Using the reflection and the transmission coefficients at the reference plane T shown in Fig. 1, the equivalent circuit parameters of the T and lattice networks are represented by only reactances, since a dielectric post is assumed to be lossless.

III. Numerical Results

We take the example given by Marcuvitz [1], Nielsen [2], Araneta *et al.* [3], Sahalos and Vafiadis [4], Hsu and Auda [5], and Leviatan and Sheaffer [6], which computes the scattering parameters and equivalent network elements as a function of relative permittivity ϵ for a centered lossless dielectric post of $r/d = 0.05$ at $\lambda/d = 1.4$, where d is the width of the rectangular waveguide, r is the radius of the cylindrical post as shown in Fig. 1, and λ is the wavelength in free space. The results obtained agree well with Leviatan's [6]. The resonance occurs at $\epsilon = 112.5$.

Now we consider the bandpass filter which takes a minimum of the reflection coefficient R at $\lambda/d = 1.4$ for a centered lossless dielectric post of $r/d = 0.05$ and with $\epsilon = 112.5$. Fig. 2 shows the magnitude of the reflection coefficient R by the solid line, where the frequency is normalized by the cutoff frequency f_c. It is found in this case that the shunt reactance for the T network decreases with increasing frequency, while the series and cross reactances for the lattice network increase with increasing frequency. The series arm resonance occurs at $f_0/f_c = 1.42089$.

We consider the equivalent lattice circuit, where the series reactance corresponds to a resonant parallel LC network and the cross reactance to an inductor. By the relationship between the original and equivalent circuit reactances, the normalized lumped lattice circuit is obtained as

$$L_1\omega_c = 0.0233 \qquad C_1\omega_c = 21.26 \qquad \text{and} \qquad L_2\omega_c = 0.227.$$

L_1 and C_1 are the inductor and the capacitor for the parallel LC network, respectively, and L_2 is the inductor for the cross arm.

The magnitude of the reflection coefficient R calculated by the lumped lattice circuit is shown by the broken line in Fig. 2, and it agrees well with $|R|$ obtained by the CFBEM.

Next we consider the waveguide loaded with two posts shown in Fig. 3, where each post is the same as the previous one and is located in the middle of the waveguide. If there is no interaction between the posts, the equivalent circuit of the waveguide may be assumed to be a chain of previous equivalent circuits connected in cascade as shown in Fig. 4. C_l is the transmission line of length l.

Fig. 5(a), (b), and (c) shows the magnitudes of the reflection coefficient R versus f/f_c with various l/d values, where the solid and broken lines correspond to the results obtained by the CFBEM and by the equivalent circuit cascaded by previous ones, respectively. It can be seen that the two agree well at $l/d \geqslant 0.5$. In the frequency range under consideration, i.e., $1.38 \leqslant f/f_c \leqslant 1.48$, the wavelength in the waveguide λ_g is $0.5258d \geqslant \lambda_g/4 \geqslant 0.4583d$. For this example, the reflection coefficient R for the waveguide loaded with two centered posts shown in Fig. 3 seems to be evaluated fairly well at $l > \lambda_g/4$ by the equivalent circuit shown in Fig. 4. For l/d less than 0.5, disagreement occurs because the coupling between the posts was neglected in the circuit analysis.

IV. Conclusions

We show that some of the lossless dielectric post resonances in a rectangular waveguide can be physically realized by a lattice circuit and that the interaction between two posts in a rectangular waveguide can also be evaluated by the lattice circuit.

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A Statistical Method for Calibrating the Six-Port Reflectometer Using Nonideal Standards

STEPHEN P. JACHIM, MEMBER, IEEE, AND
W. DARIL GUTSCHER, MEMBER, IEEE

***Abstract*—This paper presents an alternative method for calibrating the six-port reflectometer. Through the use of a redundant set of calibration standards, an estimate of the 11 real calibration constants is determined in the minimum-mean-squared-error sense. This technique enables the user to weight the contribution of each standard to the calibration process as a function of confidence in the quality of that standard. The resulting computer algorithm is quite straightforward and provides a direct measure of the tightness of fit between the estimated six-port model and the observed data.**

I. Introduction

Since their inception, six-port networks have found applications ranging from power meters to vector network analyzers. Because of the inherent simplicity and stability of the six-port network, these applications have also covered the spectrum of

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The authors are with the Los Alamos National Laboratory, MS H827, Los Alamos, NM 87545.

