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# Attenuation Measurements

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## **Abstract**

The paper details basic definitions and equations related to attenuation and describes various measurement techniques including power ratio, voltage ratio, AF substitution and IF substitution. Various attenuator types are described and a worked example of attenuation measurement is shown, complete with a measurement uncertainty budget.

## **About the Speaker**

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## Attenuation Measurement

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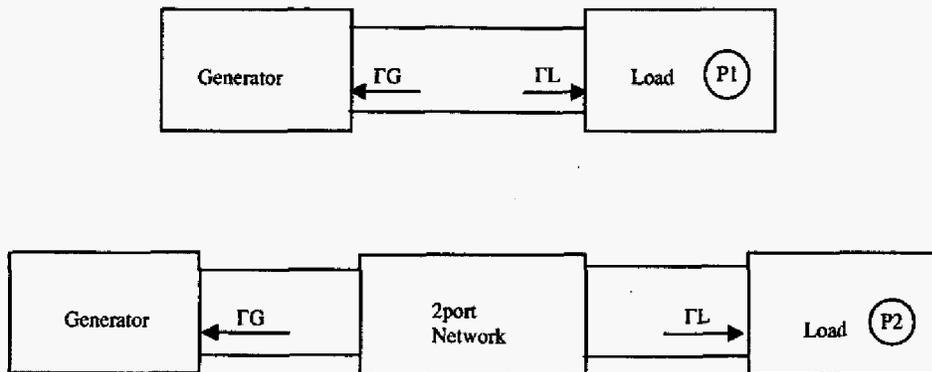
### INTRODUCTION

Accurate attenuation measurement is an important part of characterising rf or microwave circuits and devices. For example, attenuation measurement of the component parts of a radar system will enable a designer to calculate the power delivered to the antenna from the transmitter, the noise figure of the receiver and hence the fidelity or bit error rate of the system. A precision power measurement system, such as a

calorimeter described by Oldfield (1), requires the transmission line preceding the measurement element to be characterised in order to determine the effective efficiency of the system. A thermal electrical noise standard described by Sinclair (2) requires accurate attenuation measurement of the transition or thermal block between the hot termination and the ambient temperature output connector, in order to determine its excess noise ratio.

### BASIC PRINCIPLES

Diagram 1. Insertion Loss



With reference to diagram 1.

When a generator with a reflection coefficient  $\Gamma_G$  is connected directly to a load of reflection coefficient  $\Gamma_L$ , let the power dissipated in the load be denoted by  $P_1$ . Now if a two-port network is connected between the same generator and load, let the power dissipated in the load be reduced to  $P_2$ .

**Insertion loss** in decibels of this 2 port network is defined as:

$$L \text{ dB} = 10 \log_{10} \frac{P_1}{P_2} \quad (1)$$

**Attenuation** is defined as the insertion loss where the reflection coefficients  $\Gamma_G$  and  $\Gamma_L = 0$ .

Note that insertion loss depends upon the value of  $\Gamma_G$  and  $\Gamma_L$ , whereas attenuation depends only upon the 2-port network. If an attenuation measurement is made without matching both the source and load perfectly, there will be an error associated with the result. This error is called the mismatch error and it is defined as the difference between the insertion loss and attenuation.

Whence,

$$\text{Mismatch Error } M = L - A \quad (2)$$

Diagram 2. Signal Flow

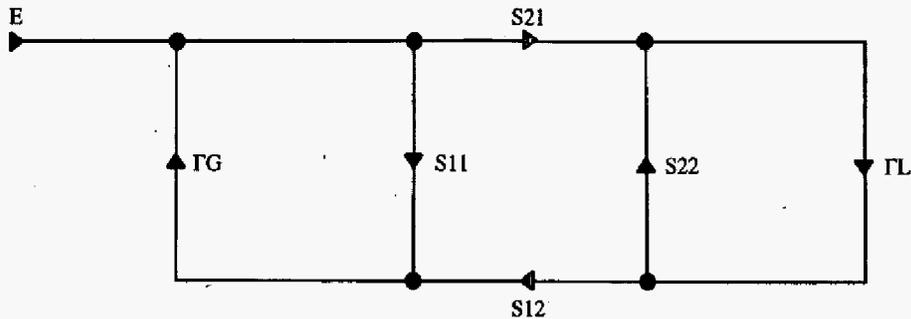


Diagram 2 shows a signal flow diagram of a 2 port network between a generator and load where:

$S_{11}$  is the voltage reflection coefficient looking into the input port when the output port is perfectly matched.

$S_{22}$  is the voltage reflection coefficient looking into the output port when the input port is perfectly matched.

$S_{21}$  is the ratio of the complex wave amplitude emerging from the output port to that incident upon the input port when the output port is perfectly matched.

$S_{12}$  is the ratio of the complex wave amplitude emerging from the input port to that incident upon the output port when the input port is perfectly matched.

From the above, the equation from Warner (3), for insertion loss is given as:

$$L = 20 \log_{10} \frac{|(1-\Gamma_G S_{11})(1-\Gamma_L S_{22}) - \Gamma_G \Gamma_L S_{12} S_{21}|}{|S_{21}| \cdot |1-\Gamma_G \Gamma_L|} \quad (3)$$

It can now be seen that the insertion loss is dependent upon  $\Gamma_G$  and  $\Gamma_L$  as well as the S parameters of the 2 port network. When  $\Gamma_G$  and  $\Gamma_L$  are matched, ( $\Gamma_G$  and  $\Gamma_L = 0$ ), then equation 3 is simplified to:

$$A = 20 \log_{10} \frac{1}{|S_{21}|} \quad (4)$$

where A = attenuation dB

From equation 2, the mismatch error M is the difference between equations 3 and 4.

$$M = 20 \log_{10} \frac{|(1-\Gamma_G S_{11})(1-\Gamma_L S_{22}) - \Gamma_G \Gamma_L S_{12} S_{21}|}{|1-\Gamma_G \Gamma_L|} \quad (5)$$

Note that all the independent variables of equation 5 are complex. In practice it may be difficult to measure

the phase relationships where the magnitudes are small. Where this is the case, and only magnitudes are known, the mismatch uncertainty is given as a maximum and minimum limit thus:

$$M(\text{limit}) = 20 \log_{10} \frac{1 \pm (|\Gamma_G S_{11}| + |\Gamma_L S_{22}| + |\Gamma_G \Gamma_L S_{11} S_{22}| + |\Gamma_G \Gamma_L S_{12} S_{21}|) / 1 \pm |\Gamma_G \Gamma_L|}{1} \quad (6)$$

If the 2 port network is a variable attenuator then the mismatch limit is expanded from equation 6 to give:

$$M(\text{limit}) = 20 \log_{10} \frac{1 \pm (|\Gamma_G S_{11e}| + |\Gamma_L S_{22e}| + |\Gamma_G \Gamma_L S_{11e} S_{22e}| + |\Gamma_G \Gamma_L S_{12e} S_{21e}|) / 1 \pm (|\Gamma_G S_{11b}| + |\Gamma_L S_{22b}| + |\Gamma_G \Gamma_L S_{11b} S_{22b}| + |\Gamma_G \Gamma_L S_{12b} S_{21b}|)}{1} \quad (7)$$

Where suffix b denotes the attenuator at zero or datum position, (residual attenuation) and suffix e denotes the attenuator incremented to another setting, (incremental attenuation).

## MEASUREMENT SYSTEMS

Many different and ingenious ways of measuring attenuation have been developed over the years, and most methods in use today embody the following principles:

1. Power ratio
2. Voltage ratio
3. AF substitution
4. IF substitution
5. RF substitution

### Power Ratio Method

The power ratio method of measuring attenuation is perhaps one of the easiest to configure.

Diagram 3. Power Ratio

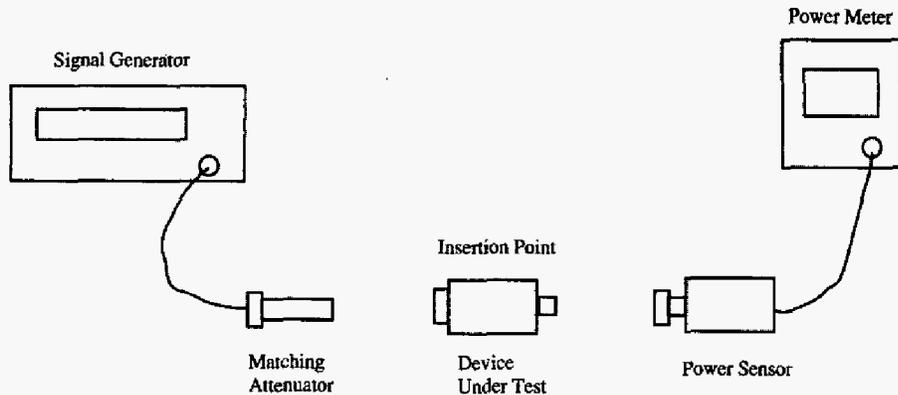


Diagram 3 represents a simple power ratio configuration. First, the power sensor is connected directly to the matching attenuator and the power meter indication noted P1. Next, the device under test is inserted between the matching pad and power sensor and the power meter indication again noted P2. Insertion loss is then calculated using:

$$L \text{ dB} = 10 \log_{10} \frac{P1}{P2} \quad (8)$$

Note that unless the reflection coefficient of the generator and load at the insertion point is known to be zero, or that the mismatch factor has been calculated and taken into consideration, measured insertion loss and not attenuation is quoted.

This simple method has some limitations:

1. Amplitude stability and drift of the signal generator.
2. Power linearity of the power sensor.
3. Zero carry over.
4. Range switching and resolution.

**Amplitude drift of the signal generator.** The drift of the signal generator rf output during the measurement process will have a directly proportional effect on the measurement accuracy.

