

# Microwave Attenuation Measurement by Time-Interval Ratio

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**Abstract**—A new method of measuring microwave attenuation using time or frequency as the reference is described. It is a power ratio type of system in which power flowing through the unknown attenuator is controlled by a constant pulsewidth, variable period, and pulse-modulated p-i-n diode switch. This power is detected by a thin-film thermocouple (TFT), the output of which is compared with that of a similar reference thermocouple whose input is a small constant fraction of the source power. The duty cycle required for equal thermocouple outputs is simply related to the unknown attenuation. The method is independent of detector linearity, and when compared with precise measurement standards, the system was found to be capable of measuring attenuation of up to at least 20 dB with an accuracy of  $\pm 0.001$  dB/10 dB.

## I. INTRODUCTION

METHODS of measuring microwave attenuation have been thoroughly reviewed by Warner [1] who has also reported [2] recent improvements to the UK national microwave standards. Usually the microwave energy is sensed either by a linear detector (e.g., mixer diode) or by a square law detector (e.g., bolometer). Substitution systems are widely used and contain reference standards such as a piston attenuator, an inductive voltage divider, or a Kelvin-Varley resistive divider, but in all such systems accuracy is dependent on detector linearity.

The time-interval ratio method to be described [3]–[5] uses time or frequency as the reference, which can, in principle, be determined with great accuracy. Development of this system which uses pulse modulation, has depended on the availability of high-speed digital counting techniques which are essential for precise measurement. The idea of attenuation measurement by pulse duty cycle has been independently proposed by F. H. Gale [6] and R. J. Satchell [7], but the only results obtained with this type of system of which the authors are aware were published as recently as 1983 [8] and 1984 [9].

## II. METHOD

In the Time-Interval Ratio System the detector must respond to mean power to make its output independent of the pulse-modulation duty cycle for a given mean power input. Such a square law detector limits the dynamic range of measurement to less than would be obtained with a superheterodyne system, but has the advantages of simplic-

ity and wide-frequency coverage. Thin-film thermocouple (TFT) detectors are used in preference to Wollaston wire bolometers or thermistor detectors because of their greater sensitivity and stability, as well as a longer thermal time constant.

The essential components are shown in Fig. 1 in which a frequency and amplitude stabilized microwave source supplies power through a directional coupler to a reference channel and to a measurement channel containing the device under test (DUT) together with the usual isolators and matching units. The signals in the two channels are detected by TFT power heads and the difference between the dc outputs of these is made zero by varying the duty cycle of a pulse-modulated p-i-n diode switch in the measurement channel. The duty cycle of this pulse modulation is precisely determined by digital counting techniques and the difference in attenuation between two settings of the DUT is given by the inverse ratio in dB of the respective pulse duty cycles required for zero output.

Fig. 1 also illustrates how the measurement can be computer controlled.

## III. THEORETICAL PERFORMANCE

Let the measurement parameters be as indicated in Fig. 1, where

$A$	attenuation, dB, through DUT,
$A_a$	difference in attenuation, dB, of DUT between calibration and reference settings,
$L$	fixed attenuation, dB, of other components in the measurement channel,
$P_i$	power entering DUT,
$P_o$	power leaving DUT,
$P_r$	power received by detector in reference channel,
$P_m$	mean power received by detector in measurement channel,
$P_n$	rms noise equivalent power for one detector referred to detector output,
$T$	switching pulse period,
$t$	switching pulse length,
$R_a$	noise equivalent input resistance of amplifier,
$R_m = R_r$	noise equivalent resistance of detectors, assumed equal,
$g$	fraction of incident power transmitted through the p-i-n diode when passing RF,

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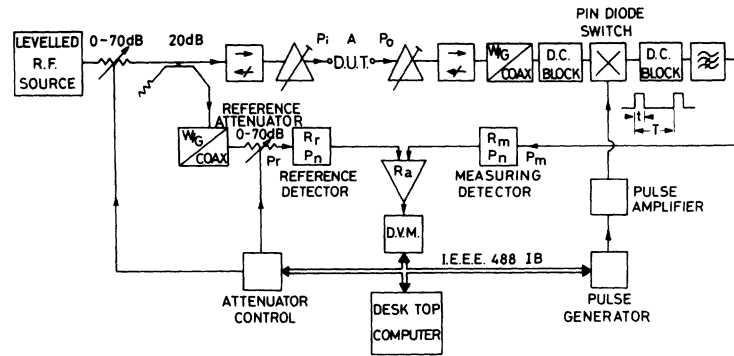


Fig. 1. Attenuation measurement by time-interval ratio block diagram.

