Comparison of Subspace and ARX Models of a Waveguide's Terahertz Transient Response After Optimal Wavelet Filtering

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Abstract-A quasi-optical deembedding technique for characterizing waveguides is demonstrated using wide-band time-resolved terahertz spectroscopy. A transfer function representation is adopted for the description of the signal in the input and output port of the waveguides. The time-domain responses were discretized and the waveguide transfer function was obtained through a parametric approach in the z-domain after describing the system with an AutoRegressive with eXogenous input (ARX), as well as with a state-space model. Prior to the identification procedure, filtering was performed in the wavelet domain to minimize both signal distortion, as well as the noise propagating in the ARX and subspace models. The optimal filtering procedure used in the wavelet domain for the recorded time-domain signatures is described in detail. The effect of filtering prior to the identification procedures is elucidated with the aid of pole-zero diagrams. Models derived from measurements of terahertz transients in a precision WR-8 waveguide adjustable short are presented.

Index Terms—Identification algorithms, multimode waveguide characterization, signal processing, terahertz spectroscopy.

I. INTRODUCTION

THIS PAPER discusses quasi-optical broad-band measurements of the scattering parameters, attenuation coefficients, and characteristic impedances of waveguide com-

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ponents at submillimeter-wave frequencies performed using a terahertz-transient spectrometer.

Although there are inherent advantages in broad-band measurements using very short terahertz pulses as the individual reflection signatures of the device-under-test are in the time domain and, therefore, can be directly isolated and gated out, the frequency bands within which the device-under-test exhibits monomode operation must be identified. A further difficulty that arises is that most of the power output from the wide-band pulsed source occurs at frequencies (e.g., 1–3 THz) where waveguides designed for lower frequencies can support a number of modes. At frequencies where multimoded propagation can occur, a meaningful analysis is only possible if the extent to which each waveguide mode has been excited is known. Finally, the power output of the source over the single-moded bandwidth of the waveguide is often rather low, leading to a poor signal-to-noise ratio in each frequency bin of the complex insertion loss function.

In this paper, the problems associated with poor signal-to-noise ratio and multimoded propagation are addressed by placing the waveguide characterization problem in a system identification framework. We introduce autoregressive and state-space models as efficient modeling tools of linear processes such as energy dissipation in a waveguide test piece. These tools, however, require low-noise data sets, which might not be realizable within practical integration time scales. Wavelet pre-filtering is well suited to the denoising of the nonstationary data sets obtained in terahertz transient spectrometry [1]–[3]. Since the commonly used wavelets such as the db4 family are not necessarily optimal in describing an experimental data set, we introduce a process for tailoring the wavelet transform to the signal to be analyzed in order to maximize the amount of energy in the wavelet coefficients kept in the filtering process. The algorithm is then used to process terahertz transient reflection signatures from a WR-8 adjustable waveguide short. The identification models provide the weights and attenuation for each mode propagating in the structure. Pole-zero diagrams are used to show the effect of filtering on the calculated mode parameters.

II. EXPERIMENTAL SETUP

The experimental setup for generating and detecting terahertz radiation is shown in Fig. 1(a). A mode-locked Ti:sapphire



Fig. 1. (a) Experimental setup for waveguide characterization. (b) Time-domain reflection signatures for five shorted WR-8 waveguides differing in length by 1 mm. The signals are vertically offset for better visualization.

laser producing optical pulses with durations of \sim 140 fs is used to resonantly excite the lowest interband transitions of an In-GaAs emitter, which is located at the focal point of a parabolic reflector [4]-[6]. The optical pulse generates an electron-hole plasma and the accelerated carriers generate a pulse of terahertz radiation that is horizontally polarized along the direction of the surface field of the InGaAs emitter. Terahertz pulses emitted by the photoconductive element are transmitted through a vertical grid and the propagating terahertz beam, which has a Gaussian transverse amplitude distribution, is focused by the second parabolic mirror to feed at normal incidence, in a co-polar manner, the open port of the waveguide under test. The optics are designed to give a frequency-independent beam waist at the test port. The effective beam-waist size and location to optimally feed the waveguide will change in frequency, but will be independent of the location of the short in the waveguide. Thus, at any given frequency, the coupling coefficient between the beam and waveguide remains constant regardless of the position of the short. Upon 180° reflection at the back short, the modified terahertz pulse exits the waveguide and is focused to a time-gated ion-implanted silicon-on-sapphire photoconductive dipole antenna [7].

A sequence of measurements was performed on a commercially available precision WR-8 waveguide adjustable short. The time-domain interferograms corresponding to five different positions of the backshort at 1-mm spacings are shown in Fig. 1(b).

