Time-Domain Measurements with the Hewlett-Packard Network Analyzer HP 8510 Using the Matrix Pencil Method

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***Abstract*—The HP 8510 time-domain network measurements are frequency-domain measurements transformed to the time domain using the inverse Fourier transform, the objective being to discriminate various scattering centers. This computational technique benefits from the wide dynamic range and the error correction of the frequency-domain data, but requires a frequency-domain response measured over a wide frequency range to give useful resolution in the time domain. The generalized pencil of function (GPOF) method, also known as the matrix pencil method, provides for much higher resolution than the Fourier techniques. A comparison of the two methods is given for the example of the Beatty standard.**

I. Introduction

The Hewlett-Packard network analyzer HP 8510B has built-in options for time-domain measurements. Two modes of measurements can be performed: time band-pass and time low-pass. The time low-pass mode simulates traditional time-domain reflectometer (TDR) measurement, which gives the response of the device to a step or an impulse stimulus. The time band-pass mode is a general-purpose time-domain mode that allows any frequency-domain response to be transformed to the time domain. The result is the impulse response. Parameters obtained are the time delay and amplitude of an impulse. The rise and fall times of the impulse are inversely proportional to the frequency bandwidth. If the response consists of two or more impulses that are close to each other by twice the rise time, all one sees in the time-domain response is a single impulse. The two discontinuities then cannot be distinguished.

The problem with conventional FFT techniques is that, if there are two impulses located at τ_1 and τ_2 representing reflections from different discontinuities, then in order to resolve them it is necessary to have data in the frequency domain up to a bandwidth of $1/(\tau_2-\tau_1)$ with $\tau_2>\tau_1$.

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However, this limitation of the FFT can be avoided if one uses a parametric technique. For the parametric technique one tries to fit a sum of complex exponentials to the frequency-domain data. Then the transform in the time domain would consist of impulses. Theoretically, to resolve two impulses in the time domain, all one needs in the frequency domain are eight samples of the measurements. These measurements need not be over a bandwidth of $1/(\tau_2-\tau_1)$. However, because of noise in the data, more than eight samples are necessary in practice. Hence by using parametric estimation techniques it is possible to go beyond the limitations of Fourier techniques and provide a general framework for the gating procedures which are the strength of the time domain.

The particular parametric estimation scheme to be used is the generalized pencil of function technique (GPOF). Not only is this technique computationally very efficient; of the many existing parametric techniques available in the signal processing literature, it has the least statistical variance of the estimates in the presence of noise [1], [2]. From a statistical point of view, the GPOF has a smaller variance, as shown in [3], than the Prony-type method [9].

In this paper the GPOF is used to improve the resolution of the HP 8510B data in the time domain instead of the conventional Fourier techniques [8].

II. The Numerical Method

The vector network analyzer standard outputs are complex reflection or transmission S parameters as a function of frequency. In the case of reflection, $S_{11}(f)$ can be seen as a general response function, $R(s)$, given in the Laplace transform domain as

$$R(s)=F(s)H(s) \tag{1}$$

where $F(s)$ and $H(s)$ are the arbitrary exciting function and the Laplace transforms of the system's impulse response for $t>0$, respectively, and s is the complex frequency. In the case of the driving function being an impulse,

$$F(s)=1. \tag{2}$$

By expressing the transfer function as an infinite set of pole singularities, according to the theory of complex variables,

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$H(s)$ is assumed to be written in a residue series as

$$H(s)=\sum_{i=1}^{M}\frac{A_i}{s-s_i},\qquad \mathrm{Re}[s_i]<0 \tag{3}$$

where s_i is a ith complex pole of $H(s)$, and A_i is the corresponding residue. When the inverse Laplace transform of (3) is taken, the time-domain response, $r(t)$, is written as

$$r(t)=\sum_{i=1}^{M}A_ie^{s_it}. \tag{4}$$

Further, by solving for complex poles and residues from frequency response function we are able to directly reconstruct the time-domain response without having to perform an inverse Fourier transform.

For time-domain responses of the form

$$r(t)=\sum_{i=1}^{M}a_i\delta(t-t_i) \tag{5}$$

the corresponding complex frequency-domain representation is

$$S_{11}(s)=\sum_{i=1}^{M}a_ie^{-st_i}. \tag{6}$$

Discretizing frequency and writing $s=k\Delta f$, we have

$$y(k)=S_{11}(k)=\sum_{i=1}^{M}a_ie^{(-j2\pi\Delta ft_i)k}+n_k \tag{7}$$

where $k=0,1,\cdots,N-1$, and n_k stands for additive noise introduced in the measurement. As mentioned before, a_i are the complex residues. Let $z_i=e^{-j2\pi\Delta ft_i}$ for notational brevity; then z_i are the poles in the Z plane. Following the GPOF method [1], we are defining matrices Y_1 and Y_2 as

$$Y_1=\begin{bmatrix} y_0 & y_1 & \cdots & y_{L-2} & y_{L-1}\\ y_1 & y_2 & \cdots & y_{L-1} & y_L\\ \vdots & & & & \\ y_{N-L-1} & y_{N-L} & \cdots & y_{N-3} & y_{N-2}\end{bmatrix} \tag{8}$$

and

$$Y_2=\begin{bmatrix} y_1 & y_2 & \cdots & y_{L-1} & y_L\\ y_2 & y_3 & \cdots & y_L & y_{L+1}\\ \vdots & & & & \\ y_{N-L} & y_{N-L+1} & \cdots & y_{N-2} & y_{N-1}\end{bmatrix}. \tag{9}$$

Matrices Y_1 and Y_2 can be written as

$$Y_1=Z_1AZ_2 \tag{10}$$

$$Y_2=Z_1AZ_0Z_2 \tag{11}$$

where

$$Z_1=\begin{bmatrix} 1 & 1 & \cdots & 1\\ z_1 & z_2 & \cdots & z_M\\ \vdots & & & \\ z_1^{N-L-1} & z_2^{N-L-1} & \cdots & z_M^{N-L-1}\end{bmatrix}$$

$$Z_2=\begin{bmatrix} 1 & z_1 & \cdots & z_1^{L-1}\\ 1 & z_2 & \cdots & z_2^{L-1}\\ \vdots & & & \\ 1 & z_M & \cdots & z_M^{L-1}\end{bmatrix} \tag{12}$$

$$Z_0=\begin{bmatrix} z_1 & 0 & \cdots & 0\\ 0 & z_2 & \cdots & 0\\ \vdots & & & \\ 0 & 0 & \cdots & z_M\end{bmatrix}\quad\text{and}$$

$$A=\begin{bmatrix} a_1 & 0 & \cdots & 0\\ 0 & a_2 & \cdots & 0\\ \vdots & & & \\ 0 & 0 & \cdots & a_M\end{bmatrix}. \tag{13}$$

