# Time-Resolved Measurements Using Conventional Analog Network Analyzer

R. C. Woods

***Abstract*—A new method of making precise time-resolved measurements using an unmodified analog network analyzer is described. This technique is most useful in wafer probing unpackaged acoustic devices in production conditions or in characterizing development acoustic devices that are packaged to allow easy physical access to the device and which may, therefore, exhibit high levels of RF feedthrough.**

***Index Terms*—Acoustic devices, automated measurements, network analysis, RF feedthrough, surface acoustic wave (SAW) devices, time-resolved measurements, wafer probing.**

## I. Introduction

Conventional network analyzers are often used in the transmission mode for network analysis of components used for signal-processing functions. Lumped-element filters and similar components are easily characterized using this technique, but problems arise in testing acoustic devices such as surface acoustic wave (SAW) filters. Conventional or traditional analog analyzers, typified by the Hewlett-Packard 8510 or 8754, produce an RF signal of swept frequency and detect the amplitude of the signal output by the device-under-test (DUT). Plotting output against frequency gives the insertion loss of the DUT. This method is often inadequate for testing acoustic devices (and other devices where the output is significantly delayed compared to the input) because, in many cases, there is a stray signal directly radiated from the input connectors of the DUT and coupled to the DUT's output that bears no relation to the acoustic properties of the DUT. To determine the properties of the DUT, it is often useful to be able to measure the acoustic signal isolated from the radiation signal. The sum of the direct signal (usually a broad-band feedthrough signal or "breakthrough" of near-zero group delay and approximately constant phase shift) and the delayed acoustic signal will often represent a high amplitude fast ripple on the final plot on the network analyzer. As an example, Fig. 1 shows the result of using a conventional network analyzer to test a simple SAW filter, with no particular care taken to reduce the feedthrough signal. The fast ripple on this plot is a result of the network analyzer summing the radiation feedthrough and the delayed acoustic signals which, therefore, oscillate in and out of phase as the frequency is swept, and is not representative of the acoustic performance of this filter when packaged such that feedthrough is reduced.

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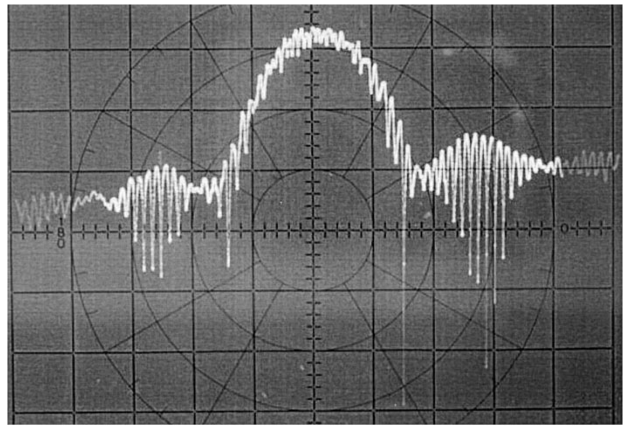

Fig. 1. Result of using a conventional network analyzer (HP8754A) to test a simple SAW filter (center frequency $f_o = 48$ MHz, response nominally proportional to $\{\sin[(f - f_o)\tau]\}/(f - f_o)$, vertical 10 dB per division, horizontal 5 MHz per division).

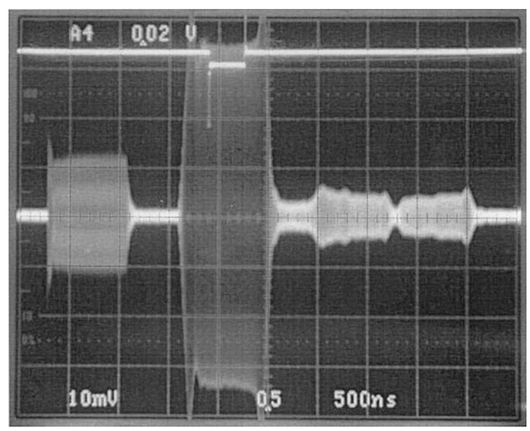


Fig. 2. Top: typical control pulse applied to linear gate. Bottom: output from device used to obtain Fig. 1, showing clearly both feedthrough (300–1100 ns after start of trace) and delayed SAW signal (1600–2500 ns after start of trace). Later signals are miscellaneous acoustic reflections. (Horizontal 500 ns per division both traces.)

The feedthrough is seen clearly in the bottom trace of Fig. 2, showing a typical output signal from the DUT when fed by a short burst of RF.

