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# SPECTRUM AND NETWORK ANALYZERS

J.M. Byrd*) and F. Caspers

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*) LBNL, Berkeley, CA 94720

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# SPECTRUM AND NETWORK ANALYZERS

J. M. BYRD
*Lawrence Berkeley National Laboratory, Berkeley, CA 94720*
*JMByrd@lbl.gov*

F. CASPERS
*CERN, Geneva, Switzerland*
*Fritz.Caspers@cern.ch*

**Abstract**

Spectrum and network analyzers (SPAs and NWAs) are found in every microwave laboratory. This report describes the basic principles of operation of a SPA and NWA and also describes several simple example measurements using each of these instruments.

## 1 Introduction

Spectrum and network analyzers are commonly found in microwave laboratories and are the primary instruments for characterizing radiofrequency (RF) and microwave systems for use in accelerators. The NWAs and SPAs in use today are the result of over half a century of development. The capabilities of these instruments are remarkable. The best commercially available SPAs today have a dynamic range of greater than 100 dB (corresponding to 10 orders of magnitude in power) and over six orders of magnitude in frequency. NWAs can span over five to six orders of magnitude in frequency with up to 0.1 Hz resolution. Furthermore, automation and built-in software analysis have greatly increased their and ease of operation. However, because of the specialized nature of these instruments, many accelerator scientists and engineers are not aware of the capabilities of these instruments. This paper presents an introduction to the basic principles of these devices, as well a specialized jargon, particularly for accelerator–related measurements in a microwave laboratory.

## 2 Spectrum analyzers

### *2.1 Basic principle of superheterodyne detection*

The basic function of a spectrum analyzer is to determine the amplitude of the frequency components of a time domain signal. For a simple sinusoidal signal, this could be easily done by measuring the frequency and amplitude of the signal on an oscilloscope. However, for a signal containing multiple frequencies, this becomes difficult, especially when the frequency components have greatly differing amplitudes. A narrowband tuneable filter could be tuned through the frequency range of interest and the power measured at each filter frequency. However, with the technical difficulty of making a tuneable filter, it is much simpler to have a filter with a fixed frequency and to translate the signal frequency to the filter frequency. The frequency translation is accomplished by a process known as mixing, or multiplying the signal with a local oscillator (LO) signal in a device known as a mixer. A mixer is a passive element made up primarily of diodes configured in a way to produce the product of two signals at the RF and LO ports. Multiplication in an ideal mixer generates signals at the sum and difference of the frequencies of the input signal and LO. The LO frequency is chosen such that the difference component of the signal is at an intermediate frequency (IF) which is then filtered by a bandpass filter with adjustable bandwidth. By sweeping the frequency of the LO, the difference signal from the mixer, including the information in the input signal, is swept through the IF filter. This is known as the heterodyne principle. The SPAs discussed here use a superheterodyne receiver.

A schematic diagram of a basic spectrum analyzer[1,2] is shown in Fig. 1. The input signal is passed through a lowpass (LP) filter which limits the frequency range of the SPA. The input signal is then mixed with a variable frequency LO to the IF filter. The envelope of the IF signal is detected and used to control the vertical deflection of the cathode ray tube (CRT). The LO frequency is swept through the frequency range of interest and is also used to control the horizontal deflection of the CRT. In this way, the screen shows the signal power as a function of frequency. The horizontal axis is also marked in units of time as a reminder that the heterodyne process only measures a given frequency component as it sweeps through that frequency. By the way, a generic radio receiver uses the same heterodyne principle, except without the flexibility of a SPA.

In single range SPAs, the IF is chosen to be above the highest frequency of the SPA. For example, in Hewlett–Packard SPAs with frequency sensitivity up to 2.9 GHz, the IF is chosen to be 3.6 GHz. In this case, the LO frequency would need to range from 3.6 to 3.6+2.9 GHz to achieve the specified frequency range. The bandwidth of the IF filter would determine the frequency resolution of the measurement. Because narrow IF bandwidths are difficult to achieve in the gigahertz range, most spectrum analyzers use several IF stages to mix down to an IF where the narrowest filtering takes place. Shown in Fig. multiheterodyne is a schematic showing several IF stages along with a swept frequency LO at the first IF stage and fixed LOs at subsequent stages. One useful feature of some SPAs is that the signal at one of the IF stages (usually the 21.4 MHz) can be made available on the front or rear panel for

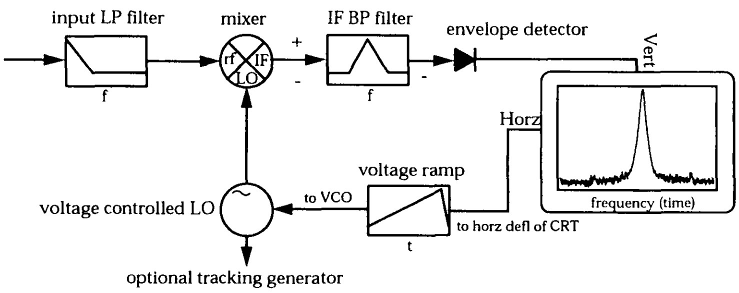


Figure 1: Simplified schematic of a superheterodyne spectrum analyzer.

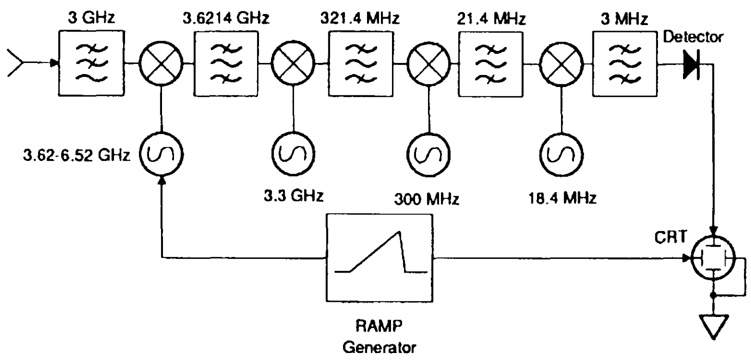


Figure 2: Schematic of a multiple IF stage SPA.

use in some other application. Note that the phase information in the original signal is still available in the IF signal and is only lost after passing through the envelope detector.