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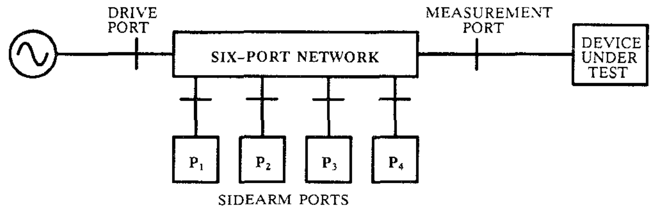

Fig. 1. A general six-port network configured as a reflectometer, where P_1 through P_4 are power indicators.

metrological precision. The topic of this paper will be limited in scope to the six-port reflectometer, but the analysis can be extended to other instruments.

Both iterative and deterministic methods have been presented in the literature for determining the calibration constants associated with the six-port reflectometer [1]–[6]. In all cases to date, however, the parameters of each calibration standard are assumed to be known absolutely, i.e., without error. In certain applications, however, some or all of these standards may be characterized with limited accuracy. Under such circumstances, a statistical estimate of the calibration constants can be computed based on the relative quality of the calibration standards used. A redundant, or overspecified, set of calibration standards can serve to reduce the calibration error induced by the use of imperfectly characterized standards.

II. Fundamental Theory

A general six-port network configured as a reflectometer is shown in Fig. 1. Hoer [8] first put forth the following defining equation for a linear network of this type:

$$\Gamma = \frac{\sum_{i=1}^{4} (F_i + jG_i) P_i}{\sum_{i=1}^{4} H_i P_i} \tag{1}$$

where Γ is the reflection coefficient of the device under test; F_i, G_i, and H_i are the real calibration constants of the six-port network; and P_i is the power measured at ith side arm port.

Owing to the ratio nature of this computation, one of these 12 real constants can be chosen arbitrarily, leaving 11 constants that uniquely define the response of the reflectometer. Here, it will be assumed that

$$H_4 = 1.$$

Equation (1) can be rewritten as

$$\sum_{i=1}^{4} (F_i + jG_i) P_i - \Gamma \sum_{i=1}^{3} H_i P_i = \Gamma P_4. \tag{2}$$

Somlo and Hunter [5] used the above set of 11 equations to develop an iterative calibration procedure, solving for the 11 real constants using the known properties of six or seven complex standards. More recently, they have described a deterministic calibration method using five complex standards [7]. The procedure under discussion here generates a statistical estimate of the 11 constants in the linear model of (2) through measurement of an arbitrarily overspecified set of calibration standards. In other words, the process incorporates measurements of six or more complex reflection standards. Statistical methods have been described by Engen [9] and Herscher and Carroll [10] which utilize the redundancy inherent in the six-port calculations to minimize the effects of power measurement errors wherein the six-port calibration constants are assumed to be known exactly. The work presented here can be considered complementary to the above in that statistical estimates of the calibration constants are determined in the presence of imperfect standards.

III. Analytical Development

On the basis of (2), the following weighted squared error ϵ can be defined [11]:

$$\epsilon = \left(\frac{1}{2}\right) \sum_{k=1}^{m} W_k \left| \sum_{i=1}^{4} (F_i + jG_i) P_i^k - \Gamma^k \sum_{i=1}^{3} H_i P_i^k - \Gamma^k P_4^k \right|^2$$

$$= \left(\frac{1}{2}\right) \sum_{k=1}^{m} W_k \left[\left(\sum_{i=1}^{4} F_i P_i^k - r^k \sum_{i=1}^{3} H_i P_i^k - r^k P_4^k \right)^2 + \left(\sum_{i=1}^{4} G_i P_i^k - x^k \sum_{i=1}^{3} H_i P_i^k - x^k P_4^k \right)^2 \right]. \tag{3}$$

Note that k is an index, rather than an exponent, where k is the index of the calibration measurement; m is the total number of calibration measurements; W_k is the weight assigned to the kth calibration measurement; r^k is the real part of the kth reflection standard; and x^k is the imaginary part of the kth reflection standard.

