A Multistate Reflectometer in Dielectric Guide for the frequency range 75-140GHz

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Abstract - *In the paper an implementation of a four-port Multistate Reflectometer using dielectric guide for the frequency band 75–140GHz is presented along with all its necessary components. The paper focuses on the realisation of a phase shifter as this, the key component, does not till now have an equivalent for use with dielectric waveguides. The characteristics of the final instrument are then measured at intermediate frequencies and a calibration performed at 94GHz. Measurements were performed at the National Standards Division at R.S.R.E. in Malvern U.K., and subsequent results are in good agreement with theory.*

I. Introduction

The use of dielectric waveguides for high microwave frequencies in place of metallic rectangular waveguides has been suggested for many years. Besides the problem of manufacturing very small metallic rectangular waveguides for use above 90GHz and the increasing attenuation due to the skin effect and surface roughness there is also the problem of attenuation and repeatability in the connectors. This latter problem is particularly acute when measurements are made using standard impedances. Following the successful six-port reflectometer made in dielectric guide[1, 2] the next project at Kent involves a multistate reflectometer in dielectric guide. A multistate reflectometer has the advantage over a six-port reflectometer in that it requires two less power meters. In order to achieve this, the device also needs a means of changing phase. The present paper describes a novel broadband phase shifter which has been designed for use in the multistate reflectometer. The whole device will then enable the user to make very accurate measurements of phase and amplitude in the range 75–140GHz for not much more than the cost of a swept source.

II. General Reflectometer Theory

A reflectometer is an n-port, ($n \geq 4$), linear waveguide junction which can direct electromagnetic energy from a source at port 1 to the measurement port 2. If it is assumed only one mode is present at each port, though not necessarily the same one, then this makes it possible to represent the waves incident and reflected from the device by an appropriate scattering matrix. Consider the 4-port circuit with wave labelling shown in *Fig 1.* The incident and reflected waves are labelled a_i and b_i respectively with $i = 1..4$ for the four accessible ports. In this circuit the incident wave a_i is detected by the power detector P_3, while P_4 receives a sample of the reflected wave from the D.U.T., plus another from the phase shifter.

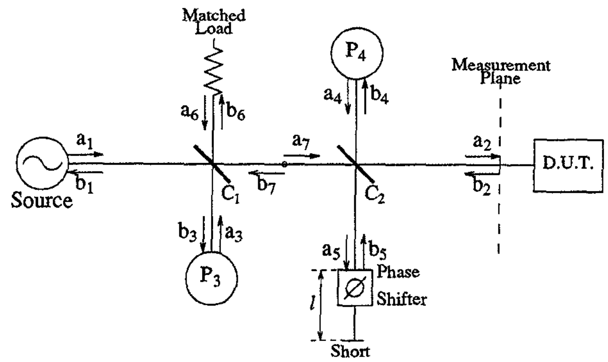

Figure 1

Schematic of a Multistate Reflectometer.

The scattering matrix for the above circuit can thus be written as $[b] = [S]_k[a]$ for each state k of the reflector. By assuming perfect isolation at one of the coupler ports of C_1 and C_2 the overall scattering matrix for the multistate can be written in terms of the reflection and transmission coefficients of the couplers as

$$[S] = \begin{bmatrix} -n_2^2 m_1^2 e^{-2j(\theta_l - \phi_k)} & m_1 m_2 & jn_1 & jn_2 m_1 m_2 e^{-2j(\theta_l - \phi_k)} \\ m_1 m_2 & 0 & 0 & jn_2 \\ jn_1 & 0 & 0 & 0 \\ jn_2 m_1 m_2 e^{-2j(\theta_l - \phi_k)} & jn_2 & 0 & m_2^2 e^{-2j(\theta_l - \phi_k)} \end{bmatrix}$$

for each state k of the reflector, where the coupling coefficient $n = Ne^{-j\theta}$, the transmission coefficient $m = Me^{-j\theta}$, and N and M are the amplitude coefficients with $N^2 + M^2 = 1$. The phase change produced by each reflector state is ϕ_k, and θ_l is the phase shift due to the physical length l of the phase shifting network. In general the pair of quadrature couplers C_1 and C_2 are dissimilar and thus the subscripts of the reflection and transmission coefficients correspond to the particular coupler. It can further be shown that the ratio of powers at the measurement and reference detectors is given by the bilinear relationship

$$\left(\frac{P_4}{P_3}\right)_k = \left|\frac{d_k\Gamma + e_k}{c_k\Gamma + 1}\right|^2 \qquad (1)$$

where c_k, d_k, and e_k are the calibration constants to be determined, and Γ is the reflection constant of the D.U.T.