Maintaining the location of the waveguide relative to the beam and using an adjustable short to provide different test waveguide lengths ensured that a constant degree of coupling



Fig. 2. Two-channel filter bank. Blocks H and G represent a low-pass and a high-pass filter, respectively and $\downarrow 2$ denotes the operation of dyadic downsampling. The decomposition can be carried out in more resolution levels by successively splitting the low-pass channel.

was maintained between the terahertz beam and the test piece throughout the measurement sequence. The terahertz pulse produced by the photoconductive antenna has frequency components up to 3 THz.

III. WAVELET PRE-FILTERING

A. Filter Bank Formulation of the Discrete Wavelet Transform

Prior to the identification procedures, the time-domain signatures reflected by the short at the end of each waveguide length were detrended by subtracting the mean. Such a detrending was performed to avoid the need to include an offset term in the identification model [8]. Wavelet filtering was performed because it is known to produce better results than conventional filters when the signal is nonstationary [9], [10]. The wavelet filtering process was performed using a filter bank formulation of the wavelet transform [11]. Each time-domain signature was represented by a data vector \mathbf{x} of length J, where the *n*th element of \mathbf{x} , denoted by x_n , represents the measured signal at the *n*th sampling instant. The filter bank transform can be regarded as a change in variables from \Re^J to \Re^J performed according to the following operation:

$$t_k = \sum_{n=0}^{J-1} x_n v_k(n), \qquad k = 0, \dots, J-1$$
(1)

where t_k is a transformed variable and $v_k(n) \in \Re$ is a transform weight. It proves convenient to write the transform in matrix form as

$$\mathbf{t}_{1\times J} = \mathbf{x}_{1\times J} \mathbf{V}_{J\times J} \tag{2}$$

where $\mathbf{x} = [x_0x_1 \dots x_{J-1}]$ is the row vector of original variables, \mathbf{t} is the row vector of new (transformed) variables, and \mathbf{V} is the matrix of weights. Choosing \mathbf{V} to be unitary (i.e., $\mathbf{V}^T \mathbf{V} = \mathbf{I}$), the transform is said to be orthogonal and it, therefore, consists of a simple rotation in the coordinate axes (with the new axes directions determined by the columns of \mathbf{V}). As described in the wavelet literature [12]–[14], the discrete wavelet transform can be calculated in a fast manner by using a finite impulse response (FIR) filter bank structure of the form depicted in Fig. 2. It is worth noting that general *M*-channel FIR filter bank decompositions could also be employed in this context [15], but the scope of this paper will be restricted to two-channel filter banks, as in Fig. 2.

In this filter bank, the low-pass filtering result undergoes successive filtering iterations with the number of iterations k chosen by the analyst. The final result of the decomposition of data vector \mathbf{x} is a vector resulting from the concatenation of row

TABLE I CONVOLUTION PROCEDURE FOR LOW-PASS FILTERING SHOWING RESULTS BEFORE AND AFTER DYADIC DOWNSAMPLING

<i>x</i> ₀	<i>x</i> ₁	 <i>x</i> _{2<i>N</i>-1}	<i>x</i> _{2<i>N</i>}	 x_{J-1}	<i>x</i> ₀	x_1		X _{2N-2}	Before	After
h_{2N-1}	h_{2N-2}	 h_0							c_0'	
	h_{2N-1}	 h_1	h_0					c_1'	c_0	
			÷						÷	÷
				h_{2N-1}	h_{2N-2}	h_{2N-3}		••	<i>c</i> _{<i>J</i>-2} ′	
					<i>h</i> _{2<i>N</i>-1}	<i>h</i> _{2<i>N</i>-2}		h_0	СЈ-1'	<i>CJ</i> /2–1

vectors $\mathbf{c}(k)$ and $\mathbf{d}(k)$ (termed approximation and detail coefficients at the *k*th scale level, respectively) in the following manner:

$$\mathbf{t} = [\mathbf{c}(k)|\mathbf{d}(k)|\mathbf{d}(k-1)|\dots|\mathbf{d}(1)].$$
 (3)

with coefficients in larger scales (e.g., $\mathbf{d}(k), \mathbf{d}(k-1), \mathbf{d}(k-2), \ldots$) associated with broad features in the data vector, and coefficients in smaller scales (e.g., $\mathbf{d}(1), \mathbf{d}(2), \mathbf{d}(3), \ldots$) associated with narrower features such as sharp peaks.

Let $\{h_0, h_1, \ldots, h_{2N-1}\}$ and $\{g_0, g_1, \ldots, g_{2N-1}\}$ be the impulse responses of the low-pass and high-pass filters, respectively. Assuming that filtering is carried out by circular convolution, the procedure for generating the approximation coefficients from the data vector **x** is illustrated in Table I. The convolution consists of flipping the filtering sequence and moving it alongside the data vector. For each position of the filtering sequence with respect to the data vector, the scalar product of the two is calculated (with missing points in the filtering sequence replaced with zeros). For instance, if N = 2, the third row in Table I shows that $c_1' = x_1h_3 + x_2h_2 + x_3h_1 + x_4h_0$. Dyadic downsampling is then performed to c_{2i}' to generate coefficients c_i . The detail coefficients d_i are obtained in a similar manner by using the high-pass filtering sequence.