**Power linearity of the power sensor.** The modern semiconductor thermocouple power sensor embodies a tantalum nitride film resistor, shaped so that it is thin in the centre and thick at the outside such that when rf power is absorbed, there is a temperature gradient giving rise to a thermoelectric emf. The rf match due to the deposited resistor is extremely good and the sensor will operate over a 50 dB range (+ 20 dBm to - 30 dBm) but there is considerable departure from linearity, (10% at 100mW) and it is necessary to compensate for this and the temperature dependence of the sensitivity by electronics.

Diode power sensors, described by Cherry and Oram et al (4) may be modelled by the equation

$$P_{in} = kV_{dc} \exp(yV_{dc}) \quad (9)$$

where:

$P_{in}$  is the incident power

$V_{dc}$  is the rectified dc voltage

$k$  and  $y$  are constants which are functions of parameters such as temperature, ideality factor and video impedance.

At levels below  $1 \mu\text{W}$  (-30 dBm) the exponent tends to zero, leading to a linear relationship between diode output voltage and input power, (dc output voltage proportional to the square of the rms rf input voltage). For power levels above  $10 \mu\text{W}$  (-20 dBm) correction for linearity must be made. Modern (smart) diode power sensors embody an rf attenuator preceding the sensing element. The attenuator is electronically switched in or out in order to maintain best sensor linearity over a wide input level range. They have a claimed dynamic range of 90 dB (+20 dBm to -70 dBm) and when told the frequency of operation, corrections for power linearity, frequency response and temperature coefficient are made automatically using data stored within the sensor e<sup>2</sup>prom.

Experiments by Orford and Abbot (5) show that power meters based on the thermistor mount and self compensating bridge are extremely linear. Here the thermistor forms one arm of a wheatstone bridge, which is powered by dc current, heating the thermistor until its resistance is such that the bridge balances. The rf power changes the thermistor resistance but the bridge is automatically re-balanced by reducing the applied dc current. This reduction in dc power to the bridge is called retracted power and is directly proportional to the rf power absorbed in the thermistor. The advantage of this system is that the thermistor impedance is maintained constant as the rf power changes. Thermistors however, have a slower response time than thermocouple and diode power sensors and have a useful dynamic range of only 30 dB.

**Zero carry over.** Due to the electronic circuits, small errors may occur when a power meter is zeroed on one range and then used on another.

**Range switching and resolution.** Most power meters operate over several ranges, each of approximately 5 dB. With a near full scale reading, the resolution and

noise of the power meter indication will be good. However, the resolution and noise contribution at a low scale reading may be an order worse. (This should be borne in mind when making measurements and all effort made to ensure that the power meter is near full scale for at least one of the two power measurements P1,P2.)

Diagram 4. Dual Channel Power Ratio

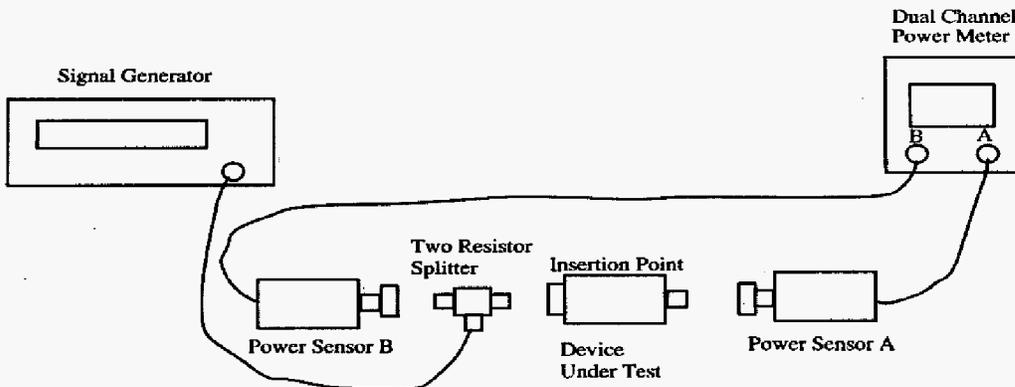


Diagram 4 shows a dual channel power ratio attenuation measurement system. First, power sensor A is connected directly to the two-resistor splitter and the power meter indication noted PA1, PB1. Next the device under test is inserted between power sensor A and the two-resistor splitter and again the power meter indication noted PA2, PB2. Insertion loss may now be calculated using:

$$L \text{ dB} = 10 \log_{10} \frac{PA1 \cdot PB2}{PB1 \cdot PA2} \quad (10)$$

The dual channel power ratio method has two advantages over the simple system of diagram 3.

1. The signal level is constantly monitored by power sensor B, reducing the error due to signal generator rf output level drift.
2. Using a two-resistor power splitter or high directivity coupler will improve the source match, Coster (6, 7).

The fixed frequency performance of this system may be improved by using tuners or isolators to reduce the generator and detector mismatch.

A system has been described by Stelzried and Reid et al (8, 9), that is capable of measuring attenuation in waveguide, WG22 (26.5 GHz to 40 GHz) with an uncertainty of  $\pm 0.005$  dB/10 dB, to 30 dB.

The author has assembled a measurement system similar to that shown in diagram 4, with the equipment computer controlled through the general-purpose interface bus (GPIB). The system operates from 10 MHz to 18 GHz and has a measurement uncertainty of  $\pm 0.03$  dB up to 30 dB, and  $\pm 0.06$  dB to  $\pm 0.3$  dB from 30 dB up to 70 dB. Return loss of the device under test is measured using an rf bridge at the insertion point. Insertion loss is then measured as described above and the mismatch error calculated so that attenuation may be quoted.

Diagram 5. Scalar Network Analyser

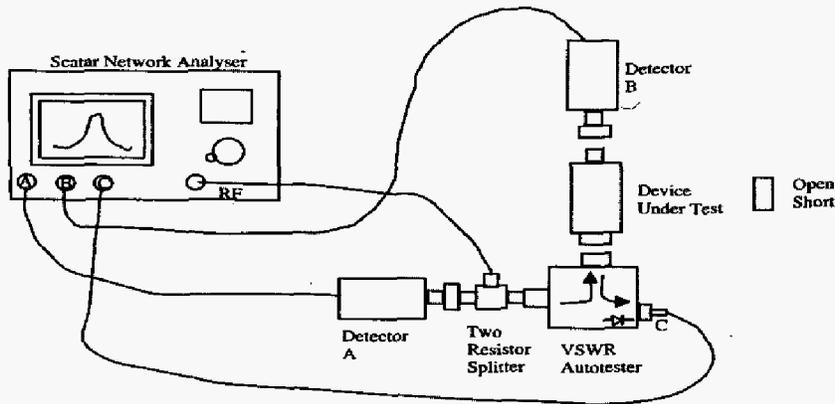


Diagram 5 shows a scalar network analyser attenuation and vswr measurement system. The analyser comprises a levelled swept frequency generator, three detector channels and a display. The dc output from the detectors is digitised and a microprocessor is used to make temperature, frequency response and linearity corrections for the detectors. Mathematical functions such as addition, subtraction, and averaging may be performed, making this set up very fast and versatile. The system has a claimed dynamic range of 70 dB (+20 dBm to -50 dBm), although this would be reduced when using the two-resistor splitter or padding attenuators, due to the maximum available output level from the rf sweep generator.

Attenuation measurement uncertainty varies according to the frequency and applied power level to the

### Voltage Ratio Method

Diagram 6. Voltage Ratio

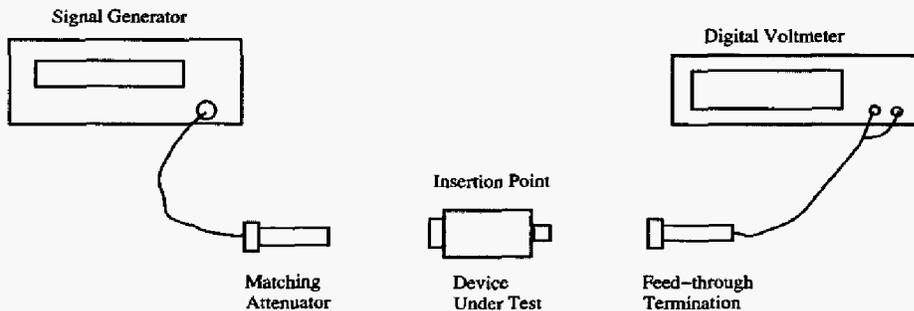


Diagram 6 represents a simple voltage attenuation measurement system where a DVM (digital voltmeter), is used to measure the potential difference across a feed-through termination, first when it is connected directly to a matching attenuator, V1, and then when the device under test has been inserted, V2. Insertion loss may be calculated from:

detectors but is generally  $\pm 0.1$  dB to  $\pm 1.5$  dB from 10 dB to 50 dB. Although this may not be considered as being highly accurate, the system is fast and can identify resonances which may be hidden when using a stepped frequency measurement technique. It is also more useful when adjusting a device under test.

An important consideration for all the above power ratio (homodyne) systems is that they make wide band measurements. When calibrating a narrow band device, such as a coupler or filter, it is important that the signal source be free of harmonic or spurious signals as they may pass through the device unattenuated and be measured at the detector.

$$L \text{ dB} = 20 \log_{10} \frac{V1}{V2} \quad (11)$$

(Note that the generator and load impedance are matched.)

This simple system is limited by the frequency response and resolution of the DVM as well as

variations in the output of the signal generator. The voltage coefficient of the device under test and resolution of the DVM will determine the range, typically 40 dB to 50 dB from dc to 100 kHz. A major contribution to the measurement uncertainty is the

linearity of the DVM used, which may be typically 0.01 dB / 10 dB for a good quality eight digit DVM. This may be measured using an inductive voltage divider, and corrections made.

