

Short Papers

A Proposed Microwave and Millimeter-Wave Spectrum Analyzer

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Abstract—A system is proposed which performs real-time spectrum analysis of large-bandwidth radio frequency signals. The apparatus allows the simultaneous monitoring of all frequencies within the band of interest, and operates in the presence of multiple frequencies. The approach also overcomes the slowness and ambiguity of scanning methods.

I. INTRODUCTION

This paper concerns itself with the problem of performing spectrum analysis over more than a few gigahertz of bandwidth when continuous coverage of multifrequency inputs is required. Specifically, a method is proposed which may accomplish this goal. A review of the primary existing methods suggests that such a capability is hard to obtain.

The most common method of spectrum analysis employs a superheterodyne receiver which is swept over the frequency band of interest (see [1, ch. 5]). This method accommodates a multitude of frequencies in the signal, allowing each to be recognized independently. The frequencies are studied sequentially, so that pulsed signals are easily missed, as such signals may be off when the receiver visits their frequencies.

Instantaneous frequency measurement (IFM) devices measure relative phase between the signal and a delayed version of the signal. Such systems cover all frequencies simultaneously, but can only tolerate one frequency in the signal at any given time. Various improvements made to this basic approach have allowed only modest input diversity. The early history of IFM techniques may be found in [2]; for a review of current IFM approaches see [1, ch. 6].

Several acoustic methods offer simultaneous coverage of the entire band and can accommodate multiple frequencies. The most popular acoustic approach is to form a compressive receiver [1], [3] from dispersive acoustic delay lines [4], [5]. Nonlinearly tapped acoustic lines have also been used [6]. These approaches require electronic processing of the filtered output at rates exceeding the signal bandwidth—a fact which makes large bandwidths very difficult to obtain.

Acoustooptic methods [7]–[11], which exploit light propagation to implement a Fourier transform, also allow multiple frequencies and cover the entire band at once. They too are constrained in bandwidth, but by their input transducers. Practical devices are limited to a few gigahertz.

A device which overcomes the objections cited above is the Fourier spectrometer, which mechanically scans out the autocorrelation function of the input signal. The autocorrelation function is then Fourier transformed to produce the power spectrum.

The Fourier spectrometer consists of a radio frequency Michelson interferometer in which one reflector is movable, and a detector which is placed at the output of the interferometer. For any position of the movable reflector, the detector responds to the signal plus a delayed replica of the signal. The detector's slow, square-law response causes the detector output to be a measure of the signal's autocorrelation at the delay between the two signals. By moving the reflector, the delay is varied and the autocorrelation is thereby scanned out. The Fourier spectrometer usually uses free-space propagation rather than guided propagation. The Fourier spectrometer approach suffers from two major drawbacks. First, it is mechanically scanned, making it slow and prone to mechanical failure. Second, the scanning introduces an ambiguity which causes signal modulation to manifest itself as spurious frequencies in the output. The Fourier spectrometer was originally developed for the spectral analysis of optical disturbances, and this continues to be its primary area of application; the interested reader may consult [12] and [13] for an in-depth discussion of the Fourier spectrometer as an optical device.

This paper proposes what is thought to be a new system which obtains the power spectrum of a high-bandwidth signal. This system, like the Fourier spectrometer, calculates the signal power spectrum from the signal's autocorrelation function, and reaps the advantages of parallel computation of all spectral components, accommodation of multifrequency inputs, and ability to operate on large bandwidths. The main difference of the current approach is that it efficiently calculates all of the autocorrelation values in parallel, thereby allowing rapid computation and avoiding the gross ambiguities introduced by scanning.

II. APPARATUS DESCRIPTION

Fig. 1 illustrates one implementation of the system. The RF input signal passes through an isolator before entering a detector/waveguide assembly (described below). The isolator prevents reflected RF power produced by the assembly from passing back to the source.

Fig. 2 shows a detail of the detector/waveguide assembly shown in Fig. 1. The RF signal enters the adapter and passes into one end of a rectangular waveguide. The opposing end of the waveguide is terminated by a shorting plate. A signal propagates down the waveguide, strikes the shorting plate, is completely reflected, and propagates back toward the source end of the waveguide. Upon entering the waveguide adapter, the reflected wave passes into the cable and is dissipated in the isolator. (An alternative arrangement is to use a waveguide isolator instead of a coaxial isolator.)