An economical technique for SPAs to extend their frequency range is called harmonic mixing. In this technique, harmonics of the LO are generated and also mixed with the input signal. The principle is similar to that of the single-range SPA discussed above except that several additional mixing products are generated and care must be taken to not confuse them. One step in this direction is that a much lower IF of 321.4 MHz is typically used. Some of the disadvantages of harmonic mixing include decrease in the sensitivity due to an increase in conversion loss in the mixer and an increase in phase noise of the LO at higher harmonics. The decreased sensitivity of a wideband SPA with

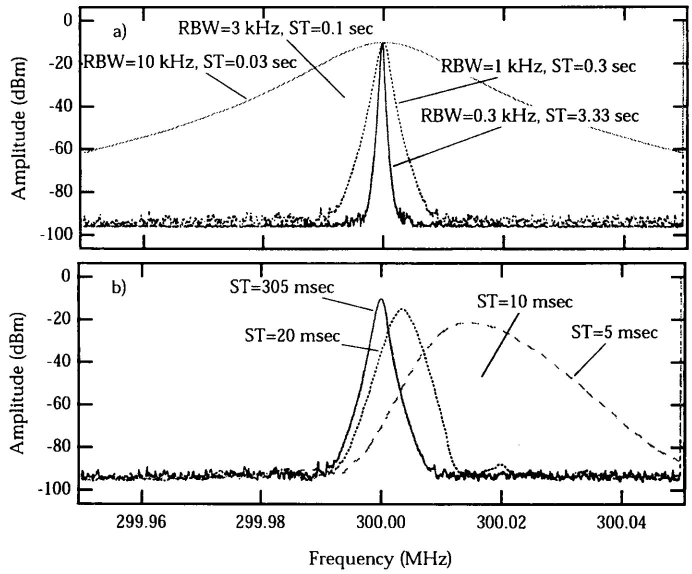


Figure 3: a) SPA measurement with variable resolution band width (RBW) and b) fixed RBW and variable sweep time.

harmonic mixing is evident when viewing the full span with no input signal. The noise floor appears to increase in steps at each harmonic band. It is useful to point out some of the common misconceptions and limitations of SPAs. Although the SPA purports to measure the frequency content of a signal, the measurement is performed in the time domain (as are all measurements) and thus is only an approximation of the actual frequency content. Furthermore, because the SPA only measures the power at a given frequency, the phase of a given frequency component relative to others is lost and thus the measured frequency spectrum is not sufficient for reconstructing a time domain representation of the signal. Superheterodyne SPA measurements are also not well suited for either single-shot signals or signals with low duty cycles since the average signal power is very low. Discrete Fourier Transform methods on digitally recorded data are more appropriate in this case.

### 2.2 *Resolution, bandwidth, and sweep time*

One of the primary specifications of a SPA is its ability to resolve signals of different frequencies, known as the resolution bandwidth (RBW). This is determined by the total bandwidth of the IF filters, usually determined by the final stage filter. Consider a single frequency input to the SPA. As the mixing product of the input signal and the LO is swept across the IF bandpass filter, the shape of the filter mapped out. The measured signal is a convolution of the width of the filter and the natural width of the signal. Clearly, the RBW must be small compared with the natural width of the signal to observe it. The general rule is that two equal amplitude signals can be resolved at a frequency separation equal to the resolution bandwidth. One of the assumptions in measuring the power at a given frequency as it is swept past the IF filter is that the signal has reached steady-state. For example, consider the response of the IF filter when it is subjected to a sinusoidal signal which is suddenly turned on. The output signal will grow to a steady-state in a time approximately equal to the inverse bandwidth of the IF filter. Narrow bandwidths reach steady-state in a longer time than wider bandwidths. For the SPA to accurately measure the signal amplitude, the signal must be swept past the IF filter slowly compared to the response time of the filter. Shown in Fig. 3a is an example of a 300 MHz sinusoidal input signal (the -10 dBm calibration signal provided on many SPAs) measured with variable resolution bandwidths. The sweep time is adjusted automatically to compensate for the changing RBW. Fig. 3b shows the same signal with a fixed RBW but variable sweep times.

The optimum sweep time (ST) for a given RBW and frequency span can be found by equating the time spent within the RBW for a given frequency span with the risetime of the IF filter. The expression is

$$ST = k\frac{Span}{RBW^2} \tag{1}$$

where $k$ is a constant relating the filter risetime to the resolution bandwidth and depends in detail on the filter shape. For Gaussian filters it is about 2.5. Usually, the RBW and sweep time are automatically adjusted according to the frequency span such that a reasonable resolution is achieved. However, higher resolution can usually be achieved but at the expense of longer sweep times. Several SPAs offer the option of digitizing the IF signal to achieve even lower RBWs. The video bandwidth (VBW) is another adjustable parameter on most spectrum analyzers. It is a bandpass filter on the video signal observed on the screen. Usually, the RBW and VBW are equal.

It is also interesting to note the variation of the noise floor of the signal as the RBW is varied as is seen in Fig. 3a. For a constant background noise signal density as a function of frequency, more noise power will pass through the IF bandpass filter as its bandwidth is increased. Therefore, to observe low level signals barely above the instrument noise, it is necessary to use narrow RBWs, implying longer measurement times. Another source of noise inherent in the SPA is amplitude and phase noise of the LO in the SPA receiver. In the previous section, it was assumed that the LO was a pure sinusoid multiplied with the input signal. Random phase or amplitude modulation of the LO can lead to an apparent noise about a signal which may hide another signal.