### The Inductive Voltage Divider.

Diagram 7. Inductive Voltage Divider

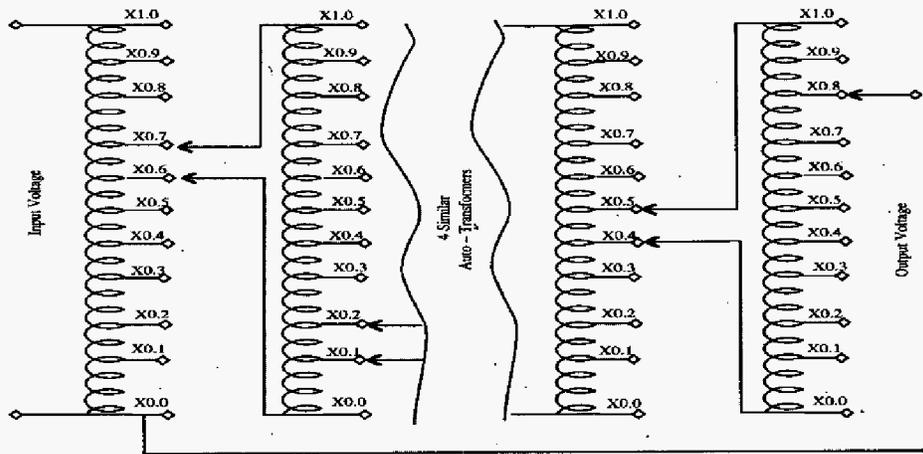


Diagram 7 is a simplified circuit diagram of an eight-decade IVD, (inductive voltage divider) described by Hill and Miller (10). This instrument is an extremely accurate variable attenuation standard operating over a nominal frequency range of 20 Hz to 10 kHz. (Special instruments have been constructed to operate at 50 kHz and 1 MHz.) It consists of a number of auto-transformers, the toroidal cores of which are constructed of wound insulated Supermalloy tape. (Supermalloy is an alloy is 70% Ni, 15% Fe, 5% Mo and 1% other elements, and has a permeability >100,000 and hysteresis <1 J/m<sup>3</sup>/Hz at 0.5 tesla).

The windings of a decade divider may be constructed by taking 10 exactly equal lengths of insulated copper wire from the same reel and twisting them together to form a rope, which is wound around the core, (double layer on the inner circumference, single layer on the outer circumference). Nine pairs of the ten strand cable are joined together, forming 10 series-aiding coils. High quality switches are required between the stages and care must be taken with the physical layout to avoid leakage and loading by the later decades.

The attenuation through a perfect IVD set to give a ratio D is given by:

$$A = 20 \log_{10} \frac{V_{in}}{D V_{in}} = 10 \log_{10} \frac{1}{D} \quad (12)$$

The overall error,  $\epsilon$ , in an inductive voltage divider may be defined as:

$$\epsilon = \frac{V_{out} - D V_{in}}{V_{in}} \quad (13)$$

Where:

$V_{in}$  is the input voltage, D is the indicated ratio and  $V_{out}$  is the actual output voltage.

$\epsilon = \pm 4 \times 10^{-8}$  for a commercially available 8 decade IVD, or  $\pm 10^{-6}$  to  $\pm 10^{-7}$  for a 7 decade IVD.

There are three important voltage ratio attenuation measurement systems in use at the National Physical Laboratory, Teddington UK

Diagram 8. Automated Voltage Ratio System

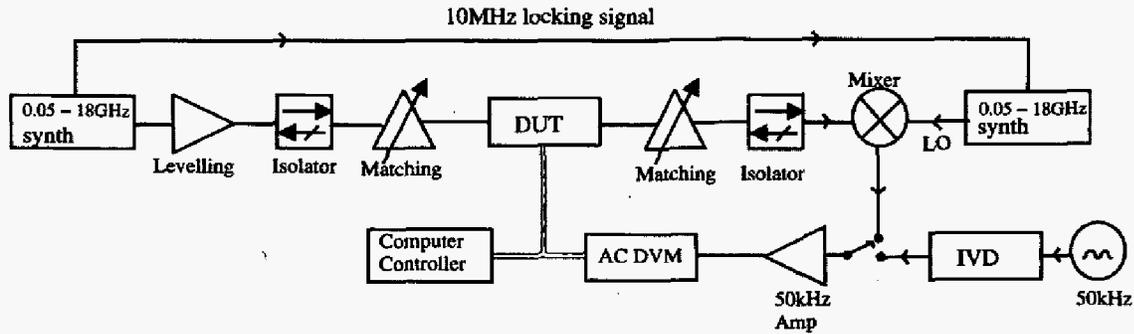


Diagram 8 is a computer controlled voltage ratio system described by Warner and Bayer et al (11, 12, 13) designed to calibrate the programmable rotary vane attenuators used at NPL in the Noise Standards Section. Two wide band synthesized signal generators are phase locked to a stable external reference frequency and their rf frequencies adjusted for a 50 kHz difference. RF from the signal generator is levelled and isolator and tuners used to match the

system, presenting a generator and load reflection coefficient better than 0.005 to the device under test. The 50 kHz IF signal is amplified and then measured on a precision AC DVM. The ac to dc conversion linearity of the DVM is periodically measured using a stable 50 kHz source and IVD. The non-corrected linearity is approximately 0.012 dB / 20 dB and the corrected linearity 0.0005 / 20 dB.

Diagram 9. Gauge Block Voltage Ratio System

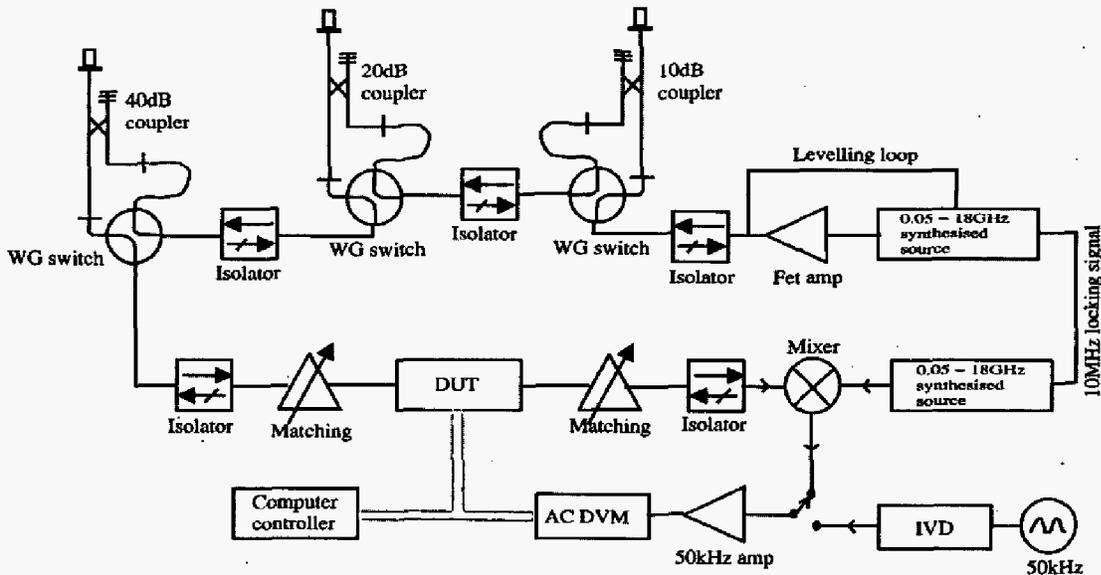


Diagram 9 is a voltage ratio system similar to that of diagram 8 but with the addition of a gauge block attenuators described by Warner and Herman (14) formed by three waveguide switched couplers in cascade. The gauge block may be switched from 0 dB to 70 dB in 10 dB steps, giving the system a 90 dB dynamic range. The standard deviation of 10 repeated measurements at the same attenuation value is below 0.0005 dB to 40 dB, 0.001 dB to 70 dB and <0.002 dB to 90 dB. Attenuation may be found from:

$$A_{DUT} \text{ dB} = \{20 \log_{10} (V_1 / V_2)\} + A_{gb} + C \quad (14)$$

Where:

