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An Improved Circuit for Implementing the Six-Port Technique of Microwave Measurements

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***Abstract*—In a companion paper, circuit design criteria were developed which lead to optimal performance in applying the six-port technique to the measurement of microwave parameters. A circuit which approximately satisfies these new design goals is described. Together with its several variants, it promises to become, in many applications, the "preferred" six-port circuit.**

I. Introduction

THE RESPONSE of a six-port measuring system is contained in the reading of four power meters $P_3 \cdots P_6$, which in the general case may be written,

$$P_3 = |A|^2 |b|^2 |\Gamma_l - q_3|^2 \quad (1)$$

$$P_4 = |D|^2 |b|^2 |1 - \Gamma_l \Gamma_g|^2 \quad (2)$$

$$P_5 = |E|^2 |b|^2 |\Gamma_l - q_5|^2 \quad (3)$$

$$P_6 = |G|^2 |b|^2 |\Gamma_l - q_6|^2. \quad (4)$$

Here $|b|$, Γ_l represent, respectively, the emergent wave amplitude, and complex reflection coefficient at the output measurement plane. These quantities, if not the measurands of interest, play a major role in their determination. The remaining parameters are determined by the properties of the six port.

In a companion paper [1] the following design criteria were developed: $\Gamma_g = 0$, $|q_6| = |q_5| = |q_3| \equiv |q|$ while the arguments of these last three terms *differ* by $\pm 120°$. (In the immediate context, the absolute values of the arguments are of no interest.) The optimum choice of $|q|$ may be expected to lie in the range 0.5–1.5, while $|A|$, $|D|$, $|E|$, and $|G|$ are scale factors which determine the power levels at the respective power detectors. As a means of illustration, it is convenient to consider a specific problem. Let it be required to design a six-port circuit for calibrating bolometer mounts and measuring reflection coefficient, where the detectors $P_3 \cdots P_6$ are also of the bolometric type. Moreover, let it be further stipulated that the circuit is to be broad-band and, in order to reduce the power requirements at the input, inherently lossless. It is the purpose of this paper to describe a six-port circuit which approaches these design goals.

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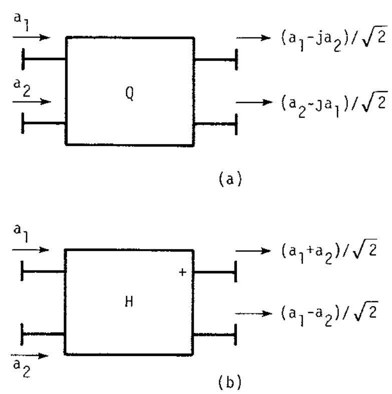

Fig. 1. The basic modules for a six-port circuit. (a) Quadrature hybrid. (b) 180° hybrid.

In today's art, the broadest frequency coverage is provided by stripline components. Here bandwidths of 10 : 1, or more, are not uncommon. The basic circuits thus available include quadrature hybrids, 180° hybrids, power dividers,[1] and directional couplers. In waveguide parlance, a quadrature hybrid is a 3-dB (4-port) directional coupler, while an 180° hybrid is an E–H tee. These two hybrids and the relationships which exist among the incident and emergent wave amplitudes are as shown in Fig. 1. Ideally, these devices are lossless and matched at all ports.

[1] For the purpose of this paper, a "power divider" is a 180° hybrid with one arm terminated internally.

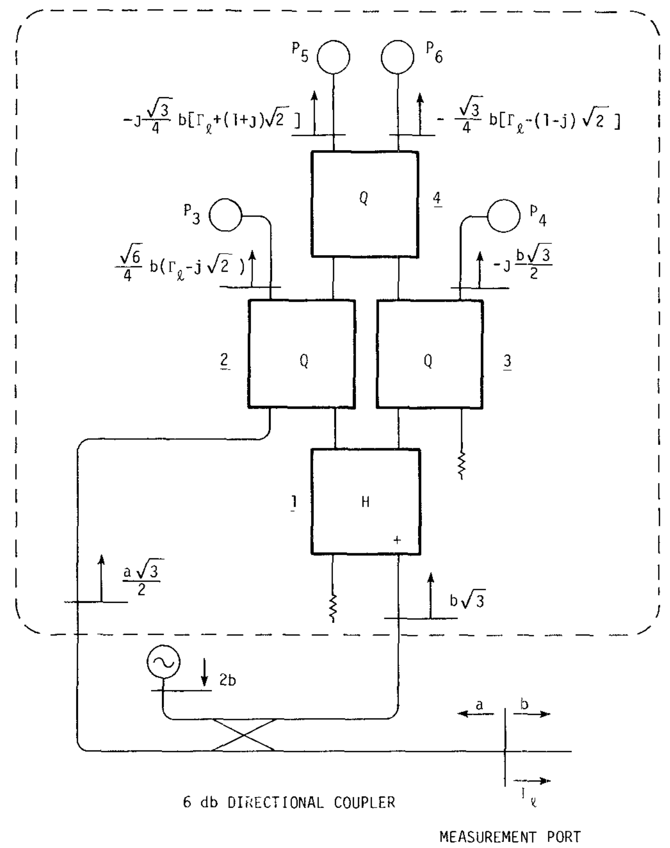

Fig. 2. A proposed six-port circuit.