### *2.3 Example measurement: Harmonic distortion of a nonlinear element*

Many of the RF and microwave circuits used in accelerators are designed to operate in their linear range. Therefore it is important to characterize the distortion of a signal due to the nonlinear response of a circuit element. For a weakly nonlinear element, the output voltage for a given input voltage can be expanded in a power series

$$V_{out} = k_0 + k_1 V_{in} + k_2 V_{in}^2 + k_3 V_{in}^3 + \ldots \tag{2}$$

where $k_0$ is the DC response, $k_1$ the linear gain, and $k_n$ the $n$th order nonlinearity. The expansion is usually truncated at third order. For a sinusoidal input signal, the output signal contains products of the input with itself generating signals at harmonics of the input frequency. One common technique for characterizing the nonlinearity is to use a two-tone input signal given by[1]

$$V_{in} = A_1 \sin\omega_1 t + A_2 \sin\omega_2 t \tag{3}$$

The output voltage can be written as

$$\begin{aligned} V_{out} &= c_0 + c_1 \sin\omega_1 t + c_2 \sin\omega_2 + c_3 \sin 2\omega_1 t + c_4 \sin 2\omega_2 t \\ &+ c_5 \sin 3\omega_1 t + c_6 \sin 3\omega_2 t \\ &+ c_7 \sin(\omega_1 + \omega_2)t + c_8 \sin(\omega_1 - \omega_2)t \\ &+ c_9 \sin(2\omega_1 + \omega_2)t + c_{10} \sin(2\omega_1 - \omega_2)t \\ &+ c_{11} \sin(2\omega_2 + \omega_1)t + c_{12} \sin(2\omega_2 - \omega_1)t \end{aligned} \tag{4}$$

where the coefficients $c_n$ are determined by $A_1$ and $A_2$. In addition to signals at the input frequencies, the output contains signals at frequencies, $|n\omega_1 \pm m\omega_2|$, where $n$ and $m$ are integers. This is commonly referred to as intermodulation distortion. The amplitude of the intermodulation products are determined by the amplitudes of the input signals, $A_1$ and $A_2$, raised to the powers $n$ and $m$. The sum $|n|+|n|$ is called the order of distortion. The order of intermodulation distortion which is observed is always smaller for lower input signals. Expressed in logarithmic terms, the second and third order terms increase 2 and 3 dB per dB of increase in input signal.

In practice, the intermodulation distortion is usually specified to third order. To characterize it, the two input tones are chosen close together in frequency in the center of the frequency response of the device. Shown in Fig. 4 is a spectrum of a with two input signals at 0 and -30 dBm at frequencies of 10 and 11 MHz. The signal harmonics are clearly seen. The intermodulation products appear as sums and differences of harmonics of the two input signals. Each signal is marked with the index n and m. Shown in Fig. 5 is a plot of the amplitudes of the first, second and third order products as a function of increase in input amplitude of each tone. Eventually the first order signal stops increasing as the amplifier reaches its linear limit known as gain compression. The point where extrapolation of the first order response to a level where it intercepts with the second and third order response is known and the second and third order intercept (SOI or TOI). The linearity of RF and microwave components is often specified with the SOI and/or the TOI.

The SPA is well-suited for this type of measurement because it is designed to receive signals at any frequency as opposed to the NWA discussed below whose receiver is meant to respond at the excitation frequency.

## 3 Network analyzers

Network analyzers (NWAs) are used in such a variety of RF and microwave measurements that they are roughly equivalent in utility to an oscilloscope in an electronics laboratory. Their basic purpose is to measure the response of a system to an applied sinusoidal input over a range of frequencies. The capabilities of modern NWAs vary widely but essentially all consist of a signal source and two input channels as shown in Fig. 6. In this simple schematic, the NWA is configured to measure the transmission of a test signal through the device under test (DUT) in channel A relative to the applied signal in channel R. NWAs with the ability to measure the relative phase of the signals are called vector NWAs (VNWAs). Scalar NWAs can measure only the relative amplitude and are typically less expensive. SPAs equipped with a signal source synchronized to the LO, called tracking generators, can be used to make scalar transmission measurements as shown in Fig. 1. This section discusses the principle of operation of vector NWAs and how they can be used to make useful measurements.

Figure 6 shows a NWA configured to measure either the transmitted or reflected signal from the DUT. A swept frequency signal source (R) is used to

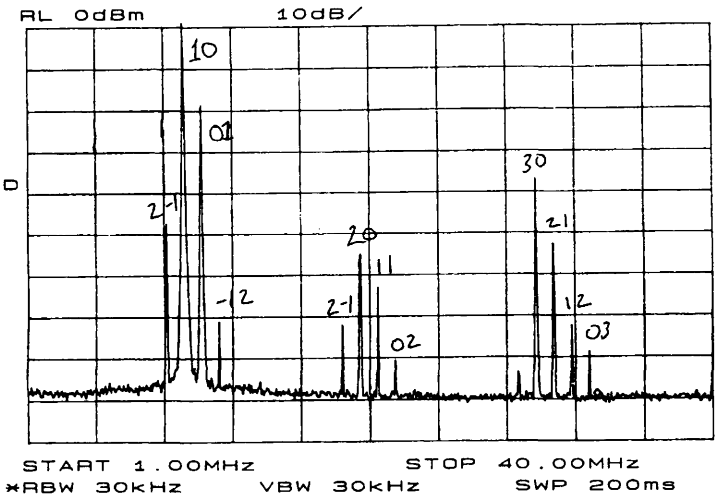


Figure 4: Spectrum analyzer measurement of intermodulation distortion in an amplifier. The amplifier input is a sum of 10 and 11 MHz signals at levels of 0 and -30 dBm, respectively. Intermodulation products are notated with indices $m$ and $n$.

