variations in the power level of the source. The power input to $P0$ is held approximately constant so that

$$P0_h \approx P0_l$$

where the subscripts h and l again refer to the high and low attenuation settings of the DUT, respectively. If the response of detector $P0$ is assumed to be linear over the small range covered by the difference between responses $P0_h$ and $P0_l$, then this difference creates no error.

For one measurement series, ten measurements of A are made in the sequence

$$A_l,\ A_h,\ A_l,\ A_h,\ A_l,\ A_h,\ A_l,\ A_h,\ A_l,\ A_h$$

giving nine values of ΔA. The effect of any drift in RF power is greatly reduced by measuring nine insertion loss differences. The above process can be repeated many times to provide a good statistical basis for the measurements and thereby help to eliminate any random error effects.

III. 1.25-MHz Detector Design

The dc response of detector $P3$ must be nearly linear with changes in input RF power. Since no commercial detectors are available at 1.25 MHz with the necessary linearity and attenuation of the RF signal in the power meter leads, a single 50-Ω thermistor bead design was selected as the more efficient linear detection scheme (Fig. 2). This detector is used in conjunction with an NBS Type IV self-balancing dc-substitution RF power meter modified to bias a 50-Ω detector [4], [5]. The NBS Type IV power meter is designed to change the thermistor bea bias current so that the thermistor always maintains th same temperature (resistance). In addition, the detecto are placed in the linearity measurement system (LMS) housing where the temperature is held to +0.01°F. This design provides good sensitivity. However, it is difficult to filter the RF signal from the power meter leads due to the nature of the single bead. The total RF signal should appear across the bead in the ideal case. Other considerations in the detector design include thermal stability and any forms of RF leakage into or out of the detector.

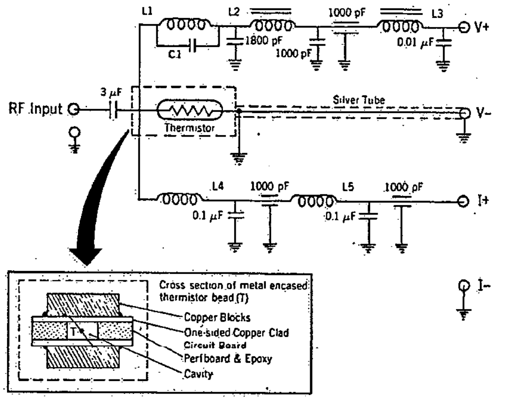

Fig. 2. Circuit diagram of single-element thermistor mount. Notes: 1) Al capacitors are metallized polyester film. Feedthrough capacitors are ceramic. Value of $C1$ is determined in text. 2) Inductors $L1$, $L4$, and $L5$ are air core toroids. $L2$ and $L3$ are ferrite core toroids.

Several ideas are incorporated in the detector to eliminate these problems. LC filter sections are inserted to cut down the RF signal in the power meter leads. No ferrite material (such as a ferrite core inductor) is used near the thermistor, since ferrite components experience changes in impedance values with changes in the RF level. These impedance changes lead to nonlinearities in the detector response. Therefore air core inductors are inserted near the thermistor. Ferrite core inductors are allowable in sections following the first filter section because the RF level is sufficiently reduced and renders any impedance variation negligible. The parallel LC filter is tuned to resonate at 1.25 MHz. A special thermistor mounting structure has been developed (shown in the inset of Fig. 2). This structure consists of an electrically insulated doughnut and two copper blocks, one on either side of the doughnut. The thermistor is placed inside the doughnut "hole" and is encapsulated in an air pocket by the addition of the copper blocks. The copper blocks also provide a large amount of thermal mass so that the entire structure cannot experience rapid changes in temperature. The thermal time constant is much longer than the time required to do a single measurement cycle. Power-off drifts in the dc power meter output of the detector response of approximately 5 μV over 10 min have been obtained. Double-sided copper-

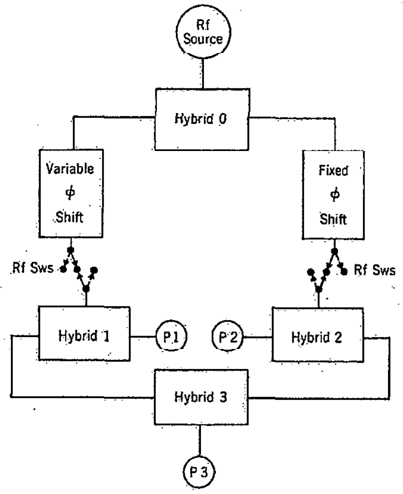

Fig. 3. Simplified block diagram of linearity measurement system.

clad fiberglass boards are used in the exterior detector housing and in the internal compartment walls which separate various filter sections. This type of construction and the use of capacitive feedthroughs in the compartment walls reduce the amount of RF leakage through the detector by at least 120 dB (measured with a lock-in amplifier). These physical and electrical construction considerations provide a stable linear detector with low RF leakage.

IV. Linearity Measurement System

We have stated that our detector is linear or nearly linear. For our measurement system and measurement sequence the extent of the linearity of the detector must be determined. If the nonlinearity of a general detector response is of the form

$$P(\text{true}) = KP_m(1 + k_1P_m + k_2P_m^2 + \cdots) = f(P_m)$$

with P_m being the measured power and the K's being constants, then this type of nonlinearity is easily determined by measurements of the ratio of two detector responses. If the nonlinearity is of the form

$$P(\text{true}) = [f(P_m)]^\gamma$$

then the exponent γ causes problems in the power ratios if both power detectors have the same value of γ. This exponential type of error due to the possible cancellation of γ in ratio measurements requires the use of a more complicated linearity measurement technique.