There are occasions (e.g., final production device measurement) when the raw output signal is a better measure of the overall device performance than is the purely acoustic signal. However, during device development, or during wafer probing of production devices with no possibility of feedthrough reduction, it is often desirable to measure the acoustic signal level isolated from the feedthrough signal. A number of techniques have been described previously for performing measurements of this type.

In the first technique [1], a commercial network analyzer is not used. Instead, a linear gate is used to time gate the required output signal. The gate can be arranged to be open during the arrival of the wanted acoustic signals and to be closed during the unwanted RF feedthrough. Under computer control, the gated output amplitude is then compared to that produced by calibrated programmable attenuators. This system has considerable flexibility, not least in the choice of RF input pulse length, which may be short compared to the acoustic length of the DUT, so that individual acoustic signal paths can be selected for examination at the output, or long so that the output examined is the sum of all the acoustic components, but excluding the RF feedthrough. The main disadvantage of this system is that a considerable amount of custom programming and some interfacing between the controlling computer and hardware is required to enable computer-controlled measurements to be made. Also, the measurement at each frequency takes several seconds so that each device takes a minimum of several minutes to characterize completely. When correctly set up, the system accuracy depends largely upon the precision of the standard attenuators used for comparison with the DUT and the relays used for RF switching.

In a variation of this method [2], a conventional network analyzer has been used together with some primitive gating to provide some aspects of time-resolved measurement. The main disadvantage of this method is that it will not operate with all network analyzers since it relies upon the network analyzer to average the returned signal over the period of the pulsing, and this will be done to uncontrolled and largely unknown extents by different network analyzers. Also, if the RF switches are not perfectly balanced (as is inevitably, in practice, the case), there is some feedthrough of the receiver gating signal directly into the network analyzer input. If the gating pulse has much larger amplitude than the RF signal from the DUT (which is usually the case), then this feedthrough may swamp the wanted signal from the DUT (and, in extreme cases, may damage the network analyzer input).

A further technique [3] is to use a digital vector network analyzer, typified by the Hewlett-Packard 8753. The impulse response output from the DUT in the time domain is obtained by a Fourier transform of frequency-domain data. The initial feedthrough signal may then be masked out, and the remaining signal then reverse transformed for display in the frequency domain. This instrument has the advantage of not requiring any particular custom programming by the user, but where the feedthrough is considerably greater in amplitude than the acoustic signal (often the case with components under development), the results will be marred by the imperfect dynamic range of the receiver and by imperfect digitizing since the important information regarding the acoustic signal may be several orders of magnitude lower than the dominating feedthrough signal. To retrieve an acoustic signal buried in a much greater feedthrough signal requires considerable precision in the receiver, digitizing processing, and display processing in the network analyzer. As a result, this method has only a limited dynamic range in cases where the feedthrough signal is large. The choice of gating functions in the time domain also involves a degree of compromise since low sidelobe levels are only obtainable for relatively slow gating functions (and, conversely, fast gating implies high sidelobe levels so that large feedthrough levels will not be adequately suppressed).

In a related method [4], the impulse response is produced directly by using an impulse generator, the resultant DUT impulse response is time gated using a fast switch, and the gated waveform is examined in the frequency domain using a spectrum analyzer. This method suffers from relatively poor signal-to-noise ratio.

An effective though complex method of using an unmodified conventional network analyzer with external components to produce pulsed operation has been published previously [5] and uses three interacting feedback loops. The principle of operation is to convert the pulsed and gated signal from the DUT to a continuous wave (CW) signal for presentation to the network analyzer. Part of the complexity of this method is necessary to maintain phase integrity between the pulsed and CW signal, which is not required in the present application where only signal levels (insertion loss) must be measured. Also, the accuracy of this method depends upon one feedback loop being fast enough to be able to track the incoming pulse amplitude. In addition, it is required that pairs of electronic attenuators are well matched over the frequency range required and over a wide attenuation range, and that several power splitters, amplifiers, and other components have flat frequency responses unless further error correction is used. These requirements are in addition to those imposed if phase measurements are also to be made.

Finally, the use of sophisticated error correction [6] by calibrating the fixture holding the DUT may enable the effects of the fixture to be annulled. However, this procedure can only be used when the DUT fixture is itself well characterized or can be accurately calibrated, and cannot be used in the case of ad-hoc fixtures where the DUT cannot be removed or in wafer probing where the probe positions are not well defined. Also, this method does not permit time resolution to isolate the various acoustic signals contributing to the final response.