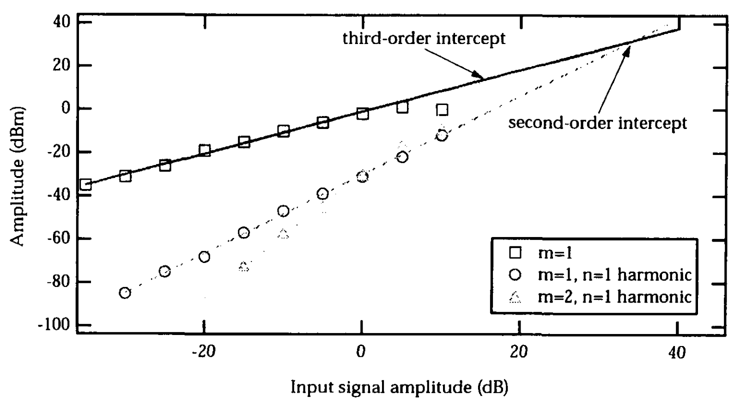


Figure 5: Plot of first, second and third order intermodulation products measured on an amplifier illustrating the intercept points.

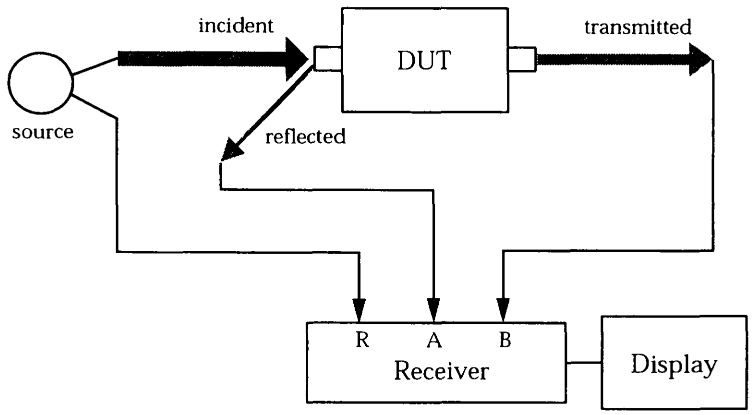


Figure 6: Schematic diagram of a NWA measuring the ratio of reflected and transmitted signal with respect to the incident signal.

generate a wave along a transmission line (coaxial cable or waveguide) which is transmitted to the DUT. Some fraction of the incident wave is reflected, transmitted, and attenuated or amplified by the DUT. From the ratio of the of the reflected wave to the incident (A/R), quantites such as the standing wave ratio (SWR), reflection coefficient, impedance and admittance, and return loss of the DUT can be determined. From the ratio of the transmitted wave to the incident (B/R), the gain or insertion loss, the transmission coefficient, the insertion phase and group delay of the DUT can be found. Note that both the reflected and transmitted ratios are complex quantities implying that both the relative amplitude and phase are measured. Thus, it is possible to characterize a 2-port device by measuring the transmitted and reflected components of a signal incident to a device as a function of frequency.

### *3.1 NWA test sets and receivers*

The directional sense of the forward and reflected signals is measured using a test set. A test set consists of a combination of directional couplers and switches configured as shown schematically in Fig. 7. A simple directional coupler works by combining signals coupled from a transmission line such that signals traveling one direction constructively interfere and signals in the opposite direction destructively interfere. An example of such a test set is shown in Fig. 7. Depending on the NWA model, the test set is either separate or integral to the NWA. The signal source itself can also be either separate or an

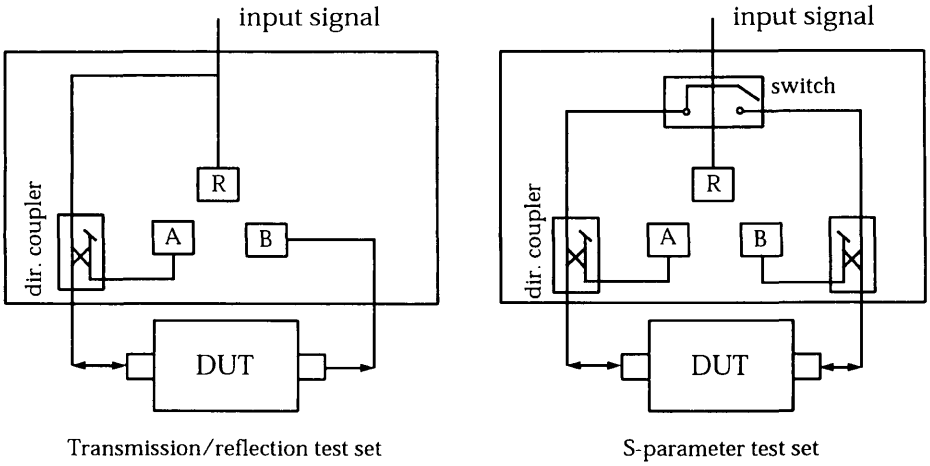


Figure 7: Schematic diagram of a NWA test set. Left) Setup for a reflection measurement from port 1 or a transmission from prot 1 to 2. Right) Setup for a full two–port measurement.

integral part of the NWA, depending on the level of precision required. For example, the HP8510 series NWA uses a separate signal source which allows the customer to choose the precision and stability of the source needed.

Shown in Fig. 8 is a schematic diagram of the receiver. The harmonic mixing is similar to that described in the superheterodyne receiver for a SPA except that it is applied to two channels at the same time. A further requirement is that the relative phase of the two signals be maintained. This is the function of the phase locked loop on the LO shown on the reference channel. The automatic gain control loop shown on the IF is to maintain optimal signal levels for subsequent components. Because the receiver for a NWA is expected to respond at only the frequency of the applied signal source, the receiver design is considerably less sophisticated than that of a SPA and typically has less sensitivity than a SPA.