Then we create matrix pencil Y_2-zY_1, which can be represented as a generalized eigenvalue problem

$$Y_2-zY_1=Z_1A(Z_0-I\lambda)Z_2. \tag{14}$$

That is, when $M\leqslant L\leqslant N-M$, then the poles z_i, $i=1,\cdots,M$, are the generalized eigenvalues of the matrix pencil Y_2-zY_1. Also, the rank of the matrix pencil is equal to the number of signal poles, M, unless $z=z_i$. Generalized eigenvalues of the matrix pencil, i.e., signal poles, are obtained by using a singular value decomposition (SVD) algorithm. We consider the matrix product $Y_1^+Y_2$, where Y_1^+ is the pseudo inverse of Y_1, and is defined by

$$Y_1^+=Y_1^H\left(Y_1Y_1^H\right)^{-1} \tag{15}$$

where superscript H denotes the conjugate transpose operation of a matrix. Note that

$$Y_1Y_1^+=I. \tag{16}$$

Referring to (10) and (11), one can write

$$Y_1^+Y_2=Z_2^+A^{-1}Z_1^+Z_1AZ_0Z_2 \tag{17}$$

where A^{-1} denotes the regular inverse. Then

$$Y_1^+Y_2=Z_2^+Z_0Z_2. \tag{18}$$

There exist vectors $\boldsymbol{p}_i$, $i=1,\cdots,M$, such that

$$Y_1^+Y_1\boldsymbol{p}_i=\boldsymbol{p}_i \tag{19}$$

and

$$Y_1^+Y_2\boldsymbol{p}_i=z_i\boldsymbol{p}_i. \tag{20}$$

Vectors $\boldsymbol{p}_i$ are then called the generalized eigenvectors of the matrix pencil Y_2-zY_1. Singular value decomposition [7] is used to compute the pseudo inverse matrix Y_1^+, namely

$$Y_1 = \sum_{i=1}^{M} \sigma_i \boldsymbol{u}_i \boldsymbol{v}_i^H \tag{21}$$

or

$$Y_1 = UDV^H \tag{22}$$

then

$$Y_1^+ = VD^{-1}U^H \tag{23}$$

where $U = [\boldsymbol{u}_1, \cdots, \boldsymbol{u}_M]$, $V = [\boldsymbol{v}_1, \cdots, \boldsymbol{v}_M]$, and matrix D is given as

$$D = \begin{bmatrix} \sigma_1 & 0 & \cdots & 0 \\ 0 & \sigma_2 & \cdots & 0 \\ \vdots & & & \\ 0 & 0 & \cdots & \sigma_M \end{bmatrix}. \tag{24}$$

U and V are matrices of so-called left and right singular vectors, respectively. Since

$$Y_1^+ Y_1 = VV^H \tag{25}$$

substituting (23) into (20) and using (19) we obtain

$$VD^{-1}U^H Y_2 VV^H \boldsymbol{p}_i = z_i \boldsymbol{p}_i \tag{26}$$

and then left multiplying both sides by V^H we have

$$V^H VD^{-1}U^H Y_2 VV^H \boldsymbol{p}_i = z_i V^H \boldsymbol{p}_i. \tag{27}$$

Note that $V^H V = I$ and $i = 1, \cdots, M$. Let us define matrix Z and vectors $\boldsymbol{z}_i$ as

$$Z = D^{-1}U^H Y_2 V \quad \text{and} \quad \boldsymbol{z}_i = V^H \boldsymbol{p}_i. \tag{28}$$

Then (27) is reduced to a square matrix eigenvalue problem

$$(Z - z_i I)\boldsymbol{z}_i = 0. \tag{29}$$

The size of the matrix Z is $M \times M$, and z_i and $\boldsymbol{z}_i$ are eigenvalues and eigenvectors of Z, respectively. If the number of poles, M, is not known, it can be estimated from the singular values

$$\sigma_1 \geqslant \sigma_2 \geqslant \cdots \geqslant \sigma_M \geqslant \cdots \geqslant \sigma_{\min(N-L,L)} \tag{30}$$

since $\sigma_{M+1} = \cdots = \sigma_{\min(N-L,L)} = 0$ for noiseless data. In the case of noisy data y_k, the largest M singular values of Y_1, $\sigma_1, \cdots, \sigma_M$, should be chosen in (21), and the resulting Y_1^+ is called the truncated pseudo inverse of matrix Y_1.

Computer code that solves for signal poles out of given frequency data has been developed in FORTRAN. A Hewlett-Packard HP9000/370 workstation running the HP-UX operating system is used as a computational platform. The input data acquisition from network analyzer HP 8510 is performed using HP-UX Basic, via IEEE-488 interface bus (HP-IB). First, network analyzer stimulus parameters are properly set and appropriate calibration is performed; then the reflection coefficient $S_{11}(k)$ is transferred to the computer. The GPOF method is applied, and the time-domain impulse response is displayed on the screen.

III. Experimental Results

As an example consider the Beatty standard, presented in Fig. 1. There are two impedance step discontinuities at the standard. If the output port is terminated with 50 Ω and the standard is excited with an impulse at port 1, there will be reflections due to the discontinuities. From a theoretical point of view there will be four dominant reflections, as shown in Fig. 2. The expected amplitude and time delay of the impulses are given in Table I.

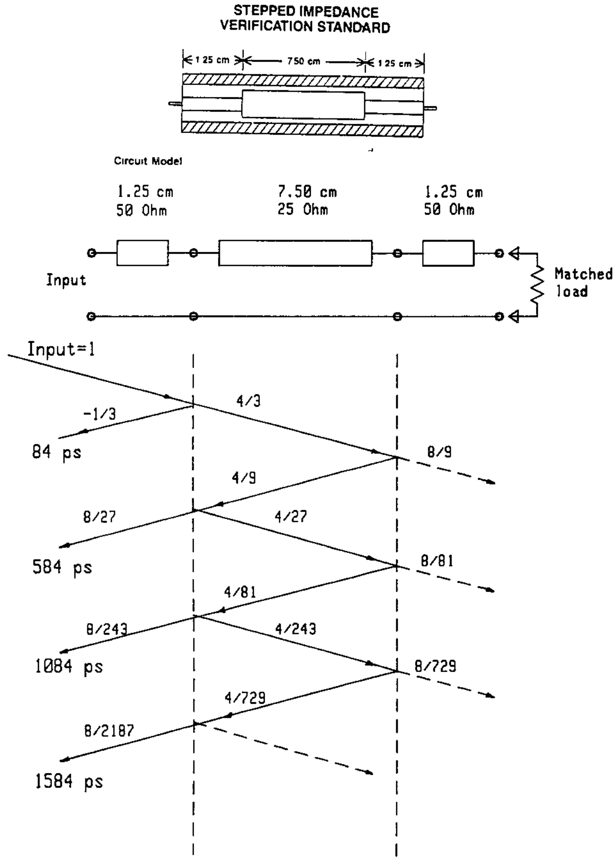

Fig. 1. Multiple reflections from the Beatty standard terminated with a matched load.