Sensitive detectors are placed along the length of the waveguide at regular intervals in a line parallel to the direction of radio frequency propagation. These detectors measure the RF power in the waveguide by removing an infinitesimal amount of said power.

There are present, at each point in the waveguide, two contributions to the electromagnetic field corresponding to the forward- and backward-propagating waves, respectively. These contributions are both replicas of the input signal, but are displaced

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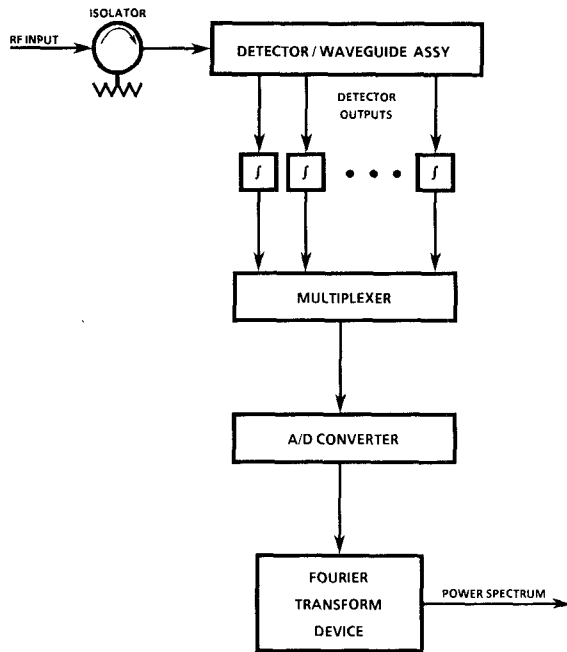


Fig. 1. System schematic. The detector/waveguide assembly is pictured in Fig. 2

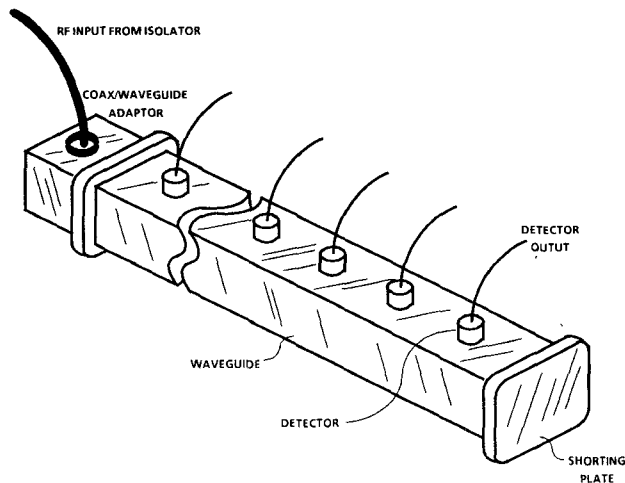


Fig. 2. Detector/waveguide assembly of Fig. 1.

in time with respect to each other. Thus, the averaged output of any of the detectors produces a term proportional to the signal's autocorrelation evaluated at the delay between the two contributions seen at the detector site.

The n detector outputs enter integrators (Fig. 1). The outputs of these integrators form the input of a multiplexer, which produces a serial stream of analog values. This stream is formed by sequentially selecting each detector output in turn and sending its value to the output. The multiplexed samples are then digitized.

A Fourier transform device collects n input samples, one from each detector output, and performs a discrete Fourier transform on them. The resulting output stream represents the power spectrum of the signal in terms of spatial frequency in the waveguide assembly.

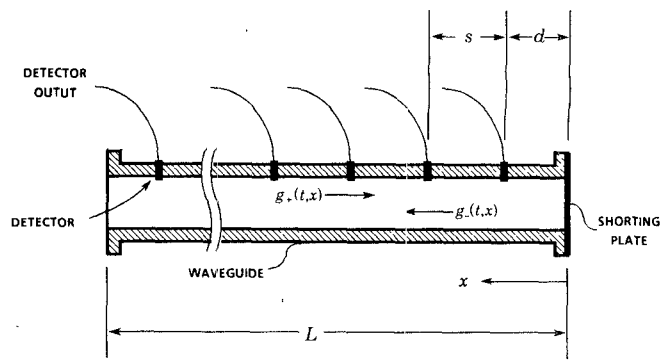


Fig. 3. Detector/waveguide assembly of Fig. 1. The coax/waveguide adapter is not shown

III. THEORY OF OPERATION

The claims made above will now be examined in more detail with reference to Fig. 3.