$h$	fraction of power emerging from the p-i-n
	diode at the fundamental frequency,
$A_p$	attenuation (isolation), dB, of p-i-n diode in
	OFF-state.

Primed and double-primed values, e.g.,  $P'$  and  $P''$ , represent measurements made at the reference and unknown settings of the DUT.

The mean powers reaching the detector in the measurement channel at the two settings of the DUT are given by

$$P'_m = P_i \cdot 10^{-L/10} \cdot 10^{-A'/10} \left( \frac{g'h't'}{T'} + \frac{(T' - t')}{T'} \cdot 10^{-A_p/10} \right) \quad (1)$$

and

$$P_m'' = P_i \cdot 10^{-L/10} \cdot 10^{-A''/10} \cdot \left( \frac{g'' h'' t''}{T''} + \frac{(T'' - t'')}{T''} \cdot 10^{-A_p/10} \right). \quad (2)$$

Hence

$$A'' - A' = A_a$$

$$= 10 \log_{10} \left[ \frac{P_m' T'}{P_m'' T''} \left\{ \frac{t'' g'' h'' + (T'' - t'') \cdot 10^{-A_p/10}}{t' g' h' + (T' - t') \cdot 10^{-A_p'/10}} \right\} \right].$$

If  $P_m$ ,  $g$ ,  $h$ , and  $t$  are approximately constant, the attenuation error due to  $A_p$  is given by

$$\delta A_a = 10 \log_{10} \left[ \frac{T'}{T''} \left\{ \frac{1 + \frac{T'' - t}{ght} \cdot 10^{-A_p/10}}{1 + \frac{T' - t}{ght} \cdot 10^{-A_p/10}} \right\} \right] - 10 \log_{10} \frac{T'}{T''}$$

from which

$$\delta A_a \approx -4.343 \ 10^{-A_p/10} \frac{T'}{t} \frac{(1 - 10^{-A_a/10})}{gh}. \quad (4)$$

When  $A_p$  is large the attenuation measurement error is

$$\delta A_a = 4.343 \left\{ \frac{\delta P'_m}{P'_m} - \frac{\delta P''_m}{P''_m} + \frac{\delta T'}{T'} - \frac{\delta T''}{T''} + \frac{\delta t''}{t''} - \frac{\delta t'}{t'} + \frac{\delta g''}{g''} - \frac{\delta g'}{g'} + \frac{\delta h''}{h''} - \frac{\delta h'}{h'} \right\} \quad (5)$$

where  $P'_m = P''_m = P_r$  at balance,  $\delta P'_m$  and  $\delta P''_m$  represent noise fluctuations of the true values of  $P'_m$  and  $P''_m$ , and  $\delta T'$  and  $\delta T''$  represent errors in time-interval measurement.

It can be shown that the rms equivalent noise input power to an amplifier with two TFT detectors in series, compared with a single detector, is increased by a factor  $(R_r + R_m + R_a)/(R_m + R_a)$ . In (5)

$$\delta P'_m = \delta P''_m = \frac{R_r + R_m + R_a}{R_m + R_a} \cdot P_n.$$

If the pulse-modulation duty cycle is set by varying the period  $T$ , while maintaining the pulse length  $t$ ; constant, then  $t' \approx t''$  and  $\delta t''/t'' - \delta t'/t' = \delta t/t$ .

Similarly,  $\delta g''/g'' - \delta g'/g' = \delta g/g$  represents any fractional change in the mean power transmission through the p-i-n diode between measurements.

Also  $\delta h''/h'' - \delta h'/h' = \delta h/h$  represents any fractional change in power in the operating bandwidth between measurements caused by a change in harmonic content.

The maximum error  $\delta \hat{A}_a$  is given by

$$\delta \hat{A}_a = 4.343 \cdot \left\{ 2 \frac{\hat{P}_n}{P_r} \left( \frac{R_r + R_m + R_a}{R_m + R_a} \right) + \frac{\delta \hat{T}'}{T'} + \frac{\delta \hat{T}''}{T''} + \frac{\delta t}{t} + \frac{\delta g}{g} + \frac{\delta h}{h} \right\} \quad (6)$$

where peak values are indicated by  $\hat{\cdot}$ .