III. Calibration

The simplicity of the hardware for the multistate is offset by the fairly complex procedure necessary for the calibration of this instrument. This is mainly as a result of less than ideal components. In general only three calibration standards will suffice as there are three complex unknowns c_k, d_k and e_k to be determined.

However the number of standards can be reduced, for example if the coupler C_1 was assumed to have perfect isolation from the reflected wave of the D.U.T., then the constant $c_k \to 0$ of (1), and the ratio P_4/P_3 becomes a linear function of Γ thus only needing two standards for calibration. Similarly if the same assumptions are placed on the coupler C_2 such that P_4 only receives a fraction of the reflected wave, then $e_k \to 0$ and

$$P_4/P_3 = |d_k\,\Gamma|^2$$

This is the basis of the tuned reflectometer[3] which is adjusted to meet the above conditions. In reality these conditions cannot be assumed and so the bilinear relationship (1) must be used for calibration.

It has been shown[4] that the number of calibration standards can be reduced if the arrangement in *Fig. 1* is reorientated such that the constant c_k of (1) becomes independent of the reflectometer state k. If the multistate is now reconfigured, as in *Fig. 2* below, then it is clear that the power detector P_3 becomes autonomous to the reflector movement and so (1) becomes,

$$\left(\frac{P_4}{P_3}\right)_k = \left|\frac{d_k\Gamma + e_k}{c\Gamma + 1}\right|^2$$

and hence only three loads are required as c is now only dependent on the scattering coefficients of the invariant junction 2, 3, 5. For a more rigorous analysis reference is made to the paper by Oldfield et.al.[4] which also includes a typical set of calibration constants for the multistate.

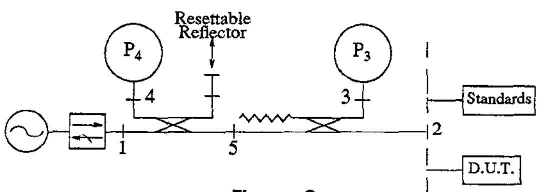

Figure 2

Diagram showing Multistate Reflectometer configuration for calibration in terms of three loads.

IV. Couplers

The use of proximity couplers for such an application is clearly unsuitable as the very nature of these frequency dependent evanescent fields make the coupler a narrow band device, and therefore unacceptable for use in the multistate. A more novel design introduced by Collier and Hjipieris[5] in 1985 is based upon a quasi-optical approach and results in a high isolation broad-band coupler. It employs the use of a dielectric film mismatch positioned at 45° to the direction of propagation to obtain any required power split between the through and coupled ports. The amount of coupling is determined by the width of the dielectric film and the relative effective dielectric permittivity. The design equations for determining these values are given in the mentioned paper. Commercial availability and ease of machining governed the choice of the coupling film. Stycast HiK was used for both couplers, and ground down to a thickness of 0.38mm. This acts as a 3.52dB coupler at the centre frequency, and results in only a 0.5dB variation at both ends of the operating band. The coupling films have a 10mm × 10mm cross sectional area to cater for the evanescent fields of the guides.

V. The Phase Shifter

Realization of this component with dielectrics has originated from work previously done on coupled guides[6], by employing the resonator action theorem of adiabatic invariance[7, 8] as applied to guides of circular cross-section. The basis of phase shifting arises from the fact that on a system of identical coupled guides the coupled mode will have a phase which differs from that of a single guide by the coupling constant c, depending on the separation of the guides. Clearly though moving a guide close to another to attempt phase shifting is unsuitable as energy coupled to the second guide will be lost as radiation from the ends of the guide. Similarly ways of supporting and moving the guide will also be lossy and difficult in practice. However the theory behind image guides provides us with a convenient means of exploiting this phase shifting phenomenon.

It is well known that an image guide dielectric of height b and width 2a will propagate modes that are similar to that of a free space unbounded dielectric with double the height, but not necessarily of similar mode designation. Hence moving the ground plane to infinity will have the effect of changing the phase of the mode from that of an image guide mode to one of an unbound dielectric mode. Thus the ground plane effectively produces an "image" of the dielectric and is equivalent to a coupling configuration when the guides are touching. Bracey et.al.[8] have confirmed this equivalence, and provided expressions for circular dielectric guides and dielectric slabs by formulating appropriate radiation pressure expressions at the shorting wall interface. A simpler formulation is provided here for phase shifting in a dielectric rod of rectangular cross sectional area. Since the model is based upon the Knox and Toulious[9] approximation, the relative accuracy of the equations presented depend solely on the accuracy of the aforementioned transcendental solutions.

