Understanding Vector Network Analysis

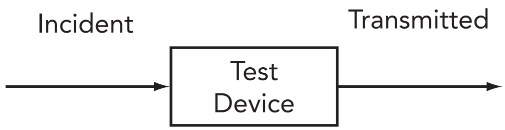
www.anritsu.com

www.tehencom.com

In this Understanding Guide we will introduce the basic fundamentals of the Vector Network Analyzer (VNA). Specific topics to be covered include phase and amplitude measurements, scattering parameters (S-parameters), and the polar and Smith chart displays.

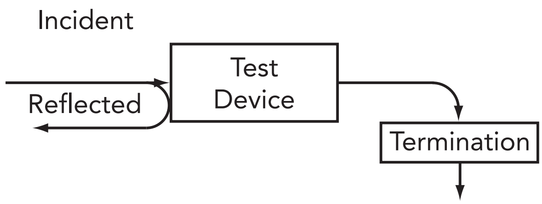
VNA Basics

The VNA measures the magnitude and phase characteristics of networks, amplifiers, components, cables, and antennas. It compares the incident signal leaving the analyzer with either the signal that is transmitted through the test device or the signal reflected from its input. Figure 1 and Figure 2 illustrate the different types of measurements that the VNA can perform.



- Gain (dB)
- Insertion Loss (dB)
- Insertion Phase (degrees)
- Transmission Coefficients (S_{21}, S_{12})
- Complex Transmission Components (Magnitude and Phase)
- Electrical Length (m)
- Electrical Delay (s)
- Deviation from Linear Phase (degrees)
- Group Delay (s)

Figure 1. The VNA can make a wide range of transmission measurements.



- Return Loss (dB)
- Reflection Coefficients (S_{11}, S_{22})
- Reflection Coefficients vs Time (Fourier Transform)
- Impedance (R+jX)
- SWR

Figure 2. The VNA can make numerous different reflection measurements.

VNAs are self-contained, fully-integrated measurement systems that include additional measurement capabilities such as time domain and group delay. The system hardware consists of the following:

- An analyzer
- Precision components required for calibration and performance verification
- Optional use of synthesizers as a second source
- Optional use of power meters for test-port leveling and calibration

The VNA internal system modules perform the following functions:

Source Module - The source module provides the stimulus to the device under test (DUT). The frequency range of the source and test set modules establish the frequency range of the system. The frequency stability of the source is an important factor in the accuracy (especially phase accuracy) of the network analyzer. Some VNA sources operate in either analog-sweep mode or step-sweep mode. Analog-sweep mode provides a faster measurement time. However, since the signal is not locked to a stable reference, measurement stability—especially phase - will suffer. For proper measurement accuracy, the VNA should always be operated in a step sweep, phase-locked condition. The difference is critical enough that all Anritsu VNAs provide only a step-sweep mode and do not provide the ability to unlock the source.

Test Set Module - The test-set module routes the stimulus signal to the DUT and samples the reflected and transmitted signals. The type of connector used is important, as is the "Auto

Reversing” feature. Auto Reversing means that the stimulus signal is applied in both the forward and reverse directions. The direction is reversed automatically. This saves the engineer from having to physically reverse the test device to measure all four S-parameters. It also increases accuracy by reducing connector repeatability errors. Frequency conversion to the Intermediate Frequency (IF) range also occurs in the test-set module.

Analyzer Module - The analyzer module receives and interprets the IF signal for phase and magnitude data. It then displays the results of this analysis on a high-resolution display screen. This display can show all four S-parameters simultaneously, as well as a variety of other forms of displayed information such as group delay, time and distance, and complex impedance information. In addition to the installed display, the engineer can also view the measurement results on an external monitor.

Network Analyzers

This discussion of networks analyzers begins with a subject familiar to most analyzer users: scalar network analysis. After showing comparisons, the fundamentals of network analyzer terminology and techniques will be presented. This discussion serves as an introduction to topics presented in greater detail later in this chapter and will touch on the concepts of:

- Reference delay
- S-Parameters: what they are and how they are displayed
- Complex impedance and Smith charts

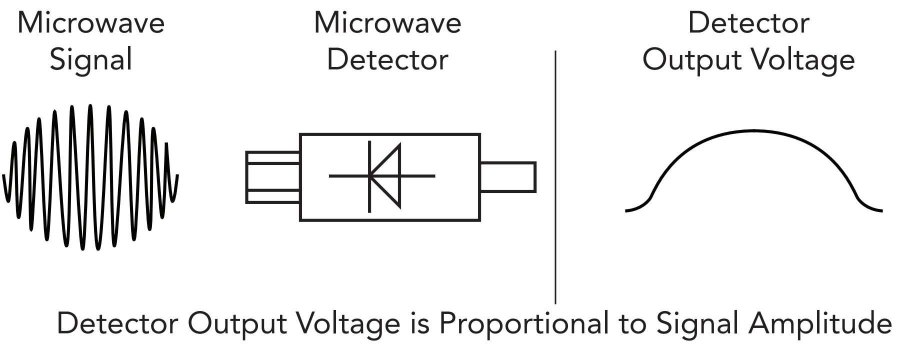


Figure 3. Scalar network analyzers measure microwave signals using the detection process shown here.

Scalar Analyzer Comparison

VNAs do everything that scalar analyzers do, plus they add the ability to measure the phase characteristics of microwave devices over a greater dynamic range and with more accuracy. If all a VNA added was the ability to measure phase characteristics, its usefulness would be limited. While phase measurements are important, the availability of phase information provides the potential for many new complex-measurement features, including Smith charts, time domain and group delay. Phase information also allows greater accuracy through vector-error correction of the instrument's own mismatches so that the instrument's own uncorrected characteristics do not influence the actual DUT response.

Now consider the scalar network analyzer (SNA), an instrument that measures microwave signals by converting them to a DC voltage using a diode detector (Figure 3). This DC voltage is proportional to the magnitude of the incoming signal. The detection process, however, ignores any information regarding the microwave signal's phase. Also, a detector is a broadband-detection device which means that all frequencies, the fundamental, harmonics, sub-harmonics, and any other spurious signals within the bandwidth of the detector, are detected and simultaneously displayed as one signal. This may add significant error to both the absolute and relative measurements.

In a VNA, information regarding both the magnitude and phase of a microwave signal is extracted. While there are different ways to perform this measurement, the method employed by the Anritsu series of VNAs is to down-convert the signal to a lower IF in a process called harmonic sampling. This signal can then be measured directly by a tuned receiver. The tuned receiver approach gives the system greater dynamic range due to its variable IF-filter bandwidth control. The system is also much less sensitive to interfering signals, including harmonics.

VNA Fundamentals

The VNA is a tuned receiver (Figure 4).

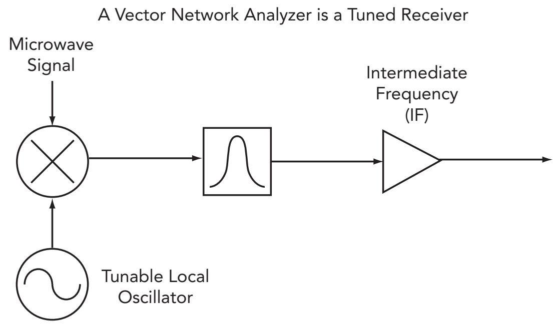


Figure 4. The network analyzer is a tuned receiver.

The microwave signal is down converted into the passband of the IF. To measure the phase of this signal as it passes through the DUT, a reference is needed for comparison. If the phase of a signal is 90 degrees, it is 90 degrees different from the reference signal (Figure 5). The VNA reads this as –90 degrees, since the test signal is delayed by 90 degrees with respect to the reference signal. The phase reference can be obtained by splitting off a portion of the microwave signal before the measurement (Figure 6).

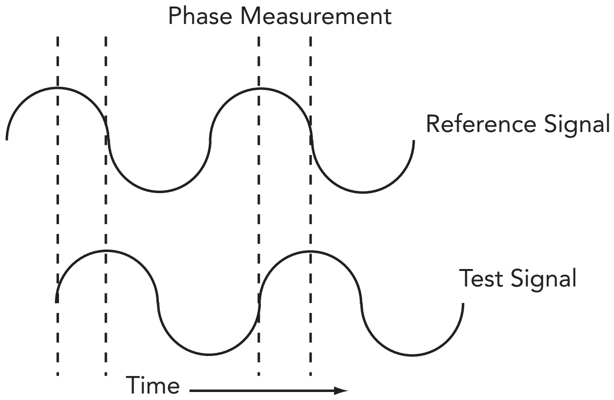


Figure 5. Shown here are signals with a 90-degree phase difference.

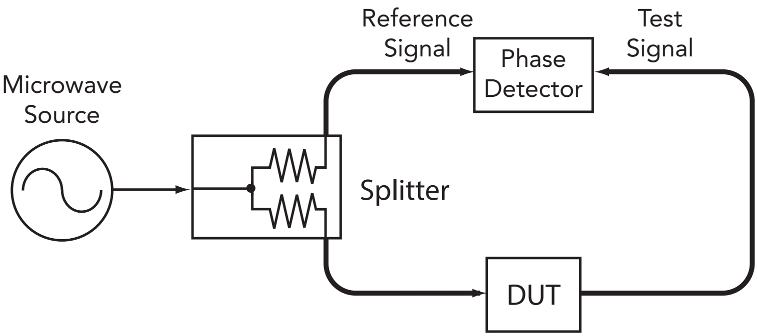


Figure 6. Splitting the microwave signal to obtain the phase reference.

The phase of the microwave signal, after it has passed through the DUT, is then compared with the reference signal. A network-analyzer test set automatically samples the reference signal so no external hardware is needed.

Consider the case where the DUT is removed and a length of transmission line is substituted (Figure 7). Note that the path length of the test signal is longer than that of the reference signal. How does this affect the measurement?