### III. EXPERIMENTAL SYSTEM

### A. Practical Values

The peak noise power  $\hat{P}_n$ , specified for the standard dc chopper amplifier (Marconi-Sanders 6460) is about 6 nW at maximum sensitivity. Addition of further amplification with increased smoothing reduced the peak noise to about 1 nW.

The pulse generator (Hewlett-Packard 5359A) has a specified accuracy of  $\pm 1$  ns with resolution  $< 1$  ns, with the contribution from the specified time base error of  $\pm 2$  ns/s rms for pulse periods of  $< 10^{-2}$  s being negligible. The output from this pulse generator was amplified by a VMOSFET switching transistor to provide +10 mA and

–10 V as required by the p-i-n diode switch (Microwave Associates ML17280-13). Direct coupling to the p-i-n diode was achieved by isolating the latter from the adjacent coaxial lines by dc blocks. Because of amplifier and p-i-n switch limitations the effective RF pulse length may differ from that of the pulse generator. The pulse length  $t$ , which should be independent of duty cycle, was not changed during an individual calibration, and no variation,  $\delta t$ , was conclusively detected at different duty cycles.

Attenuation through the p-i-n diode when passing RF at 10 GHz was measured to be about 1.0 dB. Variation with temperature was about 0.001 dB/°C, and with drive voltage about 0.1 dB/V. Any variation of p-i-n diode drive voltage with duty cycle would introduce a systematic error which would be represented in the value of  $g$ . No such variation was observed.

The attenuation (isolation) of the p-i-n diode switch  $A_p$  was  $> 80$  dB, and with 10 mW power at 10 GHz incident on the p-i-n diode no second harmonic could be detected with a spectrum analyzer sensitivity of  $-80$  dBm, showing the harmonic power generated to be negligible.

The TFT detectors (Marconi 6422) have a maximum power rating of 1 mW and a thermal time constant of about 10 ms. Since the peak-to-peak alternating component of the measuring thermocouple output voltage is about 40 percent of the dc level for a thermal time constant of 10 ms and a pulse period of 4 ms, and the difference between the measuring and reference thermocouple output voltages is amplified by a high-gain dc amplifier, it is important that there should be sufficient filtering in the early stages of the amplifier to prevent overload or nonlinearity effects. Measurements confirmed that nonlinearity effects could only be detected if the alternating component of the amplifier input voltage was increased to over 20 times the actual maximum value.

In the experimental arrangement the attenuation in the measurement channel with the DUT at the reference position was about 9 dB due to losses in the p-i-n diode, high-pass filter, isolators, matching units, dc blocks, coaxial to waveguide transformer, and coaxial line. The attenuation in the reference channel was about 32 dB consisting of 20 dB in the directional coupler, a 10-dB reference attenuator, and the remainder being in the coaxial to waveguide transformer and coaxial line.

With 32-dB attenuation in the reference channel and  $P_r = 10 \mu\text{W}$ , the leveled input power  $P_i \approx 16$  mW. If the pulse length is fixed so that  $t = 20 \mu\text{s}$ , then at the reference setting of the DUT,  $T' \approx 4$  ms, and with  $A_a = 20$  dB added attenuation in the DUT,  $T'' \approx 40 \mu\text{s}$ . Substituting these figures in (6), with  $\delta \hat{T} \leq 1$  ns,  $R_m = R_r = 200 \Omega$  and  $R_a = 1500 \Omega$ , gives

$$\delta \hat{A} = [\pm 1.0 \times 10^{-3} + 4.34(\pm 2.5 \times 10^{-7} \pm 2.5 \times 10^{-5})] \text{ dB}.$$

The first term represents random noise and is the most significant. Thus the worst combination of errors produces a maximum measurement error of  $\pm 0.0011$  dB when

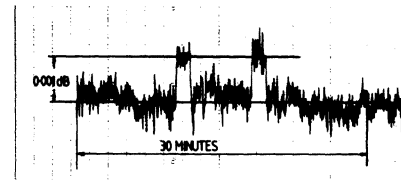


Fig. 2. System stability and resolution.

$A_a = 20$  dB. Also, substituting  $A_p = 80$  dB and  $gh \approx 1$  in (4) shows that leakage through the p-i-n switch in the off position contributes an attenuation measurement error not exceeding  $-0.00001$  dB for  $A_a = 20$  dB and proportionately more for larger values of  $A_a$ .