$A_{gb}$  is the gauge block attenuation dB

$C$  is the ac DVM correction dB

Diagram 10. Dual Channel Voltage Ratio System

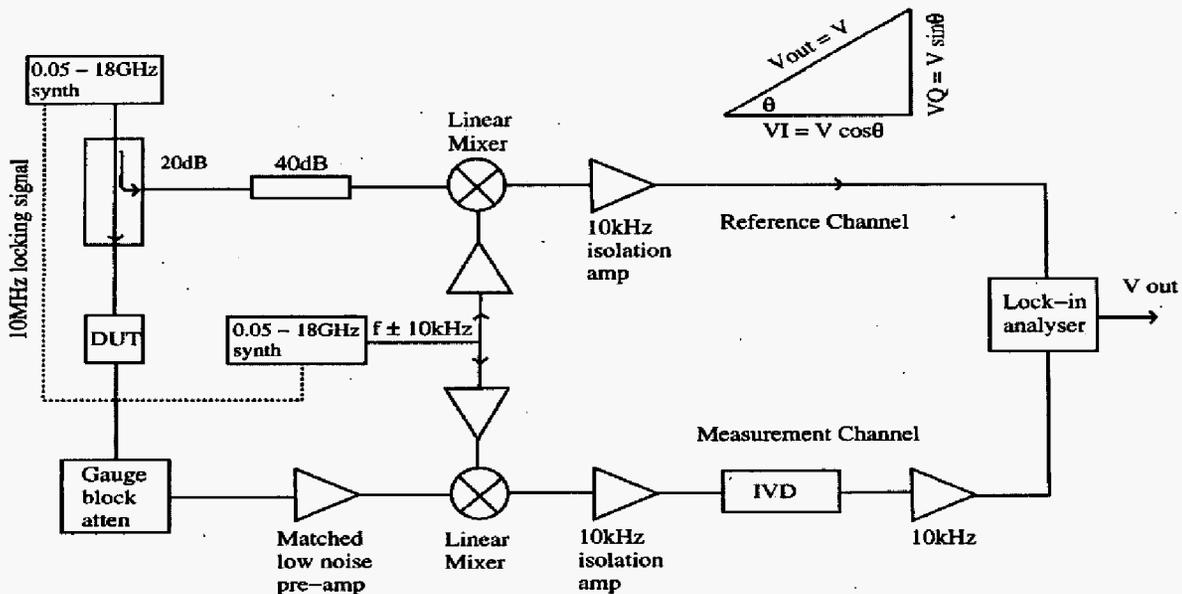


Diagram 10 shows a dual channel voltage ratio system developed by NPL Warner and Kilby et al (15, 16) to provide extremely accurate attenuation measurements for the calibration of standard piston attenuators. It operates over the frequency range 0.5 MHz to 100 MHz, is fully automatic and has a dynamic range of 160 dB without the need for noise balancing. The upper channel provides a reference signal for a lock-in analyser whilst the measurement process takes place in the lower channel. The lock-in analyser contains in-phase and quadrature phase sensitive detectors (PSDs), whose output ( $VI = \cos \psi$  and  $VQ = \sin \psi$ ) are combined in quadrature to yield a direct voltage given by:

$$\begin{aligned} V_{out} &= \{VI^2 + VQ^2\}^{0.5} \\ &= \{(V \cos \psi)^2 + V^2 \sin^2 \psi\}^{0.5} \\ &= V \end{aligned} \quad (15)$$

Thus,  $V_{out}$  is directly related to the peak value,  $V$ , of the 10 kHz signal emerging from the lower channel and is independent of the phase difference,  $\psi$ , between signals in the upper and lower channels.

Attenuation may be found from:

$$A_{DUT} = \{20 \log_{10} \frac{V_{out1}}{V_{out2}}\} + A_{gba} + A_{ivd} \quad (16)$$

Where:

$V_{out1}$  and  $V_{out2}$  are the output voltages for the datum and calibration setting.

$A_{gba}$  and  $A_{ivd}$  are the attenuations in dB removed from the gauge block and IVD when the device under test is inserted.

This dual channel system also yields the phase changes that occur in the device under test. From the vector diagram in diagram 10 we have:

$$\tan \psi = VQ / VI$$

$$\text{hence, } \psi = \arctan (VQ/VI) \quad (17)$$

Diagram 11. A Commercial Attenuator Calibrator

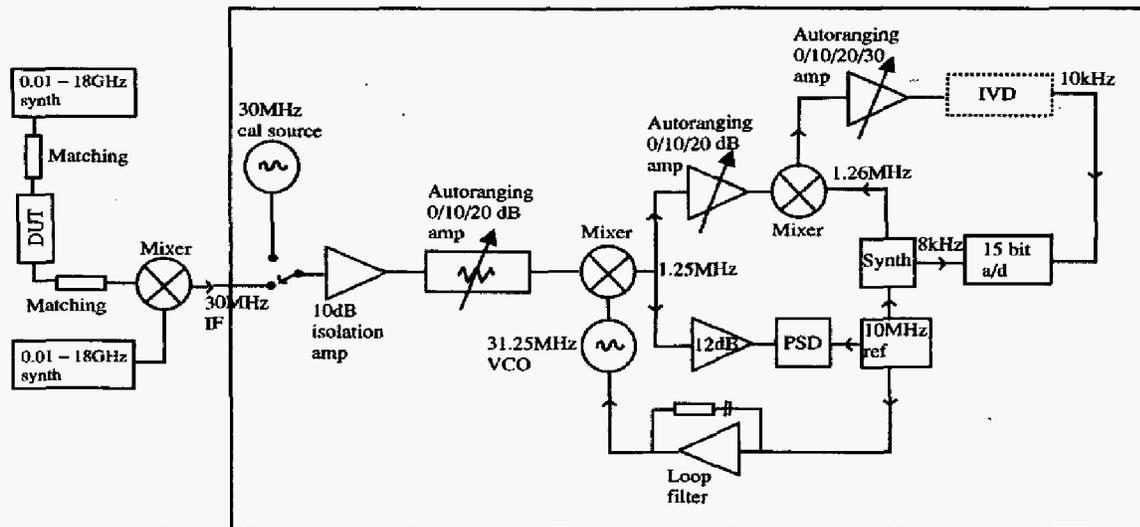


Diagram 11 is a simplified diagram of a commercially available voltage ratio attenuator calibrator from Weinschel (17). The instrument is a microprocessor controlled 30 MHz triple conversion receiver and may be used at other frequencies with the addition of an rf generator, local oscillator and mixer.

The 30 MHz input signal is routed through a 10 dB isolation amplifier to a series of pin-switched attenuators, which are automatically selected to present the first mixer with a 30 MHz signal at the correct level for low noise linear mixing. A phase locked loop referenced to the instrument 10 MHz xtal oscillator provides the correction for the first voltage controlled local oscillator (31.25 MHz), resulting in a first IF of 1.25 MHz. The first IF is routed through a variable gain amplifier, having switched gains of 0, 10 and 20 dB, to the second mixer. The signal is then mixed with a 1.26 MHz local oscillator derived from the 10 MHz reference to form a second IF of 10 kHz. This 10 kHz signal is applied to a second variable gain amplifier, having switched gains of 0, 10, 20 and 30 dB. The variable gain amplifiers and auto-ranging attenuator are microprocessor controlled in the auto-ranging sequence involved in making a measurement.

The final conversion is made by a sample and hold amplifier within the 15 bit analogue to digital converter. The sample and hold amplifier is clocked at 8 kHz. The sampling of the 10 kHz signal every 8 kHz provides a quasi-mixing action that results in a 2 kHz signal with four samples per cycle. This analogue signal is then converted to digital form.

Before use, the instrument is calibrated by switching the 30 MHz receiver input to a 30 MHz internal oscillator. Each attenuation and gain block is automatically calibrated against the a/d converter and the calibration results stored in ram to be used during the measurement process. The linearity of the a/d converter is paramount to the overall accuracy of the system and this may be checked by inserting a precision IVD into the 10 kHz path preceding the a/d.

This system has a sensitivity of -127 dBm and measurement accuracy of ± 0.015 dB, + 0.005 dB per 10 dB to 80 dB and ± 0.1 dB per 10 dB from 80 dB to 105 dB.

Chart 1. VM7 A/D Calibration

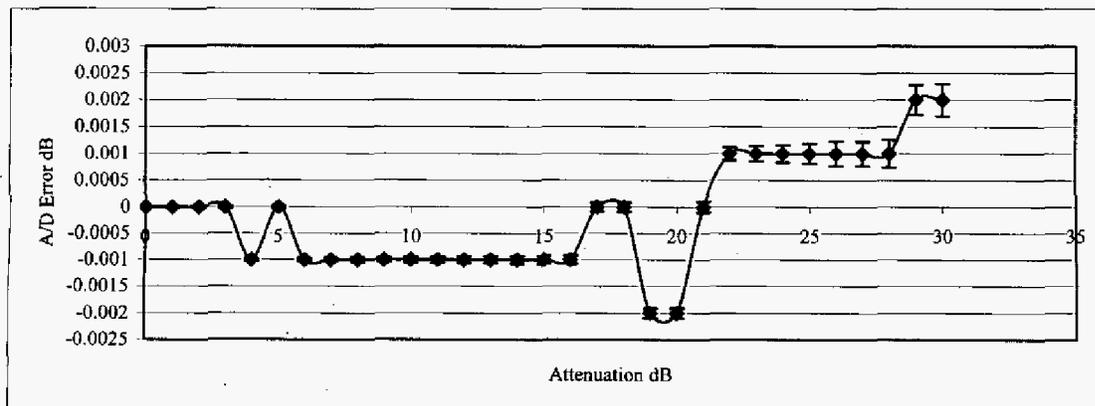


Chart 1 shows the results of the A/D calibration using an eight decade inductive voltage divider matched to the system. The 10 kHz signal is held between +14 to -

2 dBm on all but the last range, where it may cover +14 to -12 dBm.