If the approximation \mathbf{c} and detail \mathbf{d} coefficients are stacked in vector $\mathbf{t} = [\mathbf{c} | \mathbf{d}]$, the wavelet transform can be expressed in the matrix form (2) with the transformation matrix given by (4), shown at the bottom of this page. A requirement for the transform to be orthogonal (i.e., $\mathbf{V}^T \mathbf{V} = \mathbf{I}$) is that the sum of the squares of each column must be equal to one and the scalar product of different columns must be equal to zero. Therefore, for a filter bank that utilizes low-pass and high-pass filters, the following conditions ensure orthogonality of the transform so that no information is lost in the decomposition process [11]

$$\sum_{n=0}^{2N-1-2l} h_n h_{n+2l} = \begin{cases} 1, & l=0\\ 0, & 0

$$g_n = (-1)^{n+1} h_{2N-1-n},$$
(5a)$$

$$n = 0, 1, \dots, 2N - 1.$$
 (5b)

If the filtering sequences satisfy these conditions, the entire structure is termed a quadrature-mirror filter (QMF) bank [12]. A QMF bank is said to enjoy a perfect reconstruction (PR) property because \mathbf{x} can be reconstructed from \mathbf{t} , which means that there is no loss of information in the decomposition process. In fact, from the relation $\mathbf{t} = \mathbf{xV}$, it follows that $\mathbf{tV}^T = \mathbf{xVV}^T$ and $\mathbf{x} = \mathbf{tV}^T$, due to the orthogonality of the transform. At this point, it should be noticed that other nonorthogonal transforms can also enjoy a PR property, provided that they are associated to a nonsingular matrix \mathbf{V} . However, the analysis in this paper will be restricted to orthogonal transforms.

Filtering in the wavelet domain consists of replacing some of the elements of t by zero so that a new vector \overline{t} is produced and then applying the inverse transform.

B. Filtering Using Adaptively Chosen Wavelets

One limitation of the procedure described for the filtering of the time-domain signatures in the wavelet domain is that the wavelets must be chosen *a priori* and are not adapted to optimally describe the experimental data set. Optimizing the transform to maximize its compression ability and, therefore, its efficiency is normally achieved by optimizing the QMF bank. The QMF bank can be described by a set of parameters that can be adjusted by any algorithm for unconstrained optimization to maximize the compression ability of the transform.

The parametrization of PR FIR filter banks proposed by Vaidyanathan as adapted by Sherlock and Monro [13] to parametrize orthonormal wavelets of arbitrary compact support

	$\begin{bmatrix} 0\\ h_{2N-1}\\ h_{2N-2}\\ h_{2N-3} \end{bmatrix}$	$\begin{array}{c} 0 \\ 0 \\ 0 \\ 0 \\ h_{2N-1} \end{array}$	•••• ••• •••	$h_{2N-4} \\ h_{2N-5} \\ h_{2N-6} \\ h_{2N-7}$	$h_{2N-2} \\ h_{2N-3} \\ h_{2N-4} \\ h_{2N-5}$	$\begin{array}{c} 0\\ g_{2N-1}\\ g_{2N-2}\\ g_{2N-3} \end{array}$	$\begin{array}{c} 0 \\ 0 \\ 0 \\ g_{2N-1} \end{array}$	•••• ••• •••	g_{2N-4} g_{2N-5} g_{2N-6} g_{2N-7}	g_{2N-2} g_{2N-3} g_{2N-4} g_{2N-5}
		÷	÷	÷	:	÷	÷	÷	÷	÷
$\mathbf{V} =$	h_0	h_2	•••	0	0	g_0	g_2	•••	0	0
	0	h_1	• • •	0	0	0	g_1	• • •	0	0
	0	h_0	• • •	0	0	0	g_0	• • •	0	0
	:	:	:	÷	÷	÷	:	÷	÷	:
	0	0	• • •	h_{2N-2}	0	0	0	•••	g_{2N-2}	0
	0	0		h_{2N-3}	h_{2N-1}	0	0		q_{2N-3}	q_{2N-1}



Fig. 3. Procedure for parameterizing wavelet filter banks by N angles.

may be used for this purpose. For a filter bank of the form shown in Fig. 2 where the conditions in (5a) and (5b) are satisfied, the transfer function of the low-pass filter in the z-domain can be written as

$$H^{(N)}(z) = \sum_{n=0}^{2N-1} h_n^{(N)} z^{-n} = H_0^{(N)}(z^2) + z^{-1} H_1^{(N)}(z^2)$$
(6)

where superscript (N) denotes that the filtering sequences have length $2N \cdot H_0^{(N)}(z)$ and $H_1^{(N)}(z)$, termed polyphasic components of $H^{(N)}(z)$, are given by

$$H_0^{(N)}(z) = \sum_{i=0}^{N-1} h_{2i}^{(N)} z^{-i}$$
(7a)

$$H_1^{(N)}(z) = \sum_{i=0}^{N-1} h_{2i+1}^{(N)} z^{-i}.$$
 (7b)

