

ANAMET
The Collected ANAlyse Notes
Volume 1 – Numbers 1 to 10

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ABSTRACT

This report is a compilation of the first ten ANAlyse Technical Notes published originally by ANAMET – the automatic network analyser metrology club coordinated and administered by NPL. The Notes have been produced over a period dating from October 1993 to March 1995. Each Note is reproduced here in its original form.

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Preface

This report contains the first ten *ANALyse* Technical Notes published by ANAMET – the automatic network analyser (ANA) metrology (MET) club coordinated and administered by NPL. The club provides a forum for people and organisations interested in RF, microwave and millimetre-wave network analysis and measurements to meet to exchange ideas and experience. The club issues regular newsletters, holds meetings biannually and organises measurement comparison exercises for its members. It has also initiated a series of Technical Notes called *ANALyse* addressing topical technical issues relevant to the club's membership. ANAMET was launched in the summer of 1993 and the *ANALyse* series began in October 1993.

Although originally intended exclusively for the ANAMET membership, it was soon realised that the information contained in the *ANALyse* Notes would be of benefit to a wider audience and so the ANAMET membership permitted NPL to issue this Report, providing a compilation of the first ten *ANALyse* Notes. The original Notes were first issued during the period October 1993 to March 1995 and each Note is reproduced here in its original form.

ANALyse 1 describes improvements to a technique for assessing the directivity error of an ANA. The improvements rely on the use of the electrical delay and internal memory in the ANA to remove effects due to the load used for the assessment. *ANALyse* 2 provides a description of the theory underlying the technique described in *ANALyse* 1.

ANALyse 3 looks at the electrical effects of mating small coaxial connectors which are mechanically compatible – specifically, the 3.5 mm connector with the *K* (2.92 mm) connector, and the 2.4 mm connector with the *V* (1.85 mm) connector. *ANALyse* 4 gives definitions for terms commonly used in RF and microwave network analysis. *ANALyse* 5 gives an overview of time domain techniques for reflection measurements. It includes the use of gating and fencing to isolate sections of the overall response.

ANALyse 6, 7 and 8 deal with data processing problems with ANAs and, in particular, measurement comparison exercises using ANAs. *ANALyse* 6 addresses the problem of processing complex data as either magnitude and phase or real and imaginary components. The additional problem of expressing phase on either a 0° to 360° scale or a $\pm 180^\circ$ scale is also addressed. *ANALyse* 7 and 8 review the use of order statistics to provide an average value (the median) and measure of dispersion (the inter-quartile range) when analysing data sets. These estimators were used during the results analysis in one of the ANAMET comparison exercises.

ANALyse 9 discusses assessing the uncertainty of a single measurement. It introduces a method of reducing the amount of effort, and hence cost, spent assessing the uncertainty of measurement results. This is an important consideration for measurement facilities offering cost-effective services to customers.

ANALyse 10 presents some additional measurements made by one of the participants during an ANAMET comparison exercise. The measurements were made using a waveguide below cut-off (WBCO) attenuation system and are compared with the ANAMET average values for the four attenuators used for the comparison.

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ANALyse

An occasional series of Notes produced for, and by, the Members of ANAMET

Number 1

October 1993

Using the memory to extract the ripple on an HP8510C

Author: J P Ide, NPL, Malvern

Introduction.

It is generally preferable on philosophical grounds to check the performance of an instrument with known devices other than those used to calibrate it. Where the residual directivity of an ANA is being evaluated using a load and air line to generate ripples using an error separation technique, it is preferable to use a load other than the calibration load. This can lead to the situation in figure 1a and 1b where considerable judgement has to be used to separate the required ripple from the characteristics of the load. This note describes a technique for separating the two traces using the internal memory of an HP8510 and the electrical delay. The results of using this technique with two differing loads are shown in figures 1c and 1d.

Method.

Calibrate the ANA using your normal technique. This must be over a sufficiently large frequency range to show the ripples and with sufficient frequency points to show fine structure.

1. Attach a load (not the one used for calibration), and measure and display the reflection parameter appropriate to the port in question (this will be either S_{11} or S_{22}) using a linear Magnitude format. Store the data by pressing [DISPLAY],[DATA >-MEMORY]. Set up the display by selecting [SELECT DEFAULT], [MATHS OPERATION] and [MINUS] before selecting [MATH].
2. Attach load via air line to the port. Select [RESPONSE MENU], [ELECTRICAL DELAY] and adjust the delay time until the displayed length is equivalent to twice the length of the air line ($1\text{ns} \approx 300\text{mm}$). Fine adjust this delay time (length) until the overall displayed trace is flattest. The residual trace remaining under these circumstances should be a true display of the ripple alone.

Conclusions.

The accompanying figure shows that two significantly different loads (one a conventional termination, the other a 20 dB attenuator terminated with a short-circuit) give very similar results. It is interesting to note that this method exposes the optimistic picture at low frequencies given by the conventional method. The usefulness of this method depends upon the resolution being sufficient to expose the full detail and even the example traces show evidence of slight discretisation problems.

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Approved on behalf of Chief Executive, NPL
by Dr T G Blaney, Head, Division of Electrical Science

ANALyse Notes continue to be issued to the ANAMET membership and, in the fullness of time, further volumes of compiled Notes are to be expected. I would like to thank all the authors who have contributed to this compilation of Notes, the members of the ANAMET Steering Committee who reviewed and approved the contents of each Note and the National Measurement System Policy Unit of the Department of Trade and Industry for supporting the ANAMET club.

It should be clearly understood that the ANALyse series of Technical Notes were produced by, and for, the members of ANAMET. They were intended for fast dissemination of technical information for discussion purposes and do not necessarily represent an official viewpoint. No responsibility is accepted by the authors, ANAMET, or NPL for any use made of the information contained in the Notes.

NICK M RIDLER

Additional publications

RIDLER, N.M. and MEDLEY, J.C. A comparison of complex scattering coefficient measurements in 50 ohm coaxial line to 26.5 GHz, *National Physical Laboratory Report DES 138*, June 1995.

RIDLER, N.M. and JONES, G.D. Comparison assesses the quality of network measurements, *Microwaves & RF*, January 1995, pp 101-104.

RIDLER, N.M. and JONES, G.D. ANAMET comparison of type-N VSWR measurements, *20th ARMMS Conference Digest*, paper no 6, University of Nottingham, March 1994.

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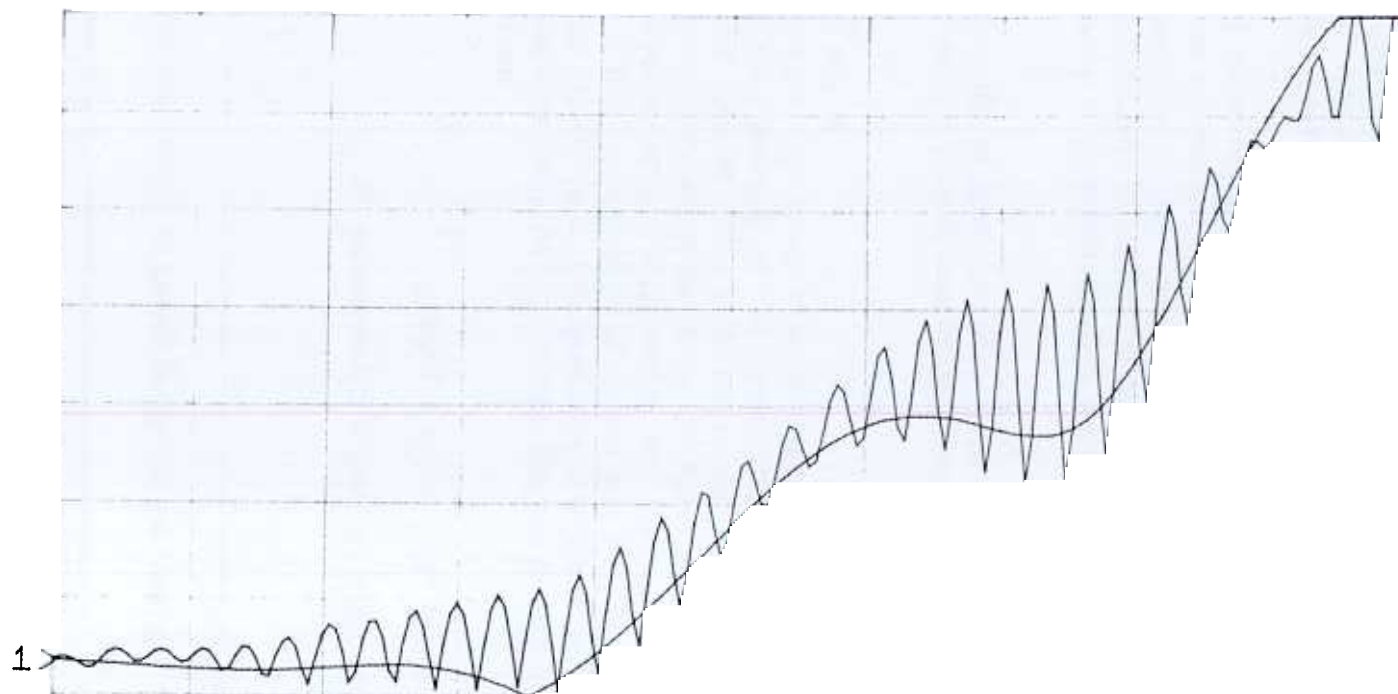


Figure 1(a): Load alone.

Figure 1(b): Load via
 300mm air line.

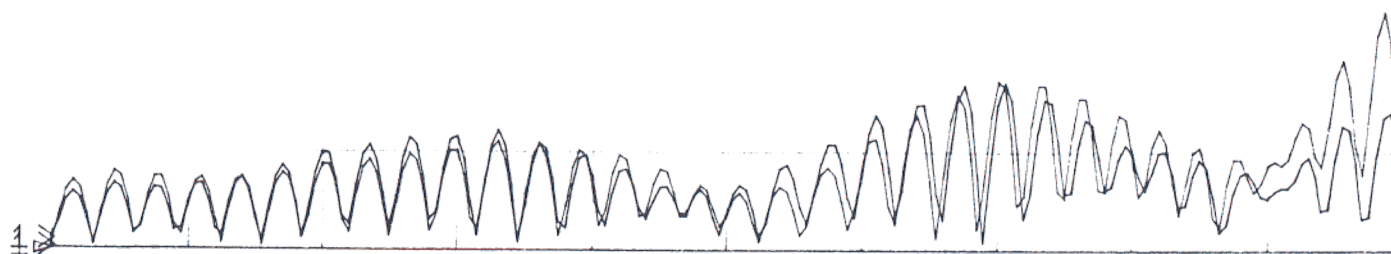


Figure 1(c): Ripple
 extracted from
 termination.

Figure 1(d): Ripple
 extracted from short-
 circuited 20 dB attenuator

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An occasional series of Notes produced by, and for, the Members of ANAMET

Number 2

February 1994

Using the memory to extract the ripple on an HP8510C
Part 2 - Why it works

Author: J C Medley, NPL, Malvern

Introduction.

The 'ripple' technique is a method of quantifying the quality of a calibration by evaluating the residual errors in the calibrated ANA system (throughout this Note the ANA is assumed calibrated). The magnitude of the residual directivity error (δ) is equivalent to the distance between the 'true' origin (O_T) of the VRC plane and that calculated by calibration (O_C).

$|\delta|$ can usually be found from the ripple on the plot of the VRC magnitude for an offset load (Method I). Unfortunately this doesn't always work. ANALyse Number 1 described Method II for finding $|\delta|$ which works even when Method I fails. This Note shows why this happens.

General idea.

Take a near-matched load. Assume it has constant VRC magnitude, $|\Gamma_T| > 0$, with respect to the 'true' origin, O_T . If this load is offset by a lossless air line, its measured VRC, Γ_M , will change with frequency and trace out a circle, centred on O_T , with radius $|\Gamma_T|$. If $|\Gamma_M|$ is then plotted against frequency, its value will ripple as Γ_M traces out the circle in the VRC plane.

Method I calculates $|\delta|$ from $|\delta| = \{\max(|\Gamma_M|) - \min(|\Gamma_M|)\} / 2$, where $\max(|\Gamma_M|)$ and $\min(|\Gamma_M|)$ are adjacent maximum and minimum values of $|\Gamma_M|$ taken from the ripple trace. Adjacent maxima and minima are used as in general $|\delta|$ changes slowly with frequency (as does $|\Gamma_T|$).

If $|\Gamma_T| \geq |\delta|$ then we have the situation shown in Figure 1. From this we can see that:

$$\begin{aligned}\max(|\Gamma_M|) &= |\delta| + |\Gamma_T| ; & \min(|\Gamma_M|) &= |\Gamma_T| - |\delta| \\ \therefore \max(|\Gamma_M|) - \min(|\Gamma_M|) &= 2|\delta|\end{aligned}$$

ie Method I gives the required answer for $|\delta|$

If $|\Gamma_T| < |\delta|$ then we have the situation shown in Figure 2. In this case:

$$\begin{aligned}\max(|\Gamma_M|) &= |\delta| + |\Gamma_T| ; & \min(|\Gamma_M|) &= |\delta| - |\Gamma_T| \\ \therefore \max(|\Gamma_M|) - \min(|\Gamma_M|) &= 2|\Gamma_T|\end{aligned}$$

ie Method I gives the magnitude of the load used, not $|\delta|$

Method II calculates $|\delta|$ from a ripple trace modified as follows:

- 1) The load is measured without an offset, giving a measured VRC Γ_0 .
- 2) The load is measured, offset by the air line, giving Γ_M as above.
- 3) The electrical delay function of the ANA is used with a delay time equivalent to twice the length of the air line, to give a modified VRC Γ_M' .
- 4) The VRC of the load (without offset), Γ_0 , is then subtracted from Γ_M' , and $|\Gamma_M' - \Gamma_0|$ plotted as the ripple trace.