### *3.2 2-port networks*

Network analyzer measurements are based on the concepts of 2-port networks[4,5]. Consider the 2-port shown in Fig. 9 with voltage $V$ and current $I$ at ports 1 and 2. Voltages at the ports can be written as a linear combination of the

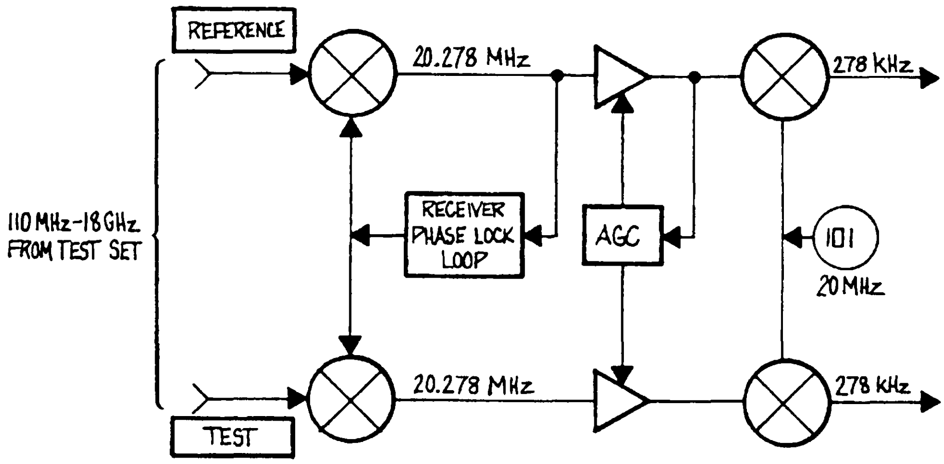


Figure 8: Schematic diagram of a NWA measuring the ratio of reflected and transmitted signal with respect to the incident signal.

currents $I_1$ and $I_2$ given by

$$\begin{aligned} V_1 &= z_{11}I_1 + z_{12}I_2 \\ V_2 &= z_{21}I_1 + z_{22}I_2 \end{aligned} \tag{5}$$

where the $Z$ matrix values are called impedance parameters. For example, $Z_{12}$ is the transfer impedance for $I_1$ equal to zero. One can similarly relate the currents $I_1$ and $I_2$ to the voltages via a matrix called the admittance parameters.

In network analyzer measurements, a 2–port device is characterized in terms of the transmission of forward and reflected waves rather than the voltages and currents at the two ports as shown in the lower part of Fig. 9. The reflected waves, $V_R$, from ports 1 and 2 can be written as a linear combination of the incident waves, $V_I$, given by

$$\begin{aligned} V_{R1} &= S_{11}V_{I1} + S_{12}V_{I2} \\ V_{R2} &= S_{21}V_{I1} + S_{22}V_{I2} \end{aligned} \tag{6}$$

The values of the matrix relating the reflected and forward waves are called scattering or $S$–parameters. For example, with port 2 terminated in the char-

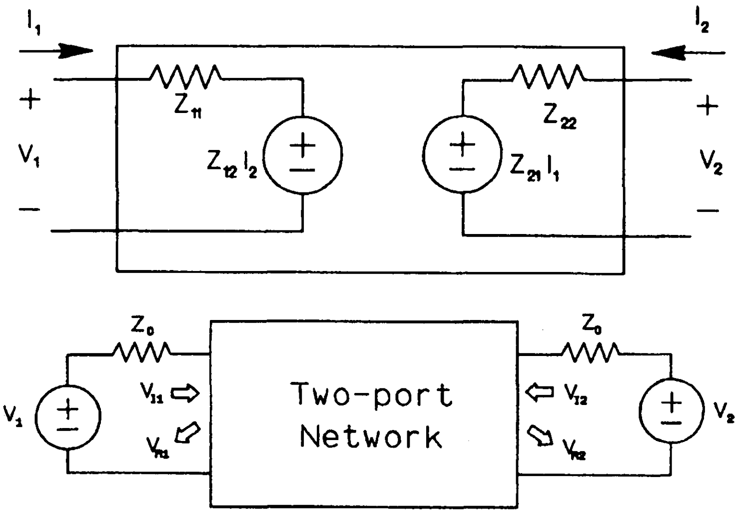


Figure 9: Upper) A 2-port network expressed in terms of the voltages and currents at the ports. These can be related by impedance or admittance parameters. Lower) The same network can be expressed in terms of forward and reflected voltage waves.

acteristic impedance, the reflection coefficient is given by

$$S_{11} = \frac{V_{R1}}{V_{I1}}, \quad S_{21} = \frac{V_{R2}}{V_{I1}} \tag{7}$$

To generalize the results, the incident and reflected voltages are usually normalized by the root of the characteristic impedance of the system, $Z_0$. The normalized reflected voltages are given by

$$\begin{aligned} b_1 &= S_{11}a_1 + S_{12}a_2 \\ b_2 &= S_{21}a_1 + S_{22}a_2 \end{aligned} \tag{8}$$

where

$$a_{1,2} = \frac{V_{I1,2}}{\sqrt{Z_0}}, \quad b_{1,2} = \frac{V_{R1,2}}{\sqrt{Z_0}} \tag{9}$$

Note that the $S$–parameters are implicitly a function of frequency. From the point of view of a NWA, a two–port device resembles the schematic shown in Fig. 10. The lower half of the figure is called a signal flow diagram.