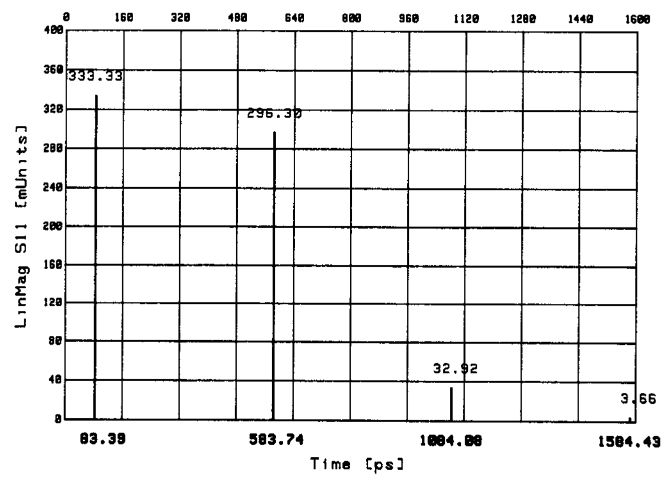

Fig. 2. Calculated impulse response of the Beatty standard.

Next we utilize the HP 8510B and measure the S_{11} parameter from 45 MHz to 18 GHz using 801 data points. The magnitude and the phase response are given in Fig. 3. Utilizing the HP 8510B internal frequency to time conversion technique, one obtains the plot in Fig. 4. The first three

TABLE I
CALCULATED IMPULSE RESPONSE OF BEATTY STANDARD TERMINATED WITH 50 Ω

Impulse	Delay Time (ps)	Amplitude
1.	83.39	333.333E-3
2.	583.74	296.296E-3
3.	1084.08	32.922E-3
4.	1584.43	3.658E-3

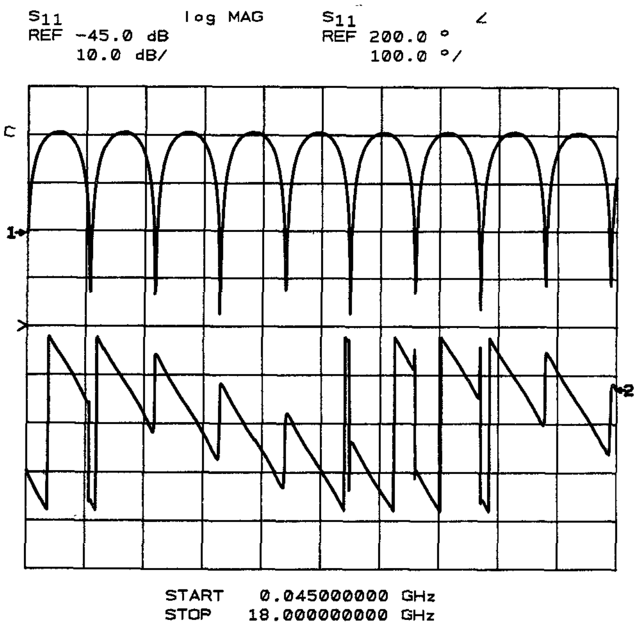

Fig. 3. The magnitude and phase response of the Beatty standard from 45 MHz to 18 GHz.

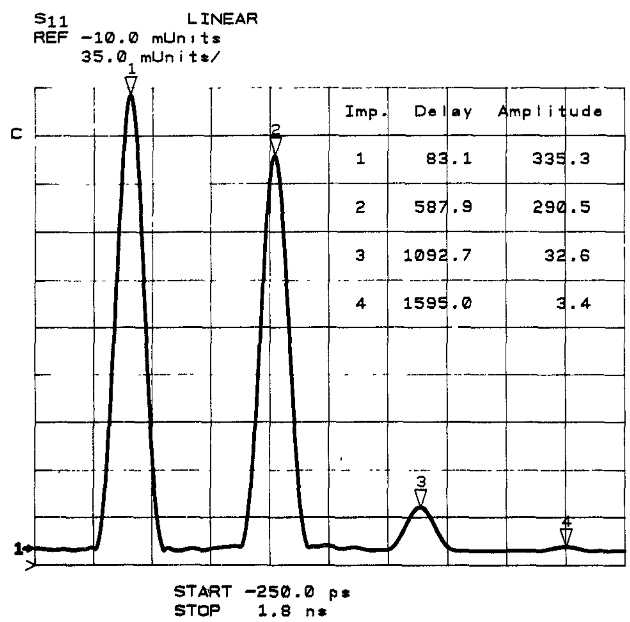

Fig. 4. Time-domain impulse response, using standard built-in bandpass option on HP 8510B, with 18 GHz bandwidth.

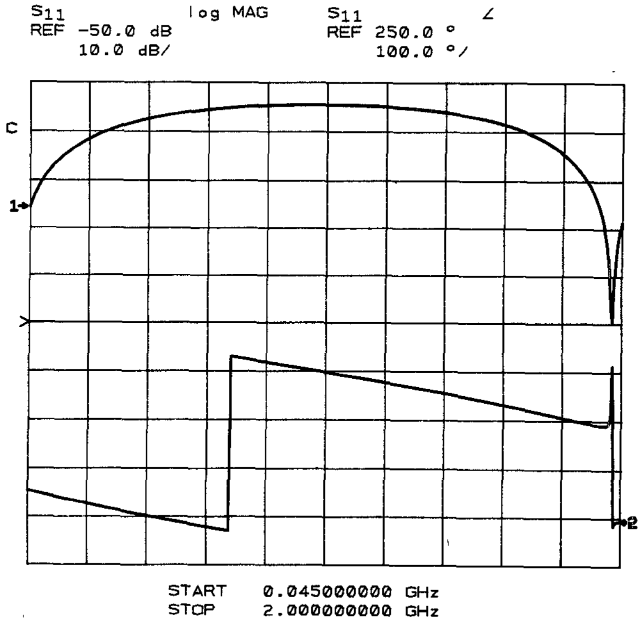

Fig. 5. The magnitude and phase response of the Beatty standard from 45 MHz to 2 GHz.