Diagram 12. IFR 2309 FFT Signal Analyser

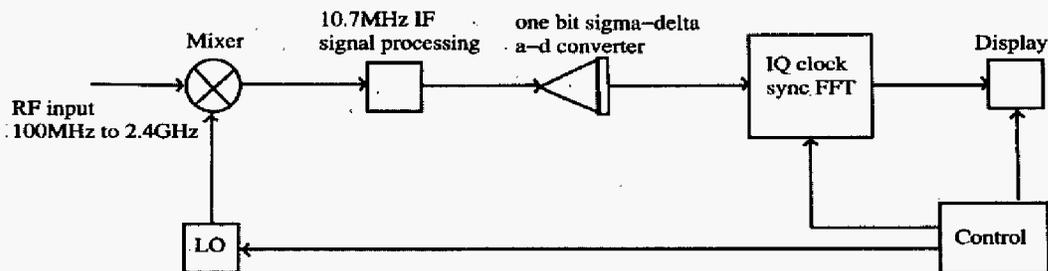
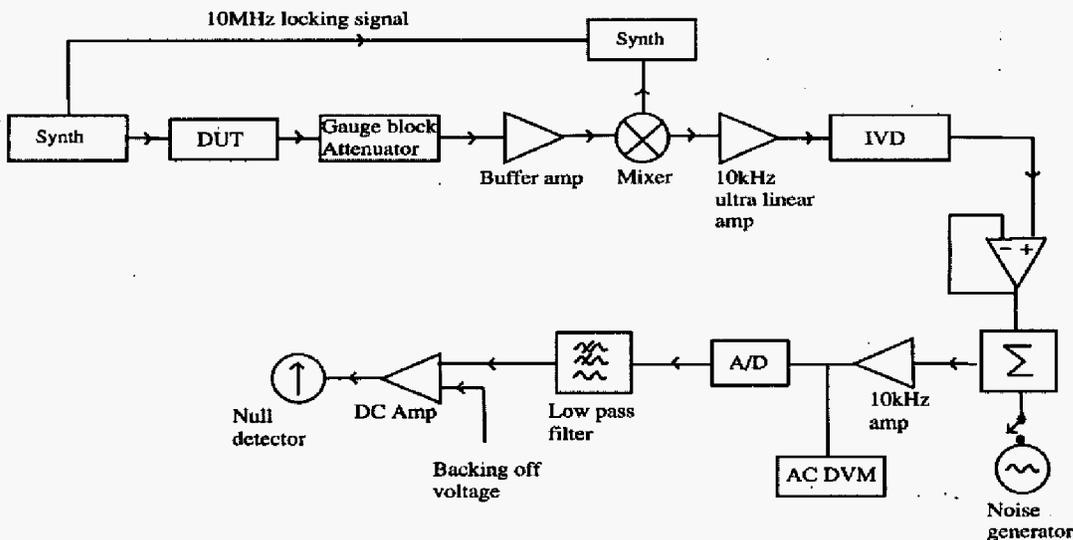


Diagram 12 is a simplified diagram of an IFR 2309 FFT Signal Analyser. Input signals are down converted to a 10.7 MHz IF and then routed through signal conditioning circuits to a 1 bit sigma-delta analogue to digital converter (better known for its use in CD players, giving low noise and high linearity). The a to d converter contains a comparator and quantizer with a feed back loop having a 2 bit delay.

The comparator compares the present input sample with the previous sample and outputs into a digital signal processor for complex analysis. The unit operates from 100 MHz to 2.4 GHz and has a maximum sensitivity of -168 dBm, resolution of 0.0001 dB and a claimed linearity of  $\pm 0.01$  dB per 10 dB.

### AF Substitution Method

Diagram 13. Audio Frequency Substitution System



In an audio frequency substitution system, the attenuation through the device under test is measured by comparison with an audio frequency standard, usually a resistive or inductive voltage divider. Diagram 13 is a simplified diagram of an AF substitution system designed and used at NPL, described by Warner (18). It is used at national

standards level over the frequency range 0.5 to 100 MHz and has a dynamic range of 150 dB. An external 10 MHz reference frequency is used to lock the two synthesiser time bases and the local oscillator is operated at a frequency difference of 10 kHz to the rf source, giving a stable 10 kHz IF after mixing. The

attenuation standard is a 7 decade precision IVD inserted between two 10 kHz tuned amplifiers.

The gauge block attenuator, which is a repeatable step attenuator, DUT and IVD are adjusted during the measurement so that the voltage to the null detector remains the same. The 10 kHz amplifiers are purpose built and are extremely linear (better than  $\pm 0.001\%$ ), are low noise and very stable. The IVD is driven from a low impedance source and its output is loaded with a high impedance. From 100 to 150 dB, noise must be injected into the system to compensate for the noise generated by the mixer. This is normally accomplished by setting the gauge block attenuator to zero, the IVD to unity and gain of the second 10 kHz amplifier for 2 volts rms output. The backing off voltage is adjusted for a null meter reading. The signal source is now switched off and the output voltage caused by the noise alone is measured on the DVM. After this, the gauge block attenuator is switched to a value,  $A_{gba}$  and the DUT moved to its datum position, the IVD is set to zero, the noise generator is switched on and adjusted to give the same output reading as before. Finally, the signal source is switched back on and the IVD is

adjusted to a ratio  $R$ , which gives a null reading on the output meter.

The attenuation change in the DUT is then given by:

$$A_{dut} = A_{gba} + 20 \log_{10}(1/R) \quad (18)$$

With great care being taken to reduce the effects of mismatch and leakage, the total uncertainty of measurement using this system is from  $\pm 0.0006$  dB at 10 dB to  $\pm 0.01$  dB at 100 dB, for 95% CL.

### IF Substitution Method

An IF substitution attenuation system compares the attenuation through the device under test with an IF attenuation standard, usually an IF piston attenuator, high frequency IVD or a box of  $\pi$  or T type resistive attenuators.

Diagram 14. Piston Attenuator

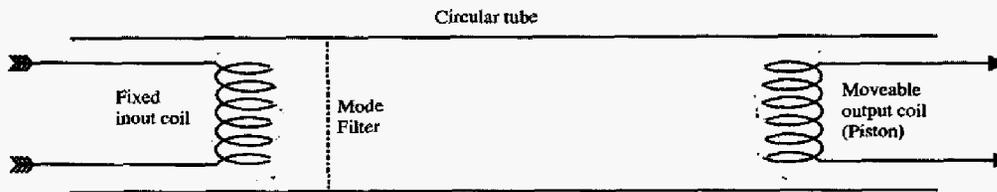


Diagram 14 shows a simplified piston attenuator arrangement normally used as the IF substitution standard. The tube acts as a waveguide below cut off transmission line. An IF signal, usually 30 or 60 MHz, is launched into the tube from the fixed input coil. A metallic grid acts as a mode filter ensuring that only a single mode, ( $H_{11}$ ,  $E_{01}$ ,  $H_{21}$ ,  $E_{11}$  or  $H_{01}$ ), is transmitted. The voltage in the output coil falls exponentially as the separation between the two coils increases. For perfectly conducting cylinder walls with a coil separation  $Z_1$  to  $Z_2$ , the attenuation in dB may be found from:

$$\alpha_p = 8.686 \times 2\pi(Z_2 - Z_1) \left\{ \left( \frac{S_{nm}}{2\pi r} \right)^2 - \left( \frac{1}{\lambda} \right)^2 \right\}^{0.5} \quad (19)$$

where:

$r$  is the cylinder radius

$\lambda$  is the free space wavelength

$S_{nm}$  is a constant dependent upon the excitation mode.

Precision engineering is required in manufacturing the piston attenuator. A tolerance of  $\pm 1$  part in  $10^4$  on the internal radius is equivalent to  $\pm 0.001$  dB per 10 dB. The cylinder is sometimes temperature stabilised and a

laser interferometer employed to measure the piston displacement.

When the piston attenuator is adjusted for a low attenuation setting, there is a systematic error resulting in non-linearity due to the interaction between the coils. This error reduces as the coils part, and is negligible at about 30 dB insertion loss.

A piston attenuator has been developed at NPL for use as a national standard, Yell (18, 19) The piston is mounted vertically and supported on air bearings to prevent contact with the cylinder wall. Displacements are measured with a laser interferometer. The cylinder is made of electroformed copper deposited on a stainless steel mandrel. This unit has a range of 120 dB, resolution of 0.0002 dB and a stated accuracy of 0.001 dB in 120 dB.

Diagram 15. Parallel IF Substitution System

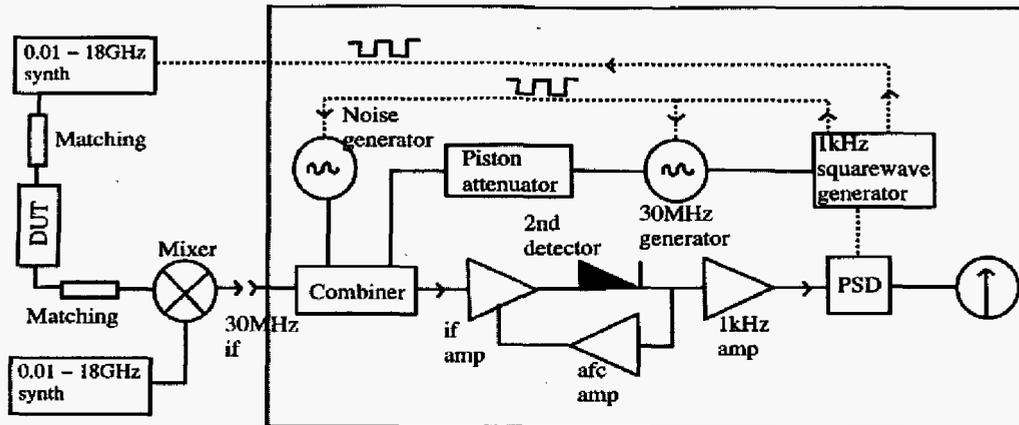


Diagram 15 shows the basic layout of a parallel IF substitution system. Here, the 30 MHz IF input signal is compared with a 30 MHz reference signal, which may be adjusted in level by using a calibrated precision piston attenuator. These two signals are 100% square wave modulated in counter phase and are fed to the IF amplifier. A tuned phase sensitive detector (PSD) is used to detect and display the difference between the two alternately received signals. This circuit is very effective and can detect a null in the presence of much noise. The system is insensitive to changes in IF amplifier gain, but an automatic gain control loop is provided, to automatically keep the detector sensitivity constant.

In operation, the two signals are adjusted to be the same amplitude, indicated by a null on the amplitude balance indicator. After the DUT is inserted, the system is again brought to a null by adjusting the precision piston attenuator. The difference in the two

settings of the standard attenuator is a measure of the insertion loss of the unknown device.

For signal levels below -90 dBm, noise generated by the mixer in the signal channel may cause an error and it is necessary to balance this by introducing extra noise into the reference channel. The system described above has a specification of  $\pm 0.01$  dB per 10 dB up to 40 dB,  $\pm 0.27$  dB at 80 dB and  $\pm 0.5$  dB at 100 dB.