Defining the polyphasic components $G_0^{(N)}(z)$ and $G_1^{(N)}(z)$ of the high-pass filter $G^{(N)}(z)$ in a similar manner, a matrix $F^{(N)}(z)$ may be defined as follows:

$$F^{(N)}(z) = \begin{bmatrix} H_0^{(N)}(z) & H_1^{(N)}(z) \\ G_0^{(N)}(z) & G_1^{(N)}(z) \end{bmatrix}.$$
 (8)

It can be shown [11], [12] that $F^{(N)}(z)$ can be factorized as

$$F^{(N)}(z) = \begin{bmatrix} C_0 & S_0 \\ -S_0 & C_0 \end{bmatrix} \prod_{i=1}^{N-1} \begin{bmatrix} 1 & 0 \\ 0 & z^{-1} \end{bmatrix} \begin{bmatrix} C_i & S_i \\ -S_i & C_i \end{bmatrix}$$
(9)

where each pair of parameters (C_i, S_i) are related to a common angular parameter θ_i as $C_i = \cos(\theta_i)$ and $S_i = \sin(\theta_i), i = 0, 1, \dots, N - 1$. It follows that the filters can be completely parameterized by N angles $\theta_0, \theta_1, \dots, \theta_{N-1}$, which can assume any value in the set of real numbers, as shown in Fig. 3.

The weights of the low-pass filter can be easily recovered from a set of angles $\{\theta_i\}$ by using the following recursive formula [11], [12]:

$$F^{(k+1)}(z) = F^{(k)}(z) \cdot \begin{bmatrix} 1 & 0\\ 0 & z^{-1} \end{bmatrix} \begin{bmatrix} C_k & S_k\\ -S_k & C_k \end{bmatrix}$$
(10)

for k = 1, 2, ..., N - 1 with

$$F^{(1)}(z) = \begin{bmatrix} C_0 & S_0 \\ -S_0 & C_0 \end{bmatrix}.$$
 (11)

Equation (10) with the initial condition of (11) provide a way to obtain the weights $\{h_i^{(k+1)}\}$ for a filter of length 2(k+1) from the weights $\{h_i^{(k)}\}$ for a filter of length 2k. To do that, one starts by writing, from (8) and (10)

$$H_0^{(k+1)}(z) = H_0^{(k)}(z)C_k - z^{-1}H_1^{(k)}(z)S_k,$$

$$k = 1, 2, \dots, N-1 \qquad (12a)$$

$$H_1^{(k+1)}(z) = H_0^{(k)}(z)S_k + z^{-1}H_1^{(k)}(z)C_k,$$

$$k = 1, 2, \dots, N-1 \qquad (12b)$$

with $H_0^{(1)}(z) = C_0$ and $H_1^{(1)}(z) = S_0$. A recursive formula for the generation of low-pass filter weights with even indexes $\{h_{2i}\}$ can then be stated by using the definitions in (7a) and (7b) to expand (12a) as

$$\underbrace{H_{0}^{(k+1)}(z)}_{i=0} \underbrace{H_{0}^{(k+1)}z^{-i}}_{i=0} = \underbrace{\left(\sum_{i=0}^{(k-1)}h_{2i}^{(k)}z^{-1}\right)}_{i=0}C_{k} - z^{-1}\underbrace{\left(\sum_{i=0}^{(k-1)}h_{2i+1}^{(k)}z^{-1}\right)}_{i=0}S_{k} \Rightarrow \\
\sum_{i=0}^{k}h_{2i}^{(k+1)}z^{-i} = C_{k}h_{0}^{(k)} + \sum_{i=1}^{k-1}\left(C_{k}h_{2i}^{(k)} - S_{k}h_{2i-1}^{(k)}\right)z^{-i} - S_{k}h_{2k-1}^{(k)}z^{-k}$$
(13)

for k = 1, 2, ..., N - 1, with $h_0^{(1)} = C_0$ and $h_1^{(1)} = S_0$. From the identity of terms with the same power of z in the last line of (13), it follows that

$$\begin{cases} h_0^{(k+1)} = C_k h_0^{(k)} \\ h_{2i}^{(k+1)} = C_k h_{2i}^{(k)} - S_k h_{2i-1}^{(k)} , & i = 1, 2, \dots, k-1 \\ h_{2k}^{(k+1)} = -S_k h_{2k-1}^{(k)} \end{cases}$$
(14a)

for $k = 1, 2, \dots, N - 1$.