The object of the procedure is to minimize (3). Thus, setting the partial derivatives of (3) with respect to each parameter to zero:

$$\frac{\partial \epsilon}{\partial F_i} = \sum_{k=1}^{m} W_k P_i^k U^k = 0, \qquad i = 1 \cdots 4$$

$$\frac{\partial \epsilon}{\partial G_i} = \sum_{k=1}^{m} W_k P_i^k V^k = 0, \qquad i = 1 \cdots 4$$

$$\frac{\partial \epsilon}{\partial H_i} = \sum_{k=1}^{m} - W_k P_i^k (r^k U^k + x^k V^k) = 0, \qquad i = 1 \cdots 3 \tag{4}$$

where

$$U^k = \sum_{i=1}^{4} F_i P_i^k - r^k \sum_{i=1}^{3} H_i P_i^k - r^k P_4^k$$

$$V^k = \sum_{i=1}^{4} G_i P_i^k - x^k \sum_{i=1}^{3} H_i P_i^k - x^k P_4^k. \tag{5}$$

Now define the following augmented matrices:

$$[\tilde{U}^k] = [P_1^k \; P_2^k \; P_3^k \; P_4^k \; 0 \; 0 \; 0 \; 0 \; -r^k P_1^k \; -r^k P_2^k \; -r^k P_3^k]$$

$$[\tilde{V}^k] = [0 \; 0 \; 0 \; 0 \; P_1^k \; P_2^k \; P_3^k \; P_4^k \; -x^k P_1^k \; -x^k P_2^k \; -x^k P_3^k]. \tag{6}$$

Restating (4) with the above definitions,

$$\sum_{k=1}^{m} W_k [P_i^k [\tilde{U}^k] + 0[\tilde{V}^k]][X] = \sum_{k=1}^{m} W_k P_4^k [P_i^k r^k + 0 x^k], \qquad i = 1 \cdots 4$$

$$\sum_{k=1}^{m} W_k [0[\tilde{U}^k] + P_i^k [\tilde{V}^k]][X] = \sum_{k=1}^{m} W_k P_4^k [0 r^k + P_i^k x^k], \qquad i = 1 \cdots 4$$

$$\sum_{k=1}^{m} W_k [r^k P_i^k [\tilde{U}^k] + x^k P_i^k [\tilde{V}^k]][X] = \sum_{k=1}^{m} W_k P_4^k [r^k P_i^k r^k + x^k P_i^k x^k] \qquad i = 1 \cdots 3 \tag{7}$$

or, alternatively, using partitioned, augmented matrices,

$$\left[\sum_{k=1}^{m} W_k\left[[C_u^k][\tilde{U}^k]+[C_v^k][\tilde{V}^k]\right]\right][X]$$

$$= \sum_{k=1}^{m} W_k P_4^k [C_u^k | C_v^k] \begin{bmatrix} r^k \\ x^k \end{bmatrix} \quad (8)$$

where

$$[X] = \begin{bmatrix} F_1 \\ F_2 \\ F_3 \\ F_4 \\ G_1 \\ G_2 \\ G_3 \\ G_4 \\ H_1 \\ H_2 \\ H_3 \end{bmatrix} \qquad [C_u^k] = \begin{bmatrix} P_1^k \\ P_2^k \\ P_3^k \\ P_4^k \\ 0 \\ 0 \\ 0 \\ 0 \\ r^k P_1^k \\ r^k P_2^k \\ r^k P_3^k \end{bmatrix} \qquad [C_v^k] = \begin{bmatrix} 0 \\ 0 \\ 0 \\ 0 \\ P_1^k \\ P_2^k \\ P_3^k \\ P_4^k \\ x^k P_1^k \\ x^k P_2^k \\ x^k P_3^k \end{bmatrix} \quad (9)$$

The simultaneous solution of (8) defines a set of calibration constants that is optimum in the minimum-mean-squared-error sense. Equation (8) is linear in the variable $[X]$ with the solution

$$[X] = \left[\sum_{k=1}^{m} W_k\left[[C_u^k][\tilde{U}^k]+[C_v^k][\tilde{V}^k]\right]\right]^{-1}$$

$$\cdot \sum_{k=1}^{m} W_k P_4^k [C_u^k | C_v^k] \begin{bmatrix} r^k \\ x^k \end{bmatrix}, \qquad m \geq 6. \quad (10)$$

This form of the solution lends itself to a convenient algorithmic implementation. As the side arm powers are measured for a given standard, the component matrices on the right-hand side of (10) are formed, weighted, and summed in succession. This process continues until all standards have been measured, whereupon the solution vector is computed.