$|\delta|$ is then calculated from $|\delta| = \{\max(|\Gamma_M' - \Gamma_0|) - \min(|\Gamma_M' - \Gamma_0|)\} / 2$, where adjacent maximum and minimum values of $|\Gamma_M' - \Gamma_0|$ are taken from the modified ripple trace.

Why this works.

Consider the situation $|\Gamma_T| < |\delta|$ shown in Figure 2. Redefine the x-axis as the line joining O_C and O_T , Figure 3 (this is just a rotation of the VRC plane about O_C which does not affect measured magnitudes but makes the Figures easier to follow). Figure 3 also defines the angles ϕ , θ , α and β .

As can be seen from Figure 3, the effect of offsetting the load is to introduce a delay of $(\theta - \phi)$ about the true origin O_T . Using the delay function of the ANA attempts to remove the delay introduced by the air line by adding a phase value of approximately $(\theta - \phi)$ to Γ_M , rotating the vector about the calculated origin O_C , to give Γ_M' (see Figure 4). Hence:

$$|\Gamma_M'| = |\Gamma_M|, \quad \arg(\Gamma_M') = \arg(\Gamma_M) + (\theta - \phi)$$

If we let $\Gamma_M' = x_M' + jy_M'$ then from Figure 4 we can see that:

$$\begin{aligned} -|\Gamma_M| \cos(\alpha + \phi) &= -|\Gamma_M| (\cos \alpha \cos \phi - \sin \alpha \sin \phi) \\ y_M' &= |\Gamma_M| \sin(\alpha + \phi) = |\Gamma_M| (\sin \alpha \cos \phi + \sin \phi \cos \alpha) \end{aligned}$$

Looking at the triangle $O_C O_T \Gamma_M$ we can deduce that:

$$\begin{aligned} &= |\Gamma_T|^2 + |\delta|^2 - 2|\Gamma_T||\delta|\cos\theta && \text{by the cosine rule} \\ \sin\alpha &= \frac{|\delta|}{|\Gamma_M|} \sin\theta && \text{by the sine rule} \\ \cos\alpha &= \frac{|\Gamma_T|^2 + |\Gamma_M|^2 - |\delta|^2}{2|\Gamma_T||\Gamma_M|} && \text{by the cosine rule} \end{aligned}$$

Substituting for $|\Gamma_M|^2$ in this last expression gives:

$$\begin{aligned} \cos\alpha &= \frac{|\Gamma_T|^2 + |\Gamma_T|^2 + |\delta|^2 - 2|\Gamma_T||\delta|\cos\theta - |\delta|^2}{2|\Gamma_T||\Gamma_M|} \\ &= \frac{|\Gamma_T| - |\delta|\cos\theta}{|\Gamma_M|} \end{aligned}$$

Substituting for $\cos\alpha$ and $\sin\alpha$ in the expressions for x_M' and y_M' gives:

$$\begin{aligned}
 x_M' &= -|\Gamma_M| \frac{(|\Gamma_T| - |\delta| \cos \theta)}{|\Gamma_M|} \cos \phi - \frac{|\delta|}{|\Gamma_M|} \sin \theta \sin \phi \\
 &= -|\Gamma_T| \cos \phi + |\delta| (\cos \theta \cos \phi + \sin \theta \sin \phi) \\
 &\quad - |\Gamma_T| \cos \phi + |\delta| \cos(\theta - \phi) \\
 y_M' &= |\Gamma_M| \frac{|\delta|}{|\Gamma_M|} \sin \theta \cos \phi + \sin \phi \frac{(|\Gamma_T| - |\delta| \cos \theta)}{|\Gamma_M|} \\
 &= |\Gamma_T| \sin \phi + |\delta| (\sin \theta \cos \phi - \sin \phi \cos \theta) \\
 &= |\Gamma_T| \sin \phi + |\delta| \sin(\theta - \phi)
 \end{aligned}$$

From Figure 3, the VRC of the load (without offset), Γ_0 , is:

$$\Gamma_0 = (-|\Gamma_T| \cos \phi + |\delta|) + j(|\Gamma_T| \sin \phi)$$

Subtracting Γ_0 from Γ_M' and looking at the magnitude of the result gives:

$$\begin{aligned}
 |\Gamma_M' - \Gamma_0|^2 &= \left(-|\Gamma_T| \cos \phi + |\delta| \cos(\theta - \phi) + |\Gamma_T| \cos \phi - |\delta| \right)^2 + \left(j(|\Gamma_T| \sin \phi + |\delta| \sin(\theta - \phi) - |\Gamma_T| \sin \phi) \right)^2 \\
 &= |\delta|^2 (1 - 2 \cos(\theta - \phi) + \cos^2(\theta - \phi) + \sin^2(\theta - \phi)) \\
 &\quad 2|\delta|^2 (1 - \cos(\theta - \phi)) \\
 \text{ie } |\Gamma_M' - \Gamma_0| &= |\delta| \sqrt{2(1 - \cos(\theta - \phi))} \\
 &= 2|\delta| \left| \sin \frac{1}{2}(\theta - \phi) \right|
 \end{aligned}$$

Hence $\Gamma_M' - \Gamma_0$ traces out a circle in the VRC plane, with radius $|\delta|$, passing through the point $\Gamma_M' - \Gamma_0 = 0$.

$|\Gamma_M' - \Gamma_0|$ is the quantity plotted on the modified ripple graph. It has the form of a full-wave rectified sine wave of amplitude $2|\delta|$ and therefore 'ripples' between 0 and $2|\delta|$ as required:

$$\begin{aligned}
 \max(|\Gamma_M' - \Gamma_0|) &= 2|\delta| ; \quad \min(|\Gamma_M' - \Gamma_0|) = 0 \\
 \therefore \max(|\Gamma_M' - \Gamma_0|) - \min(|\Gamma_M' - \Gamma_0|) &= 2|\delta|
 \end{aligned}$$

Figure

Figure 1

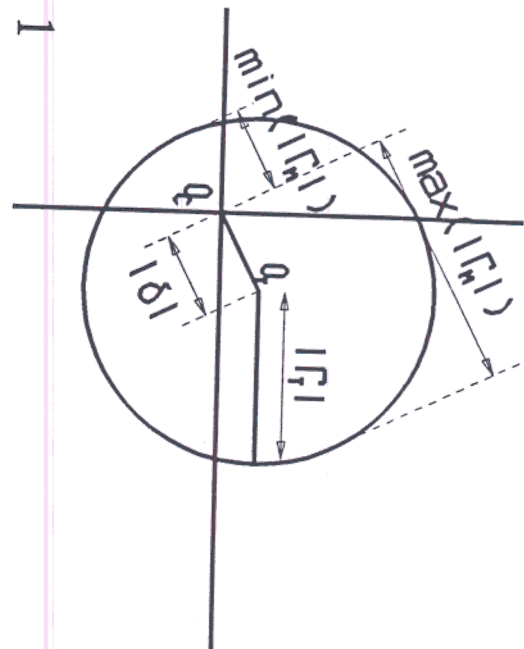


Figure 3

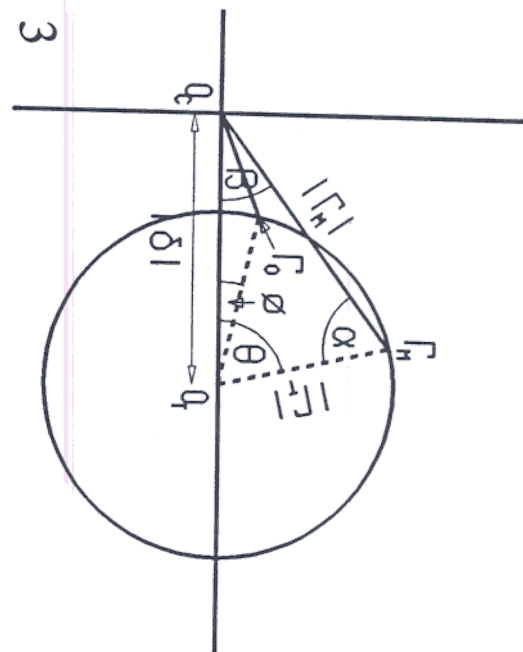


Figure 2

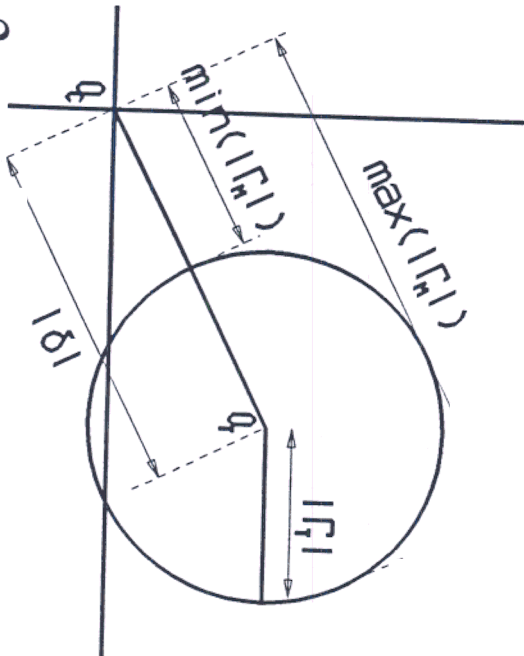
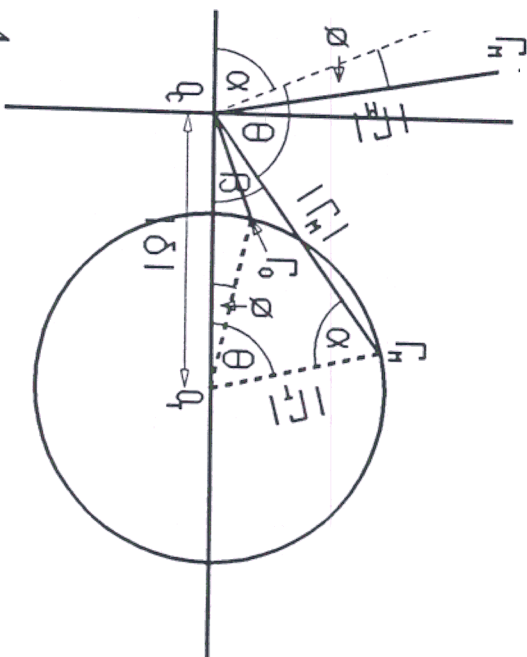


Figure 4



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Number 3

January 1994

A study of the electrical compatibility of mechanically mateable coaxial lines

Author: J P Ide, NPL, Malvern

Introduction.

Manufacturers of small coaxial connectors have agreed the mechanical dimensions of some small coaxial connectors so that they can be mated non-destructively. The fact that different series connectors can be mated without mechanical damage has led to the introduction of the phrase 'mechanically compatible'. Because both lines are nominally 50 Ω , it is assumed that mechanical compatibility equates to electrical compatibility. This note describes an investigation into the electrical compatibility of the mating of APC3.5 (3.5 mm) with K (2.92 mm) connectors and APC2.4 (2.4 mm) with V (1.85 mm) connectors. It derives an approximate method of determining the size of any effects and thus a way of estimating the significance of the effect in any particular situation.

Method.

When two lines with the same characteristic impedance, 50 Ω , but with different physical sizes meet, there are step changes in the inner conductor and outer conductor in the same direction. An approximate solution to this situation has been derived by Whinnery et al [1] who showed that each step in a conductor diameter can be treated as an equivalent capacitance and that the various equivalent capacitors can be superimposed and combined. The values for the equivalent capacitors for the two cases in question are derived in appendix A. Using these figures the effect of an additional admittance term can be calculated which gives the approximate reflection coefficient that would be generated at the junction. Figure 2 shows the magnitude of the voltage reflection coefficient vs frequency for the two cases. The traces stop at the maximum frequency of the larger line size of the pair. Figure 3 shows the effect of the reflection from the junction on a real measurement. A load fitted with a K-connector has been measured on the same system when it has been set up and calibrated as an APC3.5 system and as a K system. The solid trace shows the true reflection coefficient while the dashed trace shows the load reflection coefficient combined with the effects of the junction reflection coefficient.

Conclusions.

A method of estimating the electrical mismatch between two pairs of mechanically mateable nominal 50 Ω coaxial lines has been demonstrated. It shows that in most cases and for most of the frequency range the step capacitance caused by the mechanical discontinuities is significant and cannot be ignored. This implies that the connectors should not be treated as interchangeable where electrical performance is paramount.