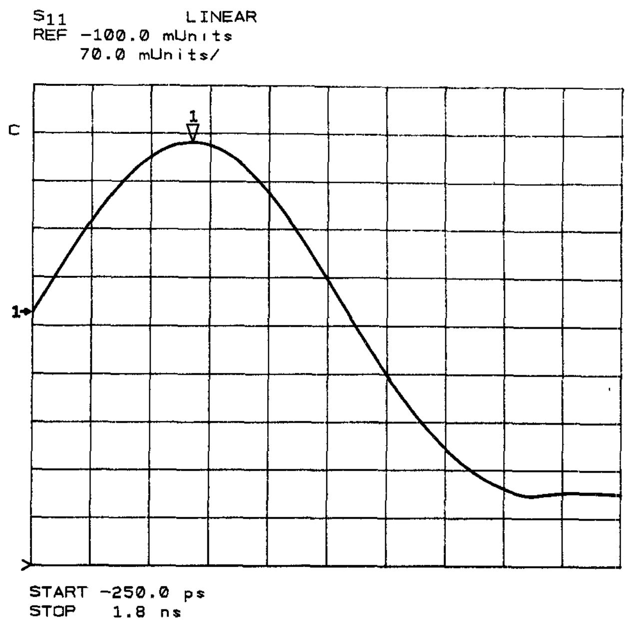

Fig. 6. Time-domain impulse response, using standard built-in bandpass option on HP 8510B, with 2 GHz bandwidth.

impulse returns are quite clear and the fourth appears where it is marked by an arrow. Comparing the measured values with the theoretical values of Table I demonstrates a reasonable agreement. Next, the bandwidth of the sweep is reduced from 18 GHz to 2 GHz. In this case the magnitude and the phase response are plotted in Fig. 5. If the inverse transform is taken with the HP 8510B internal technique, one obtains the time-domain response of Fig. 6. Observe that, as expected, no information is available about the discontinuities.

The GPOF is now applied to the same 2 GHz bandwidth data as plotted in Fig. 5. The technique computed that there are three dominant singular values from singular value analysis. So the value for the parameter M is set to be equal to 3.

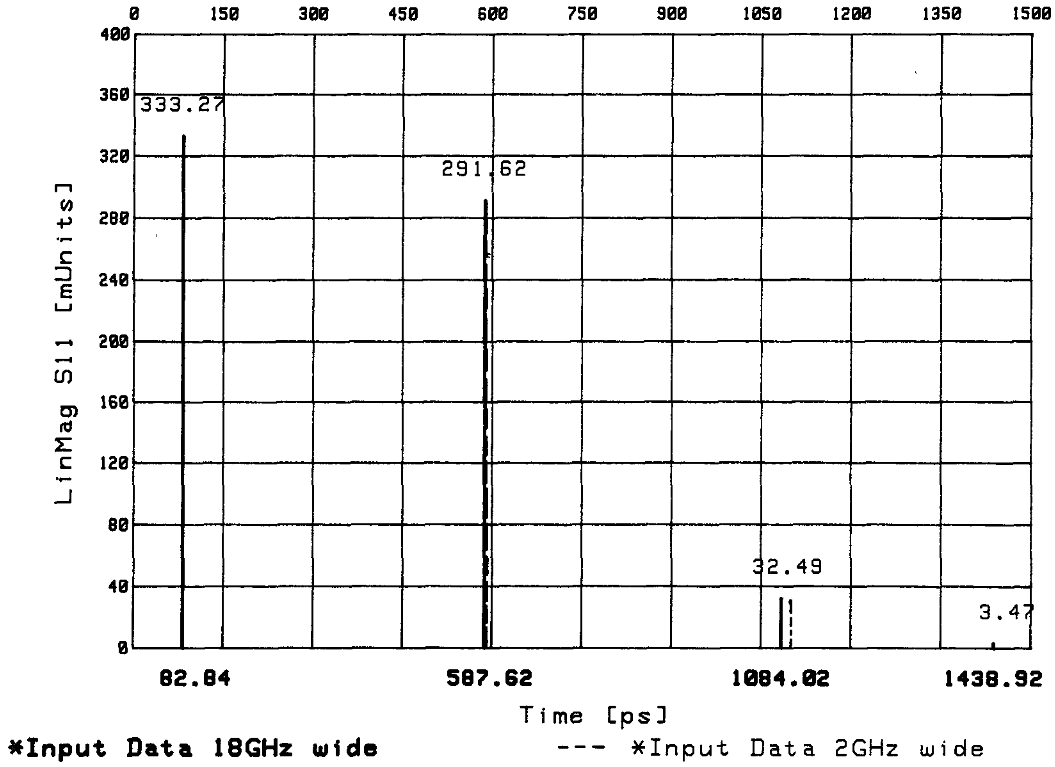

Fig. 7. Time-domain impulse response obtained using GPOF with 2 GHz (dotted) and 18 GHz (solid) bandwidth.

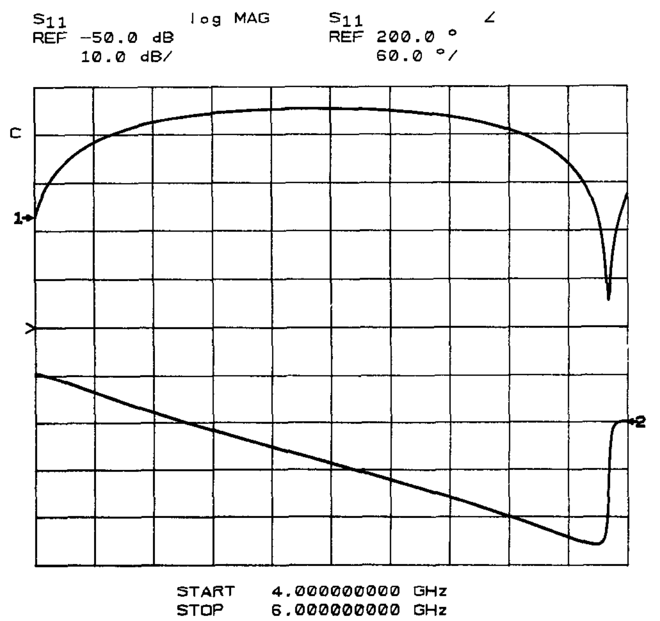

Fig. 8. The magnitude and phase response of the Beatty standard from 4 to 6 GHz.

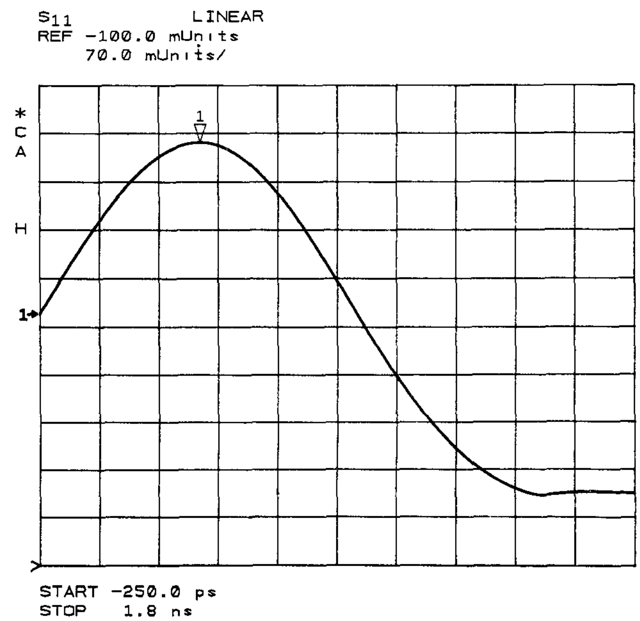

Fig. 9. Time-domain impulse response, using standard built-in bandpass option on HP 8510B, with 2 GHz bandwidth.