If the Γ^k are known to be random variables with a Gaussian distribution, then the W_k should be chosen as

$$W_k = \left(\frac{1}{\sigma_k^2}\right). \quad (11)$$

In some instances in practice, however, the σ_k may not be well known, and an assignment of the W_k based on the relative confidence in the precision of the Γ^k is appropriate [11]. Once a solution is calculated, a computation of (3) provides a measure of the tightness of fit of the estimated parameters in the model of (1). This is a useful quantitative measure of the overall quality of the calibration.

In practice, the values of the chosen reflection standards should be roughly uniformly distributed around the reflection coefficient plane. In particular, one or more of the standards should be near the center of the Smith chart. This load need not be of an extremely high quality as nonideal reflection parameters are easily entered into the calibration algorithm.

Also, at least one standard should have a reflection magnitude between 0 and 1 (i.e., 0.5) in order to properly define the mapping of the radial component in the complex plane. Failure to do so can lead to high error sensitivity in the intermediate zone. The requirement of this standard is a recognized drawback of this calibration technique, the significance of which will depend on the application. This intermediate reflection standard can be realized either with a matched attenuator followed by a short or with a quarter-wave section of mismatched line followed by a load. In either case, nonideal reflection parameters are easily accommodated.

TABLE I
TEST DATA FROM MEASUREMENTS OF ONE-PORT DEVICES ON BOTH AN EXPERIMENTAL SIX-PORT REFLECTOMETER AND A CALIBRATED VECTOR NETWORK ANALYZER

DEVICE NUMBER	SIX-PORT READINGS		ANA READINGS		DIFFERENCE MAGNITUDE
	mag.	angle	mag.	angle	
1	0.0039	64.6587	0.0001	150.4931	0.0039
2	0.0149	41.1781	0.0131	31.3281	0.0030
3	0.0112	46.5987	0.0084	36.5285	0 0033
4	0.0084	63.8371	0.0052	59.1994	0.0032
5	0.0129	88.0281	0.0113	94.7353	0 0021
6	0.0189	69.0704	0.0166	68.4408	0.0022
7	0.0175	60.1377	0.0155	58.2955	0.0020
8	0.9813	50.9346	0.9879	51.2071	0.0080
9	1.0025	−1.4789	1.0001	−1.6681	0.0041
10	0.9839	−7.0473	0.9892	−7.4213	0.0083
11	0.9931	−68.3367	0.9883	−68.1976	0.0053
12	0 9967	−130.0411	0.9850	−129.2706	0 0177
13	0.9682	170 9233	0.9788	170.6669	0.0115
14	0.9850	179.8825	1.0000	179.9746	0.0151
15	0.9612	109.9447	0.9758	110.7189	0.0196
16	0.7852	−15.4009	0.7888	−14.4742	0.0132
17	0.7844	−74.6954	0.7834	−74.6767	0.0010
18	0.7687	−135.2450	0.7794	−136.3047	0.0178
19	0.7924	163.7383	0.7793	163.0486	0.0162
20	0.7577	103.6535	0.7799	104.1699	0.0232
21	0.7690	44 6428	0.7804	45.3241	0.0147
22	0.4879	−12 4660	0.4888	−11 9930	0.0041
23	0.4787	−72.8856	0 4787	−72.0980	0 0066
24	0.4788	−134.5334	0.4751	−133.9441	0.0061
25	0.4770	164.0440	0.4807	165.1381	0.0099
26	0.4826	105.6583	0.4859	106.1331	0.0052
27	0.4880	47.5404	0.4871	47.6223	0.0011
28	0.2628	−8.5379	0.2701	−8.4530	0.0073
29	0.2517	−69.3497	0.2579	−68.4522	0.0074
30	0.2616	−133 5310	0 2545	−131.0900	0 0131
31	0.2630	168 0413	0 2619	167.0962	0 0045
32	0.2800	108.8979	0.2696	107.9918	0.0113
33	0 2806	52.3988	0.2718	50 5351	0 0126
RMS DIFFERENCE MAGNITUDE = 0.0105					

IV. EXPERIMENTAL RESULTS

The development of this calibration procedure was driven by the need to implement a six-port reflectometer in a 6.125-in EIA rigid line with a main-line peak power level of over 1 MW. Precision calculable standards are unavailable in this medium. Therefore, a set of offset shorts, loads, and mismatches will be fabricated and characterized as transfer standards.