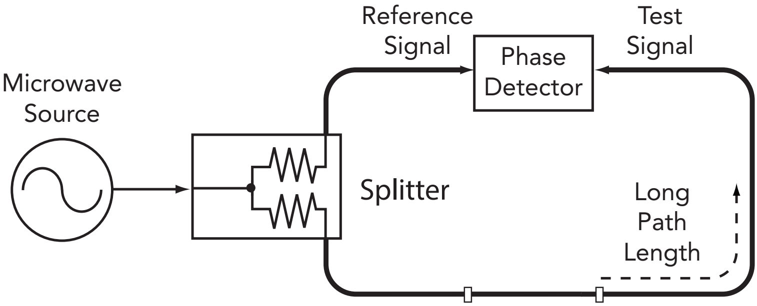


Figure 7. Pictured here is the case of a split signal where a length of line replaces the DUT.

To answer this question, assume that a measurement is being made at 1 GHz and that the difference in path length between the two signals is exactly 1 wavelength. This means that the test signal lags behind the reference signal by 360 degrees (Figure 8). It is impossible to tell the difference between one sine wave maxima and the next because they are all identical. Consequently, the network analyzer measures a phase difference of 0 degrees.

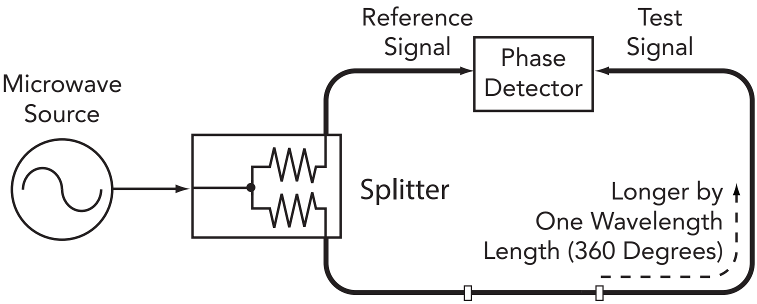


Figure 8. Depicted here is the case for a split signal where the path length differs by exactly one wave-length.

Now, consider that this same measurement is made at 1.1 GHz. Since the frequency is higher by 10 percent, the wavelength of the signal is shorter by 10 percent. The test-signal path length is now 0.1 wavelength longer than that of the reference signal (Figure 9). This test signal is: 1.1 x 360 = 396 degrees. This is 36 degrees different from the phase measurement at 1 GHz. The network analyzer displays this phase difference as –36 degrees. The test signal at 1.1 GHz is delayed by 36 degrees more than the test signal at 1 GHz.

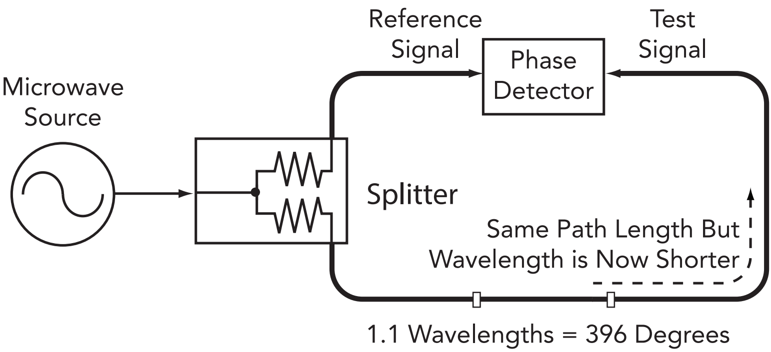


Figure 9. Depicted here is the case for a split signal where the path length differs by the same path length, but the wavelength is now shorter. Note that 1.1 wavelengths = 396 degrees.

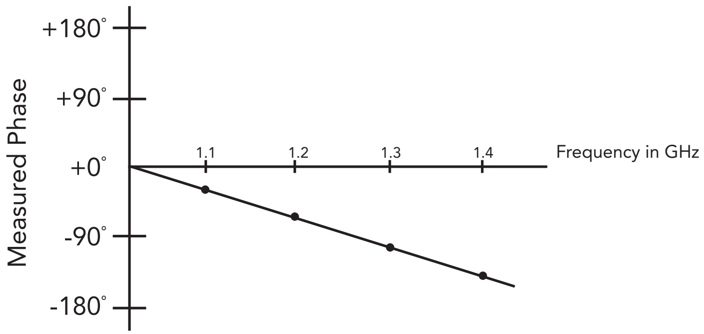


Figure 10. This graphic depicts electrical delay.

A measurement frequency of 1.2 GHz produces a reading of –72 degrees, while 1.3 GHz produces a reading of –108 degrees (Figure 10). There is an electrical delay between the reference and test signals. This delay is commonly referred to in the industry as the reference delay. It is also called phase delay. In older network analyzers, the length of the reference path had to be constantly adjusted—relative to the test path—to make an appropriate measurement of phase versus frequency.

To measure phase on a DUT, this phase-change versus frequency due to changes in the electrical length must be removed. This allows the actual phase characteristics to be viewed. These characteristics may be much smaller than the phase change due to electrical-length difference.

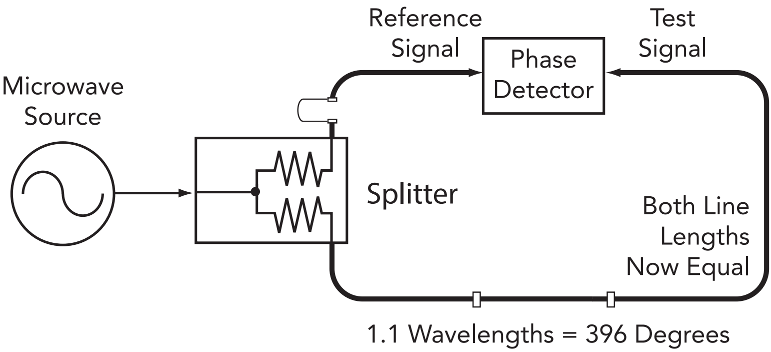


Figure 11. This graphic depicts a split signal where paths are equal in length.

The second approach involves handling the path-length difference in software. Figure 12 displays the phase versus frequency of a device. This device has different effects on the output phase at different frequencies. Because of these differences, the phase response is not perfectly linear. This phase deviation can be easily detected by compensating for the linear phase. Because the size of the phase difference increases linearly with frequency, the phase display can be modified to eliminate this delay.

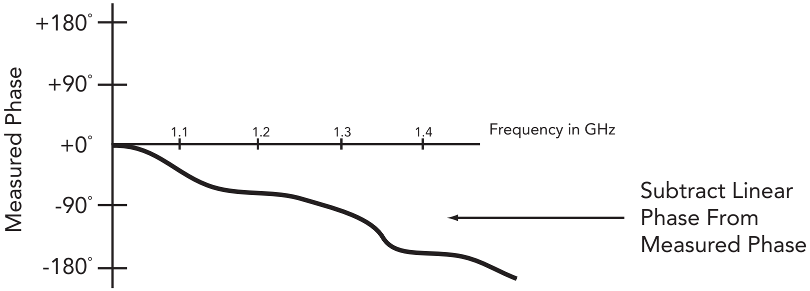


Figure 12. Here phase difference increases linearly with frequency.

Anritsu VNAs offer automatic reference-delay compensation with the push of a button. Figure 13 shows the resultant measurement when the path length is compensated.

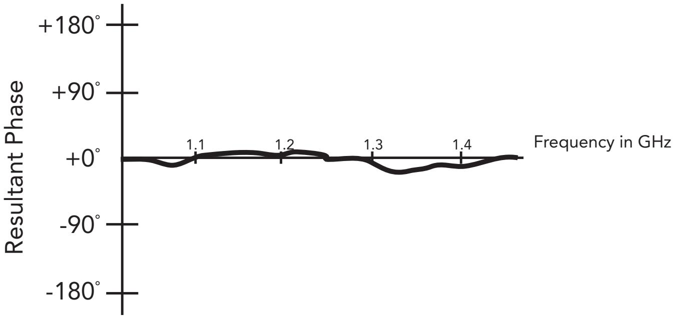


Figure 13. Shown here is the resultant phase with path length.

Network Analyzer Measurements

Now, consider measuring the DUT using a two-port device; that is, a device with a connector on each end. What measurements would be of interest? First, measure the reflection characteristics at either end with the other end terminated into 50 ohms. If one of the inputs is designated as Port 1 of the device, then this is the reference port. Next, define the reflection characteristics from the reference end as forward reflection, and those from the other end as reverse reflection (Figure 14).

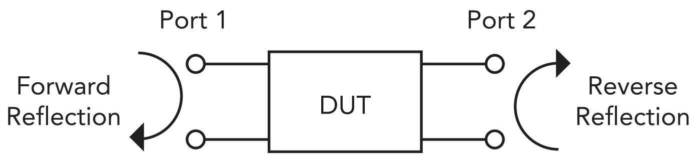


Figure 14. Forward and reverse reflection measurements are shown in this graphic.

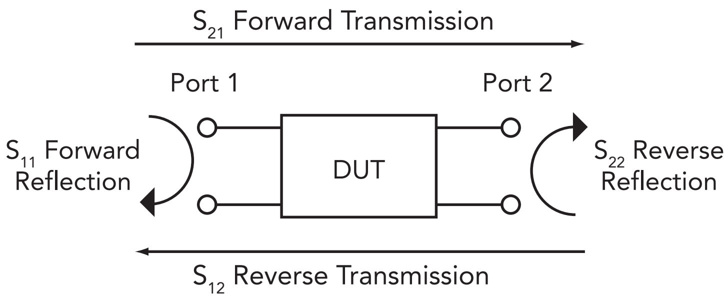


Figure 15. The four scattering parameters are shown in this figure.

Second, measure the forward and reverse transmission characteristics. However, instead of saying "forward," "reverse," "reflection," and "transmission" all the time, a shorthand can be used—S-parameters. In a typical S_{XX} nomenclature, the "S" stands for scattering, the first number is the device port which is measuring the signal and the second number is the device port which is sourcing the signal. As an example, S_{21} is Port 2 measuring, and Port 1 is sourcing or transmitting. The four scattering parameters for a 2-port device shown in Figure 15, are:

- S_{11} forward reflection
- S_{21} forward transmission
- S_{22} reverse reflection
- S_{12} reverse transmission

S-parameters can be displayed in many ways. An S-parameter consists of a magnitude and a phase. The magnitude is displayed in dB, just like the SNA, and is called the log magnitude. Another method of magnitude display is to use units instead of dB. When displaying magnitude in units, the value of the reflected or transmitted signal will be between 0 and 1 relative to the reference.