The amplitude and location of the impulses are shown in Fig. 7. Observe that the first three impulses have been identified and their positions quite accurately located, and that the fourth impulse is not visible. The GPOF is now applied to the same 18 GHz bandwidth data as shown in Fig. 3. The results are also plotted on Fig. 7. Observe that, for this example, reducing the bandwidth from 18 GHz to 2 GHz had no visible impact on the time-domain resolution of the first three impulses. This is because a parametric technique such as GPOF deals with the number of samples of data points rather than with the actual bandwidth of the data. This is the strength of the GPOF over conventional FFT techniques.

Next the frequency-domain response of the Beatty standard is generated over 4–6 GHz. The magnitude and the phase are plotted in Fig. 8. The internal HP 8510B standard

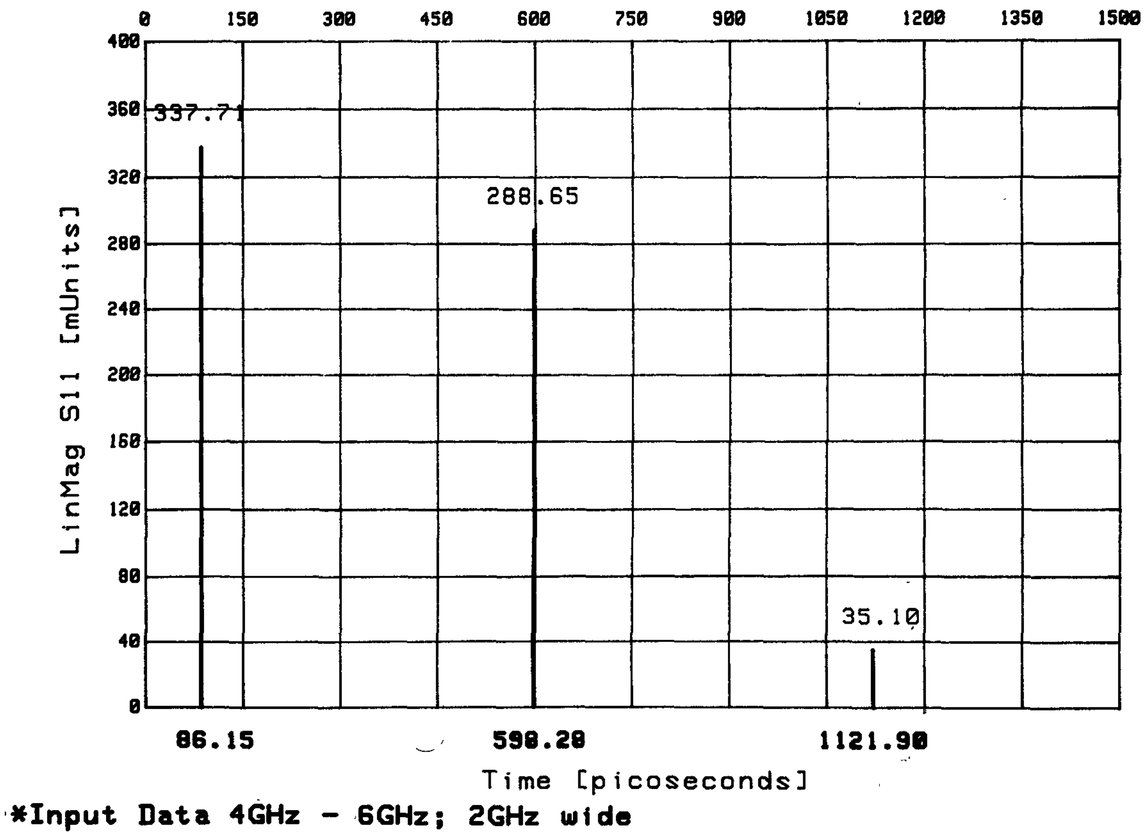

Fig. 10. Time-domain impulse response obtained using GPOF, with a bandwidth of 4–6 GHz.

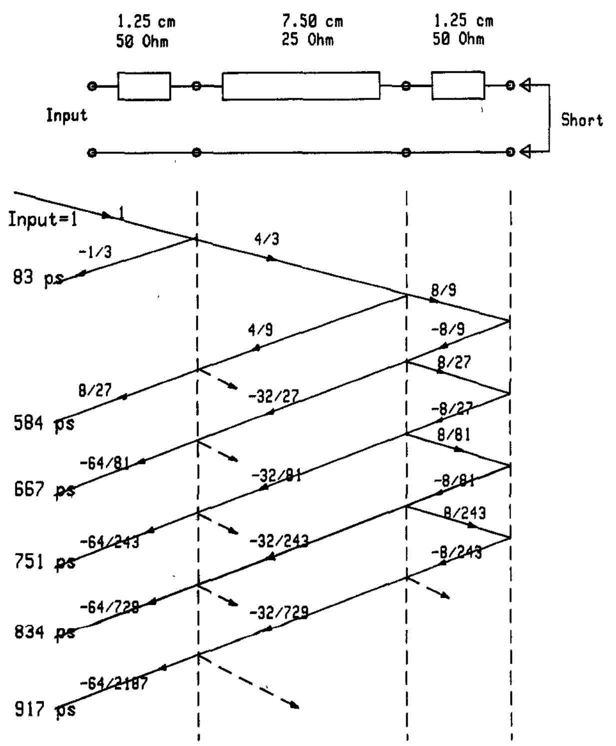

Fig. 11. Multiple reflections from the Beatty standard terminated with short.

TABLE II
CALCULATED IMPULSE RESPONSE OF BEATTY STANDARD TERMINATED WITH SHORT

Impulse	Delay Time (ps)	Amplitude
1.	83.39	333.333E-3
2.	583.74	296.296E-3
3.	667.13	790.123E-3
4.	750.52	263.374E-3
5.	833.91	87.791E-3
6.	917.30	29.264E-3
7.	1000.69	9.755E-3
8.	1084.08	29.670E-3
9.	1167.47	176.667E-3
10.	1250.87	204.486E-3
11.	1334.26	117.176E-3
12.	1417.65	48.813E-3

time band-pass response is shown in Fig. 9. Observe that no useful information about the discontinuities can be extracted. However, the GPOF technique applied to the same data gives three impulses, as shown in Fig. 10. Observe that the fourth impulse is not visible for this bandwidth data. Time delay of the first three impulses is within 3.5% of the theoretical results, while the amplitude of the first impulse is off by 1.37%, the second by 2.5%, and the third by 6.67%. The CPU time needed to compute the time-domain response by using the GPOF in Figs. 7 and 10 is of the order of 4 s on an HP9000/370 workstation.

As an even more convincing example, consider the Beatty standard terminated with a short (Fig. 11). The expected impulse response is given in Table II and in Fig. 12.