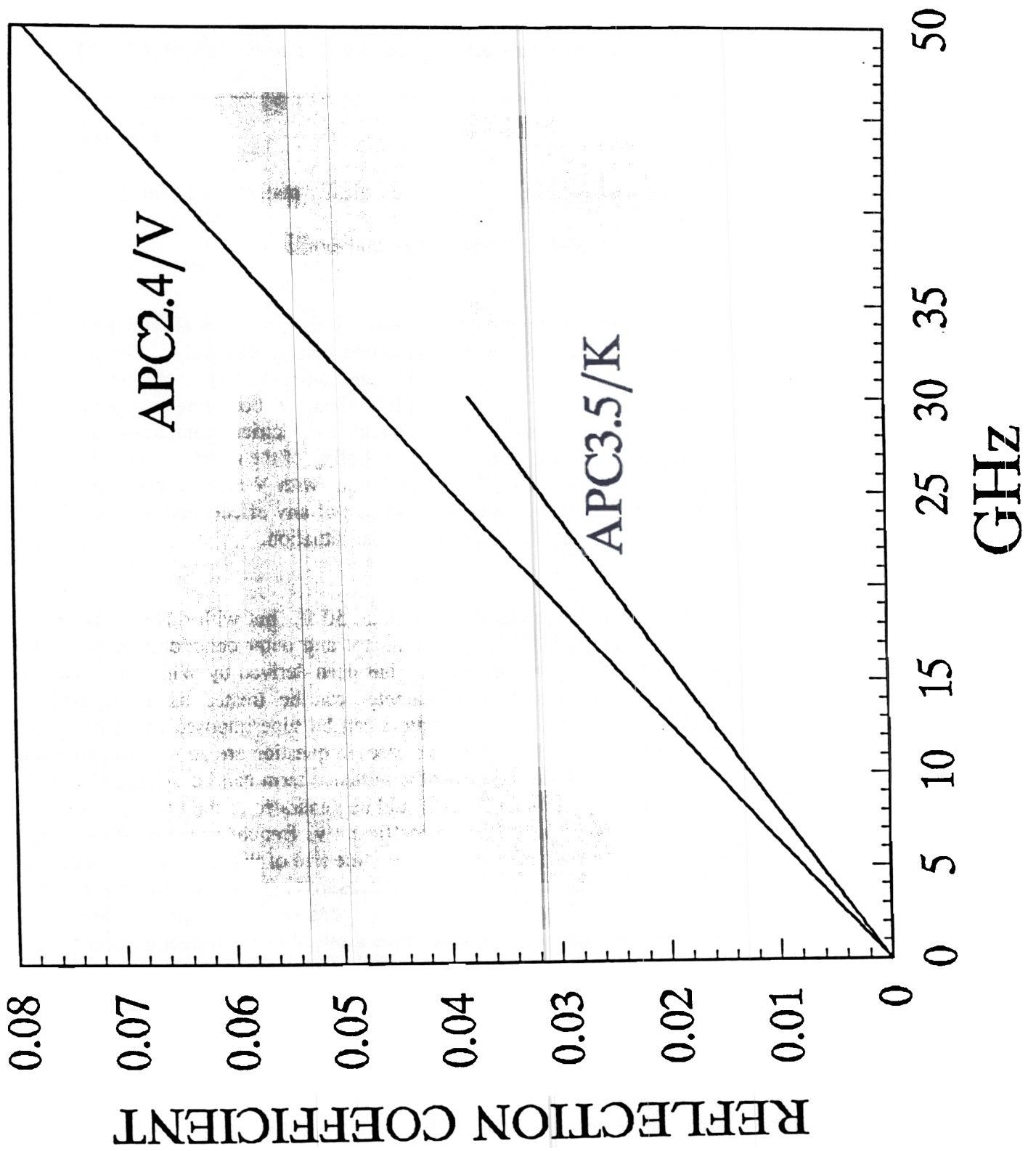


Figure 1. The predicted magnitude of reflection coefficient of the two junctions.

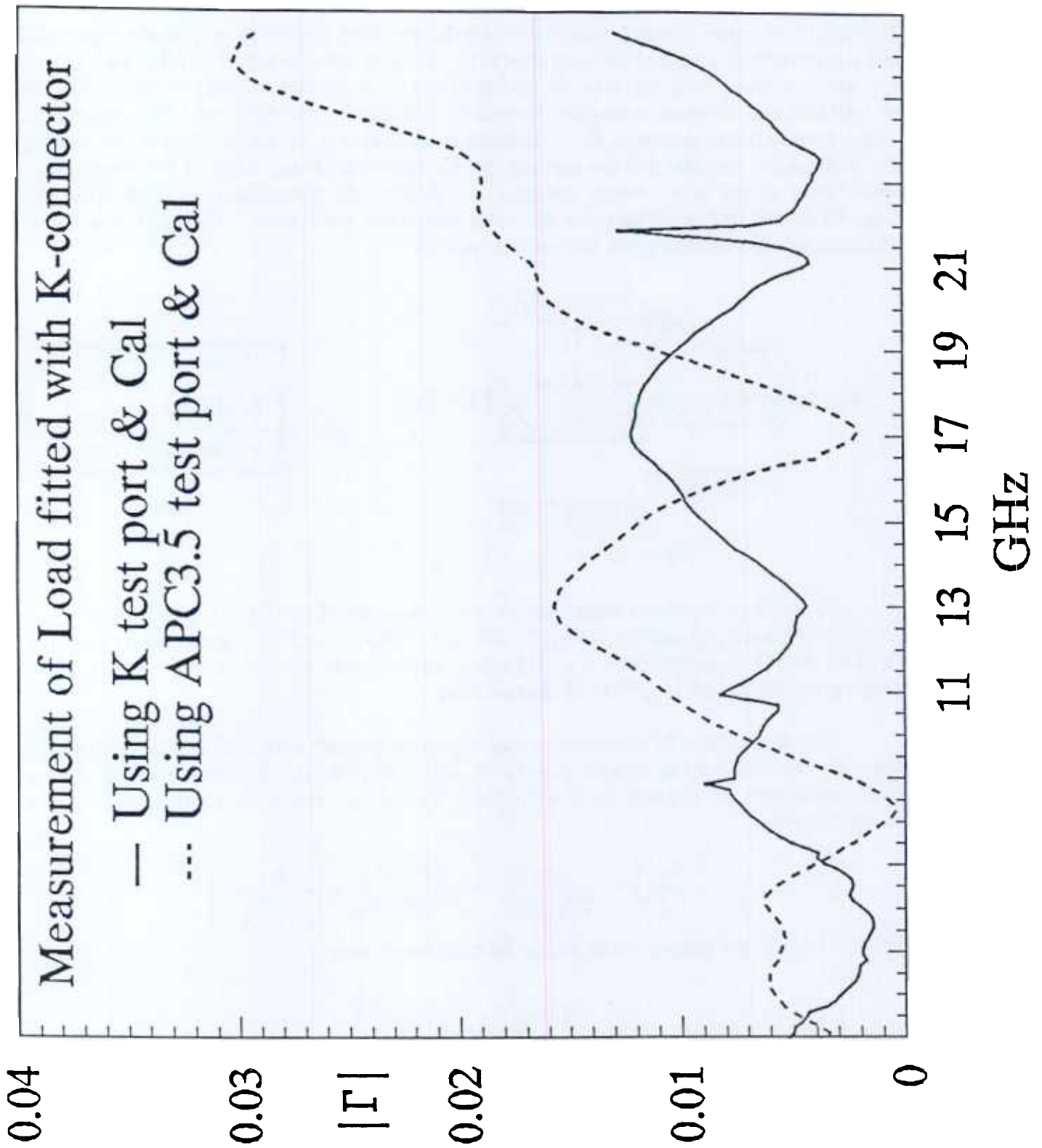


Figure 2. The measured magnitude of reflection coefficient of a load fitted with K-connectors when measured on APC3.5 and K systems.

Appendix A

This derivation by Whinnery was rediscovered by T E Hodgetts (also of NPL Malvern) who also demonstrated its application to this problem and derived the impedance/admittance transformation.

The step capacitance, C_s is derived from knowledge of the size of the step in the conductor and the equivalent length of the step (from the radius of the conductor). In the case where 3.5 mm line meets 2.92 mm line, the radius of the outer conductor changes from 1.75 mm to 1.46 mm and the inner conductor changes from 0.76 mm to 0.635 mm. C_s is calculated as the sum of the capacitance, C_o , calculated from changing the outer radius while leaving the inner radius constant and the capacitance, C_i , calculated from changing the inner radius while leaving the outer radius constant. Graphs are given in [1] which give the capacitance/unit circumference for the inner and outer conductors. Figure 3 is a cross-sectional diagram labelling the various dimensions.

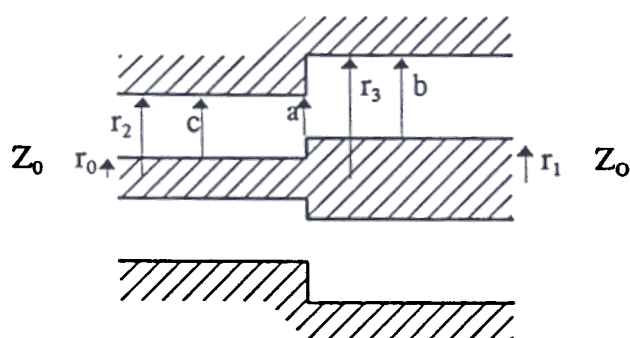


Figure 3.

Junction	C_s in fF
K/APC3.5	8.0
V/APC2.4	10.1

Table 1

In terms of the dimensions labelled on the diagram above $C_s = C_i + C_o$, where $C_i = 2\pi r_2 C'(a/c, r_2/r_0)$ and $C_o = 2\pi r_1 C''(a/b, r_3/r_1)$ when C' and C'' are the values from the relevant graphs in figures 8 & 9 of [1] using the values in brackets as coordinates. The graph gives the values in pF/cm of discontinuity.

The capacitance of the junction will appear in parallel with the impedance Z_0 of the line. By converting to admittance parameters where Y_0 is the admittance of the line and the local admittance, Y is given by $Y = Y_0 + j\omega C_s$, the voltage reflection coefficient Γ can be calculated from

$$\Gamma = \frac{Y_0 - Y}{Y_0 + Y} = \frac{-j\omega C_s}{2Y_0 + j\omega C_s} = \frac{-j\omega C_s Z_0}{2 + j\omega C_s Z_0} = \left(\frac{2j}{\omega C_s Z_0} - 1 \right)$$

From this $|\Gamma|$, the magnitude of Γ , can be calculated using

$$|\Gamma| = \left(\left(\frac{2}{\omega C_s Z_0} \right)^2 + 1 \right)^{-\frac{1}{2}}$$

This is the expression used to generate the plots in the main text from the values of the capacitance given in table 1. To ensure the estimate of the effects of the discontinuity are not underestimated, the values in the table are actually 20% higher than calculated. This is to allow for second-order frequency dependent effects which have been ignored. This implies that the above expression is useful for determining the probable size of errors but is not accurate enough to be useful in correcting those errors.

[1] J.R.Whinnery, H.W.Jamieson and T.E.Robbins. "Coaxial-Line Discontinuities", Proceedings of the IRE, Vol 32, pp695-709, November 1944.

ANALyse

An occasional series of Notes produced for, and by, the Members of ANAMET

Number 4

July 1994

Definitions of some of the terms in common use with ANAs

Author: J P Ide, NPL, Malvern

Introduction.

Any field of interest rapidly gathers its own specialists and inevitably its own language. Many of the words are borrowed from other similar interests, with slightly modified meanings and modified emphasis while others are re-invented to cover entirely new concepts. Automatic Network Analysers (ANAs) are now an essential adjunct to most RF and microwave measurements and ANA users have developed a language tailored to its peculiar strengths and weaknesses. This list of definitions contains many words with different meanings and implications to other fields but the meanings given here are couched in terms relevant to Network Analyser users. These definitions are not necessarily the most mathematically exact representation of observed behaviour but they have been chosen as the most useful and helpful meanings from the viewpoint of a user. Words used in the definitions that are highlighted in bold are themselves defined in this list.

Definitions.

ACCURACY: The closeness of agreement between the result of a measurement and its true value.

ASSURANCE: Using error separation techniques to determine the measurement errors and predict the uncertainties.

ATTENUATION: The insertion loss of a device when the source impedance and load impedance match are both exactly matched to the reference impedance.

CALIBRATION: The process of using standards to determine values for the error terms.

CHARACTERISTIC IMPEDANCE: An electrical parameter derived from the relationship between electric and magnetic fields in a medium guiding an electromagnetic wave.

NOTE - It is affected by the mechanical dimensions and the properties of the materials from which it was constructed. Each propagation mode has its own characteristic impedance.

COEFFICIENT: A physically dimensionless complex number.

NOTE - used to describe the ratio of two RF voltages or electric field strengths in both transmission and reflection and also in the mathematical modelling of a network analyser.

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CONNECTOR REPEATABILITY: The Random error that is introduced by the connectors.

NOTE - the repeatability is likely to be a function of the frequency of operation and the torque used to tighten the connector as well as the amount of wear to which both halves of the the connector pair have been subjected.

CORRECTED DATA: The values of the measurement data in an ANA after computer correction.

CROSSTALK: Undesirable signal reaching one measuring port from the other without passing through the DUT.

DIRECTIVITY: A term describing the error in a perfectly terminated reflectometer which arises from imperfect separation of signals travelling in the forward and reverse direction.