In the existing art, a broad-band circuit which yields a 120° phase shift is unknown (to the author at least!). However, one is able to achieve a broad-band 90° phase shift by means of a quadrature hybrid. This suggests some compromises in the design goals outlined above.

II. Description of Proposed Six Port

The basic configuration of the proposed six-port circuit is shown in Fig. 2. Here ideal components have been assumed, and the emergent and incoming wave amplitudes at the measurement port are designated by b and a, respectively. The wave amplitudes at other selected positions in the circuit are also shown.[2]

Comparing this circuit with the design goals, one first notes that although resistive terminations are shown at two different locations, ideally none of the signal power reaches these, so the circuit is inherently lossless. In addition, $\Gamma_g = 0$ for this circuit, while the values of q_3, q_5, q_6 are as shown in Fig. 3. As compared with the 120° objective, the differences in the angles are 135°, 90°, 135°. Moreover, while $|q_5| = |q_6| = 2$, $|q_3| = |q_5|/\sqrt{2} = \sqrt{2}$. Although this result falls somewhat short of the design objectives, these goals are more nearly achieved by this circuit than by any other which

[2] These may be confirmed by comparison with Fig. 1. Note, however, that no attempt has been made to keep track of the phase of a or b or even their phase difference in an absolute sense. The only question of importance in this context is: how does the phase difference between a and b at port 3 compare with that which exists at ports 5 and 6?

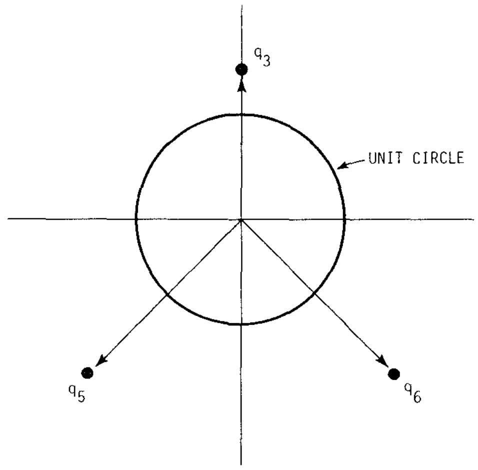

Fig. 3. Illustrating q_3, q_5, q_6 for the circuit of Fig. 2.

has been devised to date. Moreover, it appears that the theoretical loss in performance between this circuit and an "ideal" one may be small in comparison with the performance degradation which results from the use of nonideal components.

If one assumes 20 mW of power at the input, 5 mW or 1/4 of this reaches the measurement port. If the termination at the measurement port is a matched load, this power will be absorbed, while the remaining 3/4 is divided equally among the detectors $P_3 \cdots P_6$, resulting in a power level of 3.75 mW for each. If now P_4 is stabilized at this value, and a sliding short is connected to the measurement port, the value of P_3 will reach approximately 11 mW for certain short positions, while the maximums at P_5 and P_6 will be approximately 8.5 mW. The maximum dynamic range excursion at any detector is a nominal 15 dB which occurs at P_3.

The circuit thus provides a reasonably optimal set of values for $|A| \cdots |G|$, which in turn determine the distribution of the available signal power among the several detectors and measurement port.

As a variant to Fig. 2, one may replace the 6-dB directional coupler with one of 3 dB. If this is done, the power level at the measurement port is doubled, but at the expense of the power levels at $P_3 \cdots P_6$. In addition the $|q$'s$|$ are multiplied by $\sqrt{2}/2$. Because 3-dB couplers are more readily available than 6-dB couplers, this alternative may be preferred.

Apart from a constant multiplier, the "q" values are determined entirely by that portion of the circuit enclosed within the dotted lines in Fig. 2. Taken alone, this may be considered a "vector voltmeter" which has been the subject of an earlier paper [2]. As a vector voltmeter, this circuit has certain advantages for some applications. In particular, if it is stipulated that one of the four outputs be proportional to one of the input signals, this circuit requires fewer components. Moreover, as previously noted, the circuit is inherently lossless, and thus makes better use of the power input. Finally, the "q" spacing is 90°, 135°, 135°, which represents an improvement over the 90°, 90°, 180° as is usually the case in other circuits.