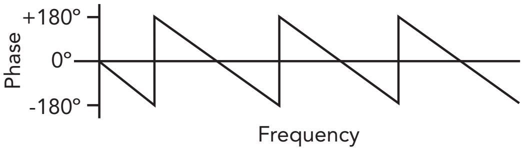


Figure 16. This waveform depicts linear phase with frequency.

Phase can be displayed as "linear phase" (Figure 16). As discussed earlier, it's impossible to tell the difference between one cycle and the next. Therefore, after going through 360 degrees, the engineer ends up back at the beginning. Displaying the measurement from –180 to +180 degrees is a more common approach. This method keeps the display discontinuity removed from the important 0 degree area which is used as the phase reference.

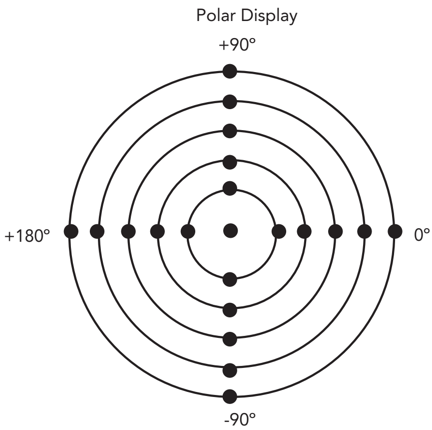


Figure 17. An example of a polar display.

There are several ways in which all the information can be displayed on one trace. One method is a polar display (Figure 17). In this display, the radial parameter (e.g., distance from the center) is magnitude, while the rotation around the circle is phase. Polar displays are sometimes used to view transmission measurements, especially on cascaded devices (e.g., devices in series). The transmission result is the addition of the phase and log magnitude (dB) information of each device's polar display.

As previously discussed, the signal reflected from a DUT has both magnitude and phase. This is because the impedance of the device has both a resistive and a reactive term of the form R+jX, where R is the real or resistive term and X is the imaginary or reactive term. The j, which is sometimes denoted as i, is an imaginary number and is the square root of –1. If X is positive, the impedance is inductive. If X is negative, the impedance is capacitive.

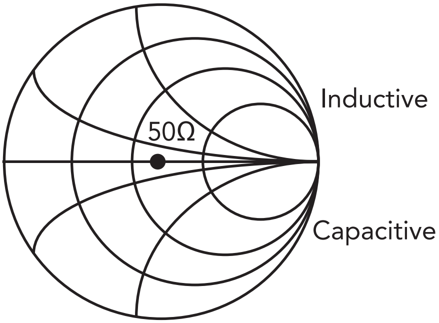


Figure 18. An example of a Smith chart.

The size and polarity of the reactive component X is important in impedance matching. The best match to a complex impedance is the complex conjugate. This complex sounding term simply means an impedance with the same value of R and X, but with X of opposite polarity. This term is best analyzed using a Smith chart which is a plot of R and X (Figure 18). Displaying all of the information on a single S-parameter requires one or two traces, depending on the format desired. A very common requirement is to view forward reflection on a Smith chart (one trace), while observing forward transmission in log magnitude and phase (two traces). In addition, it is also very helpful to simultaneously display other displays such as Time Domain and Group Delay while monitoring the S_{11} and S_{21} parameters. This requires a VNA with a complex display capability such as the any of the Anritsu benchtop VNAs.

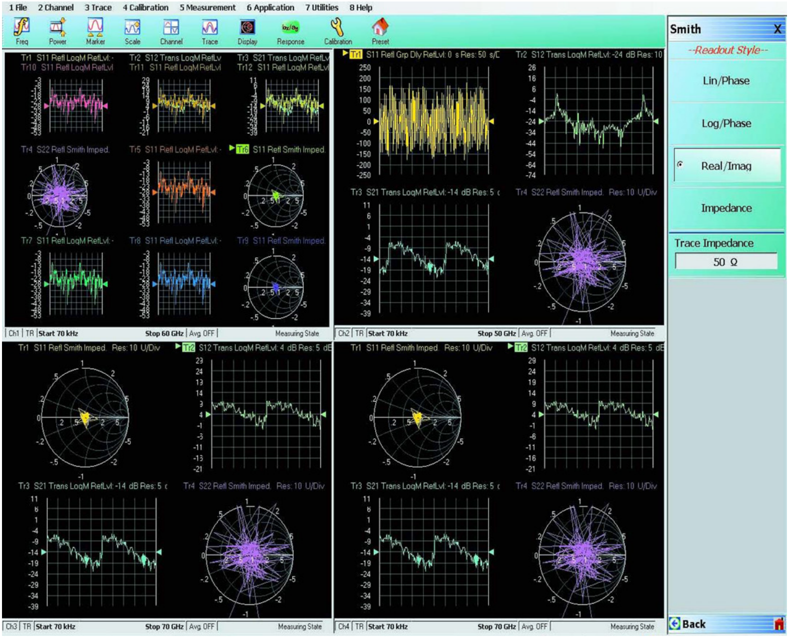


Figure 19. Shown here is the VectorStar multiple channel and multiple trace display.

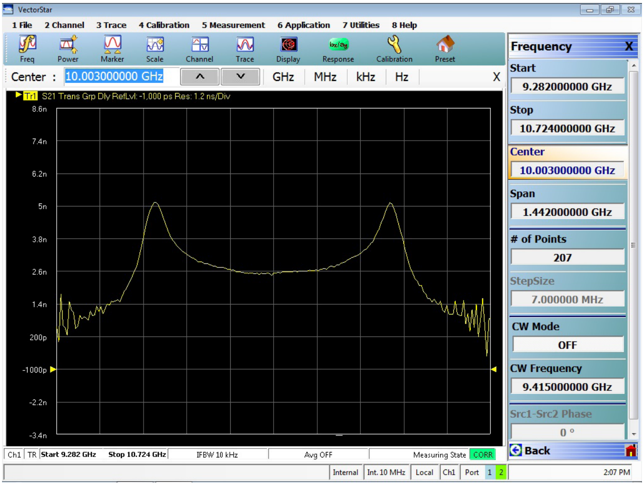


Figure 20. The VectorStar display of a group delay measurement of a wireless communication filter.

The VectorStar and ShockLine series VNAs have the ability to configure one to sixteen channels. Each channel can thought of as an independent VNA. Therefore, a channel in the VectorStar or Shockline can be configured for a specific frequency range, calibration type, power level, and IF-filter bandwidth setting. Additional channels can be configured within the VNA to help facilitate testing in multiple setup parameters, up to the 16 channels available. The VectorStar or Shockline can also be configured to display all of the active channels in a pattern that is most useful to the user.

With multiple channels configured, the Anritsu VNA sequentially sweeps from one calibrated setup and measurement condition to the next. Each of the channel displays can be configured to accept up to 16 traces. All of the traces in each channel can likewise be configured for the most beneficial display pattern. Finally, each of these traces can be configured for an appropriate display, depending on the type of data being displayed (Figure 19). For instance, Trace 6 can be set up to provide S_{11} performance of the device displayed on a Smith chart, Trace 7 can be set up for S_{11} with a Time Domain display, and Trace 12 can be set up for an S_{21} display on a Log Magnitude and Phase graph.

Another important parameter that can be measured with phase information is group delay (Figure 20). In linear devices, the phase change through the DUT is linear-with-frequency. Thus, doubling the frequency also doubles the phase change. An important measurement, especially for communications system users, is the rate of change-of-phase versus frequency (e.g., group delay). If the rate of phase-change versus frequency is not constant, the DUT is nonlinear. This nonlinearity can create distortion in communications systems.

Measurement Error Correction

Since microwave signals can be measured in both magnitude and phase, it is possible to correct for six major error terms:

- Source test-port match
- Load test-port match
- Directivity
- Isolation
- Transmission frequency response
- Reflection frequency response

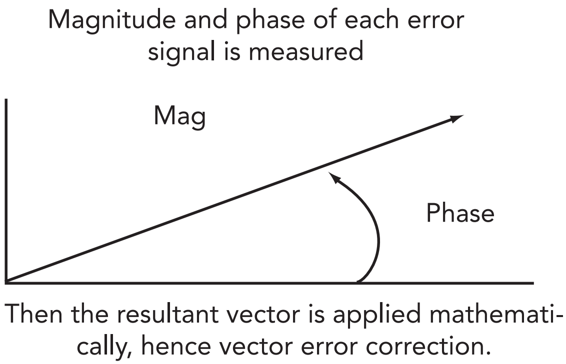


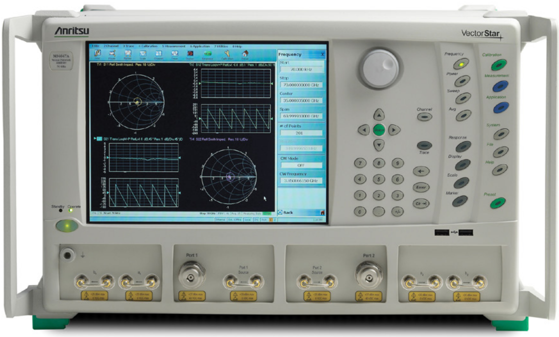
Figure 21. Magnitude and phase information of each error signal is measured and available for removal during the measurement.

It is possible to correct for each of these six error terms in both the forward and reverse directions, hence the name 12-term error correction. Since 12-term error correction requires both forward and reverse measurement information, the test set must be reversing. “Reversing” means that it must be able to apply the measurement signal in the forward or reverse direction automatically, without having to disconnect the calibration components or DUT.