NOTE - A more specific definition is the ratio (in decibels) of the power emerging from the coupled port, for a given signal propagating "only" in the forward direction of the main arm, to the power emerging from the coupled port with the same signal propagating only in the reverse direction in the main arm.

DUT: Device Under Test

DYNAMIC ACCURACY: Accuracy as a function of the amplitude of a measured signal when all possible phase values are included.

DYNAMIC RANGE: The ratio of the maximum indication to the minimum indication.

NOTE - This ratio is usually expressed in decibels (dB). The maximum indication is often related to a stated level of compression in the detection system and the minimum indication is usually determined by the noise floor.

EFFECTIVE ERROR TERMS: The difference between the values for the error terms determined during calibration and their true values.

NOTE - This can be due to a combination of imperfect calibration standards, connector repeatability, instrument repeatability and errors in the model used to describe the system. See also Residual Errors.

ELECTRICAL LENGTH: The amount by which the phase of a signal is changed as a travels through a device.

NOTE - The value may be given in absolute or relative degrees, mechanical length for a specified dielectric, or in time.

ERROR: The difference between the measured value and the true value of a quantity.

ERROR CORRECTION: The process of using the values for the error terms, determined through calibration, to correct the raw data to give a more accurate measurement.

NOTE - Although the measurement accuracy will be improved by calibration, there are limits to the amount of improvement possible so the measurement will still be imperfect. See Effective Errors.

ERROR SEPARATION: A technique for extracting error information about an ANA by changing the phase relationships between the test port and the DUT and examining the results over a range of frequencies.

NOTE - This is also known as the ripple technique because the errors are determined by examination of the ripples in the results. For accurate determination the phase relationship must change by at least 180°.

ERROR TERMS: A set of coefficients describing the performance of the measuring instrument. Values for these terms can be evaluated by calibration.

NOTE - Some of these error terms have specific names, eg Directivity, Source match, Isolation.

FREQUENCY DOMAIN: Measurements made where the independent variable is frequency.

NOTE - This is used to determine the overall frequency response of a device, especially useful for narrow band devices such as filters and tuners.

INSTRUMENT REPEATABILITY: The random error that is introduced by the measuring instrument.

ISOLATION: An error term describing the transmission paths through an ANA when both ports are perfectly terminated.

LOAD MATCH: A measure of the closeness of the impedance of a test port to the reference impedance when the test port is being used to terminate a DUT.

MATCH: A measure of the closeness of the impedance of a termination to the reference impedance of the medium.

NOTE - Voltage Reflection Coefficient, the Voltage Standing Wave Ratio and Standing Wave Ratio are some of the expressions used to describe this term.

NOISE FLOOR: The power level detected with no signal applied to the measurement device.

NOTE - Generally the signal level which results from noise generated internal to the measurement system or is unavoidable at the input of the system.

RAW DATA: The values of the measurement data in an ANA before computer correction.

REFERENCE IMPEDANCE: An electrical parameter to which measurements of the scattering coefficients are referred.

NOTE - This is usually, but not necessarily, the characteristic impedance of a lossless version of the transmission line.

REFERENCE PLANE: The location to which zero phase is referenced at all frequencies.

NOTE - In a connector it is usually the outer conductor mating surface.

RESIDUAL ERROR TERMS: See effective errors.

RESOLUTION: The smallest increment in the measured value that can be distinguished.

NOTE - It is important that resolution and accuracy are not confused. The resolution is set by the instrumentation while the accuracy is set by the calibration.

RETURN LOSS: The logarithmic measure of the ratio of the power reflected from a DUT to the power incident on that DUT.

NOTE - It can also be defined in terms of the amplitudes of reflected and incident voltages.

SCATTERING COEFFICIENTS: The set of transmission and reflection coefficients describing a multi-port network.

SOURCE MATCH: A measure of the closeness of the impedance of a test port to the reference impedance when the test port is being used as a source for a DUT.

STANDARD: A device whose electrical characteristics are known either by measurements at a standards laboratory or by theoretical prediction from its mechanical characteristics.

TIME DOMAIN: Measurements made where the independent variable is time.

NOTE - Because time and length are so closely bound together, this method can be used to either determine the physical location of reflections or the size of reflections at a physical location.

TRACKING ERROR: The departure from unity of the proportionality between the indicated value and the measured value.

TRUE VALUE: The value of a quantity which the measurement attempts to determine.

NOTE - This is the value that would be obtained by a perfect measurement. In practice the true value is unknown and unknowable unless it is a defined value.

TYPES OF MEASUREMENT ERRORS: Errors can be separated into two types for ease of analysis.

RANDOM: A non-repeatable, non-systematic error.

SYSTEMATIC: An error which is fixed and repeatable - an inbuilt bias.

UNCERTAINTY: The limits about a measured value between which the true value of a quantity is estimated to lie.

VALIDATION: The use of items whose performance is known from previous measurements to check the performance of the calibrated instrument.

VERIFICATION: The use of standards, different from those used in the calibration process, to check the performance of the calibrated instrument.

References.

1. ISO(1993), 'International vocabulary of basic and general terms in metrology', ISO, Geneva, Switzerland.
2. F Jay (ed): 'IEEE Standard dictionary of electrical and electronic terms', ANSI/IEEE Standard 100-1988.
3. British Standard Glossary of 'Terms Used in Metrology'. BSI BS5233: 1986.

ANALyse

An occasional series of Notes produced by, and for, the Members of ANAMET

Number 5

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Time Domain Measurements Using a Reflection Analyzer

Author: Chris Potter, Marconi Instruments, Stevenage

Introduction

The time domain analysis software built into Vector Reflection Analyzers such as the Marconi Instruments 6210 allows the frequency domain reflection coefficient data to be transformed into the time domain. The measurement trace can be interpreted as reflection coefficient against time (or distance) into the device under test. It is possible to identify the location of impedance discontinuities and imperfections within the Device Under Test (DUT) from the time taken for the measuring signal to reach the discontinuity and return to the test port. This aids fault-finding and gives insight into DUT operation. Furthermore, it is possible to edit the time domain response, and return to the frequency domain, showing the frequency response of the edited DUT.

Types of Time Domain Measurements

There are two types of time domain transform, Low-Pass and Band-Pass. The names come from the range of frequencies that have been used for measurement, as if a filter had been applied to an infinite range of frequencies. The Band-Pass transform uses a non-zero start and stop frequency. The Low-Pass transform is used when the measurement data points are extrapolated down to DC. For this reason, Low-Pass mode may only be used on broadband DUTs where there is a usable performance down to DC. This allows the characteristic impedance of transmission lines and the capacitive or inductive nature of discontinuities to be quantified. Alternatively, Band-Pass mode is used with band-pass DUTs, such as waveguide. However, the lack of DC information means that only the magnitude of individual reflections can be determined in the time domain. It is not possible to determine whether a discontinuity is inductive or capacitive, or whether a transmission line is too high or too low in characteristic impedance.

For both types of time domain analysis, the frequency points have to be carefully chosen for the mathematics to work properly. Therefore the choice of sweep mode affects the time domain modes which can be selected. Band-Pass time domain mode is available after a calibration using the Start and Stop Frequency sweep mode or the Waveguide sweep mode.

The Waveguide sweep mode is used to enable physical distance to be displayed in a waveguide run, by fully correcting for the effects of waveguide dispersion. If Low-Pass time domain is required, Low Pass sweep mode must first be selected before calibration and measurement. This is because the frequency points are chosen to lie on a harmonic sequence stretching from DC to the highest frequency. Unlike the frequency domain measurement, it is possible to display any time domain region without the need to re-calibrate the instrument. The only limit is the maximum time range, which is determined by the interval between measurement frequency points.

With the Low-Pass time domain mode, the *Real* display format is used. This allows the reflection coefficient of the DUT to be displayed as a function of distance or time into it. Note that the display may be negative as well as positive. A value of +1 corresponds to an open circuit, and a value of -1 corresponds to a short circuit. If this were band-pass mode, the two responses would be indistinguishable. A steady trace above the zero line indicates that the characteristic impedance of that section of transmission line is too high. An upward transient indicates inductance, and a downward transient indicates capacitance, as shown in figure 1.

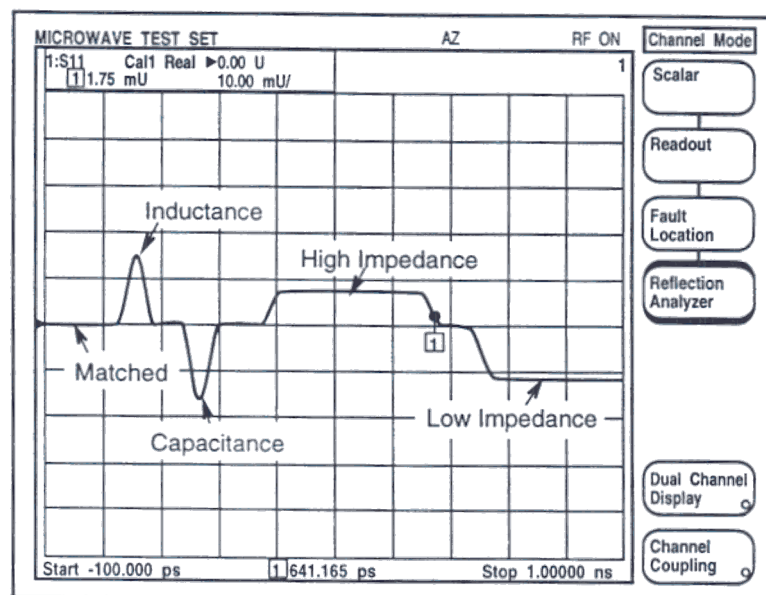


Figure 1 : Low-Pass Time Domain Step Responses

This information allows the operator to locate the problem within a circuit, and predict the type of corrective action necessary. A useful approximation in a 50 Ω system is that 1 $\Omega \cong 10$ mU for small reflections. For example, -20 mU $\cong 48 \Omega$.

Gating and Fencing

This is a powerful editing technique that allows particular sections of the DUT to be isolated or removed from the overall response. A mathematical function is applied to the measurement to make the chosen sections of the DUT appear matched. The editing is done in the time domain, and the result may then be transformed back into the frequency domain. This allows the effects of individual discontinuities to be isolated and evaluated, so that the section of a DUT to be improved can be determined with confidence.

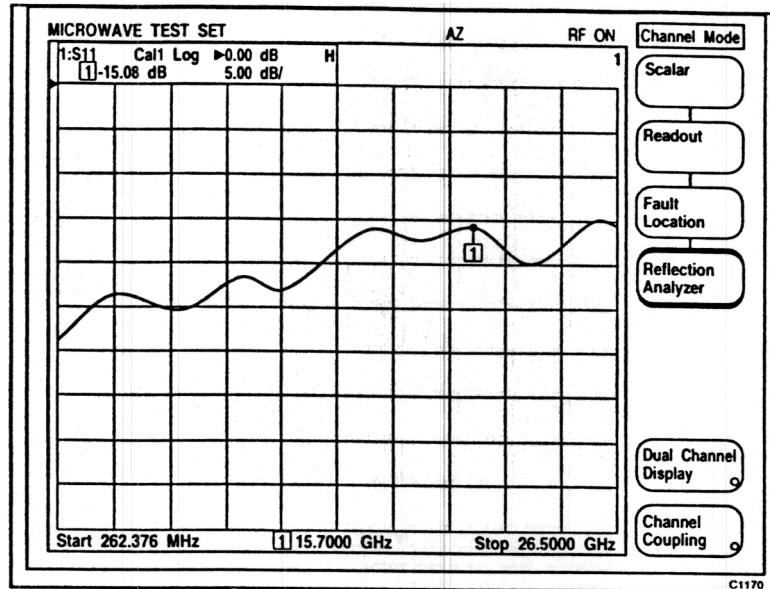


Figure 2 - Typical DUT return loss

For example, figure 2 shows the frequency domain return loss of a typical DUT. Time domain may be used to further analyse its behaviour. Figure 3 shows the Low-Pass step response of the DUT. A capacitive discontinuity is clearly visible at 230 ps delay. Note that this corresponds to the time taken for the excitation step to travel from the test port to the discontinuity. Therefore, in air, this corresponds to a distance from the test port of 69 mm, assuming the speed of light to be 3×10^{11} mm/s.

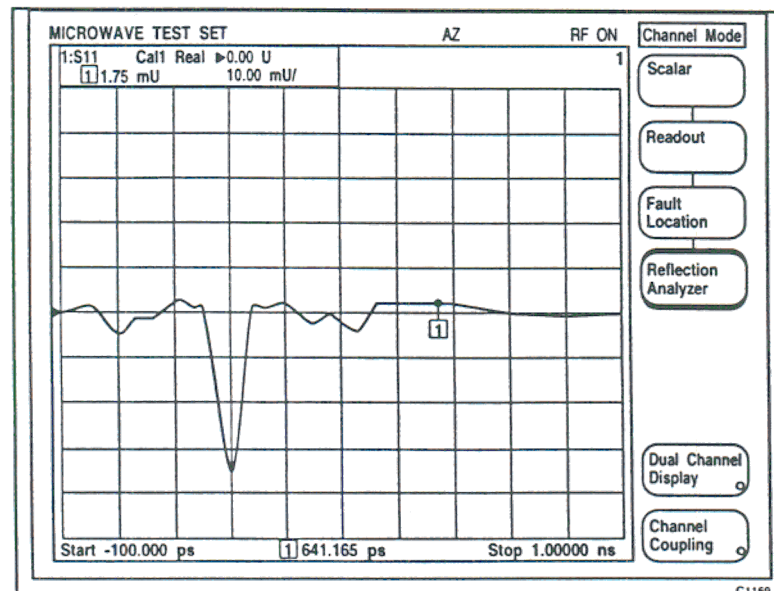


Figure 3 - Low-pass step response of DUT

Gating is a function which retains only the time domain data within the gated region, and mathematically matches the rest of the trace outside the region. Fencing is the inverse of gating, which matches the fenced region and keeps the trace unchanged outside the region. Figure 4 shows the fencing markers positioned around the region of interest and the fence function turned on.