### RF Substitution Method

With the RF substitution method, the attenuation through the device under test is compared with a standard microwave attenuator operating at the same frequency. The standard microwave attenuator is usually a precision waveguide rotary vane attenuator or microwave piston attenuator.

Diagram 16. Rotary Vane Attenuator

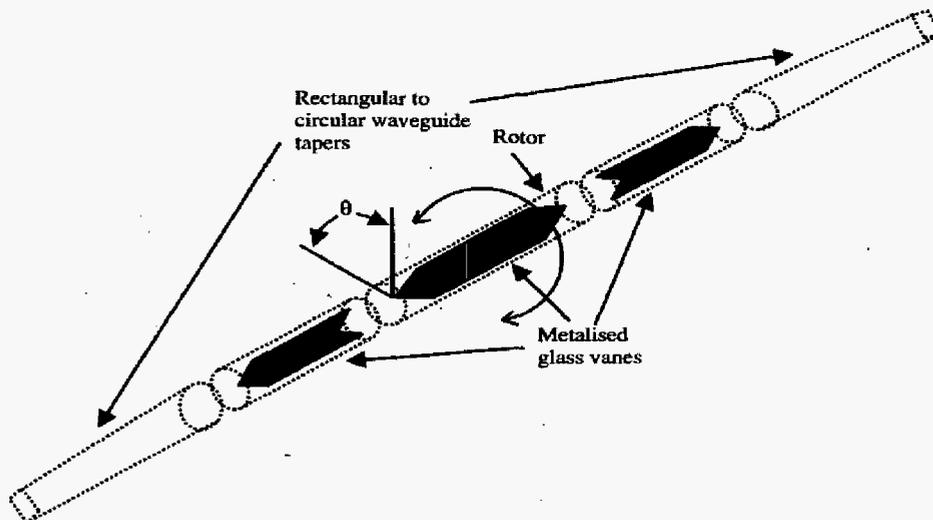


Diagram 16 shows the basic parts of a waveguide rotary vane attenuator, described by Banning (21). It consists of three metalised glass vanes, the two end vanes fixed in a direction perpendicular to the incident electric vector and a third vane, which is lossy and able to rotate, set diametrically across the circular waveguide. When the central vane is at an angle  $\theta$  relative to the two fixed vanes, the total attenuation is given by:

$$Arva = 40 \log_{10} (\sec\theta) + A_0 \quad (20)$$

where:

$A_0$  is the residual attenuation when all three vanes lie in the same plane ( $\theta = 0$ )

The rva is extremely useful as the attenuation is almost independent of frequency and there is little phase change as the attenuation is varied. As can be seen from equation 20, attenuation is not a linear function of the central vane angle. A change in  $\theta$  from  $0.000^\circ$  to  $0.615^\circ = 0.001$  dB, but a change in  $\theta$  from  $86.776^\circ$  to  $88.188^\circ$  changes the attenuation from 50 dB to 60 dB. A high quality rva can have an attenuation range of 0 to 70 dB, dependent upon the attenuation of the central vane and the vane alignment error. The accuracy of a commercially available rva is in the order of  $\pm 2\%$  of reading or 0.1 dB whichever is the larger. Rotary vane attenuators have been made at NPL with a display resolution of  $0.0001^\circ$  and they follow the  $40 \log_{10} (\sec \theta)$  law to within  $\pm 0.006$  dB up to 40 dB.

Diagram 17. Series RF Substitution

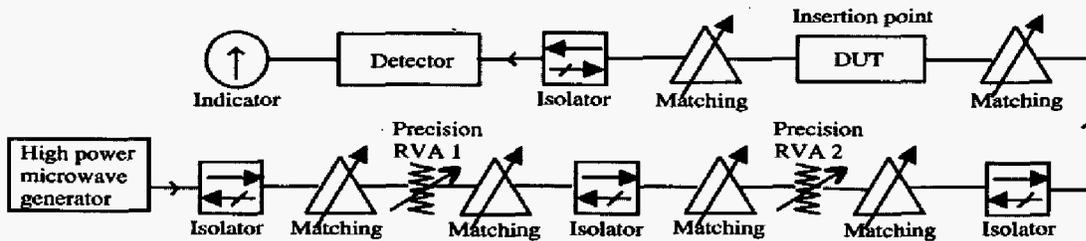


Diagram 17 is a block diagram of a very simple series rf substitution system, employing two precision rvas. As attenuation through the DUT is increased, the attenuation in the rvas is adjusted from the datum setting to keep the detector output constant. The difference between the new rva setting and the datum setting is equal to the loss through the DUT. This system could be enhanced by square wave modulation

the signal source and using a synchronous detector. By using two precision rva attenuators it is possible to obtain a dynamic range of 100 dB. A very stable high power signal generator is necessary and the output power must be within the power coefficient limits of the microwave components used. Great care must be taken to reduce RF leakage.

## The Automatic Network Analyser

Diagram 18. Vector Network Analyser

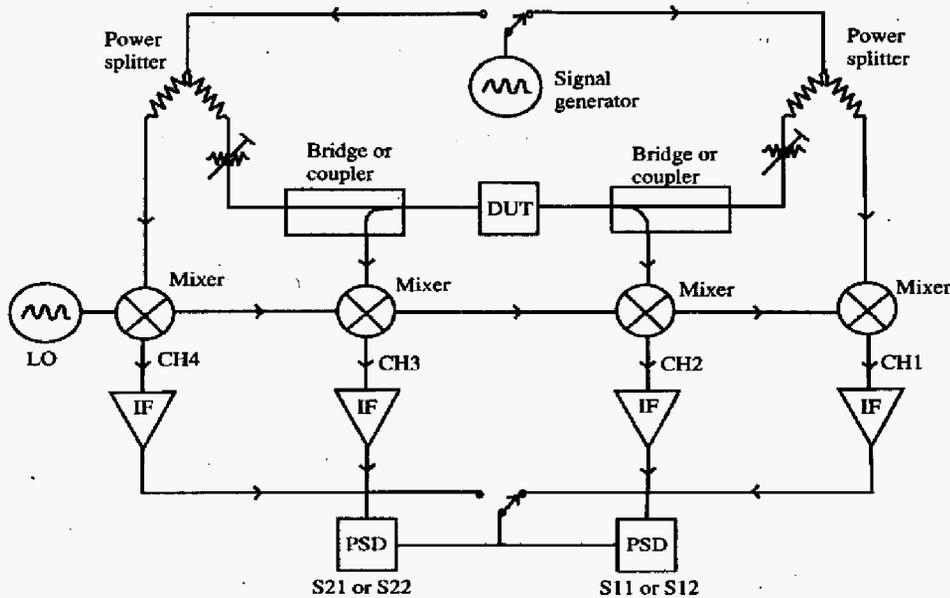


Diagram 18 is a greatly simplified block diagram of a commercially available microwave network analyser. The instrument is basically a microprocessor controlled two-channel superhetrodyne receiver, employing an ultra linear phase sensitive detector to detect magnitude and phase difference between the channels. The signal generator rf output is split into a reference channel and a measurement channel, which consists of a high directivity directional coupler or rf wheatstone bridge, capable to splitting the forward and reflected waves to and from the device under test.

With reference to diagram 18, when both switches are positioned to the right, channel 1 provides the reference signals for the phase sensitive detectors, channel 2 gives the real and imaginary parts of  $s_{11}$ , and channel 3 yields the real and imaginary parts of  $s_{21}$ . With the switches to the left, channel 4 provides the reference signals for the phase sensitive detectors, channel 2 gives the real and imaginary parts of  $s_{12}$  and channel 3 yields the real and imaginary parts of  $s_{22}$ . Thus, all four s parameters may be determined without reversing the device under test.

In use, the instrument is treated as a black box, where known calibration standards, such as open and short circuits, air lines and sliding or fixed termination are measured at the test ports and the results stored and used to correct measurements made on the device under test. With the device under test removed and port 1 connected to port 2, the instrument will measure the complex impedance of its generator and detector. This data is stored and used in the equation for calculating attenuation.

These instruments are extremely fast and versatile and results may be displayed in frequency or time domain, using a fast Fourier transform process. They are very accurate when measuring reflection coefficient but have a limited dynamic range for attenuation measurement, due to the available signal source level, harmonic mixing and channel crosstalk or leakage. The network analyser bridge or coupler sensitivity decreases with frequency, (roll off), and so one may find an rf network analyser, operating from 300 kHz to 6 GHz, a microwave network analyser, operating from 50 MHz to 50 GHz and a millimetric wave network analyser, operating from 75 GHz to 110 GHz.

Typical measurement uncertainties are:

### RF Vector Network Analyser

Attenuation	300kHz	3GHz	6GHz
0 dB	±0.03 dB	±0.04 dB	±0.05 dB
50	±0.11	±0.09	±0.12
75	±1.65	±1.34	±1.68

### Microwave Vector Network Analyser

Attenuation	50MHz	26.5GHz
0 dB	±0.03 dB	±0.11 dB
50	±0.05	±0.21
80	±0.60	±3.50

## IMPORTANT CONSIDERATIONS WHEN MAKING ATTENUATION MEASUREMENTS

## Mismatch Uncertainty

Mismatch between the generator, detector and device under test is usually one of the most significant contributions to attenuation measurement uncertainty. With reference to equation 6/7, these parameters must be measured in order to calculate the attenuation of the DUT from the measured insertion loss. It may be possible to measure the device under test and the detector by using a slotted line, rf bridge or scalar/network analyser. The signal generator used may have an active levelling loop or switched attenuator in its rf output stage, making it difficult to determine the generator output impedance accurately.