To accomplish this error correction, measurement of the magnitude and phase of each error signal occurs during calibration of the VNA (Figure 21). There are a number of different types of calibration processes some of which will be discussed in further detail in Understanding VNA Calibration guide. Once the magnitude and phase information of the error vectors are obtained during calibration, they are mathematically removed from the measurement signal - a process termed vector-error correction.

Summary

Compared to the SNA, the VNA is a much more powerful analyzer. The major difference is that the VNA adds the ability to measure phase, as well as amplitude. With phase measurements come S-parameters, which are a shorthand method for identifying forward and reverse transmission and reflection characteristics. The ability to measure phase introduces two new displays, polar and Smith chart. It also adds vector-error correction to the measurement trace. With vector-error correction, errors introduced by the measurement system are compensated for and measurement uncertainty is minimized. Phase measurements also add the capability for measuring group delay, which is the rate of change-of-phase versus frequency (e.g., group delay). All in all, using a network analyzer provides for a more complete analysis of any test device.



VNA Overview

This section is designed to introduce the reader to the fundamentals of VNA architecture, measurements, specifications, and performance. In it, the major functional analog blocks in a RF/microwave VNA are examined, along with the current and past technologies which are used in these blocks. This content will serve as the foundation on which more advanced concepts will later be developed.

VNA Architecture

In its most basic form, the objective of a VNA is to capture S-parameter data. Most VNA's today are designed for use with 50 ohm impedance DUT's. The basic requirements for capturing S-parameter data include:

- One or more signal sources with, at minimum, controllable frequency and a sufficiently-clean spectral tone for making a measurement. Sources with controlled power are preferred.
- Directional devices - whether physically or computationally directional - for separating incident and reflected waves at the ports.
- A means of switching RF signals when there are fewer sources than ports, or either more or fewer receivers than ports.
- One or more receivers, usually with down converters, to take incident and reflected waves down to some convenient IF for processing.
- An IF section and digitizer to transform the converted wave amplitudes into a useful form for computation.

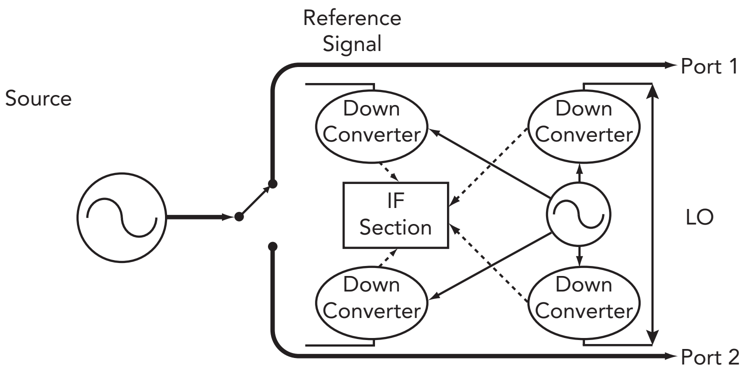


Figure 22. Shown here is one possible VNA block diagram, illustrating the key blocks to be discussed in this section.

A simplified block diagram of a common VNA is shown in Figure 22. Many variations on this structure are possible. For example, one source could be used for each port instead of switching a single source between two ports, or the coupling devices could be repositioned. The diagram therefore, is simply intended to point out that the functions listed above are generally all present in one form or another and play a role in overall system performance.

Sources

It might be tempting to view the VNA source somewhat simplistically—as a synthesizer or sweeper somehow synchronized with a local oscillator (LO) to produce the desired transceiver behavior. In practice, there are many competing demands on performance and consequently, the source is based on a complex series of decisions. Speed, for example, is often times a critical component of the modern VNA, while spectral purity is typically not as critical since the engineer knows what frequency he/she is trying to measure. The engineer cannot deviate too far, however, because the phase noise of the sources involved affects trace noise, spurring the sources to produce undesired receiver responses which impacts net dynamic range. Similarly, the source is impacted by the desired application. Because the VNA may also be used to make mixer measurements and multi-tone analysis, any architectural choices pertaining to the source become more constrained (e.g., spectral purity again becomes important).

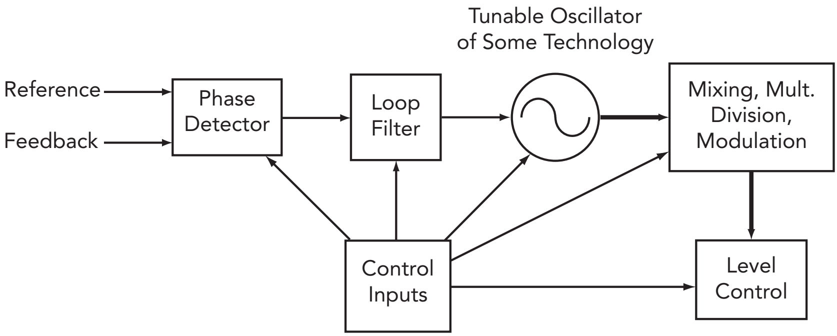


Figure 23. A generic source block diagram illustrates some of the architectural choices to be made by the engineer.

A simplistic block diagram of a source is shown in Figure 23. The diagram is extremely general since there are many possible structures. Note that if large portions of the source are digitally generated this diagram is not accurate. In the diagram, the reference comes from the system's crystal oscillator or synthesizer, while feedback comes from either the tunable oscillator (e.g., a classical phase-locked loop (PLL)) or elsewhere in the system. The loop filter may be static or of dynamic bandwidth and gain. The oscillator output may be frequency-converted or modulated. In some cases, the source may not even be locked—although this decision incurs an accuracy penalty.

Using Figure 23 as a reference, the first issue to be determined is the global source architecture. The very first VNAs essentially relied on sweepers (that may have locked an initial point) and an analog ramp to perform the sweep. This type of structure can be quite fast and devoid of some spectral artifacts, but there is a downside. Controlling the timing relationships between the source, LO and acquisition is challenging. Doing so over temperature and aging can be even more difficult, requiring sophisticated internal calibration structures. Changing the sweep dynamics in mid-measurement (e.g., changing averaging) disturbs the timing relationship and can lead to measurement distortions. Integrating more complicated applications (e.g., mixer measurements which require external sources) adds additional challenges.

A fully-synthesized approach to the source architecture avoids most of these problems but comes at the expense of control complexity. Making a fully-synthesized version fast and with good spectral purity, requires very careful loop design and even more sophisticated control electronics. While this section will not cover the details of PLL and synthesizer design, the following key points can aid in this discussion:

- In general, the wider the loop bandwidth, the faster the settling time and therefore, the more phase noise.
- From a measurement point-of-view, the settling of the final receiver IF is more important than the independent settling of the source and LO. If the source and LO settle together, faster net measurement times are possible; although the engineer must be on frequency to within a certain tolerance.
- Generally the loops settle faster, to within a fixed tolerance, for smaller frequency steps. For larger steps, dynamic loop response changes can help.

At some point in the process, locking is usually required. Therefore, the next issue to be resolved is where to perform the locking. The simplest approach is to treat both the source and LO as separate synthesizers, with there own integrated PLLs and with shared reference frequencies at some level. Another possibility involves locking through the receiver—essentially locking the LO to the IF or locking the source to the IF, sometimes referred to as follower mode and source-locking, respectively. Since the IF becomes the locking reference, this approach reduces the individual loop complexities and leads to a clean received signal. But, because one receiver is the locking parent and hence, no longer has meaningful phase information, this approach complicates the application space. Also, the source and LO must have a fixed offset for the loop to close, making mixer, intermodulation distortion (IMD) and other translating measurements quite difficult.

The various locking paths are shown in Figures 24A, 24B and 24C. In each case “Ref B” is a secondary reference usually derived from the main system frequency reference (reference A). In some cases, the two references can be the same signal but this is not necessary or common.

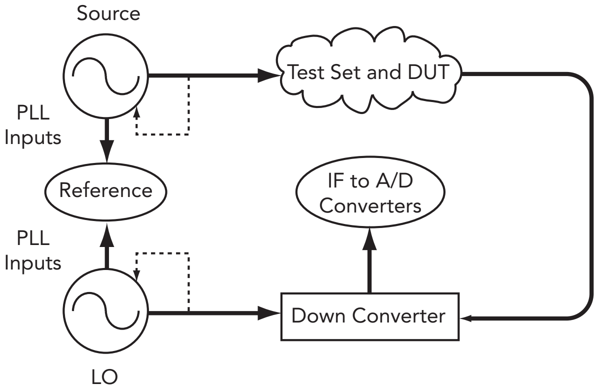


Figure 24A. This diagram illustrates a fully-synthesized architecture where the source and LO PLLs are semi-self-contained. The source and PLLs may be linked together at any level and even derived from each other, but they do not involve the system receiver.

Many decisions have to be made with regard to the individual PLLs. While most of these decisions are beyond the scope of this section, a few comments are worth noting. Historically, Yttrium-iron-garnet (YIG) oscillators have been used as the source in source-locking architectures like the one shown in Figure 24B. While the phase noise of such oscillators is quite good, they tend to be slow (at least in broadband configurations). More recently, VNAs have emerged in which all sources are based on varactor-tuned, voltage-controlled oscillators (VCOs) and can therefore tune much faster. The trade-off is degraded phase noise at offsets that is much larger than the loop bandwidth. As VCO technology improves, these differences have been shrinking.

The fine-tuning structure of these loops has also changed in recent history. Fractional-N structures are very popular and can offer fine-tuning resolution with decent spurious and noise performance. Wide-bandwidth, direct-digital synthesizers have also become more common recently, thanks to their ever-improving spurious performance. The fine-tuning capabilities of such structures are needed since the VNA tuning resolution must typically be on the order of 1 Hz.