Consider a vertically polarised E^y_{11} mode propagating in an unbounded dielectric of permittivity ε_r and width 2a. The three main propagating regions are labelled I, II, III, and corresponding field distributions are shown in *Fig. 3a*. Let a reference point x = 0 be at the centre of the dielectric. Now say an infinite electric shorting wall is brought close enough so as to perturb the evanescent fields in region III, then if we assume the fields at the shorting wall go to zero and further postulate that the maxima in region II of the dielectric is displaced by a small amount D in order to satisfy Maxwells equations for this perturbation, then solving the boundary conditions by matching the Electric and Magnetic fields give for the antisymmetric mode

$$\alpha = k_x \tan(k_x a - D) \tag{2a}$$

$$d = a + \frac{1}{\alpha}\tanh^{-1}\left[\frac{\tan(k_x a - D)}{\tan(k_x a + D)}\right] \tag{2b}$$

In the above equations α is the transverse decay constant, d is the distance to the metal wall, and

$$k_x^2 = (\varepsilon_r - 1)k_0^2 - \alpha^2 \tag{3}$$

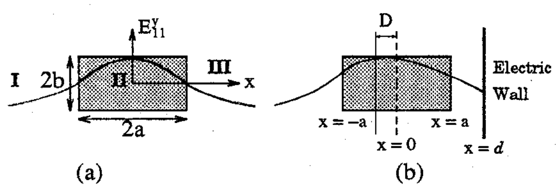

Figure 3

Field distribution of normal and perturbed E_{11}^Y mode.

The treatment for the symmetric mode of the same polarisation when the magnetic fields are parallel to the metal wall can be analysed by moving the metal wall in the y-plane onto the broad face of the dielectric rod. For the orthogonal polarisation the procedure is the same except that a and b must be interchanged. The equation pairs (2) must be solved along with (3) at each frequency to determine the change in phase for a given perturbation.

In *Fig. 4* a typical characteristic of a phase shifter is shown for a dielectric guide of dimensions 2.54mm × 1.27mm operating at 110GHz. Clearly a greater variation in D can be obtained for the symmetric mode and thus a larger degree of phase shift is obtainable with this configuration. However if a rod with unity aspect ratio, (a = b), were used the phase shift for both even and odd modes would be equal.

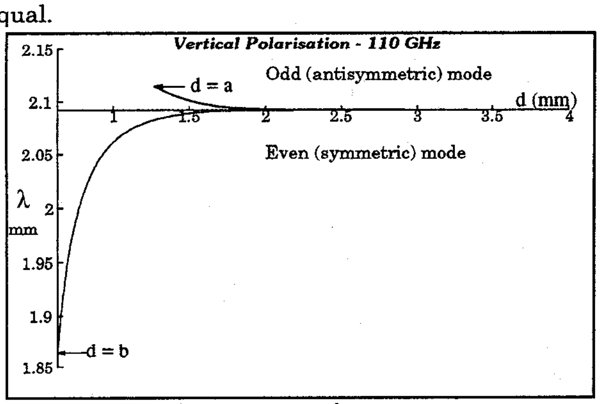

Figure 4

VI. Final Assembly

The correct choice of dielectric medium is by far the most significant factor in the implementation of this instrument. For the multistate operating in the frequency range 75–140GHz guide dimensions, flexibility, material loss, and surface finish all have to be taken into consideration.

In line with the already demonstrated capabilities in the six-port device[1], cross–linked polystyrene was used for the four-port as its low loss and excellent surface finish properties are ideal for high frequency measurements where small surface perturbations can lead to excessive loss. Guide dimensions were also chosen to match that of conventional metallic guide, Wg27 at W-band, as this could provide low loss transitions and extra stability at the points of transition. Mode conversion from the dielectric E_{11}^Y mode to the metallic guide TE_{01} mode was achieved with the use of standard 20dB horns, (flares), by inserting the dielectric into the guide aperture thus providing both support and low loss. Unger[10] has shown that this is the optimum arrangement to have in which the dielectric totally fills the throat of the horn with a gradual taper in both the E and H planes. For the short circuit the dielectric guide is butt-end shorted onto a slide with a gold deposit. Similarly the absorber consists of a gradually increasing tapered slide but this time positioned parallel to the direction of propagation. In reality the load is inserted into a slit within the dielectric guide so as to reduce its frequency dependence. Finally the enclosing box is lined with RAM‡ to absorb any stray radiated fields.

A diagram of the final device is shown below, in *Fig. 5*, with all internal dimensions given. Implementation is directly from the reconfigured diagram of *Fig. 2* for calibration in terms of three loads.