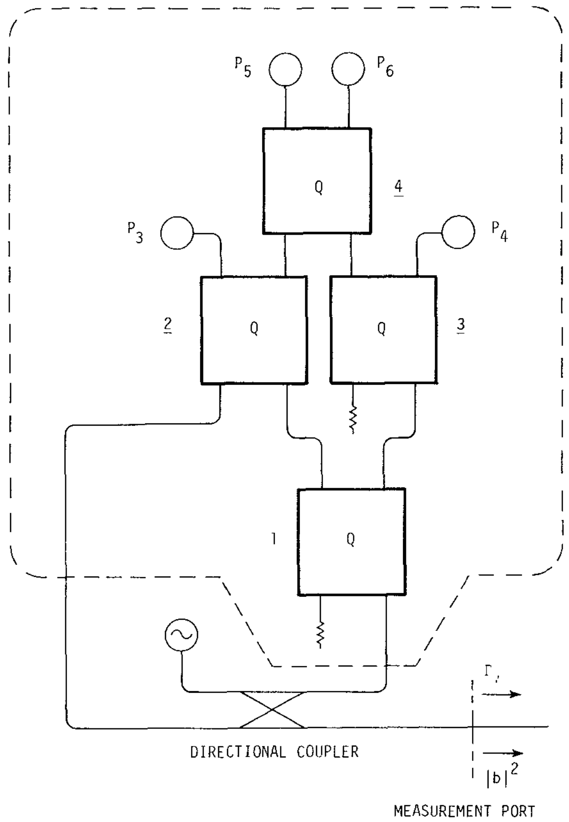

Fig. 4. A six-port circuit using only quadrature hybrids.

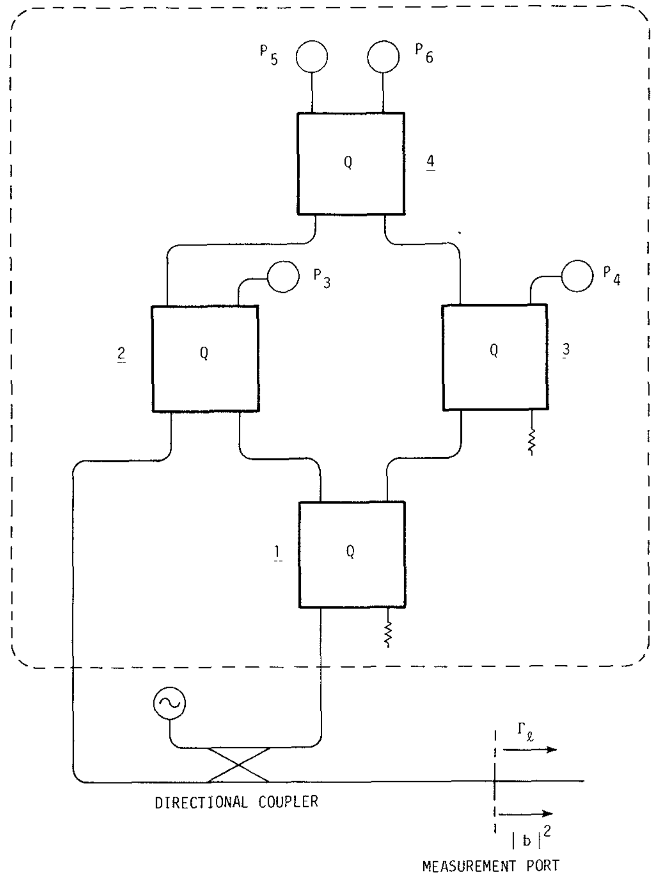

Fig. 5. An alternative to the circuit of Fig. 4.

There are a number of possible variants to the "vector voltmeter" part of the circuit; some of these include:

1) Change the hybrids at positions 2 and 3 from Q to H types.

2) Alternatively, or simultaneously with 1), interchange the Q and H hybrids at positions 1 and 4.

3) Interchange the connections at the inputs and/or outputs of both hybrids at positions 2 and 3.

4) Interchange the connections at the input to the hybrid at position 1.

Another variant is shown in Fig. 4. Here all hybrids are of the quadrature type. A variant of the circuit of Fig. 4 is shown in Fig. 5. Although the performance of each of these circuits is, ideally, identical to that already described, in actual practice, because of nonideal components, one may be preferred over another.

In addition to the ratios between the q's, the ratios among the $|A|$, $|D|$, $|E|$, $|G|$ are also determined by the vector voltmeter portion of the circuit. By contrast, and except as subsequently noted, the absolute magnitudes of all of these terms are determined by that portion of the circuit outside the dotted line.