Let $g(t)$ denote the signal as introduced to the waveguide by the adapter. This signal may be expressed in terms of its Fourier transform $G(f)$:

$$g(t) = \int_{-\infty}^{\infty} G(f) e^{i2\pi ft} df.$$

It will be supposed that the frequencies of the signal are confined to the interval $[f_L, f_H]$ and that these frequencies propagate in the lowest order mode of the waveguide. For computational ease, it will also be supposed that the signal $g(t)$ is zero outside of the observation interval $[0, T]$. Let γ be the guide attenuation constant (also including loss to detectors), let L be the length of the waveguide, and let x be the position along the waveguide from the shorted end. In addition, let c be the speed of light, let f_c be the guide cutoff frequency, and let

$$\beta(f) = [1 - (f_c/f)^2]^{1/2}$$

be the phase constant dispersion for frequency f . The forward-propagating wave in the guide can be written as

$$g_+(t, x) = e^{-\gamma(L-x)} \int_{-\infty}^{\infty} G(f) e^{i2\pi f[t - (L-x)\beta(f)/c]} df.$$

Similarly, the backward-propagating wave may be expressed as

$$g_-(t, x) = -e^{-\gamma(L+x)} \int_{-\infty}^{\infty} G(f) e^{i2\pi f[t - (L+x)\beta(f)/c]} df.$$

(The minus sign in front of the exponential is a result of reflection by a short.) As a result of these two contributions, the instantaneous power at time t and position x is

$$P(t, x) = |g_+(t, x) + g_-(t, x)|^2 = |g_+(t, x)|^2 + |g_-(t, x)|^2 + 2\text{Re}\{g_+(t, x)g_-^*(t, x)\}.$$

On the waveguide are placed n detectors, with the i th detector located at $x = is + d$, $i = 0, 1, 2, \dots, n-1$. Each detector's output is proportional to the radio frequency power at the location of the detector. The i th such output is then integrated to produce the result

$$I_i = \int_0^T P(t, is + d) dt.$$

(A proportionality constant has been dropped.) This integration also includes the temporal response of the detector. Let

$$H(f) = \int_0^T e^{-i2\pi ft} dt$$

denote the Fourier transform of the temporal observation window.

The integrator outputs $\{I_i\}$ may be considered the sum of two contributions. The first term, which is undesired, is due to the self-square terms in P and is given by

$$\begin{aligned} I_i^s = & e^{-2\gamma(L-is-d)} \int_{-\infty}^{\infty} \int_{-\infty}^{\infty} G(f_1) G^*(f_2) \\ & \cdot H^*(f_1 - f_2) e^{-j2\pi[f_1\beta(f_1) - f_2\beta(f_2)](L-is-d)/c} df_1 df_2 \\ & + e^{-2\gamma(L+is+d)} \int_{-\infty}^{\infty} \int_{-\infty}^{\infty} G(f_1) G^*(f_2) \\ & \cdot H^*(f_1 - f_2) e^{-j2\pi[f_1\beta(f_1) - f_2\beta(f_2)](L+is+d)/c} df_1 df_2. \end{aligned}$$

The second contribution, which is desired, is due to the product of the forward- and backward-propagating waves and is given by

$$\begin{aligned} I_i^c = & -2e^{-2\gamma L} \operatorname{Re} \left\{ \int_{-\infty}^{\infty} \int_{-\infty}^{\infty} G(f_1) G^*(f_2) \right. \\ & \cdot H^*(f_1 - f_2) e^{-j2\pi[f_1\beta(f_1) - f_2\beta(f_2)]L/c} \\ & \left. e^{j2\pi[f_1\beta(f_1) + f_2\beta(f_2)](is+d)/c} df_1 df_2 \right\}. \end{aligned}$$

(The $\operatorname{Re}\{\}$ above may be dropped since $\beta(f) = \beta(-f)$, $G^*(f) = G(-f)$, and $H^*(f) = H(-f)$.)