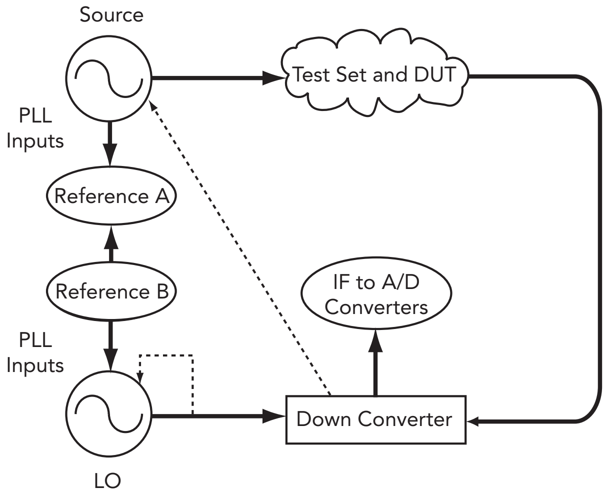


Figure 24B. A source-locking architecture is shown here. The source is locked to the IF through the down converter. Since the loop tries to maintain a clean IF, any noise on the LO is transferred, in a canceling fashion, to the source.

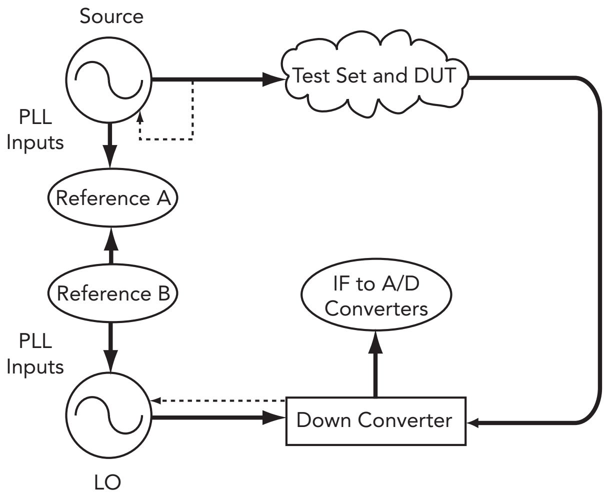


Figure 24C. A follower architecture is shown here. In this case, the LO is locked to the IF. Since the loop tries to keep the IF clean, noise on the source is transferred, in a canceling fashion, to the LO.

Another important aspect of the system's source is power control. Aside from having a vague idea of what the DUT is being driven with, swept-power measurements are increasingly important to the VNA user for measurements such as gain compression, IMD versus power and harmonics versus power. A reasonably accurate and wide-range leveling system (ALC) is therefore crucial. Complicating matters, the leveling system must be fast enough to keep up with the measurement.

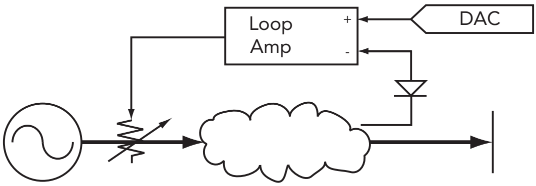


Figure 25. A very simplified ALC loop is shown here.

Leveling subsystems are used in many applications and are conceptually quite simple. As shown in Figure 25, an ALC loop employs a power detector used in the context of a negative-feedback loop. The detected output is compared to some desired reference voltage (usually from a digital-ton-analog converter (DAC)), and the result is then fed to a power modulator.

For the purposes of this example, a number of assumptions have been built into Figure 25 that are not mandatory:

A coupled detector is used for power detection.

- Although this is commonly done and the directional device improves match dependence, non-directional take-offs are sometimes used if power detection occurs far enough from the user port.
- Arrays of detectors are sometimes used for improved control range. A thermal sensor can also be used, although the speed penalty is severe.

A variable attenuator is used for power modulation.

- Amplifier bias is also sometimes used, although the engineer must watch for harmonic generation as the requested power is reduced.
- Cold field-effect transistor (FET) and positive-intrinsic-negative (PIN) diode attenuators are both popular. As with switches, PIN-diode structures often have an advantage in power handling, while FET structures perform better at low frequencies. Note that there are exceptions to this generalization. Variations and hybrids are, of course, possible.

A simple-loop amplifier, often an integrator, is used.

- Multi-stage and distributed-loop amps are often used for more control of loop gain.
- Variable poles are often used for stability in different operating modes.

The issue of loop bandwidth is an important one to consider. Since a VNA has to operate over wide frequency ranges and often times over wide power ranges, the overall loop gain is not flat. As an example, consider the attenuation curve of a commercially-available, voltage-variable attenuator (VVA) (Figure 26).

The slope variations in this curve represent changes in loop gain. If this gain was uncompensated, the bandwidth of the loop could become very small at some states (making the measurement slow), and very large at other states (potentially leading to oscillation). In addition to simple level-dependent gain changes, there may be other frequency-dependent gain changes.

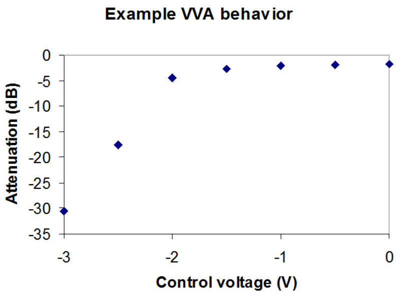


Figure 26. An example response curve of a commercial voltage-variable-attenuator is illustrated in this graphic.

Moving from a fundamental-source band to a multiplied-source band that may use a different VVA is just one example. Since detectors have non-linear responses over wide power ranges, some linearization is desirable to keep loop gain relatively flat. The modulator, or VVA, may also need special drive (e.g., high current and scaled voltage ranges); therefore additional driver stages will be required. Given these complications, the actual leveling system may look more like Figure 27.

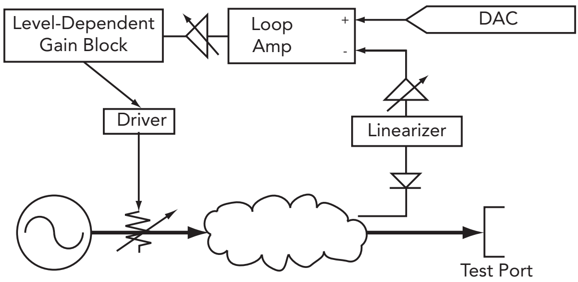


Figure 27. A more complete ALC block diagram is shown here.

Switches

RF switches are needed in VNAs for a number of reasons:

- To allow one source to drive two or more ports, saving the expense of multiple sources.
- Routing to multiple receivers; either in a multi-port scenario or to different application-specific receivers.
- Switching between different bands—either sources, local oscillators or receivers.

The demands on the switches can be extreme in terms of isolation, insertion loss, bandwidth, and in some situations, power handling/linearity. Consider a 2-port VNA in which there is a main switch (normally called a transfer switch even if it is single-pole, double throw (SPDT)) allowing one source to drive port 1 or port 2. The isolation of this switch directly translates to the raw isolation of the VNA. Likewise, the insertion loss and linearity directly affect the maximum available port power, while its bandwidth typically limits that of the VNA. For a high-performance microwave VNA, this can be a challenging combination.

In the distant past, electromechanical (EM) switches were sometimes used due to their favorable insertion loss/isolation ratio. The repeatability of these switches—typically no better than a few hundredths to a tenth of a dB at microwave frequencies—led to measurement errors which could be problematic. Also, the lifetime of most mechanical switches does not exceed 10-million cycles. Even at a slow sweep rate of one sweep per second, such a switch would last less than 3000 hours.

Today, electronic switches (e.g., a PIN diode, cold FET circuit or some combination of the two) are normally used. It is beyond the scope of this discussion to analyze the device physics in detail, but a quick summary is provided below.

PIN Diode - A PIN diode consists of heavily-doped P and N layers surrounding a relatively thick intrinsic layer. Because of this thickness, the diode's reverse-biased capacitance is quite low compared to other diode types. This results in better isolation when used in a series construction and less insertion loss in a shunt topology. When forward biased, carriers are injected into the intrinsic layer but do not recombine immediately. This causes complications at lower frequencies since the RF frequency can be on the same scale as the recombination rate and distortion occurs.

Cold FET Switch - A typical cold FET switch is just that: a Metal Epitaxial Semiconductor Field Effect Transistor (MESFET) or similar structure setup with no drain bias. When the gate is biased strongly negative, no carriers are available in the channel and the device provides reasonable isolation in a series sense. Like the PIN diode, the off capacitance (drain to source) of the cold FET is fairly low due to the geometry. Shunt-topology insertion losses are therefore low as well, although they are typically worse than with a PIN structure. The elevated capacitance can be mitigated by embedding the switch in a transmission-line structure. When the gate is near ground potential, carriers are available in the channel, along with a relatively-low series resistance. Unlike the PIN diode, the recombination time remains fast so there are few low-frequency effects. Since the engineer is usually operating against a 0-bias limit, there can be linearity issues.

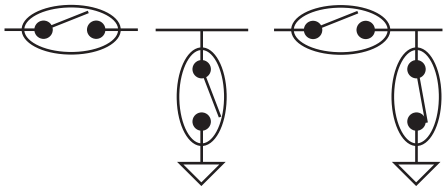


Figure 28. SPST topologies include series, shunt and series-shunt.

The common single-pole-single-throw (SPST) topologies are shown in Figure 28. It is not uncommon to have the series-shunt pair contained within a single die or package.

Which ever technology, or combination of technologies, is chosen, the issue of switch topology is critical. For simple applications requiring low isolation, a single series-shunt element per arm may be appropriate (Figure 29A). In some cases, a single element may even be used, but severe match implications on a multi-throw switch are possible. When high levels of isolation are required, more elements are often used per path (Figure 29B). There are many choices the engineer can make about the combinations of elements. Here are a few tips:

Series elements generally become less effective at higher microwave frequencies, requiring more shunt elements in that frequency range.

- Sometimes series-shunt pairs are available as a single die or cell and are often convenient to bias that way.
- Proper allocation for biasing inductors must be made, primarily for PIN switches. Their layout is critical since above 50 GHz or so, bias-circuit parasitics may contribute as much to insertion loss as the switch itself.
- Isolation may end up being limited by radiative effects, making housing design and layout quite critical.
- Switch spacing in higher-isolation structures is quite important due to the standing waves that appear between switches.
- Terminating switches are often needed which, in turn, demands the addition of a load branch at the output ports, although other approaches are possible.