Returning again to Fig. 2, one notes that the input levels to the vector voltmeter are $\sqrt{3}\,b$ and $(\sqrt{3}\,a)/2$. Basically, the parameters at one's disposal in adjusting the magnitudes of q or $A \cdots G$ are the coefficients of b,a. Let these be designated by β and α so that in Fig. 2, $\beta = \sqrt{3}$ and $\alpha = \sqrt{3}/2$. It is possible to adjust α and β by a variety of methods, but first it should be noted that the $|q|$'s are proportional to β/α while $|A| \cdots |G|$ are proportional to β. In principle, α and/or β can be made small by the use of attenuation in the respective lines. (This attenuation could also be placed at various positions inside the vector voltmeter.) Moreover, β can be made arbitrarily large by changing the 6-dB coupler to 10 dB, 20 dB, etc. At the same time α also increases but, in contrast to β, approaches unity as a limit. Moreover, it is not hard to recognize that this must also be the limit for any passive circuit. This limit on α indirectly imposes an upper limit of perhaps 2 or so on β, since otherwise the $|q|$'s would become larger than desired. This, in turn, leads to an upper limit on $|A| \cdots |G|$ and thus the signals at the power detectors.

Returning to Fig. 2, if one removes the lossless restriction, and assumes a surplus of input power, one could, as already noted, realize a small increase in α by going to a 10- or 20-dB coupler. The resulting increase in β, however, would need to be offset by an attenuator in the line which feeds the "H" hybrid. As an alternative to this attenuation, one might insert another directional coupler or power divider at this point and use some of the surplus signal to obtain a substantially larger value for P_4. In either case, however, the increase in P_3,P_5,P_6 would only be a nominal 5 or 10 percent, which makes this modification of doubtful practical interest.

If, in addition to assuming a surplus of signal input power, one removes the restriction on the signal level at the measurement port, or assumes that it is substantially larger than that required at $P_3 \cdots P_6$, this eliminates the requirement to make α as large as possible. Ordinarily, in this mode, the connections at the left side of the coupler should be reversed, and the desired signal levels and values of $|q|$ realized through the choice of coupling values and possibly attenuators.

III. Visual Display of Reflection Coefficient

In addition to obtaining a numerical output, to which the various corrections have been applied, it is frequently useful to have a real-time oscilloscope display of the results, even at a substantially reduced accuracy. For the ideal circuit of Fig. 2, one has

$$\frac{P_5 - P_6}{P_4} = \sqrt{2}\ \text{Re}\ (\Gamma_l) \tag{5}$$

while

$$\frac{P_5 + P_6 - P_3 - P_4}{2P_4} = \sqrt{2}\ \text{Im}\ (\Gamma_l). \tag{6}$$

Thus if $P_3 \cdots P_6$ are available in analogue form, and assuming the system is leveled so that P_4 is constant, one can obtain signals proportional to the real and imaginary parts of Γ_l by a simple addition.

IV. Summary

As compared with earlier six-port circuits, the one in Fig. 2, together with its variants, requires fewer components and is inherently lossless. This, in turn, reduces the power input requirements. Moreover, the "q" values, which it provides, more nearly approach the ideal. Although only a limited amount of practical experience has been realized with this circuit, to date, the preliminary results have been encouraging. In particular, this circuit has been implemented in WR-15 waveguide, and is described in an accompanying paper [3].

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A Semiautomated Six Port for Measuring Millimeter-Wave Power and Complex Reflection Coefficient

MANLY P. WEIDMAN

***Abstract*—A six-port system has been developed and applied to the precision measurement of power and complex reflection coefficient in WR-15 (50–75 GHz) waveguide. The system is automated except for frequency and switching control for the signal source. This system provides a time-saving factor of at least five as compared to a tuned reflectometer with little, if any, degradation in accuracy.**

Introduction

PRESENT state-of-the-art systems for measuring power or reflection coefficients at NBS in the WR-15 waveguide size consist of tuned three- or four-port reflectometers. The tuned system does not lend itself to broadband, stepped frequency measurements. The six-port system described in this paper does lend itself very well to stepped frequency measurements and was chosen for this reason. The system has only been evaluated at six frequencies in the range 55 to 65 GHz, but the extension of its use to broadband measurements should be straightforward. As it presently exists, it still provides a time-saving factor of at least 5 to 1 over tuned reflectometers. The measurement uncertainties of the six-port system are equivalent to those obtained using a tuned reflectometer with the possible exception of a slight reduction in the accuracy of reflection coefficient magnitude for small reflection terminations ($|\Gamma| < 0.01$).

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System Design

The six-port system illustrated in Fig. 1 was chosen as the basis for the measurement system. The basic theory for this configuration has been described by Engen [1]. All detectors,