The integration period is sufficiently long that when f_1 and f_2 allow $H(f_1 - f_2)$ to have a nonnegligible value, $\beta(f_1)$ may be treated as equal to $\beta(f_2)$ and $[f_1\beta(f_1) - f_2\beta(f_2)]L/c$ is much less than 1. Using these observations, the integrator outputs may be expressed as

$$\begin{aligned} I_i^s = & 2e^{-2\gamma L} \cosh[2\gamma(is+d)] \int_{-\infty}^{\infty} \int_{-\infty}^{\infty} G(f) \\ & \cdot G^*(f - \Delta) H^*(\Delta) df d\Delta \end{aligned}$$

and

$$\begin{aligned} I_i^c = & -2e^{-2\gamma L} \int_{-\infty}^{\infty} \int_{-\infty}^{\infty} G(f) G^*(f - \Delta) \\ & \cdot H^*(\Delta) e^{j4\pi f\beta(f)(is+d)/c} df d\Delta. \end{aligned}$$

Finally, since the signal g is constrained to be zero outside of the observation window,

$$\int_{-\infty}^{\infty} G^*(f - \Delta) H^*(\Delta) d\Delta = G^*(f)$$

so that

$$I_i^s = 2e^{-2\gamma L} \cosh[2\gamma(is+d)] \int_{-\infty}^{\infty} |G(f)|^2 df$$

and

$$I_i^c = -2e^{-2\gamma L} \int_{-\infty}^{\infty} |G(f)|^2 e^{j4\pi f\beta(f)(is+d)/c} df.$$

The n integrator outputs are fed, via a multiplexer, to a discrete N -point Fourier transform device to produce the values $\{S_k, k = 0, 1, 2, \dots, N-1\}$, where

$$\begin{aligned} S_k = & \sum_{i=0}^{n-1} I_i e^{-j2\pi k(i+i_0)/N} \\ = & \sum_{i=0}^{n-1} (I_i^s + I_i^c) e^{-j2\pi k(i+i_0)/N} \\ = & S_k^s + S_k^c \end{aligned}$$

and where i_0 is chosen as described below and the two contributions to S_k are defined in the obvious way.

First consider the desired contribution, S_k^c , which may be expressed in terms of

$$W(z) = \frac{\sin(\pi n z)}{\sin(\pi z)}$$

which results from the transform of the observation window. In terms of W ,

$$\begin{aligned} S_k^c = & -2e^{-2\gamma L} \int_{-\infty}^{\infty} |G(f)|^2 e^{j2\pi[2f\beta(f)s/c - k i_0/N]} \\ & \cdot e^{j\pi(n-1)[2f\beta(f)s/c - k/N]} \\ & \cdot W(2f\beta(f)s/c - k/N) df. \end{aligned}$$

In order to avoid sign precession, the choice $i_0 = d/s$ is made; that is, if construction constraints require that d be at least some minimum value, then d can be chosen as the least allowed multiple of s . With this choice,

$$\begin{aligned} S_k^c = & -2e^{-2\gamma L} \int_{-\infty}^{\infty} |G(f)|^2 e^{j\pi(2i_0 + n-1)[2f\beta(f)s/c - k/N]} \\ & \cdot W(2f\beta(f)s/c - k/N) df. \end{aligned}$$

The function W peaks when its argument is an integer. Around each integer, the nominal width of the center lobe is $1/n$. It may be seen that the function W serves to select which frequencies of g are permitted to contribute to S_k . For each choice of k , frequencies f satisfying

$$|2f\beta(f)s/c - (k/N + m)| \leq \frac{1}{2n}$$

for any integer m form the nominal k th output. As the frequency gets close to the edge of the band, the phase also begins to rotate.

To avoid aliasing of one frequency in the band of interest with another, the input frequency band must be chosen so that $2f\beta(f)s/c$ ranges over an interval of length less than $1/2$. Thus, the system can support approximately $n/2$ independent frequency resolution cells. These cells are examined using k in the range $0, 1, 2, \dots, N/2-1$ or in the range $N/2, N/2+1, \dots, N-1$. The size of the transform, N , may be chosen equal to n or may be made larger to obtain more, but overlapping, cells.