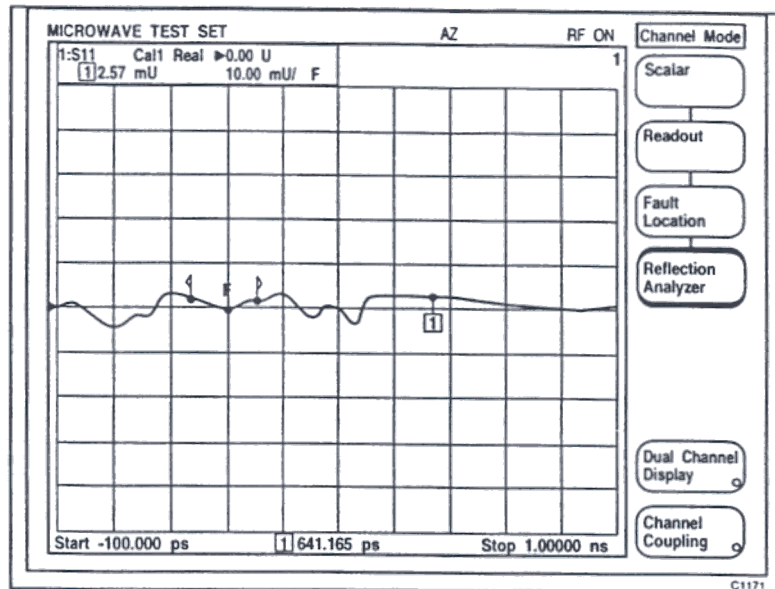


Figure 4 - Low-pass step response of DUT with fence applied

The capacitive discontinuity between the fence markers has now been matched out. It is now possible to quantify the performance improvement that could be achieved if the major impedance discontinuity within the DUT was improved. Figure 5 shows the frequency response of the fenced measurement.

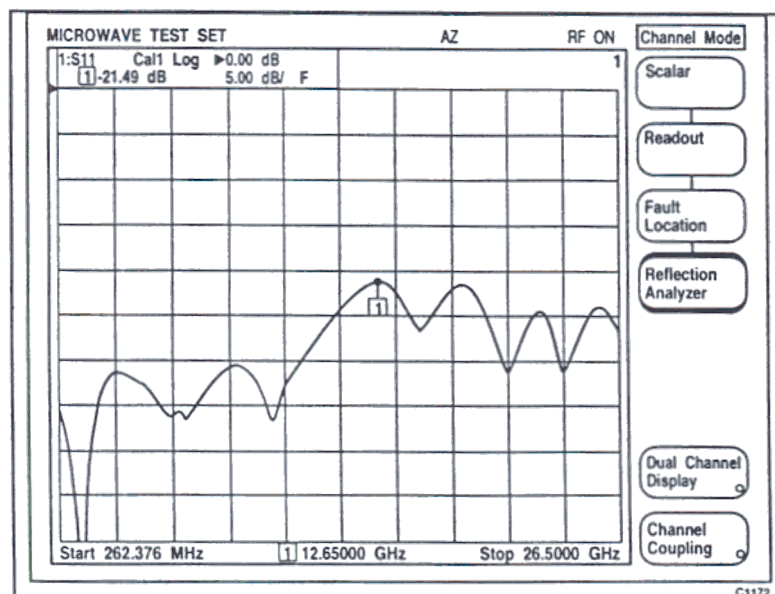


Figure 5 - Return Loss of DUT with fence applied

Conclusions

Time domain processing is a useful feature built into most Reflection Analyzers, to aid diagnosis of problems within the DUT. The precise location of each imperfection can be determined and the inductive or capacitive nature established. Gating and fencing are used to predict the resultant frequency response if the hardware performance were to be improved.

ANALyse

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Number 6

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"A Comedy of Errors!"
or
That's not what I mean by 'mean'."

Author: J P Ide, NPL, Malvern

Introduction.

The processing of measurement data can be one of the most significant parts of the measurement process and yet its importance is frequently over-looked. One of the most common mistakes is to automate a relatively simple process, test it under a range of fairly benign operational conditions and then incorporate it into the procedure and forget it. A little time later the circumstances under which the data is gathered change, disaster strikes and because the process has been automated the disaster is not recognised. An area that is particularly liable to this type of error is the apparently trivial task of finding the mean of data which are complex numbers, eg ANA data. This *Analyse* note looks at a simple example of the things that can go wrong.

The Problem

Much of the data utilised in RF and microwave measurements is complex, ie each piece of data is characterised by two values, eg magnitude and phase or real and imaginary components. These two descriptions are equally valid ways of locating the position of the required point on the measurement plane. Although these two descriptions are equivalent and therefore interchangeable under most circumstances, the average human brain seems happier working with magnitude and phase. This preference can lead to dramatic errors because performing mathematical operations on the data can have different results depending on the description used. There is an added complication in that phase information can be given in either radians or degrees($^{\circ}$). By definition there are 360° in a complete circle but convention allows an angle to be specified either between 0° and 360° or $+180^{\circ}$ and -180° .

The example is based on actual data gathered while performing measurements on a low reflection test artefact where the phase measurements are distributed around the 0° region. It has been selected because it shows very clearly the errors that can arise from finding the mean values while using an inappropriate description.

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Data Point	Magnitude	Phase 0° to 360°	Phase -180° to 180°	Real Component	Imaginary Component
1	0.0046	18.87°	18.87°	0.0043	0.0015
2	0.0036	331.44°	-28.56°	0.0032	-0.0017
3	0.0029	340.12°	-19.88°	0.0027	-0.0010
Mean Value	0.0037	230.14°	-9.86	0.0034	-0.0004

Table 1: Three data points with phase angles around 0° degrees

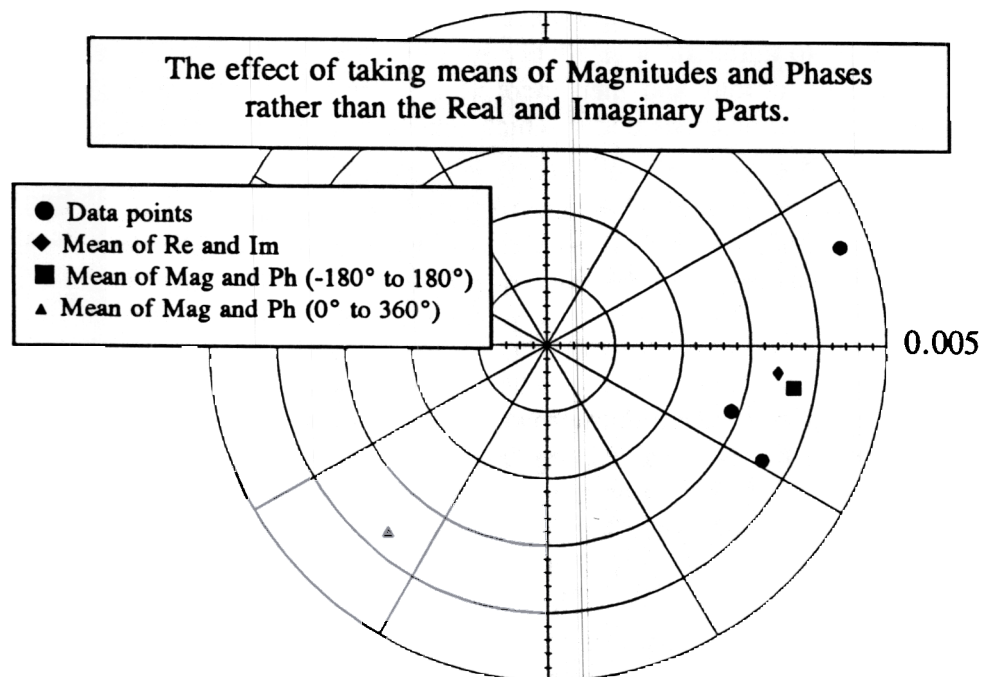


Figure showing the positions of the data points and of the variously calculated means

Recommendations

Perform any mathematical operations that involve complex numbers in Real and Imaginary format. The diagram shows that the errors produced by calculating the mean values using Magnitude and Phase can range from small to enormous but there will always be some error. This is another example where a computer is no substitute for thought.

PS. Working in radians rather than degrees does not necessarily solve the problem.

ANALyse

An occasional series of Notes produced by, and for, the Members of ANAMET

Number 7

January 1995

Making a good estimate of a measurand

Author: Nick Ridler, NPL, Malvern

Introduction

One of the fundamental tasks in measurement is obtaining an estimate of a value for a quantity being measured (the measurand). This might be for reporting the result of a measurement or for choosing a value to compare with other values, as is often the case when summarising data from an inter-laboratory measurement exercise. In both cases, the estimate is usually based on a series of repeated determinations of the measurand.

Deriving a good estimate for the value from a set of repeated measurements may, at first glance, seem to be a trivial task. A first reaction is often to calculate (using pen and paper, pocket calculator, or computer) the arithmetic mean for the set of values and use this as the best estimate of the measurand. However, the mean can sometimes be of limited use, especially if the results being analysed exhibit erratic variation, *e.g.*, when one or two results are radically different from the majority.

This situation sometimes occurs when analysing what should be similar data, from a variety of sources (inter-laboratory comparisons). It can also occur in repeat measurements made on a single system when one of the measurements is affected by an undetected gross error, such as connecting a short-circuit instead of an open-circuit during an ANA calibration.

This note examines the behavior of the mean under two different circumstances; firstly, when the results to be analysed are relatively well-behaved, and secondly when these results are 'contaminated' by an additional result having a very different value. It also looks at another statistic - the median - and compares its behavior under the same conditions. The data used in this Note was obtained during the second ANAMET measurement comparison exercise.

The mean

By far the most common estimator for the value of a measurand is the "arithmetic mean", often just called the "mean". It is calculated by simply summing the values of all the results, then dividing this sum by the number of results made. To illustrate this, we will look at some *S*-parameter phase measurements made during the ANAMET-941 comparison exercise.

Measurements of S_{11} of the 3 dB attenuator at 1 GHz, made by seven participants, were as follows; +118°, +121°, +115°, +109°, +113°, +116° and +114°. The mean value, denoted by $\bar{\phi}$, is given by;

$$\bar{\phi} = \frac{118 + 121 + 115 + 109 + 113 + 116 + 114}{7} = 115.1^\circ$$

Under normal circumstances, the mean works well as an estimator for this purpose. However, if we include a measurement made by another of the participants (who measured the phase to be -58°) and re-calculate the mean, we get;

$$\bar{\phi} = \frac{118 + 121 + 115 + 109 + 113 + 116 + 114 - 58}{8} = 93.5^\circ$$

Including this extra measurement has caused the mean to change dramatically. This indicates that the mean is struggling, under these circumstances, to make a reliable estimate of the measurand. The reason for this is that the mean is being used in, what can only be described as, unusual circumstances. In this case, what is unusual about the results is the presence of one value which is *very* different from the other values.

The median

Another estimator, for a measurand from a series of results, is the median. This is simply the middle value (in order of size) of the results. If the number of results is an even number, a unique middle value doesn't exist, so the median is the midpoint of the middle pair of values. We will use the same results above to illustrate finding the median.

For our first seven results, if we arrange them in ascending order of size, we get:

+109° +113° +114° +115° +116° +118° +121°

The middle value, or median, is +115°

For our eight results, arranged in order of size, we get;

.58° +109° +113° +114° +115° +116° +118° +121°

The middle two values are +114° and +115°, so the median is now $(114 + 115)/2 = 114.5^\circ$

The median has remained relatively unaffected by the inclusion of the very different result. We can conclude, therefore, that the median is more likely to make a more reliable estimate of a measurand when a relatively small series of results, containing one or two contaminated values, is being analysed.

Conclusions

Statistics play an important role in the analysis of results of measurements. These measurements might be made by one operator performing multiple measurements (*e.g.*, during measurement uncertainty assessments), or by many operators (*e.g.*, during inter-laboratory measurement comparisons).

It is important to choose an appropriate statistic when analysing data. The choice depends on the nature of the data (*e.g.*, so called "normal", or "abnormal", data) and the information required from the analysis (*e.g.*, an estimate of a measurand).

Ultimately, the correct use of statistics can enhance the understanding of most types of measurement data.

ANALyse

An occasional series of Notes produced by, and for, the Members of ANAMET

Number 8

February 1995

Assessing the quality of an estimated value

Author: Nick Ridler, NPL, Malvern

Introduction

It is often helpful, when reporting an estimated value, to give some indication of the reliability of the estimate so that those using the estimate can assess its usefulness. Producing a figure for the reliability of an estimate often involves interpreting the dispersion of a series of scattered values. The scatter may have been produced by a single measurement system (*e.g.*, from multiple, repeated measurements) or from multiple systems (*e.g.*, from inter-laboratory test studies).