This paper presents a simpler technique for using a conventional network analyzer in transmission mode to make time-resolved insertion-loss measurements without special programming being involved, with much simpler hardware requirements than the most sophisticated methods described above. The present method uses only some standard electronic and microwave components in addition to a standard analog network analyzer, yet retains the full flexibility given by hardware time gating.

## II. Method of Making Time-Resolved Measurements Using an Analog Network Analyzer

A block diagram of the system reported here is shown in Fig. 3. The principle of this equipment is to use the tracking generator within the network analyzer to provide an RF signal, which is then time gated in a manner similar to that used in [1]. A regular series of RF bursts is applied to the DUT, such that the pulse repetition frequency (PRF) of these RF bursts is the highest value that allows all acoustic activity to have decayed between adjacent RF bursts. A PRF around 10 kHz is usually

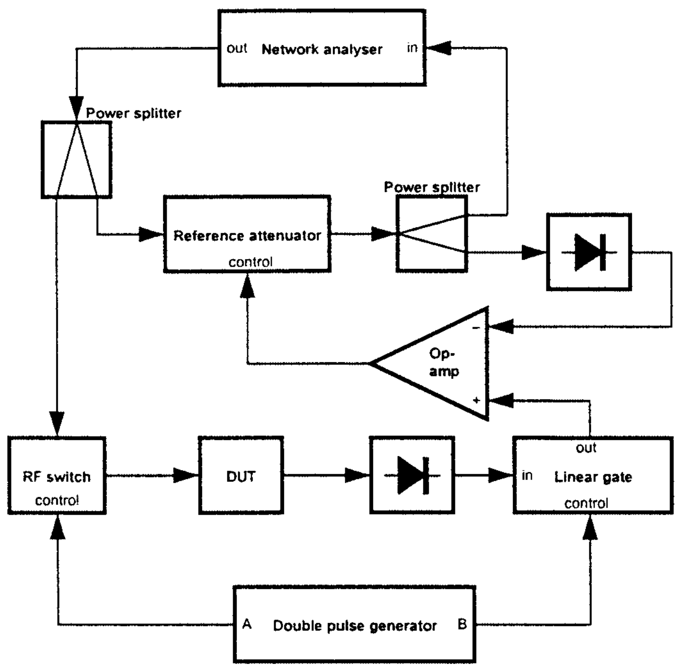


Fig. 3. Block diagram of equipment used for obtaining time-resolved network analysis using an analog network analyzer.

satisfactory. The output from the DUT is detected and applied to a linear gate triggered by a control pulse delayed after the start of the RF burst and having an independently specified duration. The linear gate functions as a "sample and hold" circuit [1] to extract only the wanted signal from the stream of signals that commences with the direct feedthrough signal simultaneous with the input RF signal, followed a microsecond or so later by the various acoustic signals from the DUT including the main signal and any undesirable reflections. The linear gate output is then compared with the output from an RF detector fed with the CW output from the tracking generator, and whose amplitude can be adjusted by a continuously variable electronic attenuator. By forming a negative feedback loop including the reference attenuator and CW detector, and controlled by the "sample and hold" circuit, the reference attenuator produces a signal at the CW detector of the same amplitude as the signal selected from the DUT. A fixed proportion of this CW signal is returned to the network analyzer for display.

The accuracy of this system depends only upon the following factors:

1) two detectors should have similar characteristics;
2) two power splitters should split the RF power with a flat frequency response over the frequency range of interest (although exact 50 : 50 splitting is not necessary, the precise ratio must be constant);
3) RF switch producing the RF bursts applied to the DUT should have a flat frequency response.

Fortunately, high-specification coaxial diode detectors, power splitters, and RF switches are readily available from suppliers of RF and microwave components so that these conditions may be met in practice. There follows a detailed description of the individual components used in this study.

### A. Network Analyzer

An unmodified Hewlett-Packard model 8754A network analyzer was used in this study. This instrument provides automated network analysis over the frequency range of 1–1300 MHz. It is fundamentally an analog instrument and all signal processing and displaying appears to the user in analog form. The frequency sweep rate must be kept sufficiently low such that the external feedback loop can accurately track the signal variations from the DUT. In practice, a sweep rate of approximately 2 Hz provided a good compromise between trace visibility, readability, and accuracy.

### B. Power Splitters

In this work, Mini-Circuits type ZFSC-2-4 power splitters were used.