If single frequency measurements are to be made then the generator and detector match may be improved by

Diagram 19. Two-Resistor Power Splitter

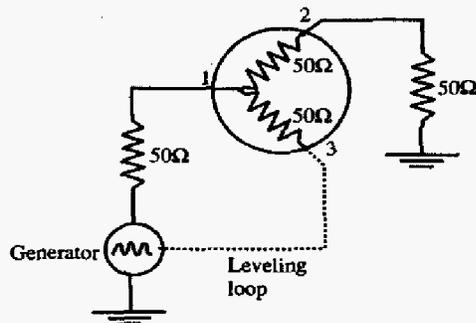


Diagram 19 shows a two-resistor power splitter, which is constructed to have a 50 Ω resistor in series between port 1 and port 2, and an identical resistor between port 1 and port 3. This type of splitter should only ever be used in a generator levelling circuit configuration as in diagram 19 and must never be used as a power divider in a 50 Ω system. (If you connect a 50 Ω termination to port 1 and port 2 then the impedance seen at port 3 is approximately 83.33 Ω.)

This power splitter has been specifically designed to level the power from the signal source and to improve the generator source match. In diagram 19, if a detector is connected to port 3 and the output of the detector is compared with a stable reference and then connected to the signal generator amplitude control via a high gain amplifier (levelling loop), then any change in the generator output at port 1 is seen across the both 50 Ω series resistors and detected at port 3. This change is used to correct the generator output for a constant level at port 3. As port 2 is connected to port 1 by an identical resistor, port 2 is also held level.

If an imperfect termination is connected to port 2 then voltage will be reflected back into port 2 at some phase. This voltage will be seen across the 50 Ω resistor and the generator internal impedance. If the

using ferrite isolators and tuners such as in diagram 8. The tuner must be adjusted for each frequency and this once tedious process has been improved by the introduction of computer controlled motorised tuners. Tuners and isolators are narrow band devices and are not practical for low rf frequencies (long wavelength) use due to their physical size. They are also prone to rf leakage and great care must be taken when making high attenuation measurements with these devices in circuit.

For wide band measurements, the generator match may be improved by using a high directivity directional coupler or a two-resistor splitter in a levelling loop circuit. A high quality 'padding' attenuator placed before the detector will also improve the detector match but its attenuation may restrict the dynamic range of the system.

generator internal impedance is imperfect then some of the voltage will be re-reflected by the generator, causing a change in level at port 1. This change in level is detected at port 3 and used to correct the generator to maintain a constant output level at the junction of the two resistors. Port 1 is perceived as a virtual earth, thus when the levelling circuit is active, the impedance measured at port 2 will effectively be the impedance of the 50 Ω series resistor.

Two resistor splitters are characterised in terms of output tracking and equivalent output reflection coefficient. A typical specification for a splitter fitted with type 'N' connectors would be ± 0.2 dB tracking between ports at 18 GHz, and an effective reflection coefficient of ± 0.025 to ± 0.075 from dc to 18 GHz. The splitter has a nominal insertion loss of 3 dB between the input and output ports.

In keeping with the UKAS document M3003 (22), the equation recommended for calculating mismatch uncertainty for a step attenuator is:

$$M = 8.686 / (2)^{0.5} \cdot [|\Gamma_G|^2 \cdot (|S_{11a}|^2 + |S_{11b}|^2) + |\Gamma_L|^2 \cdot (|S_{22a}|^2 + |S_{22b}|^2) + |\Gamma_G|^2 \cdot |\Gamma_L| \cdot (|S_{21a}|^4 + |S_{21b}|^4)]^{0.5} \quad (21)$$

where:

$\Gamma_G$  and  $\Gamma_L$  are the source and load reflection coefficients.

$S_{11}$ ,  $S_{22}$ ,  $S_{21}$  are the scattering coefficients of the attenuator, a and b referring to the starting and finishing value.

The probability distribution for mismatch using this formula is considered to be normal, compared to the limit distribution of formula 6 and 7 shown earlier in this paper.

### RF leakage

In wide dynamic range measurements it is essential to check for rf leakage bypassing the device under test. This may be done by setting the system to the highest attenuation and moving a dielectric, such as a hand or metal object, over the equipment and cables. In a leaky system the detector output will vary as the leakage path is disturbed. If the measurement set up incorporates a levelling power meter, this can be used to produce a known repeatable step which should be constant for any setting of the device under test.

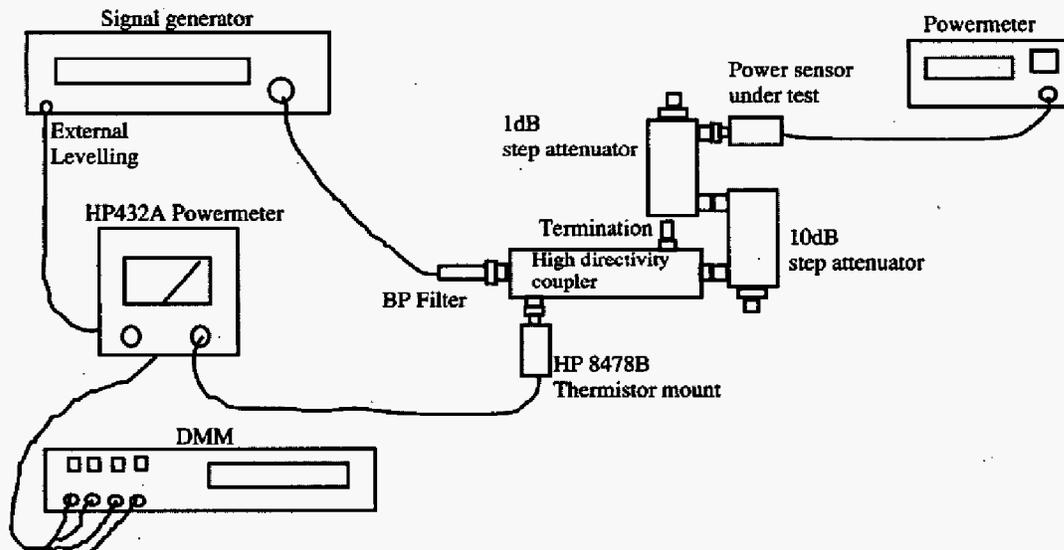
RF leakage may be reduced by using good condition precision connectors and semi rigid cables, wrapping connectors, isolators and other components in aluminium foil, keeping the rf source a distance from the detector and ensuring that there is no other laboratory work at the critical frequencies. When the measured attenuation is much greater than the leakage,  $A_1 \gg A_a$  then:

$$U_{LK} \text{ is approx } \pm 8.686 \cdot 10^{-(A_1 - A_a)/20} \quad (22)$$

where:

$A_1$  = leakage path

Diagram 20. Detector Linearity Tests



$A_a$  = DUT attenuator setting

$U_{LK}$  = uncertainty due to leakage

For example, if a leakage path 40 dB below the DUT attenuator setting is assumed, then error due to leakage could be:

Measured Step	Assumed Leakage Error
10 dB	$\pm 0.000$ dB
20	$\pm 0.000$
30	$\pm 0.000$
40	$\pm 0.000$
50	$\pm 0.000$
60	$\pm 0.001$
70	$\pm 0.003$
80	$\pm 0.009$
90	$\pm 0.027$
100	$\pm 0.087$
110	$\pm 0.275$

This contribution is a limit having a rectangular probability distribution.

### Detector linearity

This includes the linearity of the analogue to digital circuits of a digital voltmeter, square law of a diode or thermocouple detector, or mixer linearity of a superhetrodyne receiver. The linearity may be determined by applying the same small repeatable level change to the detector over its entire dynamic range. This level change may be produced with a very repeatable switched attenuator operating in a matched system.

Diagram 20 shows a variation of a technique described at the 22nd ARMMS Conference 1995 and at BEMC 1995. It involves applying a precise and repeatable 5 dB step, at various levels, to a power sensor detector or receiver under test. If the signal generator is externally levelled from the power meter recorder output as in diagram 20, then switching to a consecutive range, say 3 mW to 1 mW, produces a nominal but highly repeatable 5 dB step. The power level applied to the device under test is adjusted using the step attenuators such that the 5 dB step is applied over the instrument's operating range.

Measurements are best made starting with minimum power applied to the DUT and increasing the power to a maximum. The system may be automated using a GPIB controller to set the signal generator output and switch the step attenuator. The HP432A power meter bridge output voltages V0, V1, Vcomp are measured using a long scale digital voltmeter.

Chart 2. Power Sensor Linearity

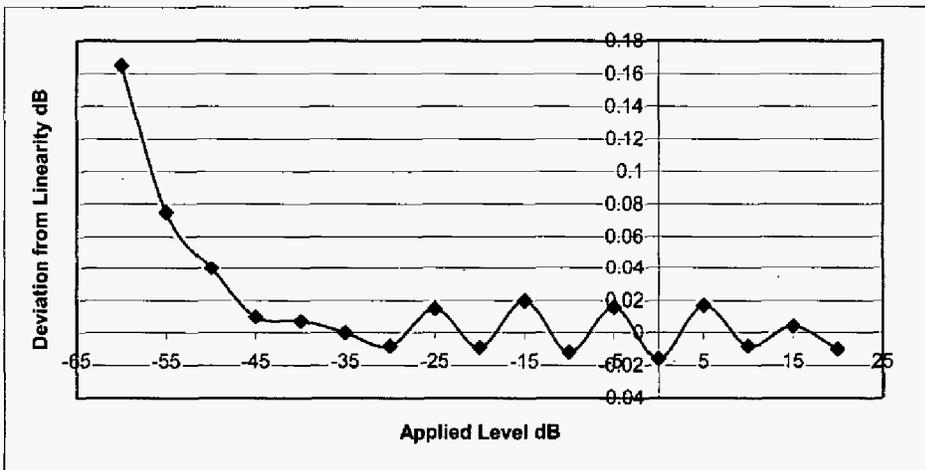


Chart 2 shows a typical diode power sensor linearity response using the measurement method described above. Note that drift and noise effect the measurement at levels less than -50 dBm.