A similar formula can be stated for the low-pass filter weights with odd indexes by expanding (12b) as

$$\begin{cases} h_1^{(k+1)} = S_k h_0^{(k)} \\ h_{2i+1}^{(k+1)} = S_k h_{2i}^{(k)} + C_k h_{2i-1}^{(k)} , & i = 1, 2, \dots, k-1 \\ h_{2k+1}^{(k+1)} = C_k h_{2k-1}^{(k)} \end{cases}$$
(14b)

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for k = 1, 2, ..., N - 1. After obtaining the low-pass filtering sequence, as explained above, the high-pass filtering sequence can be obtained by using (5b).

Condition (5a) states that the 2N weights $\{h_n\}$ of the lowpass filter are subject to N restrictions. Thus, there are N degrees of freedom that can be used to optimize the filter bank according to some performance criterion. It is worth noting that since the restrictions are nonlinear and may define a nonconvex search space, the optimization task is not trivial. We circumvent this difficulty by using the lattice structure for the filter bank, which is parameterized by N angles $\{\theta_0, \theta_1, \dots, \theta_{N-1}\}$ that can assume any real value, as shown in Fig. 3.

The problem then becomes one of unconstrained optimization in \mathbb{R}^N . The optimal filtering procedure employed in this study was aimed at maximizing the amount of energy in the wavelet coefficients kept in the thresholding process. The optimization consisted of maximizing an objective function $F(\theta)$: $\mathbb{R}^N \to \mathbb{R}$ defined as

$$F(\theta) = \sum_{k \in \mathbf{I}} p^2(k; \theta)$$
(15)

where θ is the vector of N angles that parameterize the filter bank, as explained above, $p(k;\theta)$ is the kth wavelet coefficient resulting from the signal decomposition, and I is the index set of the coefficients kept in the thresholding process. It is worth noting that I is defined on the basis of the magnitude of the wavelet coefficients before the optimization. The flexible polyhedron algorithm available in the MATLAB Optimization Toolbox was employed to search for the optimum θ using the parameters associated with the db4 wavelet as a starting point. The optimization was carried out separately for each signal.

It is possible that the objective function (15) may have local maxima different from the global maximum. In that case, the flexible polyhedron algorithm will tend to converge to the closest local maximum. However, even if the global maximum is not attained, an improvement over the original wavelet transform may still be obtained.

For comparative purposes, a nonoptimized filtering procedure, where each time-domain signal \mathbf{x} to be filtered was decomposed using a db4 wavelet filter bank with two resolution levels [16], was also performed. Hard thresholding was employed, with all wavelet coefficients with magnitude smaller than a certain fraction of the largest coefficient being replaced with zeros. The inverse transform was then applied to obtain the filtered signal shown in Fig. 4.

As can be seen from Fig. 4, a direct comparison between filtered and optimally filtered waveforms is almost impossible in the time-domain plots. Plotting, however, pole-zero diagrams after performing system identification for the terahertz-transient excited waveguide sections makes the effect of filtering more apparent.

IV. SYSTEM IDENTIFICATION

A. ARX Model

This section describes the process of obtaining the frequency response of a waveguide test piece by identifying a parametric



Fig. 4. Unfiltered, filtered, and optimally filtered time-domain responses for the 2-mm-long WR-8 waveguide section.

model from the reflection signatures for two different waveguide lengths. Let u(t) and y(t) be the reflection signatures for the shorter (length Ld, where d is a length unit) and longer (length (L+1)d) waveguides, respectively. Ratioing the spectra $U(\omega)$ and $Y(\omega)$ yields the effective complex insertion loss function $H(\omega)$ for a unit length waveguide [17] that is given from $H(\omega) = Y(\omega)/U(\omega)$, where $H(\omega)$ can be regarded as the frequency response of a linear system, which produces an output y(t) for an input u(t).

Instead of using the ratioing procedure, which is very sensitive to measurement noise, system identification techniques can be employed to obtain the complex insertion loss. Since this approach is novel in the context of waveguide characterization, results from two different identification techniques will be presented to demonstrate this concept. Firstly, $H(\omega)$ can be obtained directly through a parametric approach [8], [18] in which the unit length waveguide is described by an AutoRegressive with eXogenous input (ARX) model of the form

$$y[k] + a_1 y[k-1] + \dots + a_{n_a} y[k-n_a] = b_1 u[k-n_k] + b_2 u[k-n_k-1] + \dots + b_{n_b} u[k-n_k-n_b+1] + \varepsilon[k]$$
(16)

where $y[\mathbf{k}]$ is the response measured at time kT_s (sampling interval $T_s = 27$ femtoseconds), $\{a_1, \ldots, a_{n_a}, b_1, \ldots, b_{n_b}\}$ are model coefficients, n_a is the number of poles (equivalent to the system order), $(n_b - 1)$ is the number of zeros, n_k is a pure time delay, and $\varepsilon[\mathbf{k}]$ is a residual. $H(\omega)$ can be discretized as a z-domain transfer function $H_d(z) = Y(z)/U(z)$

$$H_{d}(z) = z^{-n_{k}} \frac{b_{1} + b_{2} z^{-1} + \dots + b_{n_{b}} z^{-n_{b}+1}}{1 + a_{1} z^{-1} + \dots + a_{n_{a}} z^{-n_{a}}}$$