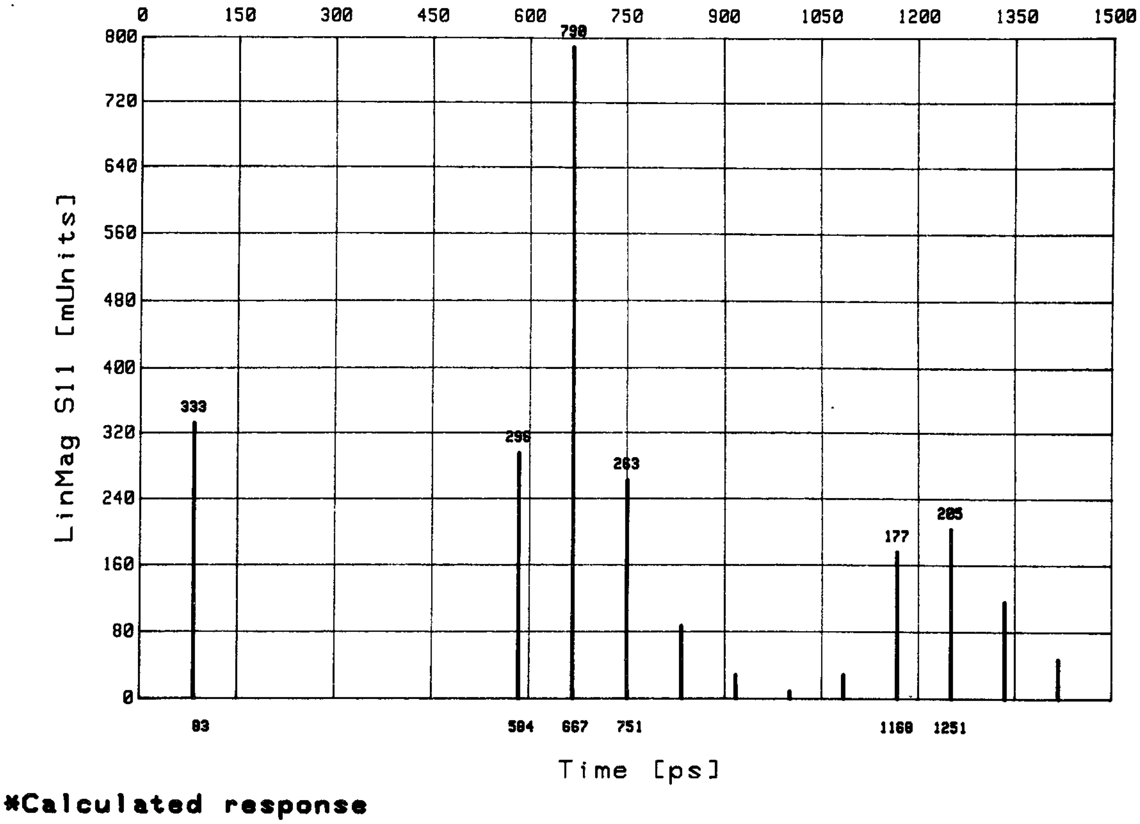

Fig. 12. Calculated impulse response of the Beatty standard terminated with short.

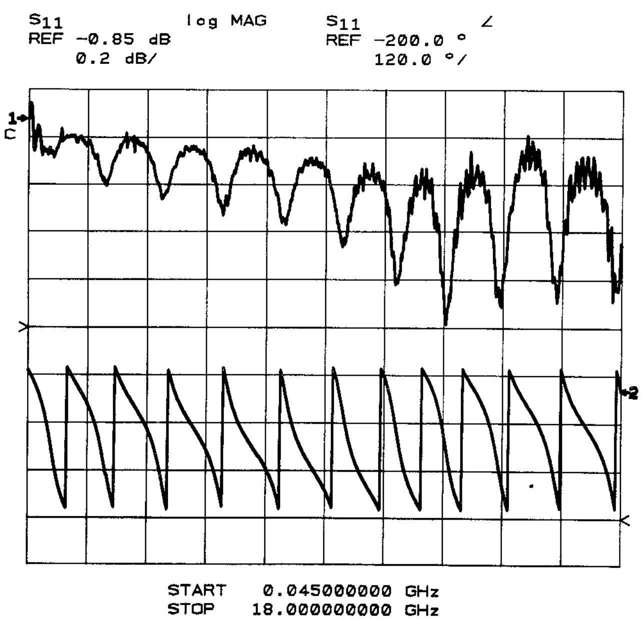

Fig. 13. The magnitude and phase response of the Beatty standard terminated with short, from 45 MHz to 18 GHz.

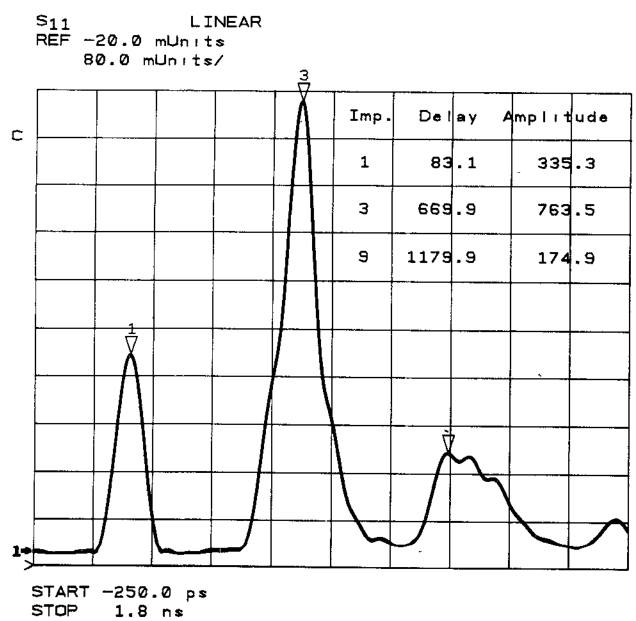

Fig. 14. Time-domain impulse response of shorted Beatty standard, using standard built-in band-pass option on HP 8510B, with 18 GHz bandwidth.

Utilizing the HP 8510B, the S_{11} parameter is measured from 45 MHz to 18 GHz, using 801 data points. The magnitude and the phase response are given in Fig. 13. Applying the HP 8510B internal frequency to time conversion technique, one obtains the plot in Fig. 14. Only impulses 1 and 3 (see Table II) are clearly observable. Fig. 15 shows unwindowed time-domain impulse response, obtained from the same 18-GHz-wide data set. Resolution is enhanced at the expense of the higher ringing and reduced dynamic range.