To determine the nonlinearity of a detector over a nominal 6-dB change in input power at 1.25 MHz, a special LMS was constructed. A simplified block diagram of the LMS is shown in Fig. 3. The LMS was fabricated by hand at NBS, Boulder, Colorado. Double-sided copper-clad fiberglass boards were used in the construction. All circuit elements are tuned to 1.25 MHz. Single-contact mercury-wetted relays were used as switches; with special compensation circuitry in the relay control leads the switches had a repeatability difference of less than 0.5 μB. All impedances presented to the outside world or to the detectors are tuned to $50 + j0\ \Omega \pm 0.2$ percent.

The nonlinearity of a detector can be determined by using different excitation conditions in the LMS. First by applying power to only one branch, then to the other, and then to both branches, the 6-dB power change detected by $P3$ is obtained. A mathematical model relating the relative phase of the two branches and the ratios of detectors $P3$–$P1$ and $P3$–$P2$ under different power excitation conditions is a sinusoidal function. The minimum value of this function is the maximum amount of nonlinearity for the $P3$ detector. The amount of this nonlinearity is 13 μB for the present detector in the LMS. This correction can be applied to data from $P3$ only when $P3$ is used to measure a change of 6 dB in input power.

V. Results

Many step or variable attenuators and voltage doublers operate specifically at 1.25 MHz. We have measured several of these devices. One of these has a switchable step of 6.0 dB. Fig. 4 shows the results of three separate test sequences on this device. Each sequence consists of 27 actual changes in the device attenuation. The circle on the plot represents the corrected mean of the 27 determinations. The error bars represent plus or minus three times the standard error of the 27 values plus an estimated 5.0-μB systematic error. The causes of systematic error include the following: 1) mismatch error due to inexact 50-Ω impedance value of the DUT and AMS; 2) impedance changes in the DUT when the amount of insertion loss is changed, and 3) changes in noise levels when the DUT is present in the circuit or when the insertion loss of the DUT is changed. The dashed lines represent variations of plus and minus 10 μB from the average of the three means. The means of the three runs vary by less than 2.0 μB.

Another set of data was taken on a voltage doubler. In this case a total of 81 determinations was made of the change in attenuation. Fig. 5 shows a plot of observation number versus the measured change in attenuation. All of the values fall within 5 μB of the mean of the 81 values. The standard error of the 81 measurements was 0.17 μB, showing that the systematic error is the dominant error. The system measurements therefore are quite repeatable.

VI. Error Analysis

The quoted systematic error is a combination of several different factors. All known elements of the AMS have been considered and areas of known significant error have been analyzed. There is not sufficient room in this paper to go into the details of this analysis, which is still in a preliminary state.

The areas considered in the error analysis follow with estimates of the respective error values stated.

Error in correction factor determined by LMS due to $P1$ and $P2$ not being identical	0.23 μB

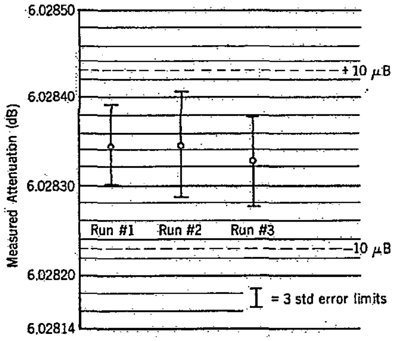

Fig. 4. Typical calibration results for a 6-dB step attenuator.

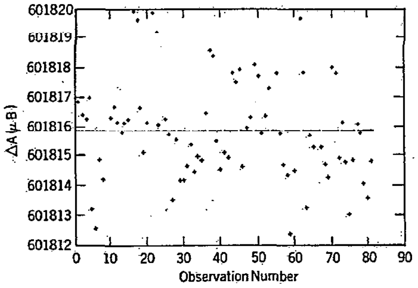

Fig. 5. Plot of measured values of ΔA (μB) for a voltage doubler versus observation number (the broken line is the uncorrected mean).

Variance in input impedance of detectors (assumed 10-percent change)	0.16 μB
Error due to RF leakage	1.0 μB
Changes in power-off voltage levels of 5 μV	0.3 μB
Characteristic impedance mismatch between the DUT states and the measurement system (0.5-Ω impedance difference)	1.0 μB

The sum of these errors is well below 5.0 μB; but due to the preliminary status of the analysis, the value of 5.0 μB for a systematic error was picked as a conservative nominal figure.

VII. CONCLUSION

High-precision attenuation measurements have been made at 1.25 MHz with the dual power detector approach and ratio measurements. The heart of this type of measurement system is a detector of known linearity characteristics. When making measurements on the order of 1.0 μB, thermal instability and RF leakage can have significant effects. Our new detector design has provided the necessary RF filtering and thermal stability. The AMS is a fairly simple concept. It has been found to make very precise attenuation determinations. Work is continuing to improve our determination of the systematic error and to extend the dynamic range of the system.

ACKNOWLEDGMENT

The authors would like to acknowledge the contributions of G. Chamberlain in helping to assemble and test the linearity measurement system and attenuation measurement system, and of G. Rebuldela and E. Livingston in preparing the text and the figures for this paper.

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