= $z^{-n_{k}} \frac{N(z)}{D(z)} = z^{-n_{k}} \sum_{i=1}^{n+m} w_{i} \frac{1 + f_{i} z^{-1}}{1 + c_{i1} z^{-1} + c_{i2} z^{-2}}$ (17)

by making $z = \exp(j\omega T_s)$. The roots of D(z) (poles of $H_d(z)$) are associated with the TE_{mn} modes of the waveguide (m + n modes in total). It is worth noting that, since the modes have an oscillatory behavior, it is reasonable to model them with secondorder terms [19], as in the right-hand side of (17).

The roots of N(z) (zeros of $H_d(z)$) are associated with the weights of each mode in the overall response of the system and $n + m = n_a/2$ (since the order of the denominator of $H_d(z)$ is n_a and there are n + m terms in the expansion, each one with a second-order denominator). Each partial fraction describes a propagating mode inside the waveguide. The *i*th mode will have a weight, resonant frequency of oscillation, and damping ratio associated with the roots of the polynomial in the *i*th denominator. It is worth noting that this is the minimal representation needed to describe a mode since the first-order term in the numerator of the partial fraction will account for the cut-on effect in the frequency domain, whereas the second-order term in the denominator will ensure that the response is attenuated at high frequencies. For modeling parsimony, no extra zeros or poles are incorporated in the description of each mode. Parameters a_i and b_i can be obtained by minimizing the mean square value of the residual $\varepsilon[\mathbf{k}]$ over the time horizon used for identification using a least squares procedure.

When choosing the model order (n_a) , one must make an assumption on the number of modes that are propagating in the structure. The value of n_k can be accurately estimated after observing the delay in the time-domain signatures recorded by the spectrometer from

$$\varphi(\omega) = \tan^{-1} \left[\operatorname{Im}(H(\omega)) / \operatorname{Re}(H(\omega)) \right]$$
 (18)

where $\varphi(\omega) = z^{-n_k}$ [17]. The choice of n_b can be made in an empirical way from cross-correlation of the outputs with the residuals. Other methods such as spectral analysis, analysis of the information matrix, Akaike's final error prediction criterion, or Akaike's information theoretic criterion [8] are also appropriate. The System Identification Toolbox of MATLAB chooses n_b on the basis of parsimony arguments in order to balance the accuracy and complexity (number of parameters to be identified) of the model.

It is worth noting that the use of physical insight in the identification procedure reduces the number of design choices that need to be done on the basis of the data themselves. If the entire definition of the model structure (including the choice of n_a and n_k) were data driven, the identification procedure would be more sensitive to artifacts and noise.

The experimental time-domain reflection signatures from the backshorts must be aligned with respect to each other before an identification model is applied to the discretized time-domain signatures. Using the known dimensions of the rectangular section of the waveguide, the cut-on frequency, guide wavelength, and theoretical phase delay at each frequency for each mode for all TE_{mn} modes propagating in the structure can be calculated [20]. To calculate the overall phase delay, vectorial addition must be performed for all the propagating modes inside the waveguide. Since each mode is attenuated to a different degree, its contribution must be weighted by a weight w_i and a factor with an attenuation constant α_i

$$H(\omega) = \sum_{i=1}^{m+n} w_i \left(e^{-\alpha_i(\omega)d/2} e^{j\varphi_i(\omega)} \right).$$
(19)

By matching the theoretical calculations for the phase delay in WR-8 waveguide samples to the recorded delay by the interferometer, it was concluded that most of the power is distributed among two modes and, thus, a fourth-order model should be adopted for the backshort experiment. In addition, an examination of the phase velocity of the received signal shows that the propagation of terahertz pulses inside the waveguide is almost dispersion free within the bandwidth of the excitation pulse. Such preferential coupling of the pulse energy to the lower order modes of the structure and the almost dispersion-free propagation observed are similar to the ones obtained by Grischkowsky's group [21] where they also used broad-band excitation in their experimental procedure to characterize plastic ribbon terahertz waveguides.