An important feature is that these cells may correspond to any range of the form $2f\beta(f)s/c \in [m, m+1/2]$ or $2f\beta(f)s/c \in [m+1/2, m+1]$, respectively. Thus, the same number of detectors, with the same spacing, can cover the frequency band $[f_L, f_H]$ or a translated band $[f_L + f_T, f_H + f_T]$. The advantage of using this latter band with $f_T > 0$ is that the fractional bandwidth may be made smaller, making component requirements less stringent and making the range of $\beta(f)$ smaller (and perhaps negligible).

Note that the relationship between k and the associated center frequency is not linear and is, in fact, given by

$$f = \left[f_c^2 + \left(\frac{c}{2s} \right)^2 \left(m + \frac{k}{N} \right)^2 \right]^{1/2}.$$

This nonlinearity may become negligible if the waveguide is operated at a sufficiently small fractional bandwidth away from cutoff. If desired, the spatial-to-temporal frequency interpretation may be made by using a lookup table.

Except near cutoff, the nominal width of a frequency bin centered at f is

$$\frac{c}{2ns\beta(f)} \approx \frac{c}{2L\beta(f)}.$$

Thus, say, a device which uses an air dielectric waveguide and which is to achieve a resolution of 100 MHz needs to be a little more than 2 m long when operated midband ($f \approx 1.5 f_c$). The use of a dielectric with higher index would allow a shorter waveguide.

Now consider the undesired output term S_k^s . Its amplitude dependence on k is of the form

$$\frac{\sinh[n(j\pi k/N \pm \gamma s)]}{\sinh(\pi k/N \pm \gamma s)}$$

which peaks when k is a multiple of N and is small elsewhere. Thus, this term only interferes with at most a few of the usable values of k . To avoid such interference, the range of inputs must be slightly curtailed, so that slightly less than $n/2$ resolution cells may be covered.

Although the system was described above using rectangular waveguide, any other guiding medium, such as stripline, would do. Dielectrics with slow propagation velocity may be used to decrease the size of the apparatus.

One of the crucial questions to be investigated is whether the detector coupling can be achieved without introducing significant reflections. Such reflections would be misinterpreted as new frequency components.

Several generalizations of the above system are possible. One such modification is to include weighting of the autocorrelation samples to provide more desirable crosstalk characteristics. By weighting these terms, the function W may be given lower side lobes at the expense of a wider center lobe.

Another observation is that the integrators need not be strict integrators as described, but can be simpler low-pass filters. (The filtering may occur naturally as the detector response.) Further, these low-pass filter outputs need not be sampled and held, but need only be sampled by the analog multiplexer.

IV. CONCLUSION

A scheme of performing spectrum analysis over large bandwidths has been presented and briefly examined. Engineering questions, such as the ability to couple energy into the detectors without causing significant perturbations and the dynamic range limits imposed by the detectors, have yet to be addressed.

The approach allows simultaneous monitoring of all frequency bins within the band of interest and, unlike IFM methods, can operate with multifrequency inputs. The proposed scheme does not suffer from the bandwidth limit imposed by acoustooptic input devices or the bandwidth limit imposed by output sampling in compressive receivers. The approach also overcomes the slowness and ambiguity of scanning methods.

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Modeling of Arbitrarily Shaped Signal Lines and Discontinuities

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Abstract—The propagation characteristics for signal lines and discontinuities embedded in a homogeneous medium and having any shape composed of steps along the Cartesian coordinates are obtained through an extension of the author's work on scattering from periodic apertures. The approach is specifically applied to high-performance computer packages, where previously employed capacitance and inductance techniques may not be appropriate. Numerical results are given for representative structures that involve signal lines, mesh planes, vias, and crossing and coupled lines.

I. INTRODUCTION

Techniques for analyzing uniform transmission lines [1], [2] and discontinuities of regular shape [3] are well known. Unfortunately, a general solution technique has not existed for nonuniform lines or 3-D discontinuities of irregular shape, although it would be invaluable for analyzing computer modules [4], where signal lines bend and numerous discontinuities are encountered in the path between chips.

A section of such a module (Fig. 1) shows a signal line situated between mesh reference planes and connected to a second signal line through a via. Vias, stubs, right angle bends, and adjacent but nontouching other lines and vias represent discontinuities. The delay and impedance of, and coupling between, these structures must be determined so reflections that cause circuits to falsely switch can be avoided [5].

Various quasi-static and quasi-TEM approaches [6], [7] may be applied, but only to pieces of the problem. A more general approach involves three-dimensional capacitance and inductance

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