### B. Stability and Sensitivity

Sensitivity to input power changes depends on how well the thermocouple detector characteristics are matched. Tests confirmed that the balance condition was not changed for reasonably small (20 percent) changes in RF power level, such changes being much greater than the expected variations of the leveled input power. The system was not unduly sensitive to frequency, being unaffected by a change of 4 kHz, but a 1-W Impatt oscillator or a TWT amplifier were both found to be unsatisfactory, possibly due to frequency components of amplitude noise outside the leveling loop bandwidth. The measurements to be described were made with a CV2346 low-voltage klystron oscillator giving about 50 mW before leveling and frequency stabilized to  $10 \text{ GHz} \pm 1 \text{ kHz}$  (Microwave Systems MOS/5) with the power contained within a 3-dB bandwidth of  $< 10$  kHz.

Although a coaxial system offers wide frequency coverage, experience has shown that it is much more sensitive to mechanical and temperature effects than a waveguide system. Consequently, to investigate the capabilities of the time-interval ratio method of attenuation measurement high-quality waveguide components were used wherever possible, and the length of coaxial cables reduced to the minimum by mounting the thermally insulated and electrically and magnetically screened power detectors close to the equipment in a temperature controlled environment. Fig. 2 shows the stability and noise fluctuation of the balance voltage and the response to a step change of 0.001 dB in the measurement channel.

### C. Automatic Measurement

A desk top computer (Hewlett-Packard HP85) was used to control the various system components via an IEEE-488 interface bus as shown in Fig. 1. During a change in output parameters, the Hewlett-Packard 5359A Pulse Generator suppresses its output. To prevent a short excessive imbalance between the two detectors overloading the amplifier and differentially heating the detectors, a switched attenuator (Hewlett-Packard 11713A) was used to reduce the power input by 70 dB during this time.

The computer was programmed to vary the pulse period  $T$ , while maintaining the pulse length  $t$  nominally constant, until balance conditions could be determined. With

TABLE I  
MEASUREMENT RESULTS

Device Under Test	Rotary Vane Attenuator				Switched Coupler	
Nominal Attenuation, dB	5	10	15	20	10	20
Measured attenuation, dB (a) with (random uncertainty for 95% confidence (dB $\times 10^{-5}$ ))	5.00038	9.99982	14.99968	19.99660	10.64461	20.38133
	(24)	(32)	(40)	(66)	(32)	(35)
	5.00093	9.99990	14.99953	19.99715	10.64479	20.38129
	(22)	(37)	(24)	(38)	(23)	(43)
	5.00092	9.99979	14.99972	19.99785	10.64482	20.38140
	(19)	(25)	(26)	(41)	(45)	(37)
Calibration, dB (b) with (random uncertainty for 95% confidence (dB $\times 10^{-5}$ ))	5.00102	9.99991	14.99946	19.99780	10.64458	20.38119
	(20)	(20)	(40)	(57)	(56)	(33)
Mean Difference, dB	0.0002	-0.0001	-0.0003	0.0021	-0.0001	0.0018

(a) Time Interval Ratio Measurement

(b) UK National Standard Measurement

the DUT set at the reference value (0 dB) and an approximate value of  $T$ , the input power was reduced by the switched attenuator until the output unbalance voltage could be measured. Three measurements were made of the unbalance voltage at slightly different values of the period  $T$ . Using these voltages and periods the value of  $T$  required for zero output was calculated from a least squares error fit to a linear relationship between voltage and period. This sequence was followed at successively increased sensitivity until a sufficiently precise value of  $T$  required for balance was obtained. The process was repeated again with the DUT set to the attenuation value to be calibrated, and the true value obtained from

$$A_a = 10 \log \frac{T'}{T''}. \quad (7)$$

At certain values of the period  $T$  the harmonic relationship of the corresponding pulse repetition frequency to the high-gain dc chopper amplifier switching frequency can cause low-frequency ( $<0.5$  Hz) beating. This effect can be observed and overcome by a small change in pulse length and pulse period.

#### IV. MEASUREMENT RESULTS

The attenuation of a precision rotary vane attenuator [1] whose angular position could be repeated to better than  $0.001^\circ$  was measured at the theoretical angular positions determined from  $A_a = 40 \log_{10} (\sec \theta)$  which corresponded to attenuation values of 5, 10, 15, and 20 dB. The leveled maximum input power was about 16 mW at 10 GHz. With the pulse length  $t$  maintained constant in the

range 7 to 20  $\mu$ s and a corresponding maximum value of the period  $T$  of up to about 4 ms, a number of sets of 10 measurements was made. Representative results of these measurements are shown in Table I, where the uncertainty values are obtained from

$$\frac{\text{standard deviation}}{\sqrt{\text{number of measurements}}} \times \text{Student } t.$$

Results obtained from measurements on a switched coupler having nominal attenuation values of 10 and 20 dB are also included in Table I.