### C. RF Switch

The requirements for the RF switch are quite demanding because, as well as having a flat frequency response, it must be capable of switching the RF burst with an on/off ratio approximately 70 dB or more and a rise time around 50 ns or less for clear time resolution with most conventional SAW devices of normal size. It is common practice to use RF double-balanced mixers [7][1] to perform this function, rather than p-i-n diode RF switches. This is largely because mixers are easier to drive than p-i-n diode switches (to turn off a p-i-n diode switch requires the application of reverse polarity current, whereas to prevent RF transmission through a mixer requires only that zero current be applied). To achieve sufficient on/off ratio, two mixers were used in series (see Fig. 4). The control pulse must be applied to the central (dc coupled) ports of each mixer. Most commercial double-balanced mixers are electrically symmetrical so that it is immaterial which of the other two ports is used as the RF input and which is the RF output. In this study, two Mini-Circuits type ZP-5X double-balanced mixers were used. Each mixer port represents a 50-$\Omega$ load so that the termination presented to the signal generator is actually 25 $\Omega$ and, hence, a fixed 6-dB pad attenuator was inserted at the mixer end of the cable connecting the mixers to the pulse generator to reduce ringing due to incorrect termination. In practice, this does not cause significant problems.

### D. Reference Attenuator

The requirements for the reference attenuator are considerably less stringent than for the RF switch since much slower rise time and nonflat frequency response can be tolerated, although the dynamic range requirement is similar to that for the RF switch. Nevertheless, two series mixers may also be used as the reference attenuator. In this study, either two Mini-Circuits type ZP-5X mixers (connected in series, as shown in Fig. 4, but omitting the 6-dB pad attenuator) or two Mini-Circuits type ZAS-1 attenuators (connected in series, also as shown in Fig. 4, but omitting the 6-dB pad attenuator) were used. If the minimum

[1]See also "Mixers application information"; Watkins-Johnson TechNotes, 2001. [Online.] Available: http://www.wj.com/pdf/technotes/Mixer_application_info.pdf

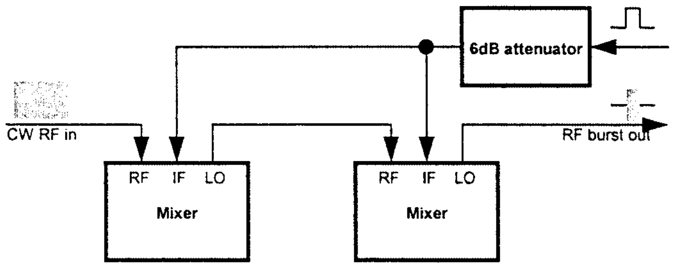


Fig. 4. Use of two mixers as a high-performance RF switch.

insertion loss of the reference attenuator is too great, then the resultant trace on the network analyzer will exhibit a "plateau," indicating that either the maximum current drive to the reference attenuator is insufficient or that the attenuation through the DUT is too small for the system to measure as set up. In this case, an extra fixed attenuator should be added in the DUT signal path.

### *E. Diode Detectors*

Since the method relies upon close tracking of the output voltages from the detectors under a wide range of input powers and frequencies, it is important that a matched pair of detectors should ideally be used. In this study, a number of different detectors were tested and used. Pairs of commercially available general-purpose coaxial detectors of similar type were found to operate satisfactorily, but in practice, custom-made detectors constructed using small discrete components in a zero-bias "voltage doubler" circuit [8] using Agilent Technologies type HSMS-2852 Schottky barrier diodes were found to outperform the commercial detectors up to at least 1.3 GHz since the power levels are quite low in this application. (Most general-purpose commercial coaxial detectors are designed for medium power input levels.) The HSMS-2852 Schottky barrier diodes have a very low forward voltage drop and are specifically intended for low-power zero-bias applications. Slightly higher sensitivity may be achieved by applying dc bias, but this was not done in this study for simplicity. It is not necessary to restrict the detectors either to their linear or to their square-law region of operation since all that is necessary is that their output voltages are matched for identical input RF powers and frequencies.

For the op-amp connections shown in Fig. 3, the detectors must give positive outputs. If negative output detectors are used, the op-amp inputs must be transposed.

The load resistors used with the detectors affect their sensitivity and output rise time; a large load resistance gives highest sensitivity, but slowest rise time, whereas a low load resistance gives poor sensitivity, but fastest rise time. In this study, both detectors were loaded by 3.3 k$\Omega$. For good matching, both detectors must, of course, be loaded with the same resistance value.