In ANALyse 7, two methods of estimating a value were investigated; the arithmetic mean and the median. This Note, which is a companion to ANALyse 7, discusses the choice of method for quantifying dispersion. Two methods are examined; the standard deviation and the inter-quartile range. The standard deviation is calculated from the estimated mean value, while the inter-quartile range is associated with the median estimate.

The data used to examine the properties of these two methods is the same as that used in ANALyse 7 and was obtained during the second ANAMET measurement comparison exercise.

Standard deviation

The calculation for the standard deviation, $s(x)$, is given by the following expression;

$$s(x) = \sqrt{\frac{1}{n-1} \sum_{i=1}^n (x_i - \bar{x})^2}$$

where x_i represents the i th individual result (of which there are a total number n), and $\bar{x}$ the estimated mean value.

To demonstrate the use of this equation, we shall calculate the standard deviation for seven S_{11} phase measurement results for the 3 dB attenuator at 1 GHz obtained during the second ANAMET measurement comparison exercise. The results were as follows; +118.2°, +120.6°, +115.2°, +108.9°, +112.9°, +116.3° and +114.0°.

First, calculate the mean;

$$\bar{\phi} = \frac{118.2 + 120.6 + 115.2 + 108.9 + 112.9 + 116.3 + 114.0}{7} = 115.157 \approx 115.2^\circ$$

then the sum of squared deviations;

$$\begin{aligned} \sum_{i=1}^n (\phi_i - \bar{\phi})^2 &= (118.2 - 115.2)^2 + (120.6 - 115.2)^2 + (115.2 - 115.2)^2 \\ &\quad + (108.9 - 115.2)^2 + (112.9 - 115.2)^2 + (116.3 - 115.2)^2 + (114.0 - 115.2)^2 \\ &= 85.79 \end{aligned}$$

and finally the standard deviation;

$$s(\phi) = \sqrt{\frac{85.79}{6}} = 3.781 \approx 3.8^\circ$$

Under normal circumstances, the standard deviation gives a useful measure of the dispersion of the results about the mean.

However, if we include the measurement made by an eighth participant (who measured the phase to be -58.3°) and re-calculate the standard deviation, we get the following.

First, calculate the new mean value;

$$\bar{\phi} = \frac{118.2 + 120.6 + 115.2 + 108.9 + 112.9 + 116.3 + 114.0 - 58.3}{8} = 93.475 \approx 93.5^\circ$$

then the modified sum of the squared deviations;

$$\begin{aligned} \sum_{i=1}^n (\phi_i - \bar{\phi})^2 &= (118.2 - 93.5)^2 + (120.6 - 93.5)^2 + (115.2 - 93.5)^2 + (108.9 - 93.5)^2 \\ &\quad + (112.9 - 93.5)^2 + (116.3 - 93.5)^2 + (114.0 - 93.5)^2 + (-58.3 - 93.5)^2 \\ &= 26\,412.24 \end{aligned}$$

and finally a new value for the standard deviation;

$$s(\phi) = \sqrt{\frac{26\,412.24}{7}} = 61.426 \approx 61.4^\circ$$

Including this extra result has caused a dramatic increase in the standard deviation (from 3.8° to 61.4°). This indicates that the standard deviation is giving a very erratic estimate for the dispersion of the results. The erratic behaviour has been caused by the inclusion of a result which is far removed from the other seven results. Such behaviour makes the standard deviation of limited use under these circumstances.

Inter-quartile range

Another method of quantifying dispersion is given by the inter-quartile range, which is the difference between the values for the upper and lower quartiles. Quartiles (like the median, discussed in

ANALyse 7) are evaluated by arranging the data to be analysed in ascending numerical order of size. The lower quartile divides the data so that 25% of the observations are below its value, the median has 50% below and 50% above, and the upper quartile has 25% above its value.

However, the ordered data cannot always be split exactly into four equal parts. This calls for more precise definitions for these terms, which are given in terms of a calculated 'depth', *i.e.*, how far to go through the data to find the required value. For n observations, the depth for the median and quartiles are as follows¹;

Median:
$$Depth = \frac{n + 1}{2}$$

Quartiles:
$$Depth = \frac{Depth\ of\ median + 1}{2}$$

The depth for the lower quartile is found by entering the data from the low end (*i.e.*, starting with the lowest value) and the depth for the upper quartile is found by entering at the high end (starting with the highest value).

The inter-quartile range is then simply the difference between these values. We will use the data already analysed (in terms of standard deviations) to illustrate finding the inter-quartile range.

Firstly, we will use our seven results which, when organised in ascending order of size, are;

+108.9° +112.9° +114.0° +115.2° +116.3° +118.2° +120.6°

The depth for the median is $(7 + 1)/2 = 4$, so the depth for the quartiles is $(4 + 1)/2 = 2.5$, *i.e.*, the quartiles are midway between the second and third values (when entering the data from either end).

$$\text{Lower quartile} = (112.9 + 114.0)/2 = 113.45^\circ$$

$$\text{Upper quartile} = (116.3 + 118.2)/2 = 117.25^\circ$$

$$\text{Inter-quartile range} = 117.25 - 113.45 = 3.8^\circ$$

Now, if we repeat the calculations using all eight results, organised in ascending order of size, we have;

-58.3° +108.9° +112.9° +114.0° +115.2° +116.3° +118.2° +120.6°

The depth for the median is $(8 + 1)/2 = 4.5$, and the depth for the quartiles is $(4.5 + 1)/2 = 2.75$, *i.e.*, the quartiles are three-quarters of the way between the second and third values (when entering the data from either direction).

For the lower quartile, the second and third values are +108.9° and +112.9°, and three-quarters of the way between these two values (in an ascending direction) is +111.9°.

For the upper quartile, the second and third values from the top are +118.2° and +116.3°, and three-quarters of the way between these two values (in a descending direction) is +116.775°.

Some textbooks give slightly different expressions for finding the depth for quartiles. However, the one used here is the same as that in the software used to analyse the data in the second ANAMET measurement comparison exercise.

The inter-quartile range is therefore $116.775 - 111.9 = 4.875^\circ \approx 4.9^\circ$

The change in the inter-quartile range when a very different result has been included in the results being analysed is relatively small (it has increased from 3.8° to 4.9°). This indicates that the inter-quartile range estimate is less vulnerable to variation due to the inclusion of a result which is very different from the other, majority, values. Such resistant behaviour makes the inter-quartile range a useful estimator of dispersion when a relatively small number of observations, containing a contaminated value, are being analysed.

Discussion

The Table below summarises the calculations for the standard deviation and inter-quartile range for the phase measurement results given in this Note.

	Calculation based on seven results.	Calculation based on eight results (with one contaminated value).
Standard deviation	3.8	61.4
Inter-quartile range	3.8	4.9

It is worth remembering that the two intervals produced by the standard deviation and the inter-quartile range have different meanings. For measurement results from a normal (or Gaussian) distribution, the standard deviation defines an interval about the mean which will contain approximately 68% of the values. Whereas the inter-quartile range defines an interval, about the median, which contains half the total number of measurement results - one quarter on either side of the median. An important distinction between the two intervals is that, for the inter-quartile range, no assumptions need to be made about the nature of the underlying distribution (*e.g.*, normal, rectangular etc). Assuming that a set of results comes from a normal distribution can be difficult to justify when the sample of results is relatively small.

A second consideration in the use of such statistics is their ease of calculation. Most pocket calculators have a facility for calculating the standard deviation for a series of measurement results, so that the calculation can be made fairly easily. The inter-quartile range requires the measurement results to be sorted before calculations can begin. This usually results in a more lengthy calculation time. However, such calculations can often be made using software spreadsheet packages on computers.

Conclusions

Statistics can be used to gain insight into the reliability of an estimated value. However, it is important to choose an appropriate statistic for the particular set of measurement results being analysed. In particular, a set of results containing a value exhibiting a gross difference from the other values is particularly problematical. Under such circumstances, a more robust method than standard deviation is desirable. There are many such methods of estimating the reliability of a value which fall into this category, but perhaps the inter-quartile range is one of the easier to compute and interpret.

ANALyse

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Number 9

March 1995

The Uncertainty of a Single Measurement !

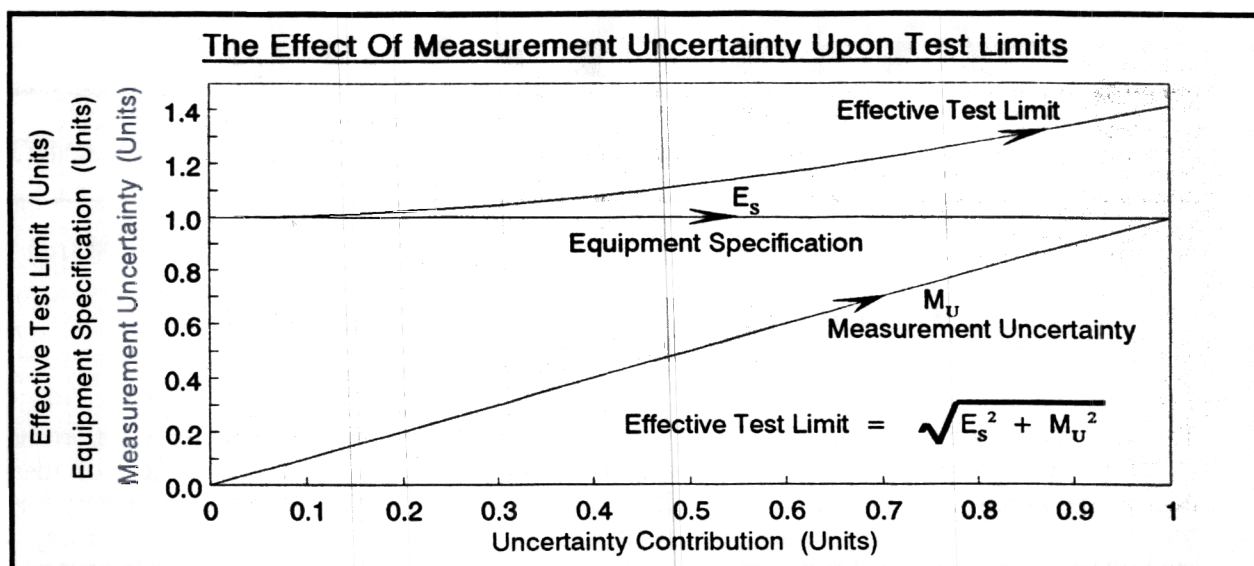
Author: Ian Instone, Hewlett-Packard Limited

The latest definitions of traceability include references to measurement uncertainty. Unfortunately, all of the recognised methods for assessing uncertainties assume that several sets of identical measurements will be performed for the sole purpose of calculating the measurement uncertainty. In practise most of the different assessment methods will arrive at a value within 10 % of any other method so the method used is not important, unless specified by the customer or other agency. The practise of performing many sets of identical measurements will increase calibration or testing costs and tie up valuable equipment during the course of these measurements. To address this issue a method is required which does not require vast quantities of measurements to be performed for the sole purpose of assessing the measurement uncertainties. Throughout this note uncertainties are referred to by their more descriptive title, random uncertainties and systematic uncertainties. Random uncertainties (U_R) are now known as Type A uncertainties, and systematic uncertainties (U_S) are known as Type B uncertainties.

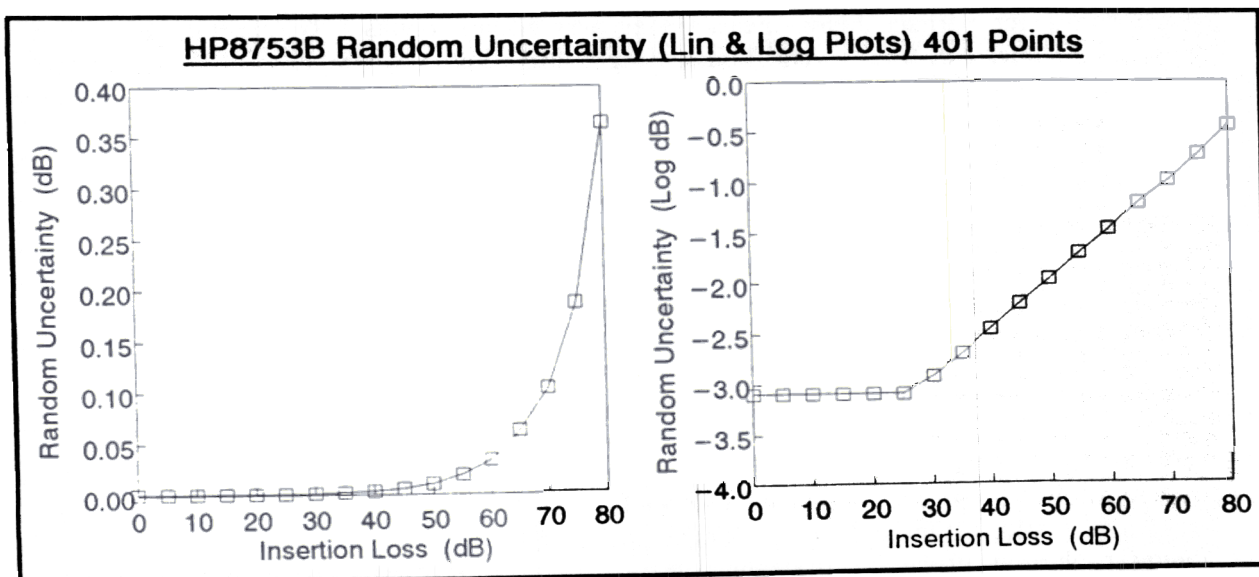
In an environment where cost plays such an important role as in calibration it is necessary to ensure that the quality and quantity of measurements performed precisely match the customers requirements. Too low a quality level will have an impact upon integrity and too high a quality level will have an adverse affect upon costs. In assessing measurement uncertainties guidance can be sought from a variety of sources, most measurement accreditation bodies publish suitable documents. Unfortunately however, these documents assume that a very few measured values will be reported upon a very simple artifact, for instance one document provides an example where only two measured values appear on the calibration certificate. In this case it may be practical to follow the guidance and report the average of many sets of measurements. In the case of complex micro-processor controlled measuring instruments, such as a network analyser, customers expect a calibration report which shows the results of measurements performed on over a dozen parameters, each parameter perhaps being measured at over fifteen points. To generate this quantity of measured values can take several hours using automation, and several days if measurements are performed manually. It is not practical to expect calibration laboratories to perform several sets of measurements on these instruments solely for the purpose of calculating an often insignificant component of the uncertainty budget. This note proposes a simplified method by which measurement uncertainties can be assessed and the safeguards which can be built into the process to monitor the quality of the assessment.