Next, the bandwidth of the sweep is reduced to the 4–6.5 GHz range. In this case the magnitude and the phase response are plotted in Fig. 16. If the inverse transform is

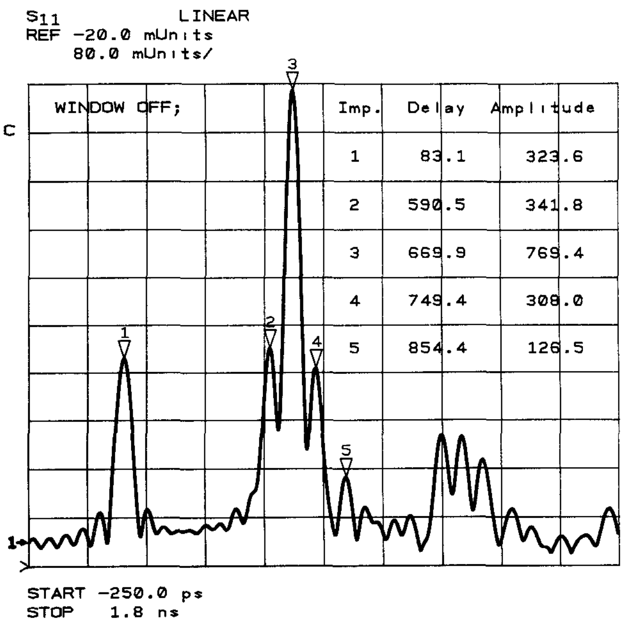

Fig. 15. Time-domain impulse response of shorted Beatty standard, using standard built-in band-pass option on HP 8510B, with 18 GHz bandwidth, windowing option turned off.

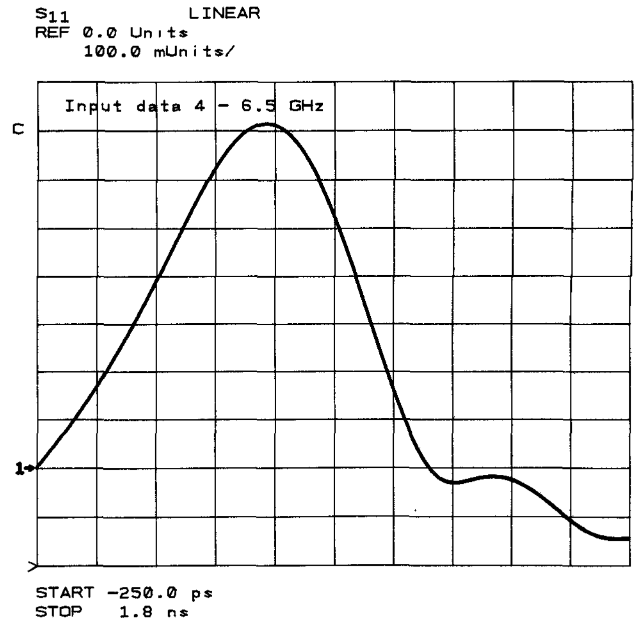

Fig. 17. Time-domain impulse response of shorted Beatty standard, using standard built-in band-pass option on HP 8510B, with a bandwidth of 4 to 6.5 GHz.

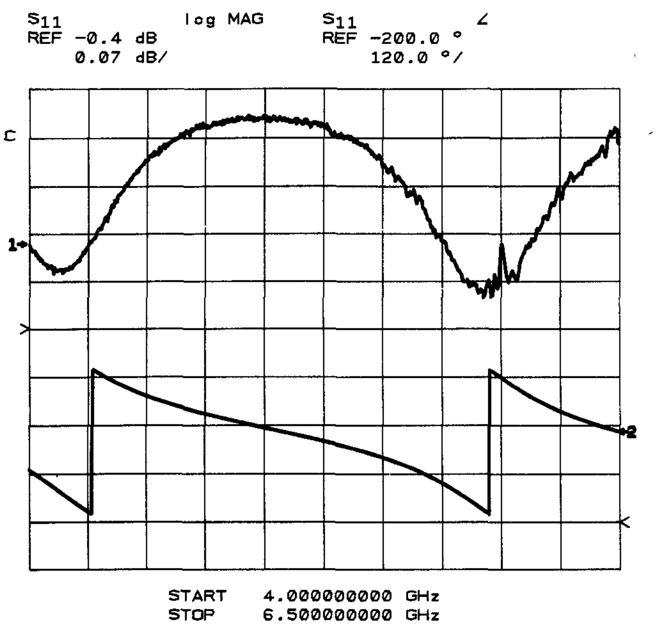

Fig. 16. The magnitude and phase response of the shorted Beatty standard from 4 to 6.5 GHz.

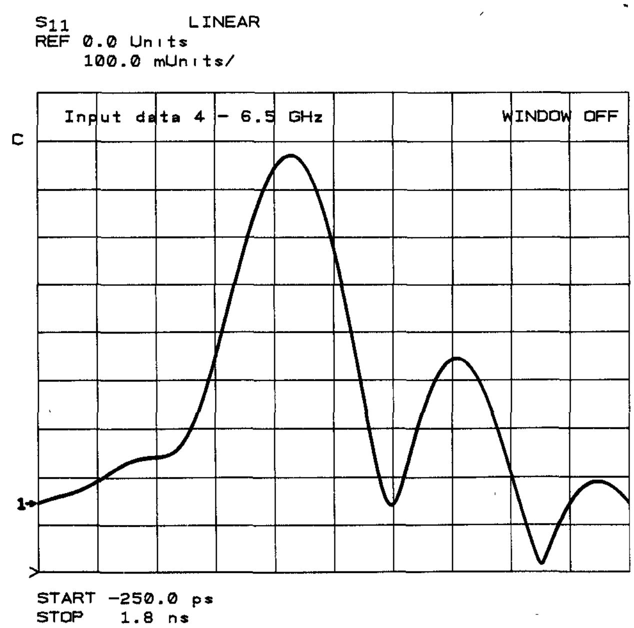

Fig. 18. Time-domain impulse response of shorted Beatty standard, using standard built-in band-pass option on HP 8510B, with a bandwidth of 4–6.5 GHz, windowing option turned off.

taken with the HP 8510B internal technique one obtains the time-domain response of Fig. 17. Fig. 18 shows unwindowed time-domain response. Observe that, as expected, no information is available about the discontinuities.

The GPOF is now applied to the same 4–6.5 GHz data as plotted in Fig. 16. The amplitude and location of the impulses are shown in Fig. 19. If the GPOF is applied to 18 GHz bandwidth data from Fig. 13, the impulse response of Fig. 20 is obtained. Observe again that, for this example, reducing the bandwidth from 18 GHz to 2.5 GHz had no visible impact on the time-domain resolution (compare Fig. 20 and Fig. 19). When Figs. 19 and 12 are compared, the time delay of the first four impulses of Fig. 19 is within 0.6% of the theoretical results, while the amplitude of the first impulse is off by 0.5%, the second by 3.1%, the third by 1.9%, and the fourth by 4.4%. Because of the memory limitations of the HP9000/370 workstation (8 MB), calculations of the impulse response in Figs. 19 and 20 are performed on a VAXstation 3500 (32 MB of memory). This particular example, which is an extreme case, involves solving for eigenvalues of a 400 by 400 matrix, and the elapsed CPU time is of the order of 1 h.