In order to test the calibration procedure described here, a low-power experiment was performed to simulate the conditions mentioned above. A set of coaxial delay lines were built with standard semirigid coaxial line using nonprecision connectors and interseries adapters. A short, a load, and 1, 3, and 6 dB attenuators were used in conjunction with these delay lines to form transfer standards that were characterized using a calibrated vector network analyzer.

The experimental six-port reflectometer was then calibrated using seven of these transfer standards. A separate set of test devices, differing in reflection coefficient from the calibration standards, was then measured on the six-port reflectometer and on the network analyzer. The test data were compared under the assumption that the network analyzer would exhibit negligible measurement error relative to that of the six-port reflectometer.

The experimental data are shown in Table I, wherein the magnitude of the vector distance between measurements on the two instruments is listed for each test device. The rms radius of the error circles over all the data points is 0.0105. This measure of calibration error is ultimately limited by influences such as uncertainty in the reflection standard parameters, connector repeatability, and measurement system linearity and stability.

V. Conclusions

A method for calibrating the six-port reflectometer with standards of limited quality using the method of minimum-mean-squares has been developed. The solution provided can be implemented in software in a straightforward manner, and a statistical measure of fitting of the derived parameters can be calculated.

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Results of Phase and Injection Locking of an Orotron Oscillator

R. W. McMillan, senior member, ieee,
D. M. Guillory, R. G. Hay, D. E. Wortman,
H. Dropkin, senior member, ieee, and
J. M. Cotton, Jr., member, ieee

***Abstract*—We describe experiments resulting in phase and injection locking of a 60 GHz orotron oscillator in pulsed and CW modes. The measured phase-locked phase noise results obtained in CW mode were −85, −95, and −105 dBc/Hz at 1, 10, and 100 kHz separation from the carrier, respectively. The null depths and asymmetry of the first maxima of the pulsed spectrum for this source were 35 dB and 2 dB (difference between power levels of first maxima), respectively, operating with a pulse width of 15 μs. At 3 μs, these quantities become 25 dB and 1 dB, respectively. The orotron was observed to injection lock in pulsed mode with an input signal 22 dB below the output power level.**

I. Introduction

The orotron is a linear-beam oscillator whose operation is based on the Smith–Purcell effect [1], in which radiation is generated when an electron beam skims the surface of a metallic diffraction grating. Operation of the orotron has been treated elsewhere [2], [3] and will not be described herein, except to say that the orotron may be tuned by varying the cathode-to-grating voltage, which modulates the electron beam velocity, or by varying the mirror spacing, which changes the resonant frequency of the cavity. For phase locking, the phase control signal is fed back to the grating, so that the physical effect of applying a phase correction voltage to the grating is to modulate the beam velocity. The orotron used in the experiments described herein has a grating period of 0.4 mm.

II. Phase-Locking Approaches

Different phase-locking methods were used depending on whether the orotron was operated in short-pulse, long-pulse, or CW mode. The tuning coefficient of the orotron was determined to be 0.2 MHz/V during these measurements, which is a fairly low value, and this lack of gain must be compensated with higher gain elsewhere in the loop. In long-pulse and CW modes a digital phase/frequency detector was used, while in short-pulse mode, a broad-band ac-coupled analog phase lock was used.

The digital phase lock used to lock the orotron in long-pulse and CW modes has been described in [4]. Although the linear circuits [5] give better phase noise performance, digital phase locks have been found to be more nearly immune to circuit variables, such as intermediate frequency (IF) levels and external interference. The approach used was conventional, with part of the orotron power coupled to a harmonic mixer, where it is compared in phase to a stable reference oscillator to generate a

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R. W. McMillan and J. M. Cotton, Jr., are with the Georgia Tech Research Institute, Georgia Institute of Technology, Atlanta, GA 30332.

D. M. Guillory is with the Hewlett-Packard Company, Boise, ID 83707.

R. G. Hay, D. E. Wortman, and H. Dropkin are with the U.S. Army Harry Diamond Laboratories, 2800 Powder Mill Road, Adelphi, MD 20783-1197.

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