It is important that a similar understanding of "calibration" exists between both the supplier and consumer of the service. At the standards laboratory level it is normal to supply a certificate which can be used as "correction" data for the equipment. With complex multi-parameter instruments it is not possible to supply enough calibration data to enable reliable corrections to be made so certificates often show the equipments calibration status along with supporting measurement data. Much of the assessment of measurement uncertainties is possible without ever having seen the instrument requiring calibration. Systematic uncertainties can be fully assessed using the paper specifications of both the measuring instruments and the unit being calibrated. This can result in a very conservative estimate of measurement uncertainty, which is however valid for all of the items being calibrated providing that they meet their specification. In addition test equipment substitutions are possible and will be accommodated within the uncertainty budget assuming that the same model is used.

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The magnitude of the systematic uncertainty (which can be calculated using any of the recognised methods) will be used as a means of determining the amount of random uncertainty assessment required. So that uncertainties can be "ignored" when determining specification compliance they should be less than one third of the equipment's specification. In the graph it can be seen that at these levels the measurement uncertainty has negligible effect when determining specification compliance providing that both the specification and the uncertainty are at the same confidence level. Similarly if it can be shown that the random uncertainty is less than one third of the systematic uncertainty then it can be assumed to be negligible. If however the random uncertainty is shown to be larger it will have to be assessed using one of the traditional methods. A quick assessment of random uncertainty can be made by repeating all measurements twice in fairly rapid succession. If the first set of measurements were at the positive extreme of random uncertainty allowance and the second set at the negative extreme then the difference between them would be twice the worst case random uncertainty. If we assume that the "worst case" is represented by 3 standard deviations then the difference between the highest value measured at any point and the lowest value will be 6 standard deviations. Since measurement uncertainties are usually quoted at 95 % confidence level (approx 2 standard deviations) then using the maximum and minimum values as the random uncertainty should produce a considerable over-estimation, assuming that the (max - min) value is the true maximum and minimum.



The example above shows that the random uncertainty follows a law and can therefore be predicted. There is some noise present at the input of the measuring system which gives rise to the horizontal plot. There is also some noise generated at the output stages which is demonstrated by the positive going line. The plots above were generated using "STATS" on an HP8753B with a measurement sample of 401 values at each point.

Or, using methods above, the random uncertainty could be estimated from only two sets of measurements using:

$$U_R = \text{Maximum}(\text{Measurement}_1 - \text{Measurement}_2)$$

Where Measurement_1 and Measurement_2 are taken from two sets of measurements performed at the same points on the same test sample using the same measuring equipment. The maximum "difference" is used to produce a conservative estimate, however if a large percentage of the differences are approximately the same value they may be a more accurate representation of the random uncertainty.

The methods described here are based upon "type testing" the first of each instrument type to be calibrated. To ensure that this method continues to produce accurate values it will be necessary to periodically audit the process by checking the random uncertainty contribution using a different test sample. Ideally, unless the procedure identifies specific items of test equipment (by serial number) then substitute equipment of the same model & type should also be used for the audit.

Total Uncertainty

The uncertainty contributions are combined in the usual way, e.g.

$$U_{\text{TOTAL}} = \sqrt{U_{S1}^2 + U_{S2}^2 + U_{S\ldots}^2 + U_R^2}$$

The systematic uncertainty contributions are usually obtained from the various measuring instrument specifications, and other uncertainties imported from other calibrations, where appropriate.

Although the specifications of the various measuring instruments might be quoted at a variety of confidence levels it is reasonable to assume that they will be for at least 95 %. To use a specification which is quoted at 99.7 % confidence level (3 standard deviations) and assume it to be at 95.5 % (2 standard deviations) will produce a very conservative measurement uncertainty and the decision to convert specifications from one confidence level to another will depend upon the accuracy ratios involved. In most cases it will be acceptable to leave the confidence levels as they are and gain additional assurance from the rather pessimistic values produced. This approach to uncertainty assessment can produce values close to those obtained from more traditional methods, but in most cases at a fraction of the cost.

Step by Step Guide:

- 1 Assess and calculate the systematic (Type B) uncertainties using traditional methods
- 2 Divide the systematic uncertainty by 3, this value is the largest allowance that can be made for random (Type B) uncertainty before it becomes "significant"
- 3 Run the entire calibration process through twice and subtract the first set of measured values from the second.
- 4 Ignoring the signs, if the values from step 3 are not "significant" no further assessment of random uncertainty is required, therefore assume U_R to be $U/3$.
- 5 If random uncertainties are significant perform more sets of measurements and calculate the random uncertainty allowance for one measurement using:

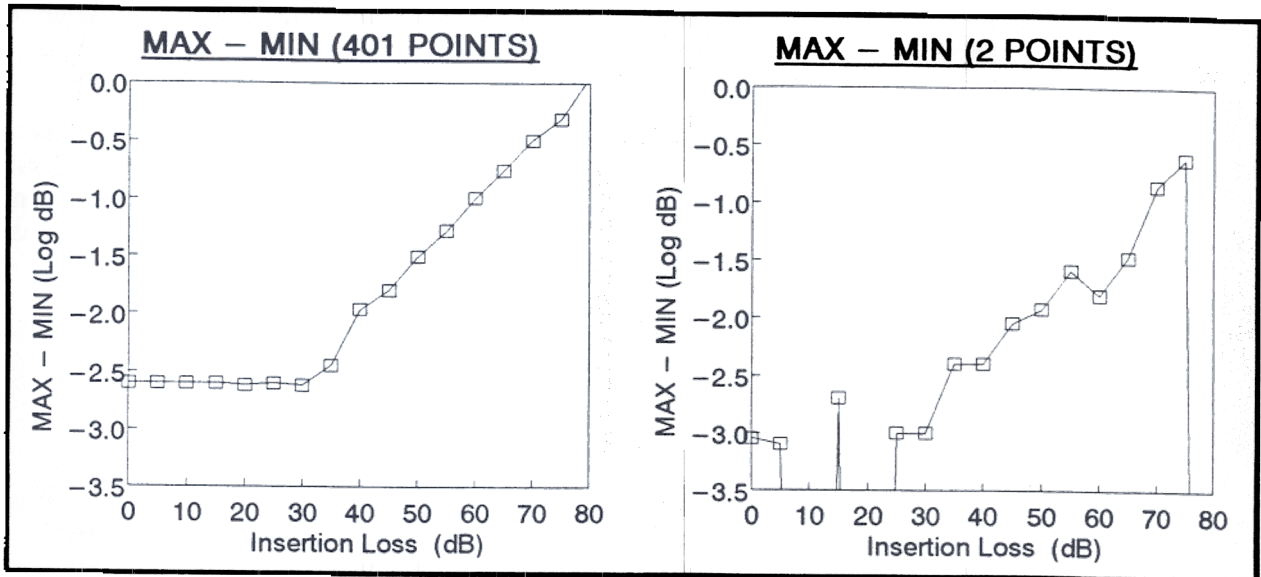
$$U_R = \frac{t \times \text{standard deviation}}{\sqrt{1}}$$

where t is chosen for the quantity of measurements performed on the "test" sample.

- 6 Calculate the total uncertainty using:

$$U_{\text{TOTAL}} = \sqrt{U_{S1}^2 + U_{S2}^2 + U_{S\ldots}^2 + U_R^2}$$

The plots above show the value of determining a "law" by which the random uncertainty varies with another, connected influence. Having established this on one unit it is feasible to reduce the random uncertainty assessment points by several orders, perhaps only assessing U_R at 20 dB intervals. This is of course still a very costly exercise so for "run-of-the-mill" specification compliance type calibrations faster methods based upon type testing a sample instrument and applying the values obtained to all of the calibrations performed is proposed.



The plots above show that the range of random uncertainty values obtained is less uniform with fewer measurement points, however they also demonstrate that we may still be able to predict a random uncertainty contribution based upon only two sets of measurements providing that some knowledge of the measuring instruments is used. Some caution may be required if this type of assessment is applied to too few points across the range.

The Uncertainty of a Single Measurement

The plots above have demonstrated that in many cases we can reduce the quantity of measurements required to assess the random uncertainties to two. However, even with two sets of measurements performed on complex instruments there could be an unacceptable cost burden. The information above needs to be translated into an uncertainty budget based on only one set of measurements being performed on the instrument submitted for calibration. If an infinite quantity of measurements were performed at every point the limits of random uncertainty would be described by the (max - min) value. The limits are often assumed to be at the (mean $\pm$ 3 standard deviations) points, therefore:

$$(\text{max} - \text{min}) = (\pm 3 \text{ standard deviations}) = (6 \text{ standard deviations})$$

For a 95 % confidence level with an infinite quantity of measurements performed at every point we calculate the (2 x standard deviation) points. In using the (max-min) method of calculating random uncertainty we have a very pessimistic assessment which needs to be balanced against the limited quantity of measurements used. Normally, when reporting the mean of several measurements the random uncertainty would be divided by the square root of the quantity so when only one measurement is performed the random uncertainty (U_R) will remain the same, e.g.

$$\text{For the mean of many measurements } U_R = \frac{t \times \text{standard deviation}}{\sqrt{N}}$$

Where "t" is Student's "t" for the quantity (N) of measurements performed

$$\text{Or, for only one measurement, i.e. } N = 1, U_R = \frac{t \times \text{standard deviation}}{\sqrt{1}}$$

Where the standard deviation is calculated from a typical measurement set and where N has become the square root of 1, because only one measurement is performed on the unit under test. Student's "t" is chosen according to the quantity (N) of measurement sets performed on the evaluation sample.

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Services Electrical Standards Centre ANAMET-941 Observations

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Introduction

The ANAMET-941 measurement comparison exercise involved measurements of S_{11} and S_{21} , magnitude and phase, of four coaxial line attenuators fitted with GPC-3.5 connectors from 1 GHz to 26.5 GHz in 1.5 GHz steps. Attenuator values were nominally 3 dB, 6 dB, 10 dB and 70 dB and eight ANAMET participants contributed results data. The Services Electrical Standards Centre (SESC) has measured the S_{21} magnitude parameters for each attenuator using an automated WBCO (Waveguide Beyond Cut-Off) IF Substitution Attenuation measurement system.

This Note compares the median of the ANAMET-941 results with the SESC WBCO results and also briefly discusses the techniques used for the WBCO measurements.

S_{21} Magnitude Observations

Attenuation measurements were performed using an automated WBCO IF Substitution System with "gauge blocking" employed for the 70 dB attenuator. Matching for a 50 ohm system was achieved using isolators with tuners and selected low-value VSWR attenuators. The total measurement uncertainty, at a confidence level of not less than 95%, for the system includes the random uncertainty calculated from the results of at least ten connector repeatability measurements at each measurement frequency and the uncertainty due to mismatch.

Figures 1, 2, 3 and 4 show the median values for the ANAMET-941 results along with the WBCO system results, with uncertainties, for the 3 dB, 6 dB, 10 dB and 70 dB attenuators, respectively.

The Table below gives the largest difference between the ANAMET median values and the WBCO system results for each attenuator. The Table also gives the largest inter-quartile range (IQR) for the ANAMET values over the full frequency range (1 GHz to 26.5 GHz). WBCO system results have total measurement uncertainties ranging from ± 0.003 dB to ± 0.015 dB at 1 GHz and ± 0.015 dB to ± 0.036 dB at 26.5 GHz, depending on the value of attenuation being measured.