B. State-Space Modeling

An alternative way to obtain $H_d(z)$ would be through a state-space modeling approach. Discrete time state-space models represent difference equations, as ARX models do, but are rearranged such that only one delay is used in the expression. This leads to the introduction of extra variables, known as state variables, which are not measured, but can, under certain conditions, be reconstructed from input–output data. For the waveguide characterization problem, a single-input single-output state space model may be identified directly from the data

$$\mathbf{x}[\mathbf{k}+1] = \mathbf{A}\mathbf{x}[\mathbf{k}] + \mathbf{B}u[\mathbf{k}] + \mathbf{K}e[\mathbf{k}]$$
$$y[\mathbf{k}] = \mathbf{C}\mathbf{x}[\mathbf{k}] + Du[\mathbf{k}] + e[\mathbf{k}]$$
(20)

where $\mathbf{x}[\mathbf{k}]$ is an *n*-dimensional state vector, u[k] is a deterministic input, y[k] is the measured output, e[k] is the output residual, D is a scalar parameter, and $\mathbf{A}, \mathbf{B}, \mathbf{C}$, and \mathbf{K} are matrices of the appropriate dimensions. A sound method for estimating the state-space model parameters using input–output data is the subspace approach [22], [23]. A more detailed description of the subspace algorithm as applied to optical systems is provided elsewhere [24]. A common implementation is the N4SID algorithm, which is available as part of MATLAB's System Identification Toolbox.

There is an equivalence between the state-space representation, and the input-output representation. Taking the z-transform of the state-space model and eliminating the state variables, it follows that

$$H_d(z) = \frac{Y(z)}{U(z)} = \mathbf{C}(z\mathbf{I} - \mathbf{A})^{-1}\mathbf{B} + D.$$
 (21)

Writing both ARX, as well as state-space models in an input/output form, a direct comparison of the two methods is possible. However, the state-space representation can be deemed more appropriate than the ARX representation in optimal estimation and optimal control frameworks, as discussed in [24].

Instead of relying on matching the observed phase delay in the interferometer with theoretical models of phase delay that take into account the waveguide dimensions to define the model order as the ARX model does, the subspace algorithm avoids



Fig. 5. Measured (solid line) and simulated (dotted line) time-domain responses using the: (a) ARX model and (b) state-space model (modeling data, d_3/d_2 case).

over-parametrization by observing the singular values plot resulting from the singular value decomposition of a Hankel matrix built from the input–output data [22]. The optimal model order is selected in such a way that the singular values for higher orders are comparatively smaller.

V. RESULTS AND DISCUSSION

The System Identification Toolbox of MATLAB v.6.1 was used to obtain models in the form of (17) and (21). Fig. 5 compares the measured and simulated time domain responses for the modeling data, which consisted of the signals for two- and three-waveguide-length units. Both the ARX and state-space models reproduced 87% of the y-variance.

Readers familiarized with control systems theory will notice that the time-domain plots of Fig. 5 resemble impulse responses of nonminimum phase systems [19]. However, one should bear in mind that those are the responses of the entire structure (including the coupling of the test piece), whereas the models considered in this paper are developed for the waveguide only (the coupling effects being ratioed out). Thus, if the nonminimum phase behavior is actually caused by coupling effects, it should not be reflected in the identified models.

For validation, the models were applied to an independent data set consisting of the responses for four- and five-wave-guide-length units. The results presented in Fig. 6 show that the model predictions are in good agreement with the measured responses. It is worth noting that the accuracy is smaller in the validation set, as compared to the modeling set (Fig. 5). Such a finding was to be expected because the validation data are not included in the model-building process. They are used instead to check the generalization ability of the model, i.e., its ability



Fig. 6. Measured (solid line) and simulated (dotted line) time-domain responses using the: (a) ARX model and (b) state-space model (validation data, d_5/d_4 case).



Fig. 7. Residual statistics for the state-space model with the d_5/d_4 validation data. (a) Autocorrelation values of the modeling residual e[k]. (b) Cross-correlation between the modeling residual and input signal.

to predict the behavior of the system in a situation that is not exactly equal to the one considered in the modeling phase [8].

Fig. 7(a) shows the autocorrelation values of the state space model residual $\sum_{\mathbf{k}} e[\mathbf{k}]e[\mathbf{k}-p]$, where p is a time lag between two points of the state-space model residual sequence. Fig. 7(b) shows the cross-correlation $\sum_{\mathbf{k}} e[\mathbf{k}]u[\mathbf{k}-p]$ between the modeling residual and input signal for different time lags p.

Fig. 8(a) and (b) shows similar residual correlation plots for the ARX model. Ideally, the autocorrelation should be zero for



Fig. 8. Residual statistics for the ARX model and the d_5/d_4 validation data. (a) Autocorrelation values of the modeling residual ε [k]. (b) Cross-correlation between the modeling residual and input signal.



Fig. 9. Calculated transmission coefficient |H(jw)| of a unit length waveguide for the backshort.

time lags different from zero (white noise) and there should be no cross-correlation between the residual and input since any correlation would indicate that part of the residual could be predicted from the input. Since the autocorrelation of the residuals is small, the assumption of whiteness in the noise can be assumed to be valid.

Fig. 9 shows the insertion loss of a unit length waveguide (as calculated both by the ARX and state-space models) after ratioing the spectra obtained with the backshort adjusted to 3 mm with that at 2 mm (denoted as the d_3/d_2 case), which is, in fact, the frequency response of the identified models. An effective propagation constant for the multimode propagation can, therefore, be calculated [17]. The observed increased insertion loss at higher frequencies is attributed to the variation of the conductivity of gold with frequency.