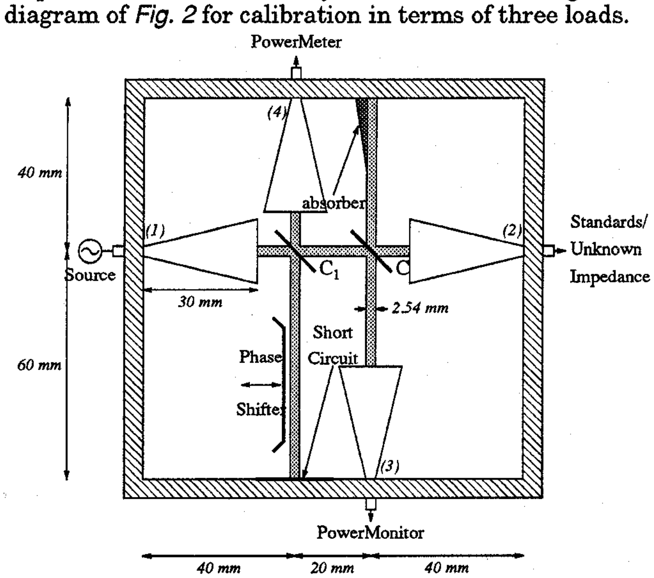

Figure 5

Dielectric Multistate Reflectometer

‡ Radiation Absorbing Material

VII. Practical Measurements and Calibration

Calibration and subsequent measurements of the multistate were carried out in a temperature controlled enviroment at R.S.R.E in Malvern, U.K. The source consisted of a frequency locked klystron to within ±15kHz, two analog power meters, a digital ammeter, and a power ratio meter which measured the ratios of the power from ports 3 and 4.

It can be shown that the bilinear realtionship (1) describes three circles often termed the "impedance locating circles" in the complex Γ plane. At a given frequency the circle centres are fixed positions determined by the scattering parameters of the circuit, while their radii are proportional to the square root of the power ratios. Thus it is clear that the intersection of these circles correspond to the reflection coefficient of the D.U.T. and subsequently that the centres of these impedance locating circles should have an even distribution around the origin in order to resolve any ambiguity. As the distribution of the circle centres is dictated by the phase shifter, an initial measure of the amount of phase shift attainable was undertaken. Up to 180° phase shift was obtained at the lower frequency for the 3cm length of metal used and were within a few percent of the predicted values.

At this stage a calibration was attempted at an intermediate frequency. Although the procedure is fully automated with data from the power meters acquired via A to D converters and stepper motors driving the phase shifter, the calibration was in this instance performed manually and values for various loads fed into a desktop computer. The loads consisted of waveguide standard short circuit and an imperfect load both measured through a series of precision spacers. Following the location of the Q-circle centres and a successful calibration the computer prompted for various test pieces to be measured. Offered up were a sliding short, two power heads, and a waveguide iris. A table of the results are shown in *Table 1* and a plot of phase for varying positions of the short is given in *Fig. 6*. It can be seen from the table that the measurements are in good agreement with the waveguide multistate, and in excellent agreement with the averaged values. Both multistate measurements were taken as a one-off and do not include the inverted measurements. They therefore do not account for the connector repeatability which is a particular problem at high frequencies.

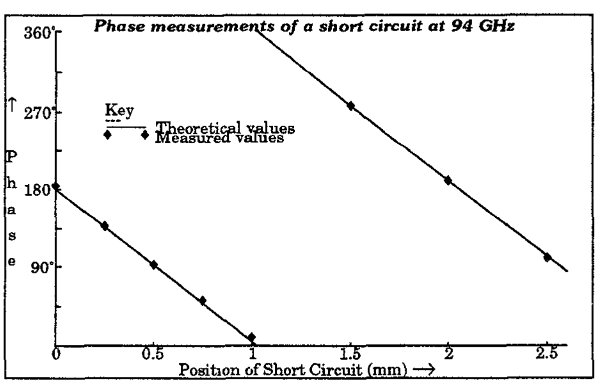

Figure 6

Test Piece	AVERAGE - Normal and Inverted Measurements		Measurements with Dielectric Multistate		Measurements with Waveguide Wg 27 Multistate	
	Mag	Phase	Mag	Phase	Mag	Phase
Iris	-	-	0.5536	299.95°	0.578	299.2°
Hughes Mount #109	0.1436	173°	0.1524	174.64°	0.138	170.2°
Hughes Mount #116	0.144	183°	0.1553	182.71°	0.139	187.6°

Table 1

Measurements on various Test pieces at 94GHz

VIII. Conclusion

Dielectric guides provide a low loss, low cost alternative to metallic waveguides. This paper describes the use of dielectric waveguides for general reflectometry and also how a phase shifter can be made for use in the multistate reflectometer to give a constant phase over an octave bandwidth. In a further design, dielectric standards, embedded power detection, and the possibility of making the phase shifting mechanism electronic via piezo-electronic transducers, have been proposed. This will make it possible to achieve broadband measurements at a far superior rate and accuracy to its metal counterparts and at a reduced cost.

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