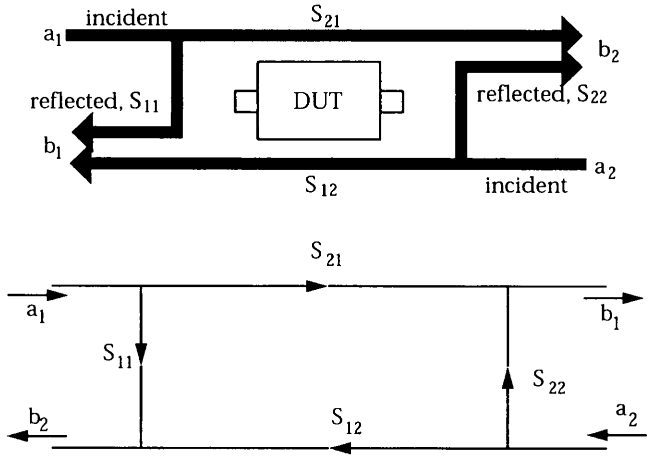


Figure 10: Upper) Schematic relationship of the S-parameters to the forward and reflected signals. Lower) The same relationship expressed as a signal flow diagram. This is often printed on the front panel of S-parameter test sets. Other parameterizations of the network such as impedance and admittance can be related to the S-parameters.

### *3.3 NWA calibration*

One of the necessities in making accurate NWA measurements is calibration. Because of imperfections in the test set, i.e. the directional couplers, attenuators, large measurement errors can arise. To minimize the effect of these imperfections, it is possible to measure the response of standards with well known responses and thus normalize subsequent measurements from the effects of the test set. Thus the calibration removes a large part of the systematic errors of the measurement but does not affect the random errors.

Consider the signal flow diagram of a reflection measurement with errors as shown in Fig. 11. The error terms $E_{rf}$, $E_{df}$, and $E_{sf}$ represent the imperfections in forward transmission, forward directivity, and the forward source, respectively. The measured reflection coefficient is related to the actual by

$$S_{11M} = E_{df} + \frac{S_{11A}E_{rf}}{1 - E_{sf}S_{11A}} \tag{10}$$

where $S_{11A}$ and $S_{11M}$ are the actual and measured reflection coefficients. Clearly the error terms can have a large influence on the measurement.

By measuring the reflection coefficient of three known loads, the values of

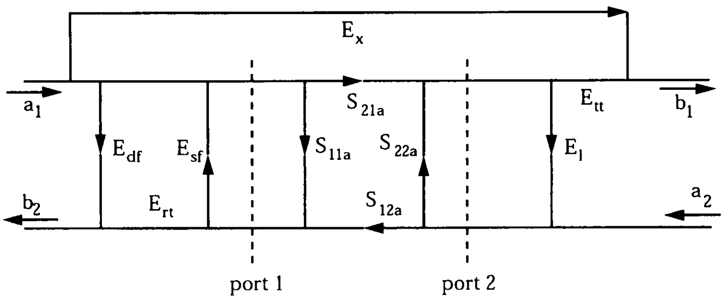


Figure 11: Signal flow diagram for a forward S-parameter measurement with error terms. $E_{sf}$=forward directivity, $E_{sf}$=source match, $E_{rf}$=reflection tracking, $E_{tt}$=transmission tracking, $E_l$=load match, $E_x$=isolation. Reverse flow has a similar diagram.

$E_{rf}$, $E_{df}$, and $E_{sf}$ can be solved at each frequency over which the measurement is made and their effect can be removed from subsequent measurements of a DUT. For example, shown in Fig. 12 is a measurement of $S_{11}$ for a standard 50 $\Omega$ termination with and without a calibration. For an ideal termination, no reflection should be present. The effect of the calibration in this case improves the measurement by 20 dB.

For a full 2-port measurement, the error model must be expanded to include errors at the receiving port, requiring calibration of each port using 3 known loads in reflection as well as a standard in transmission.

For measurements of devices with standard connectors, calibration standards such as a termination, open and short circuit are available. It is necessary to either load the response of the standards into the NWA or internally specify which calibration standards are being used. In current NWA models, it is important to remember that the calibration is valid only over the frequency range over which the calibration was performed. Although the calibration process is tedious, it is fairly well-automated and only requires the connecting of standards as indicated by the analyzer plus the maintenance of a set of standards. Future NWA models may include internal standards which are automatically switched to perform fully automated calibration.

### *3.4 Example NWA measurements*

RF cavities are one of the accelerator devices most commonly characterized using NWAs. Among the desired properties to measure are the frequencies and quality factors (Q) of the resonant modes, particularly the fundamental mode,

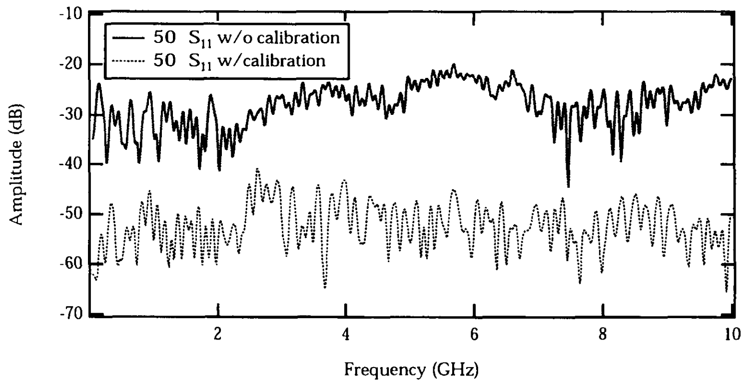


Figure 12: $S_{11}$ measurement of a 50 Ω termination with and without calibration. The calibration provides and 20 dB improvement over this frequency band.

and the match of the coupler to the fundamental mode. The most common method for measuring the mode frequency and Q is to measure transmission through the cavity (S21) using probes lightly coupled to the cavity. Shown in Fig. 13a is an example of such a measurement on the fundamental mode of a pillbox cavity. The Q can usually be determined automatically by the NWA by finding the peak width at the 3 dB points (i.e. half power points). The match of the coupler to fundamental mode can be measured by measuring the reflection of the cavity on resonance. The coupler is the means by which external power is injected into the cavity and usually takes the form of a waveguide coupled to the cavity with either a loop or an aperture. If the cavity is uncoupled to the external source, all of the input power is reflected. If the cavity presents a perfect match to the external source, no power is reflected. Shown in Fig. 13b is an example of a reflection (S11) measurement of the fundamental mode via an input loop coupler. Off resonance, the power is reflected with either a ±180 phase shift, indicating an inductive or capacitive load. It is particularly important to calibrate the reflection measurement to get an accurate measurement of the cavity coupling.