These measurements were unaffected by changing the value of the fixed pulse length in the range 7 to 20  $\mu$ s, and were also unaffected by three samples of the p-i-n diode switch. For comparison, calibration values obtained from measurements by the UK National microwave attenuation standard [2] are included.

#### V. CONCLUSIONS

A relatively simple method of measuring microwave attenuation has been described which depends on the accurate measurement of time intervals. Measurements of attenuation values of up to 20 dB of a rotary vane attenuator and of a switched coupler have agreed with similar measurements made with the most sophisticated equipment [2] within the best accepted limits of 0.001 dB/10 dB.

The system has been successfully automated for computer-controlled measurements, but the requirement for relatively long, thermal time-constant detectors means that each measurement takes about 15 min.

It has not been possible to realize the potential advan-

tages of a wide-band coaxial measurement system for precision measurement due to the unreliability and lack of constancy of coaxial components, connectors, and cable which are sensitive to temperature and vibration effects. Nevertheless, a carefully engineered Time-Interval Ratio System can provide accurate measurements of attenuation independent of more conventional methods.

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# Wide-Band Homodyne Method of Measuring Microwave Circuits

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**Abstract**—Two-arm homodyne systems which enable simultaneous and precision measurement of real and imaginary parts of parameters are investigated. The usual manual and time consuming calibration of the system, necessary for each measuring frequency is troublesome. The difficulty has been avoided by introducing a special scaling process, and then calculator correcting the systematic measurement errors within a high-frequency band. Theoretical basis of the method and the original scaling process have been presented. The results of preliminary studies, which allow the measurement error estimated at below 1 percent, have been presented.

## I. INTRODUCTION

**M**ICROWAVE homodyne detection systems are used in many measuring applications [1]–[5]. Their most important features are: very high dynamic range and sensitivity, which make it possible to measure a wide range of attenuation, reflection coefficient [4], and small variations in resonance frequency. The phase sensitivity of homodyne detection extends the measurement capability to complex parameters of the circuits under test.

A multiarm homodyne system [5], which enables simultaneous measurement of real and imaginary parts of parameters under test, requires laborious manual calibration. To obtain accurate results it is necessary to ensure constant sensitivity and symmetry of the detectors, symmetry of the microwave arms, and specified phase relations between the signals which reach each detector. These aspects will be discussed in Section II. These reasons tend to make homodyne systems narrow banded.

Many attempts have been made to eliminate this manual

calibration [1]. The use of wide-band elements and a specified configuration of arms for the entire system has not proven satisfactory. Though the frequency bands obtained amount to several or tens of percentage, the accuracy of measurements is not up to standard.

It will be shown that a description, and numerical correction of systematic errors, of homodyne systems is possible based on the analysis of a six-port technique. Simple and accessible computer technology make it possible to eliminate manual calibration, thus ensuring not only broad-band operation, but also high measurement precision.

## II. BASIC CONFIGURATION OF THE MEASURING SYSTEM

In Fig. 1 a two-arm homodyne system is depicted. Instantaneous voltage values  $e_r$  and  $e_s$  of each detection diode, produced by a microwave signal from the reference and signal arms, amount to

$$e_{ri} = E_{ri} \exp(j\omega t) \quad (1)$$

$$e_{si} = E_{si} M \exp(j\omega t), \quad i = 1, 2 \quad (2)$$

where  $E_{ri} = |E_{ri}| \exp(j\Phi_{ri})$  is a complex voltage amplitude in the reference arms,  $E_{si} = |E_{si}| \exp(j\Phi_{si})$  is a complex voltage amplitude in the signal arms, and  $\Phi_{ri}$  and  $\Phi_{si}$ , respectively, are phase angles of these voltages.  $M$  is a complex factor which depends on the mode of signal modulation. Assuming double sideband modulation of the signal (DSBWC), a shift of phase between signals  $e_{s1}$  and  $e_{s2}$  equal to  $\Delta\Phi$ , and square law operation of the detectors, we can determine the amplitudes of component variables with  $\Omega$  modulation frequency on the output of each detec-