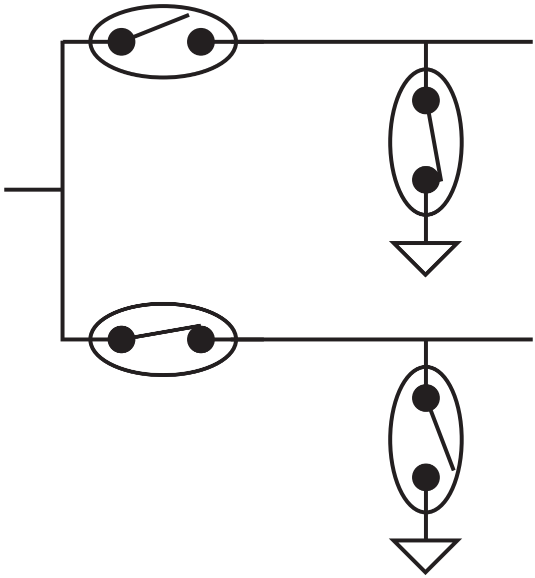


Figure 29A. A simple SPDT topology for fairly low isolation and insertion loss.

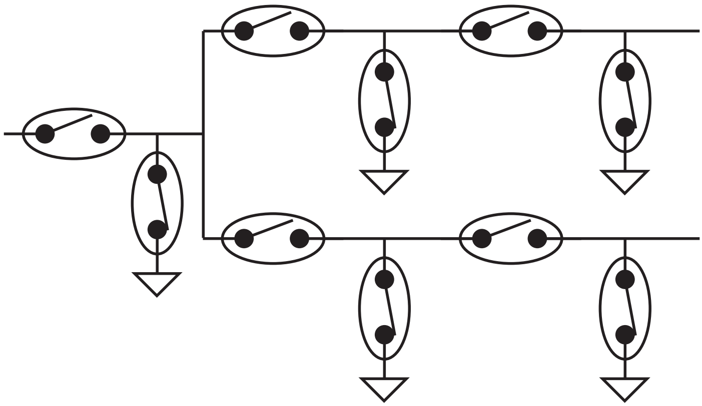


Figure 29B. A higher-isolation SPDT topology where a highly-isolating, double-off state is also desired.

An example of some of these behaviors is shown in Figure 30. Empirical curves of isolation and insertion loss for a particular pin-diode technology at 65 GHz are presented. Similar construction techniques were used in both cases, although not entirely held constant.

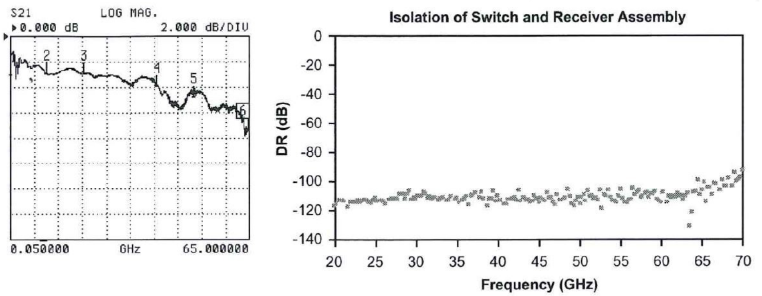


Figure 30. Example insertion loss and isolation of a broadband, high-isolation switch construction are shown here.

Directional Devices

Key to the concept of reflectometry is the directional device used to collect the incident and reflected signals. While there are a number of ways to do this, broader-band microwave VNAs tend to rely on directional couplers, while RF VNAs use a directional bridge—or some combination of the above concepts. Without delving into the theory of directional devices, a few definitions will be useful in this discussion. Figure 31 shows the port assignment required for these definitions. Note that reverse coupling is assumed in this figure.

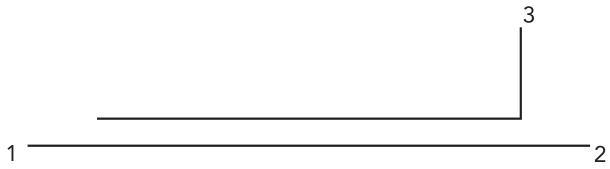


Figure 31. This graphic depicts an example of a coupler block.

Coupling - S_{32} Note that sometimes coupling is defined to include insertion loss, much like S_{31}, but with a perfect reflection connected.

Insertion Loss - S_{21}

Isolation - S_{31}

Coupling is usually dictated by the system's signal levels; subject to the constraint that directivity usually worsens (for a broadband coupler) if the coupling gets too tight. Coupling factors therefore, usually end up in the 10 to 20 dB range, although there are exceptions. Of course, a minimal forward insertion loss (to maximize available port power) and reasonable match (since this may dictate raw port match and is usually connected to directivity) are desired.

The wildcard, which is principally a function of the construction techniques and level of assembly tuning, is directivity or isolation. In view of the power of VNA calibrations, it is useful to understand the importance of these raw parameters to overall system performance.

Before exploring that question, it is necessary to revisit the concepts of residual versus raw parameters, such as directivity and source match. Raw parameters describe the physical performance of the components involved (e.g., the directivity previously described for the directional device). The residual directivity is defined as what is left after the calibration. It describes the quality of the calibration components, algorithm and the process. This concept is discussed in more detail in the calibration section of this section. The residuals at the time of the DUT measurement, describe the measurement uncertainty and not the raw parameters. An individual DUT may be sensitive to the raw parameters (e.g., an amplifier may or may not be stable for a given raw-port match on the VNA), but the measurement itself is largely invariant to them.

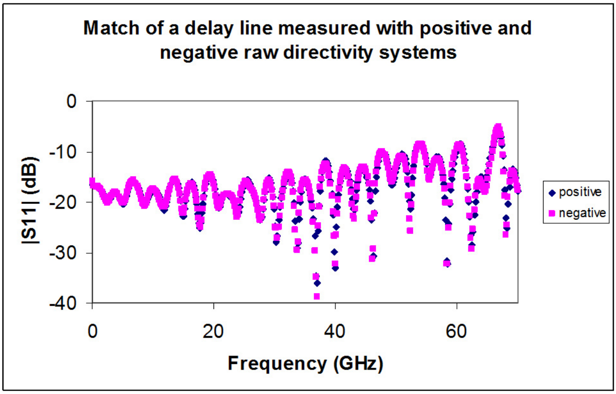


Figure 32. The impact of positive and negative raw directivities on a calibrated measurement is shown here. As long as the environment is stable, both calibrations are roughly equivalent.

To see this, consider two calibrations performed on a VNA. In the first calibration the VNA is configured normally with a raw directivity of about 10 to 15 dB across the 70 kHz to 70 GHz frequency band. The second calibration is performed with a 10 dB pad on the test port so that the raw directivity (ignoring pad mismatch, so this is an upper bound) is -5 to -10 dB. With each calibration, the return loss of the same delay line is measured. The results are shown in Figure 32 and indicate agreement—to within connector repeatability limits—even in the deepest notches. The residual directivities are therefore, nearly identical.

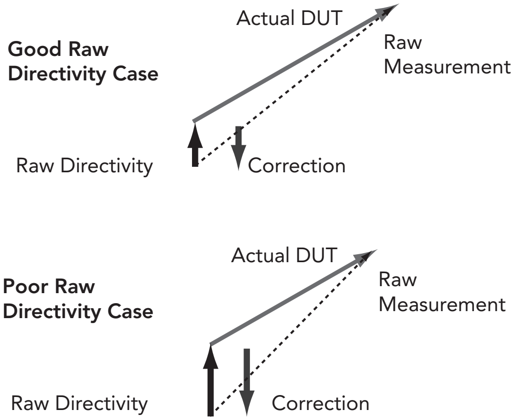


Figure 33. The mathematics of directivity correction is illustrated here for two different raw directivities.

In a practical sense, this is important since the raw parameters have a great deal of impact on the stability of the calibration. Consider the directivity correction. In a reflection measurement, the directivity error adds to the DUT's reflected wave to produce the net measurement. In the correction, the directivity is subtracted out along with other tasks being performed. If that subtraction is small in magnitude, a small drift in the actual amount of directivity will have little effect on the end result. If the subtraction is large, however, a fairly minor drift in the directivity vector results in a substantial change in the final result (Figure 33).

It is important therefore, to strive for the best directivity possible within the boundaries of the constraints. In the example of Figure 32, both measurements were made shortly after the calibrations. Had the delay line been measured several hours after the calibration, in a thermally-dynamic environment, the results might have been quite different.

Another constraint to consider is bandwidth. While the upper end is relatively easy to understand with the collapse of directivity under the wavelength limits, the low end is often misunderstood. Obviously, as the coupling section becomes electrically short, the coupling factor must typically fall and often does so at a 6-dB/octave rate. The available signal level decays rapidly and signal-to-noise becomes a problem. Directivity usually also suffers at this end, but more due to match problems than anything else.

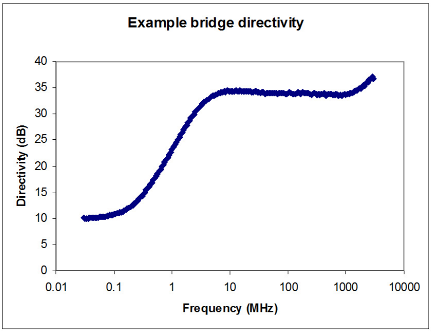


Figure 34. The directivity of a RF-bridge structure is shown here. Reasonable performance down to very low frequencies is possible with the right balun structure.

Bridges are a slightly different structure and do bear some resemblance to idea of the classical Wheatstone Bridge. The difficulty, from a RF point-of-view, is how to generate the non-ground referenced nodes. Typically this is done with a transmission-line balun, although there are other alternatives. The balun helps to set the bandwidth and parasitics of the lumped components being used.

Reasonable directivity can be maintained over large frequency ranges through proper balun design. The example shown in Figure 34 can be further optimized through the use of a more elaborate balun structure, although this optimization comes at the expense of some insertion loss.