There are numerous RF and microwave devices in any accelerator system and they are usually characterized using a NWA. One of the simplest components is a coaxial cable. Because of the skin effect on the conductors, characteristics of the coaxial line such as signal loss, and group velocity vary as a function of frequency, leading to a dispersive distorion of a signal. Although this may seem like a trivial concern, accelerator signals sometimes must be cleanly transmitted over distances exceeding kilometers, requiring innovative

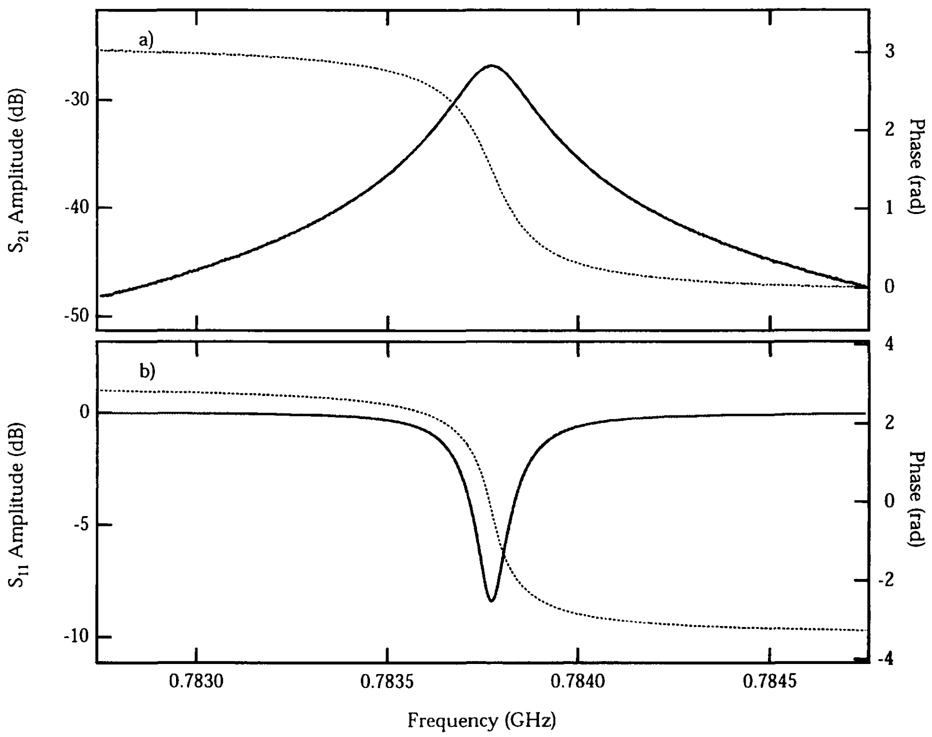


Figure 13: NWA measurements on the fundamental mode of a pillbox cavity. a) Transmission ($S_{21}$) measurement to the determine the resonant frequency and quality factor. b) Reflection (S11) measurement to measure the coupling to the cavity.

designs.

Shown in Fig. 14 is a transmission measurement of a 30 meter length of RG-233/U coaxial cable (i.e. standard BNC cable). The increase in signal loss at higher frequencies is evident as well as a nonlinear slope of the phase as a function of frequency. A delay, $\Delta T$, in the time domain creates a phase shift in the frequency domain of $\omega\Delta T$. This corresponds to a constant slope of the phase response. Because the due to a time delay is constant, is can be removed numerically. This is done in a NWA using a port extension, which acts as if one the ports were physically extending closer to the other port, effectively removing the time delay. For example, the phase response shown in Fig. 14 has an added port extension of 150 nsec, almost exactly equal to the electrical delay of the cable at low frequency. Without the port extension, the phase would change by hundreds of radians over a 1 GHz bandwidth. Note that most NWAs have a feature to measure the delay directly from the slope of the phase response. Shown in Fig. 15 is another example of a common microwave device: a bandpass filter. Among the important characteristics to measure are the bandwith, frequency rolloff above and below the bandpass, and the insertion loss.

### *3.5 Time domain analysis*

A useful feature that has been added recently to several VNWA models is the ability to transform the measured frequency response of a device into the time domain displayed as either the impulse or step response. This feature simulates the measurements usually made using time domain reflectometry (TDR). Note that the synthetic pulse technique is only applicable to linear time invariant networks. Otherwise a real pulse approach is needed.

Although the time domain display of the measurement data may be of only passing interest, the most useful feature is the ability to filter or gate the data in the time domain and subsequently transform back to the frequency domain. For example, the effect of imperfect connectors on a measurement is not clearly distinguished from the response of the device itself when viewing the measurement in the frequency domain. However, by transforming the measurement to the time domain where the reflection at the imperfect connector is evident, it is possible to remove the effect of the connector by gating out the reflection and transforming back to the frequency domain. This feature is particular useful in bench impedance measurements where the mechanical arrangement of the connection to the device often leads to step impedance mismatches which tend to obscure the effect of the device.