The pole-zero charts in Figs. 10 and 11 provide details of the resulting ARX model and state-space model (when converted to an input–output model). The dashed isoclines provide values of



Fig. 10. Pole-zero diagram for the state-space model showing the confidence regions using one standard deviation.



Fig. 11. Pole-zero diagram for the ARX model showing the confidence regions using one standard deviation.

natural frequencies ω_n and damping ratios ζ for the propagating modes. The isoclines are obtained from

$$z = \exp\left[T_s(-\zeta\omega_n \pm j\omega_n\sqrt{(1-\zeta^2)})\right]$$
(22)

where $0 \leq \zeta \leq 1, \omega_n \geq 0$ by fixing ζ and varying ω_n or by fixing ω_n and varying ζ .

In these plots, the confidence locus of the zeros is much larger than the confidence locus of the poles. This was to be expected since it is easier to estimate poles than zeros, as they are associated with the exponential decay and frequency of oscillation of the modes, whereas the zeros are associated with the weights of each mode, which are intuitively more difficult to obtain. In addition, if more modes were assumed in the model, the radii of the confidence zones of the poles and zeros would increase since the number of unknowns would be larger for the same number of observations, i.e., the confidence levels in the model decrease significantly when the model is over-parametrized.



Fig. 12. (a) Zero- and (b) pole-charts of original (\bullet), wavelet-filtered (\blacksquare), optimal wavelet-filtered (\diamondsuit) responses assuming a fourth-order ARX model.

Since the autocorrelation function of the residual for both the ARX and state-space models is almost within the one standard error bounds, it can be concluded that the models have been appropriately chosen. Some of the zeros calculated using the state-space model are outside the unit circle, which indicates that this model exhibits nonminimum phase behavior, which is not the case with the identified ARX model.

Fig. 12 provides pole and zero charts where an ARX identification model is used to obtain $H_d(z)$ after assuming two modes propagating in the structure. It was observed that the wavelet filtering process resulted in a zero that was outside the unit circle being brought to the inside. This is in accordance with our expectations since there does not seem to be a physical reason for a zero being outside the unit circle.

As stated earlier, the poles are directly related to the mode properties. Filtering increased the estimated damping in the higher frequency mode. The frequency itself remained practically unchanged. Furthermore, in the lower frequency mode, the filtering decreased the estimated frequency, whereas the damping remained unchanged.

A premise of this study is that the filtering procedure improves the signal-to-noise ratio of the measurements and, consequently, the accuracy of the identification results. However, one should bear in mind that if the filtering is too aggressive (in the sense of more wavelet coefficients being replaced with zeros), there is a risk that the identified model will be related to the filter dynamics rather than to the waveguide response. At this point, it is worth noting that the proposed wavelet optimization procedure may actually alleviate such a risk. In fact, by tailoring the wavelet transform to the denoising task, it is possible to minimize the distortion effects on the signal caused by removing the noise.

VI. CONCLUSION

Both ARX and state-space models are efficient modeling tools for linear processes such as energy dissipation in a waveguide test piece. These tools, however, require low-noise data sets, which might not be realizable within practical integration time scales. Wavelet pre-filtering is well suited to the denoising of the nonstationary data sets obtained in terahertz transient spectrometry.

One limitation of commonly used wavelets such as the db4 family is that the wavelets must be chosen *a priori* and are not adapted to optimally describe the experimental data set. We, therefore, introduced a QMF bank to optimize the compression ability of the transform before performing any thresholding in the wavelet domain. The optimal filtering procedure employed in this study was aimed at maximizing the amount of energy in the wavelet coefficients kept in the thresholding process.

Although pre-filtering of the time-domain signatures with a db4 and an optimal filter did not produce traces that would be classified as different by eye inspection, these different filtering procedures had a marked effect on the pole-zero diagrams obtained after ARX and subspace models were fitted to the data. This justifies their use in waveguide mode identification from broad-band experiments.

An extension of the subspace algorithm to bilinear systems would be required to take into account the nonlinearities of the cut-on conditions in the modes and account for evanescent fields for a more accurate model. In this case, the complex values arising in the simulations would be treated separately in matrix form in a similar manner to real valued data. The good fit, however, of both models to the time-domain data sets indicated that this was not necessary in our current analysis. In fact, since the scope of this study was restricted to the modeling of the propagating modes, the cut-on conditions and the evanescent fields were not modeled or considered in the validation of the obtained models. Different experimental protocols may be designed to focus the study on such a phenomena. In that case, the nonlinear effects would have to be properly taken into account.

Finally, it is worth noting that, although the proposed analysis in the wavelet domain has been placed within a filtering framework, there have been suggestions [25] that such transforms may also be used directly for waveguide mode identification.

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Rainer Martini, photograph and biography not available at time of publication.

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