Down Converters

While there is a great deal of content available regarding receiver design, this discussion will focus on VNA-specific topics and decisions, as well as general analyses. With a broadband-microwave VNA, the engineer must perform some means of down conversion. There are many decisions to make, including the number of down-conversion stages and with what frequency plan.

Classically, measuring receivers have used multiple up and down conversions to provide better image rejection and to allow a more flexible frequency plan for avoiding spurs. In a VNA, the image is usually less of a concern since there is one known signal present (which may be a harmonic) that the engineer wants to measure. As the application space changes to include IMD and other more complicated measurements, image-related issues become more prevalent, but there are “work-arounds.” Because sources are becoming increasingly clean and converters increasingly linear, spur problems are on the decline. Couple that with cost, complexity and again, with the situation of a single-known signal, and much of the reasoning behind multiple conversions diminishes. Depending on how the IF is implemented and for other signal processing reasons, two conversions are sometimes desired, but it has become less common these days to go beyond that. For reasons of stability, homodyne receivers have been avoided in recent years, but that may change as the adaptive-conversion circuitry used in commercial receivers improves.

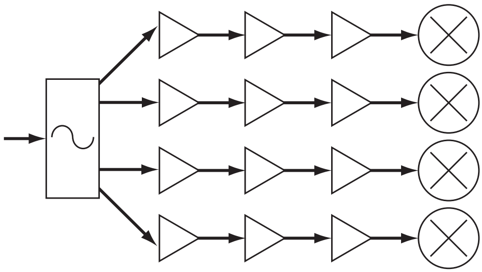


Figure 35. In this four-channel receiver architecture, a single LO is shared between four down converters. Amplifier chains are used to ensure channel-to-channel isolation. More or fewer amplifiers can be used and, in some circumstances, the engineer might even opt to use isolators or filters instead.

In an ideal scenario, the engineer is able to fundamentally down convert over the VNA's entire frequency range. This allows for the lowest spurious possibilities, the best receiver-noise figure and probably the best linearity. In a broadband-microwave system, this can get very expensive since the isolation chains must run over the full frequency range, provide enough LO power for the converter (10 to 20 dBm, typical), and provide 100 to 120 dB of round-trip isolation (Figure 35). The design engineer might opt to have a separate LO for each of the four or more converters, but this gets even more expensive and maintaining phase synchronization can be challenging (Figure 36). Some kind of harmonic-conversion process is therefore typically used to limit the required range of the LO.

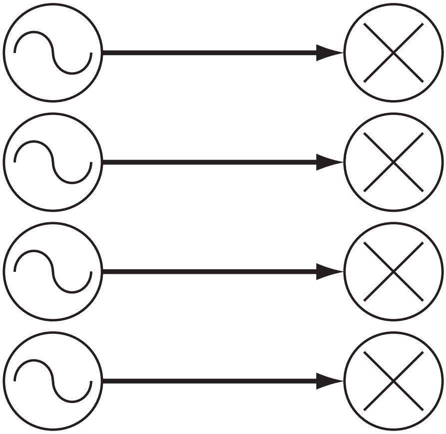


Figure 36. In this four-channel receiver, each down converter has its own LO. This can be expensive at higher frequencies and challenging in terms of having to ensure adequate phase synchronization between channels.

The common choices are mixers or samplers. Mixers are most commonly doubly balanced or double-double balanced, but many other configurations are possible. Samplers can be further subdivided into high-LO and low-LO configurations. The harmonic mixer and high-LO sampler have become quite similar in many regards and their distinctions now verge on the philosophical.

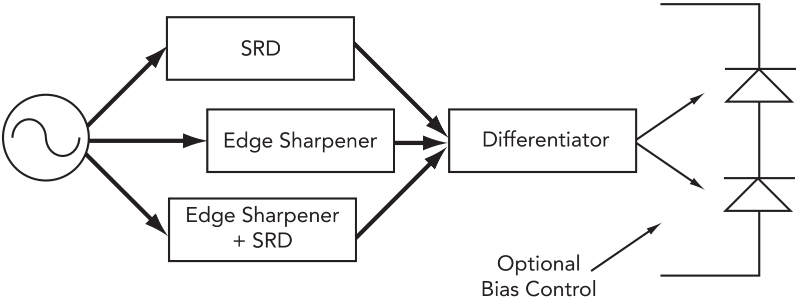


Figure 37. A flow chart of microwave-sampling structures is shown here. A sinusoidal LO feeds some edge-sharpening device which, in turn, feeds a differentiator to generate strobe pulses for the sampling diodes. When the narrow strobe hits, the RF is coupled through to the IF port. Additional bias control can be used to adapt the strobe responsivity.

In all cases, the process relies on something in the converter structure to generate a harmonic of the LO. This is mixed against the RF to produce the desired IF. In the case of the harmonic mixer, this is typically the switching action of the ring diodes and, for symmetry reasons, is usually limited to odd harmonics. Note that there are other mixer structures that favor even harmonics or even all harmonics, but those will not be discussed here. This can be viewed—in balanced mixers at least—as clipping the LO waveform to look more closely like a square wave; hence, the odd harmonics. In the limit of a perfect square wave, the harmonic content must go as 1/N, where N is the harmonics number.

In the case of a sampler, an additional nonlinearity such as a step-recovery diode, a shockline or some other construct is typically used to assist this process (Figure 37). These devices sharpen one edge of the incoming LO waveform which is then fed to a passive differentiator to produce sharp pulses. Because the order of nonlinearity is much higher, the harmonic content can be quite flat out to the bandwidth of the circuitry. This fundamental distinction between harmonic mixing and sampling, as they are normally defined, has both positive and negative impacts which will be discussed later.

The effect of harmonic content can be seen by looking at the conversion efficiency over frequency as plotted in Figure 38. For this example, the LO was limited to 10 GHz. A third harmonic mix was therefore required between 10 and 30 GHz, and a fifth harmonic mix was required between 30 and 50 GHz. The sampler conversion efficiency is flat across this range, while the mixer efficiency falls roughly as $20\log_{10}(N)$.

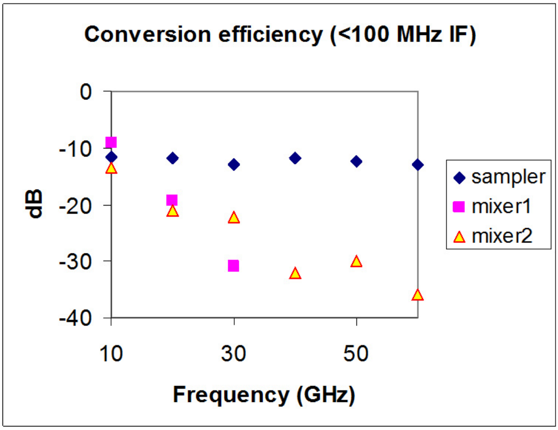


Figure 38. This graphic compares the conversion efficiency of a sampler and two example harmonic mixers. The LO is limited to 10 GHz so the harmonic mixers use harmonics 1, 3 or 5 in this picture.

The improvement in conversion efficiency that results with the sampler does not come free. Since all of the harmonics are converting with about the same efficiency, the unused ones contribute noise to the output. If the self-conversion noise is a large part of the overall noise budget, this can become an issue. It is of more concern with a low-LO sampling process. In Figure 39, an attempt is made to normalize for this change in the noise floor. In some sense, the conversion efficiency represents a signal-to-noise ratio (SNR). The flatness versus slope effect is still visible, but the performance cross-over point moves from an N=1 location to a location that is between N=2 and N=3.

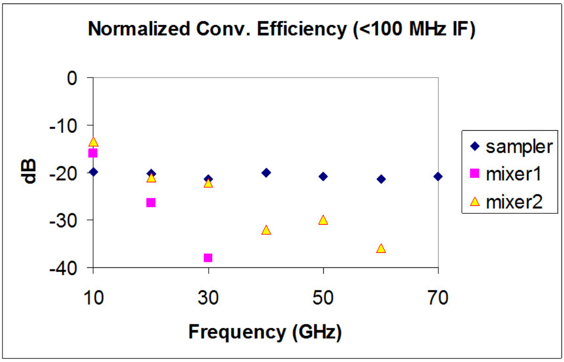


Figure 39. A normalized conversion efficiency comparison of a sampler to two example harmonic mixers is shown here. The setup is the same as in Figure 38, but the result is corrected for the different absolute-noise output powers of the different structures.

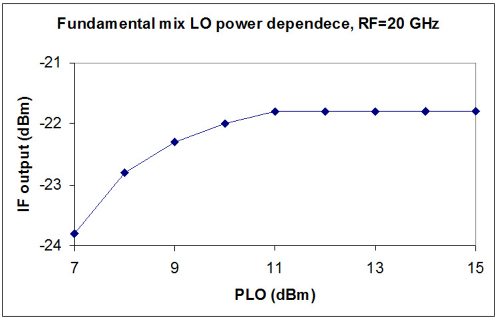


Figure 40. The dependence of IF output power on LO power for an example fundamental mixer is shown here.

For samplers and to a lesser extent, harmonic mixers, what range of LO to use is an important question. A lower LO means higher numbers of multiples will be in play which increases the noise delivered to the IF. It also increases the likelihood of spurious responses coinciding with the source and/or DUT spurs and harmonics; resulting in a more challenging frequency plan. A higher LO requires more expensive LO hardware and the system is less able to provide additional information about the incoming signal. The latter is an attribute not normally associated with the VNA, although it has been used in large-signal implementations to capture and analyze the harmonics and/or modulation components of the DUT's output signal.

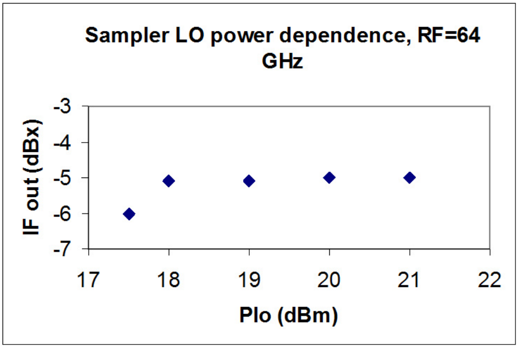


Figure 41. The dependence of IF output power on LO power for an example sampler is shown here. The result saturates for an ordinary fundamental mixer.