#### System resolution

If a power meter or measurement receiver incorporating measurement ranges is used as the detector, then it is possible that the resolution of the instrument will depend upon whether it is measuring at near full scale or low scale. A power meter for instance, may resolve to 3 decimal places at full scale but only 2 decimal places at low scale. A digital voltmeter resolution is limited by the scale of its analogue to digital converter and the resolution of

#### Detector linearity measurement uncertainty budget.

Linearity of the 5 dB step due to the thermistor and power meter =  $\pm 0.0005$  dB

Measurement repeatability =  $\pm 0.01$  dB nominal, (but should be determined by multiple measurements).

Drift is determined by experiment and dependent upon power meter range.

Leakage effect is dependent upon step attenuator setting.

Mismatch: The effective source match,  $\Gamma_e$ , remains constant for the 5 dB step and any change in load impedance is seen as contributing to the measured non-linearity.

analogue meters must be determined by experiment. The contribution due to resolution is taken as a limit having a rectangular probability distribution.

#### System noise

The system noise is a measure of the receiver or detector sensitivity, where as a measurement gets close to the sensitivity limit, thermal, shot and  $1/f$  noise creates a fluctuating reading. The detector output may be time averaged to present a more stable indication but there is a limit to applying averaging, after which, this process becomes unhelpful and may obscure the correct answer. The system noise may be reduced by increasing the generator output level, using a phase sensitive detector, or using a low noise amplifier before

the detector. The gauge block technique shown in diagram 9 is another way around this problem.

Noise is considered as a random (type A) uncertainty contribution, where multiple measurements using the same equipment set up will give different results. (For systematic (type B) uncertainties, results change if the system set up changes.)

### Stability and drift

System stability and drift are particularly important uncertainty contributions for single channel measurement systems, where any drift in generator output or detector sensitivity will have a first order effect upon the measurement results. It is always good measurement practice to allow test equipment to temperature stabilise and to measure system drift over the possible time required to make a measurement.

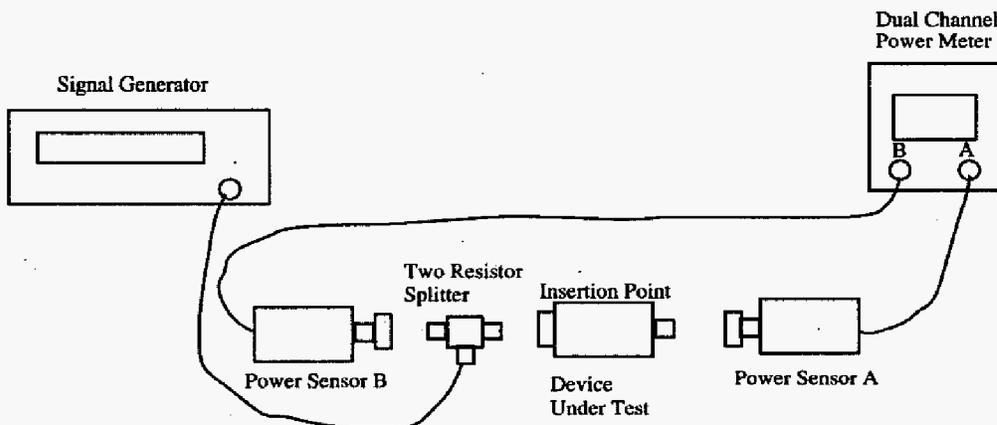
### Repeatability

The repeatability of a particular measurement may only be derived by fully repeating that measurement a number of times. Statistics, such as those described in UKAS document M3003, are used to determine the contribution to measurement uncertainty. If the device under test is a coaxial device inserted into the measurement system, it is usual to rotate the connector by 45° for each insertion. Uncertainty contributions due to the effect of flexing rf cables, mechanical vibrations, operator contribution, system noise, and rf switch contacts, to list but a few, should be fully explored.

Semi-rigid rf cables give lower leakage and attenuation/phase change than screened coaxial cables

### A WORKED EXAMPLE OF A 30 dB ATTENUATION MEASUREMENT

Diagram 21. Simple Dual Channel Power Ratio



when they are flexed. Chokeless waveguide flanges give better repeatability than choked joints. For precision measurements or where a microwave network analyser is employed it is useful to allow several people to make the same measurement in order to quantify operator uncertainty. Where an instrument or system contains mechanical microwave switches it has been found beneficial to exercise the switches several times if they have not been in use for some hours (a small software routine can do this).

### Calibration standard

Calibration standards such as the inductive voltage divider, IF piston attenuator, rf piston attenuator, rotary vane attenuator, switched coaxial attenuator and various power sensors have been covered previously in the text. These standards would normally be calibrated by the National Physical Laboratory or a UKAS Accredited Laboratory providing national traceability.

The National Standards are proven by exhaustive physical and scientific research and international intercomparison. The measurement uncertainties are usually calculated for approximately 95% confidence probability.

When using the calibration results from the higher Laboratory, the uncertainties must be increased to include the drift of the attenuator between calibrations and any interpolation between calibrated points. It is also important that the standard be used at the same temperature at which it was calibrated, particularly a piston or switched step attenuator. A 100 dB coaxial attenuator having a temperature coefficient of 0.0001 dB/°C will change by 0.04 dB for a 4°C change in temperature.

This worked example is for a simple case of a coaxial 30 dB attenuator, measured using the dual channel power ratio system shown in diagram 21. The measurement is repeated 5 times, from which the mean result and standard deviation are calculated. This example follows the requirements of UKAS document M3003.

P1A μW No DUT	P1B μW No DUT	P2A μW With DUT Inserted	P2A μW With DUT Inserted	Calculated Insertion Loss dB
10.001	10.1	10.01	10.2	30.039
10.002	10.1	10.01	10.2	30.039
9.993	10.1	9.99	10.1	30.047
9.997	10.0	9.99	10.2	30.089
9.995	10.0	9.98	10.1	30.050

\* Insertion loss is calculated from:

$$10 \log_{10} (P1A/P1B) \cdot (P2B/P2A) \quad (23)$$

The mean value of the five calculated samples  
= 30.053 dB

#### Contributions to Measurement Uncertainty Are:

**Type A random contributions,  $U_{ran}$ .** The estimated standard deviation of the uncorrected mean  
= 0.009 dB (normal probability distribution).

$$\text{This may be calculated using } (\sigma n - 1)/(n)^{0.5} \quad (24)$$

Where:

$\sigma$  is the standard deviation of the population  
 $n$  is the number of measurements

#### Mismatch contribution, $U_{mis}$

Where:

$$\Gamma_G = 0.05, \Gamma_L = 0.02, S_{11} = 0.07, S_{22} = 0.05, S_{12} \& S_{21} = 0.031$$

The measurement uncertainty of these values is taken as  $\pm 0.02$

and combining the measured value and its uncertainty in quadrature, we have:

$$\Gamma_G = 0.054, \Gamma_L = 0.028, S_{11} = 0.073 \text{ and } S_{22} = 0.054$$

Putting these values into the mismatch formula (21) we arrive at an uncertainty of :

0.026 dB (normal probability distribution)

**Detector linearity,  $U_{lin}$ .** The detector linearity was measured and found to be:

$\pm 0.02$  dB over 30 dB range (limit distribution).

No account has been taken for the measurement uncertainty of the detector linearity, as it is in the order of  $\pm 0.002$  dB. In practice, if this contribution was significant, it may be included by partial differentiation or by adding in quadrature, with the measured linearity error.

**Power meter resolution,  $U_{res}$ .** The uncertainty due to power meter resolution was determined by experiment and found to be a maximum of:

$\pm 0.03$  dB (limit distribution).

**Leakage.** The leakage was found to be less than 0.0001 dB and has not been included in the uncertainty calculations.

**Power meter range and zero carry over** are included in the power sensor linearity measurements and cannot be separated. Similarly, noise and drift cannot be separated from the random measurement of the five samples.

The following spreadsheet is a convenient way to list the uncertainly contributions and calculate the measurement uncertainty.

**Measurement Uncertainty Spreadsheet for a 30 dB Coaxial Attenuator at 10 GHz**

Symbol	U <sub>lin</sub>	U <sub>res</sub>	U <sub>mis</sub>	U <sub>ran</sub>		U <sub>eff</sub>	k	U <sub>c</sub> (Atten <sub>x</sub> )	Uncertainty
<b>Source of Uncertainty</b>	Linearity of Power Sensor	Resolution	Mismatch	$u(x_i) = \frac{\sigma_{est}}{(n)^{1/2}}$					
<b>Probability Distribution</b>	rect.	rect	normal	normal				normal	normal (k=2)
<b>Divisor</b>	$\sqrt{3}$	$\sqrt{3}$	1	1					
<b>Sensitivity Multiplier c<sub>i</sub></b>	1	1	1	1	<b>Measured Step</b>			<b>Combined Standard Uncertainty ±dB</b>	<b>Expanded Uncertainty ±dB</b>
<b>Contribution dB</b>	0.02	0.03	0.026	0.009	30.053	1386	2	0.039	0.08

Measured attenuation: 30.53 dB ± 0.08 dB

The reported expanded uncertainty is based on a standard uncertainty multiplied by a coverage factor k

= 2, providing a level of confidence of approximately 95%. The uncertainty evaluation has been carried out in accordance with UKAS requirements.

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