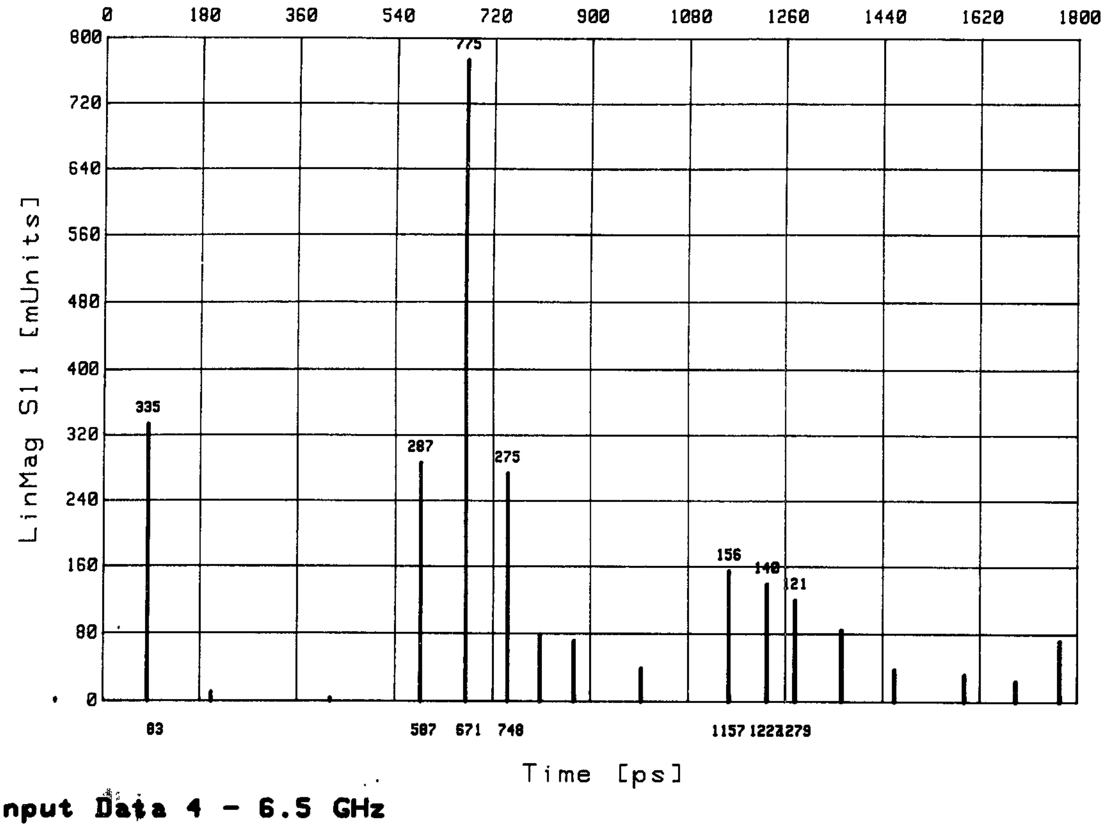

Fig. 19. Time-domain impulse response of shorted Beatty standard obtained using GPOF, with a bandwidth of 4–6.5 GHz.

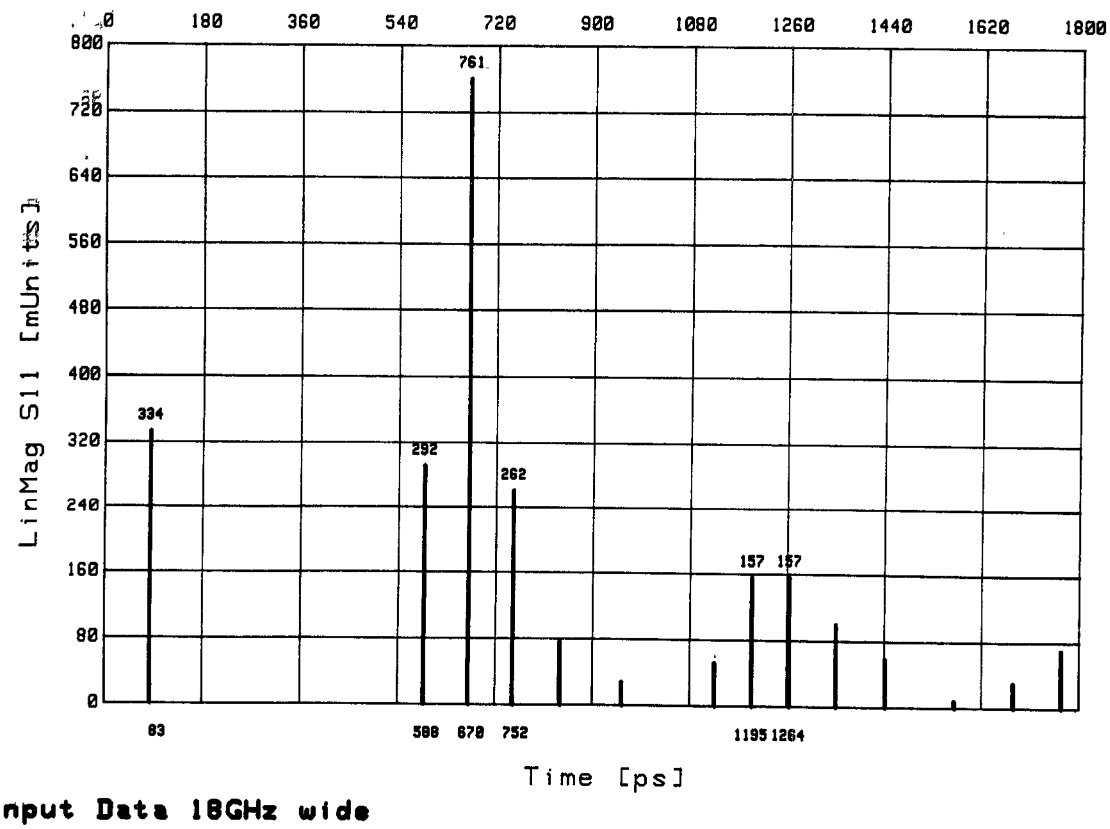

Fig. 20. Time-domain impulse response of shorted Beatty standard obtained using GPOF, with a bandwidth of 18 GHz.

IV. Conclusion

The generalized pencil of function technique (GPOF) is used for extraction of the impulse response out of limited frequency bandwidth data. The examples show that a parametric technique such as the GPOF can provide accurate, reliable results with a high degree of resolution even when the FFT-based technique fails. The method has wide application in antenna measurements, location of transmission line discontinuities, and radar cross-section measurements. It could be implemented as a standard firmware feature on advanced vector network analyzers.

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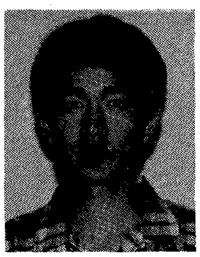

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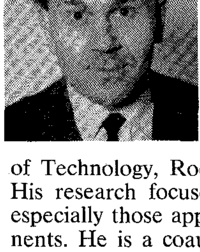
Dr. Hua has over 30 publications in the above areas.



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