Nominal value of attenuation (dB)	Largest difference between ANAMET median and WBCO values (dB)	Largest value for the inter-quartile range for the ANAMET values (dB)
3	0.024	0.031
6	0.024	0.035
10	0.021	0.027
70	0.336	1.051

Conclusions

ANAMET median values for the 3 dB and 6 dB attenuators show good agreement with the WBCO system results (see Figures 1 and 2). Measurement uncertainties, for the WBCO system, over the entire frequency range compare well with the IQR values, given in the Table. (The IQR represents the spread achieved by the four results closest to the ANAMET median value.) At 10 dB, a deviation (comparable with the IQR) between the ANAMET-941 median and the WBCO system results starts to occur (see Figure 3). At 70 dB, the largest value for the IQR has increased dramatically. An examination of the spread produced by all the ANAMET-941 results, given in the final report for the exercise, shows this spread to be more than 30 times the WBCO system total measurement uncertainty (as shown in Figure 4).

The above results demonstrate the importance of low uncertainty, NAMAS accredited, verification kit measurements obtainable from a precision attenuation measuring system.

These results may initiate discussion on the need for an attenuation measurement exercise that includes values between 10 dB and 70 dB in this increasingly important connector type and frequency range.

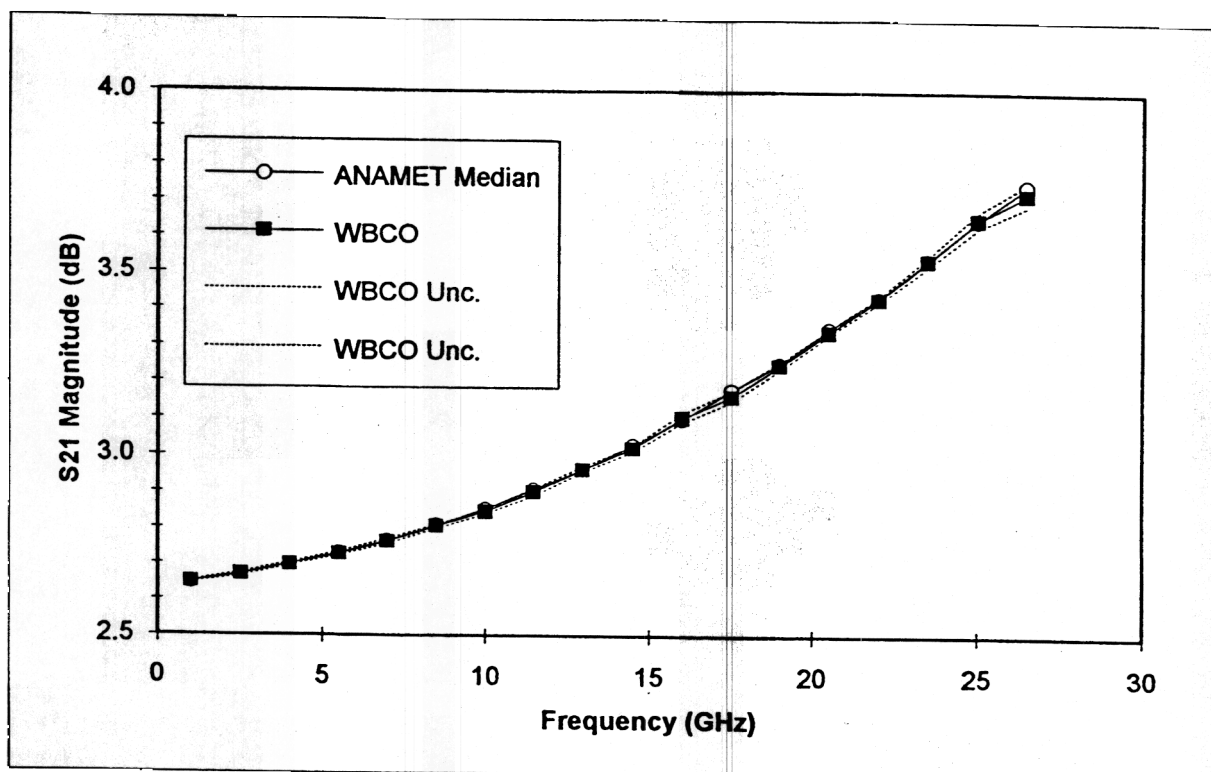


Figure 1: S_{21} magnitude results for the 3 dB attenuator

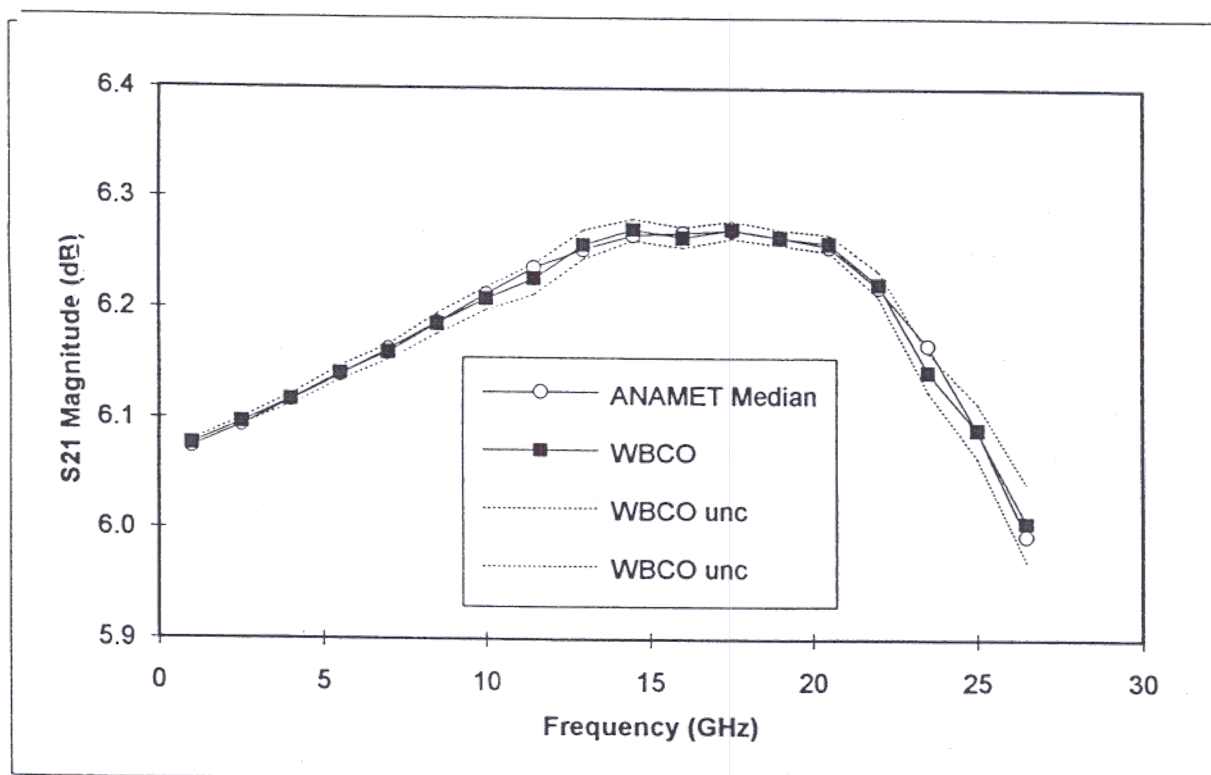


Figure 2: S_{21} magnitude results for the 6 dB attenuator

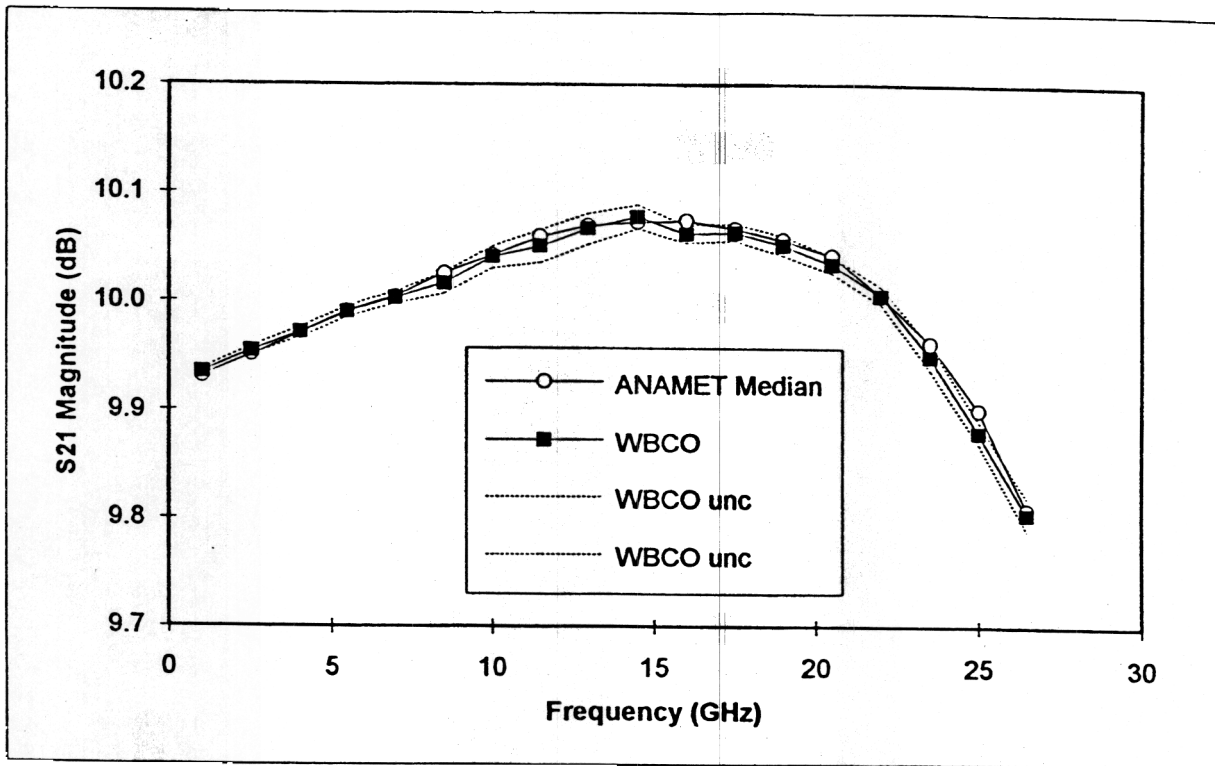


Figure 3: S_{21} magnitude results for the 10 dB attenuator

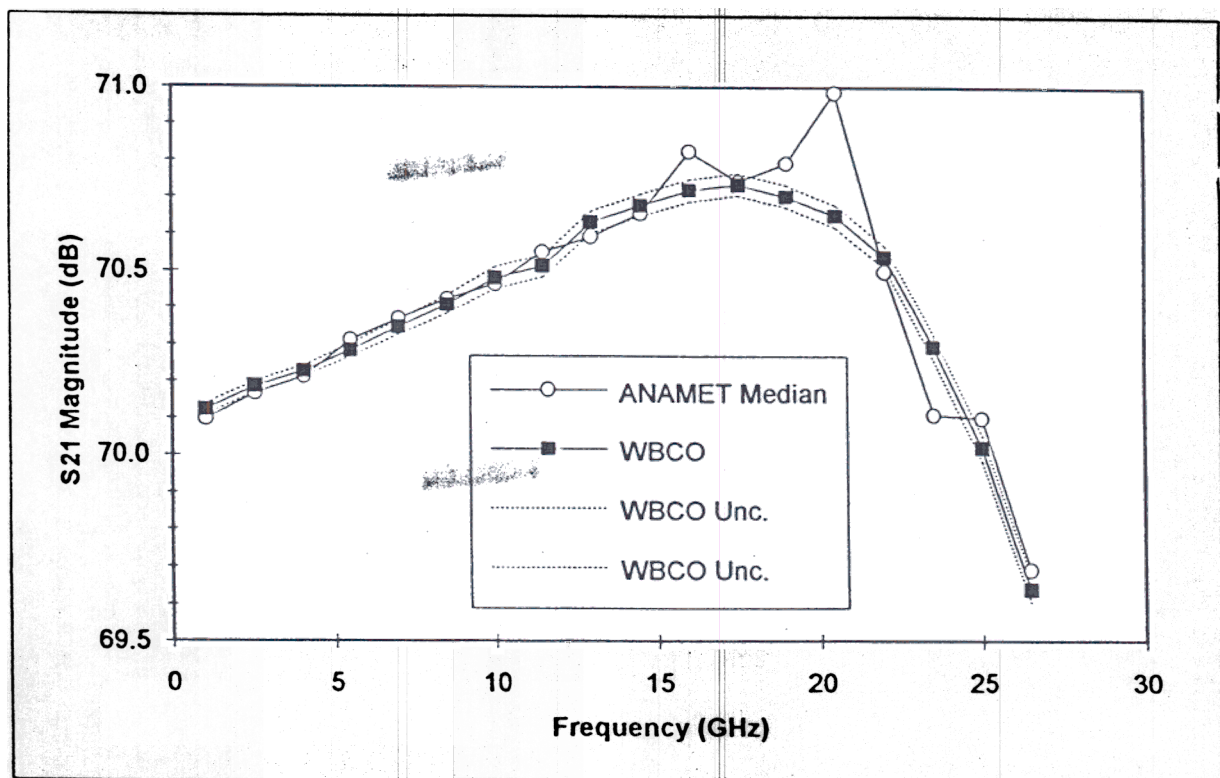


Figure 4: S_{21} magnitude results for the 70 dB attenuator