### *F. Linear Gate*

The function of the linear gate is to sample the detected output signal from the DUT at the times required, and to produce a steady output of the same amplitude. In the work reported here, a Brookdeal type 415 linear gate was used. This instrument has an overall dc gain from input to output of $\times 100$. A low averaging time-constant ($\sim$ 3 ms) was found satisfactory, as high noise levels requiring considerable averaging were not encountered.

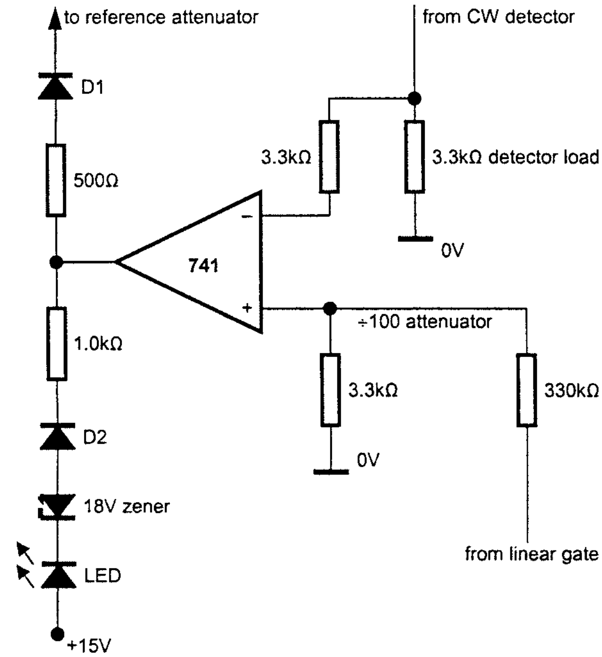


Fig. 5. Op-amp circuit used in this study.

### *G. Op-Amp Feedback Circuit*

Since the linear gate used here has a dc gain of $\times 100$, the feedback system would not operate correctly without compensating for this extra gain. This means that the signal from the linear gate must be attenuated by $100\times$ before feeding to the op-amp feedback circuit. This is accomplished by the 3.3-k$\Omega$ resistor connected from the noninverting input to ground, and the associated 330-k$\Omega$ resistor. The 3.3-k$\Omega$ resistor from the inverting input to the circuit input maintains the same Thévenin impedance at both op-amp inputs under low-impedance drive to minimize op-amp drift problems [9]. In principle the signal from the CW reference detector could instead be amplified by $\times 100$, but this increases the feedback loop gain beyond the maximum value for unconditional stability. Amplifying the reference signal by $\times 10$ and attenuating the linear gate output by $10\times$ also gives too much loop gain for stability.

The op-amp circuit used is shown in Fig. 5. The response speed of this feedback loop governs the maximum sweep speed that can be set on the network analyzer. The type 357 op-amp is much faster than the type 741 and was also tested in this circuit with little difference in performance. The response speed is largely determined by the stray capacitance of the low-frequency cables, which should, therefore, be kept as short as possible. Most other high-gain low-to-medium frequency op-amps could also be used successfully here, such as the type 356 op-amp, which offers a speed between that of the type 741 and 357. The reference attenuator requires zero current drive for maximum attenuation and, thus, it is necessary to include diode D1 to prevent negative current drive under conditions where the DUT transmits zero RF. (Negative current will not harm the reference attenuator, but it results in large RF transmission through the attenuator.) The resistor in series with D1 driving the reference attenuator was chosen to match the maximum current drive allowable for the reference attenuator

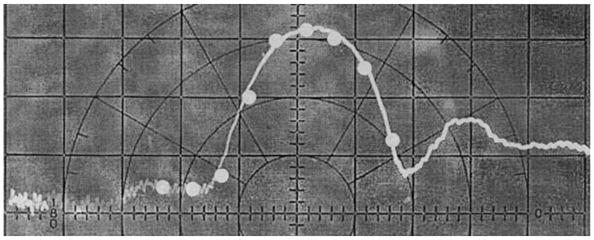

Fig. 6. Result obtained using the method described here from the same device used to obtain Fig. 1 (center frequency $f_o = 48$ MHz, vertical 10 dB per division, horizontal 5 MHz per division). White dots (◯) are manual measurements of the gated acoustic frequency response of the same device.

used at maximum voltage excursion (by coincidence, this also corresponds with the maximum current available from the op-amp used, 25 mA). The LED was included to indicate when the reference attenuator was completely turned off and the op-amp was saturated in the negative supply direction. (D2 prevents excessive reverse voltage being applied across the LED, and the Zener diode prevents it lighting until the op-amp is saturated.) In practice, after the offset null control for the op-amp (for clarity, not shown in Fig. 5) has been adjusted by shorting the two external inputs together while monitoring the op-amp output voltage, the dc output zero-shift control of the linear gate must be adjusted so that the LED is just on the threshold of lighting with no signal transmitted through the DUT.