Shown in Fig. 16 are the same NWA measurements shown in Figs. ??

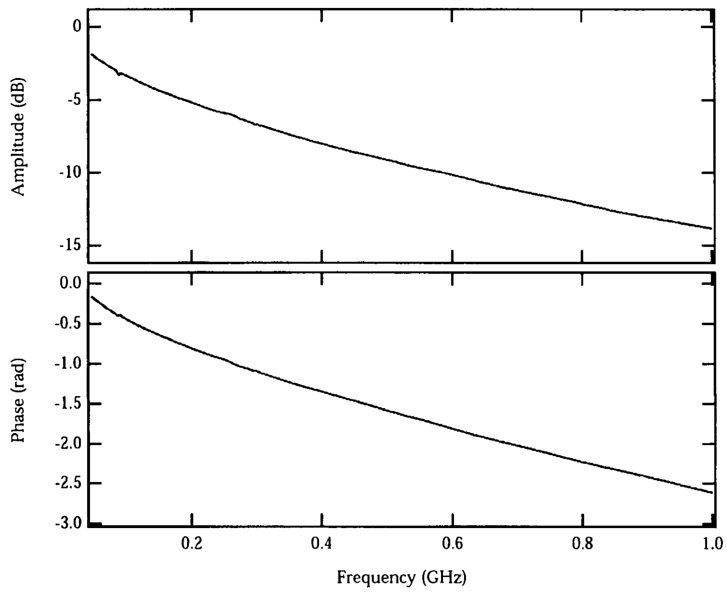


Figure 14: Transmission measurement of a 30 m standard BNC cable.

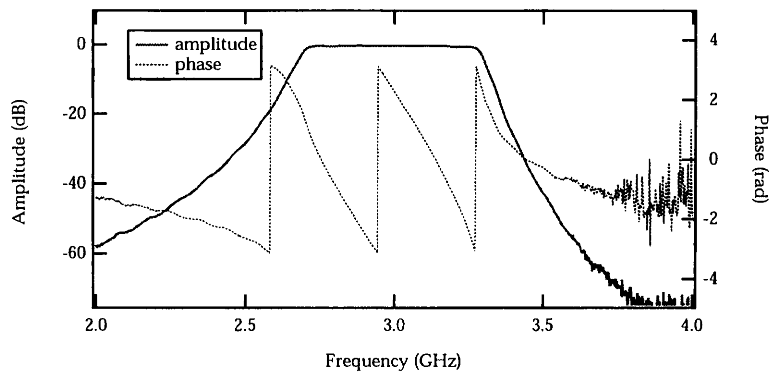


Figure 15: Transmission ($S_{21}$ measurement of a 2.7–3.3 GHz bandpass filter.

tranformed to the time domain. In Fig. 16a, is shown the simulated impulse response of the fundamental mode of an RF cavity. The high frequency component is at the resonant frequency and the slow exponential decay of the amplitude is related to the relatively high cavity $Q$. Fig. 16b shows the impulse response of the bandpass filter. The ringing signal is at 3 GHz, the filter center frequency, and the inverse of the time width of the main tone burst corresponds to the frequency width of the filter. In Fig. 16c shows the impulse response of a long cable, where the long delay of the cable has been removed using a NWA port extension. An ideal lossless cable would have an delta function impulse response. The real dispersion of a cable spreads the frequency components of the signal in time resulting in a reponse as shown.

## 4 Acknowledgements

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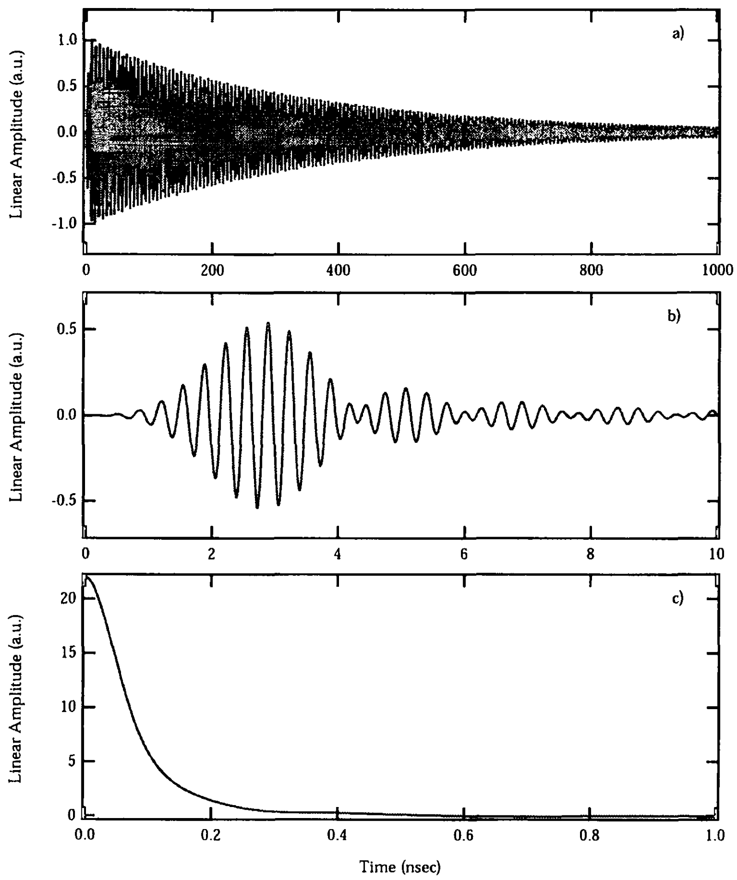


Figure 16: Examples of synthetic time domain on measured on an HP8510 NWA. a) Impulse response of the fundamental mode. b) Impulse response of a bandpass filter. c) Impulse response of a long cable with a port extension set to remove the long time delay.