### *H. Double Pulse Generator*

Two independently controlled pulses are necessary for controlling the equipment. The first pulse defines the duration of the RF burst applied to the DUT, and the second defines the time and duration of the sampling by the linear gate. Either a generator producing two independent output pulses or two separate pulse generators using a common trigger may be used. In the work reported here, two Datapulse 101 single-pulse generators were used. Pulse generation phase locked with the RF signal does not affect the operation of the system described here.

## III. RESULTS

Fig. 6 shows the result of using the equipment described above to test the same SAW device used in obtaining Fig. 1. The time gating used is as shown in Fig. 2. All traces of fast ripple are removed and the plot shows much more clearly the fundamental acoustic performance of the DUT than does Fig. 1. For comparison, superimposed on Fig. 6 is the acoustic performance of the same time-gated signal from the DUT measured manually to within 0.5 dB. The results agree within the resolution of the network analyzer used. In practice, a dynamic range of around 60 dB was achievable using the system described here.

As a further test of the system, Fig. 7(a) shows the insertion loss of a Telonic Berkeley tunable bandstop cavity filter (model TTR-190-3EE) measured using the unmodified network analyzer, and Fig. 7(b) shows, for comparison, the result obtained using the system described here. Of course, the signal from a cavity filter is not substantially time delayed and, thus, this test was conducted purely as a demonstration of the capabilities of

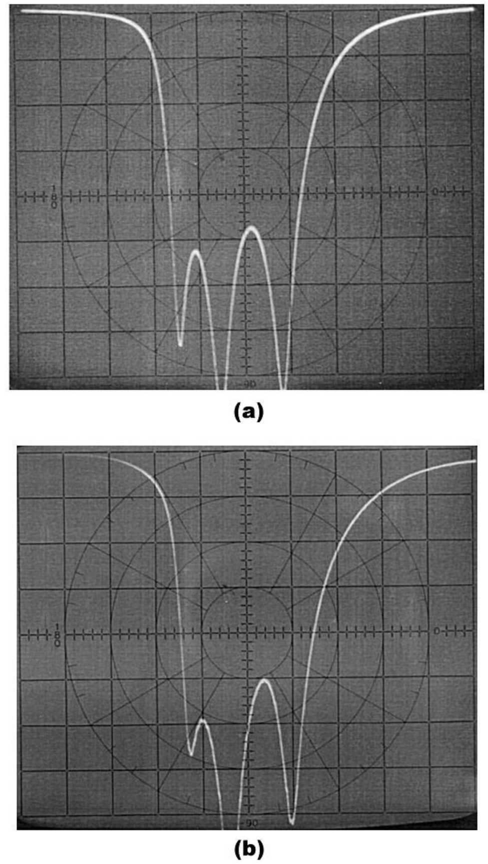


Fig. 7. (a) Insertion loss of a commercial bandstop filter measured using the unmodified network analyzer. (b) Same filter measured using the system described here, using an RF pulse length $= 5.0\ \mu$s time gated between 4.0–4.8 $\mu$s. (For both traces, center frequency $f_o = 125$ MHz, vertical 2.5 dB per division, horizontal 1 MHz per division.)

the system. However, when using a high-$Q$ filter with pulsed RF signals, ringing transients are produced at the output and, thus, the time gating must be arranged to sample the output late enough after the start of the input RF pulse (as determined by observing the output signal in time domain) to avoid artifacts caused by the pulsing. A delay of several microseconds was found satisfactory in the present case.

## IV. CONCLUSIONS

A new method of making precise time-resolved measurements using an unmodified analog network analyzer has been described and demonstrated. The technique described will be most useful in wafer probing unpackaged production components or in characterizing development devices that are packaged to allow easy access to the device rather than to achieve low levels of RF feedthrough.

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Prof. Woods is an invited member of the International Awarding Committee for Marie Curie Fellowships funded by the European Commission. He is also a member of the American Society for Engineering Education (ASEE).