Another important topic is LO-power dependence. As LO power increases, the mixer's performance saturates and its behavior becomes less sensitive to minor variations in LO power (Figure 40). Depending on the drive configuration, the same tends to be true in a sampler (Figure 41). Once a sufficiently large gating pulse is generated, further increases do little to the gating aperture. It is important to note, however, that there can be some fundamental amplitude modulation/phase modulation (AM/PM) in this process since the sharpest point of the pulse moves in time as the amplitude changes.

There is more LO-power dependence in a harmonic-mix process since true square-wave generation is only approached in a loosely asymptotic manner (Figure 42). The drive levels have to be quite high to reach a saturation level. In some cases, it may even verge on a damage level, depending on the structure and the technology. The same type of AM/PM conversion can occur here with the samplers since the edge shape is still somewhat critical to the harmonic phasing.

Linearity is a critical topic in any discussion of the VNA receiver and is normally the first converter that establishes the limit. For fundamental mixing, many rules-of-thumb exist for the third-order intercept (TOI) for various mixer topologies. A common value for doubly-balanced mixers is that the input-referred TOI point (IP3) is roughly equal to the LO power. In the examples previously cited, the analyses are somewhat more complicated by the more highly, non-linear mixing processes involved. In general though, the IP3 will degrade.

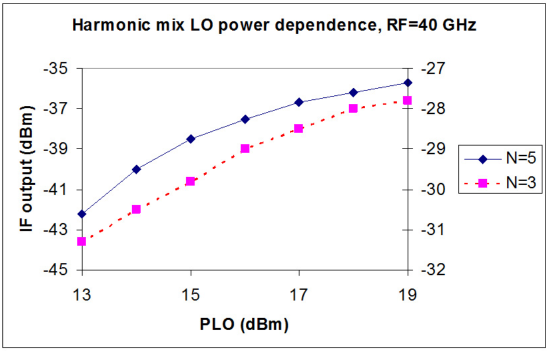


Figure 42. The dependence of IF output power on LO power for two levels of harmonic mix is shown here. Due to the required harmonic generation in the mixer and the soft conversion curve of this particular device, saturation is more difficult to achieve.

The plot of a shockline sampler being driven with 20 dBm is shown in Figure 43. It depicts a very flat intercept with frequency. The harmonic mixer results (doubly-balanced mixers driven with 17 dBm) show a quickly degrading IP3 with harmonic number (Figure 44). This might be expected since the available LO power is degrading and is limited by the mixing diodes' series resistance. Compression, another commonly quoted measure of linearity, typically tracks the intercept point in a relative sense.

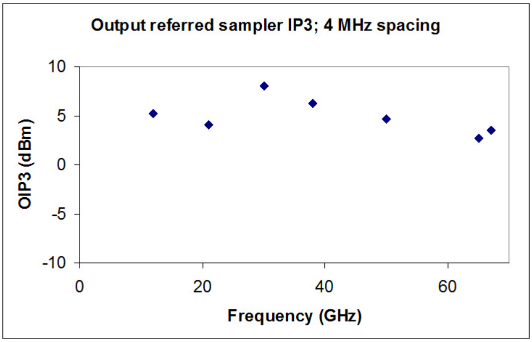


Figure 43. Example output-referred IP3 of a sampler is shown here. Since the conversion loss is on the order of 10 to 12 dB, the input-referred values are on the order of 15 to 20 dB.

A more peripheral, but still very important, linearity topic is that of spurious generation by the converter. This is a somewhat more subtle problem to understand. Consider a filter measurement where the DUT is highly reflective in the stopband. The reflected signal comes back to the stimulus port and enters its coupler and converter. Products of that signal, and multiples of the LO, emerge from the RF port of the converter and travel back toward the DUT, some of which may be in the DUT's passband (Figure 45). These products can travel to the other test coupler/converter and convert to the IF using different order multiples of the LO. This causes the DUT's stopband transmission to measure higher than it is in reality. Since these products can be randomly phased with respect to the desired IF, they appear as noise, making it difficult to diagnose as a measurement issue.

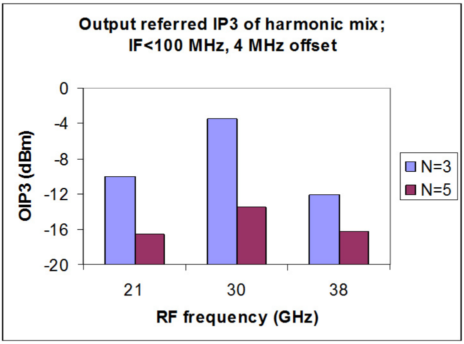


Figure 44. Output-referred IP3 for an example harmonic mixer is shown here for two different harmonic numbers (driven at +20 dBm). The results can be compared directly to Figure 43. For comparison, a doubly-balanced fundamental mixer of conventional design may have values in the range of 5 dBm.

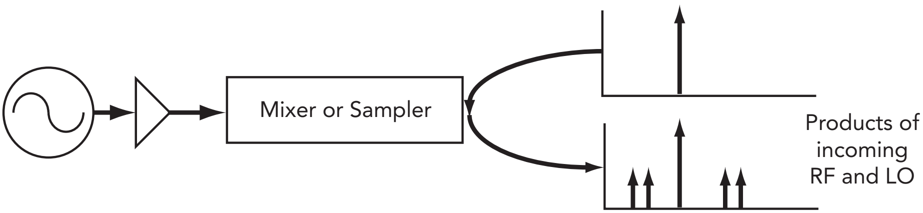


Figure 45. The concept of mixer or sampler bounce is shown here. The incoming RF signal mixes with multiples of the LO and the products re-emerge from the RF port.

As shown in Figure 46, both samplers and harmonic mixers can be susceptible to this issue, although the relative frequencies of the problem products are usually different. Typically when a harmonic mixer is operating at N>1, the lower order-related products are more problematic because the conversion efficiencies are favorable. In the case of a sampler, the conversion efficiencies are equally weighted (up to a point), so it is often those products related to the higher orders that create a problem. In the experiment of Figure 46, the LO and RF drive levels were fixed, as was the IF frequency. At each test frequency, the worst-case regenerated product was measured, in absolute power terms, and the result plotted.

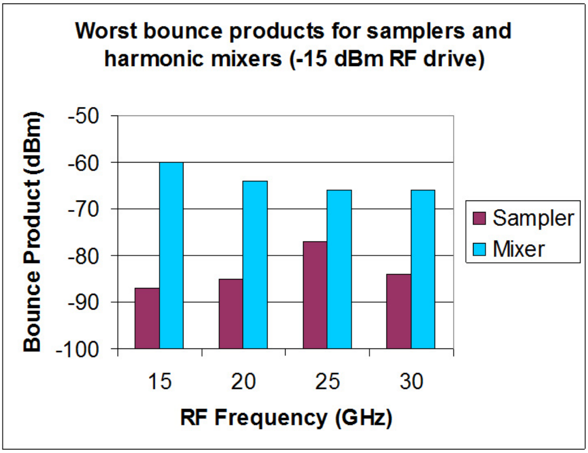


Figure 46. A comparison of bounce products for a harmonic mixer (N=3) and a sampler (N=2 to 3) is shown here. In both cases, LO drive is +20 dBm.

There are a number of ways to ameliorate this problem, such as by including padding in each sampler path, using isolation amplifiers on the sampler/mixer RF ports and turning off a test sampler/mixer (with bias or LO power) when not in use. Each of these options has its drawbacks:

- A pad reduces dynamic range.
- A RF pre-amplifier may cause compression issues or may have stability issues, since the amplifiers in the various channels are not identical.
- Turning off unused converters can increase overall measurement time.

IF Section

The IF section of any receiver typically gets the least amount of attention. Never the less, it is a critical component of the VNA's performance as this section is where the possible floor for speed, dynamic range and trace noise are often set. One of the first questions to ask is: What IF frequency, or range, should be used? If the frequency is very low, the analog-to-digital (A/D) circuitry can be simple but, as suggested by Figure 47, converted-LO phase noise becomes more of a problem (depending on the conversion structure). A very high IF frequency requires a more complex A/D structure and potentially more noise injection at the IF level, but the noise and spur contributions from the RF section are usually lower. Some applications may demand certain ranges of IF frequency (e.g., larger bandwidths needed).

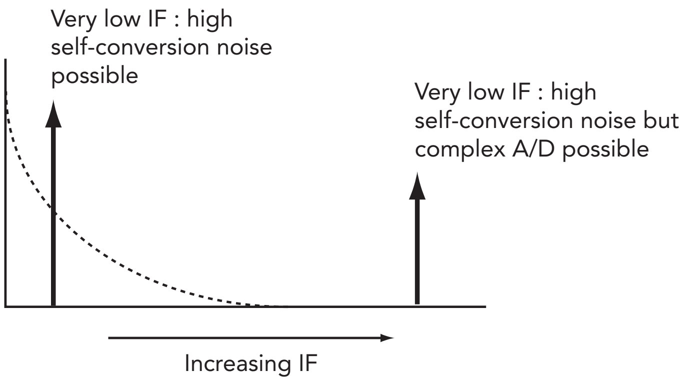


Figure 47. Provided here is a comparison of the noise effects of high and low IF frequencies.

Once the IF frequency is selected, the frequency plan for the A/D system usually comes next. Classically, an over-sampled structure would be used to allow the extraction of maximum spectral information. This requires a faster A/D clock and places more of a constraint on cleanliness and the A/D converter. Returning again to the concept of "knowing the signal" that is being measured, the design engineer could improve noise and simplicity by moving to an undersampled structure. The downside is an increase in spurious responses which may require more analog filtering. The classical difference between undersampling and oversampling is illustrated in Figure 48.

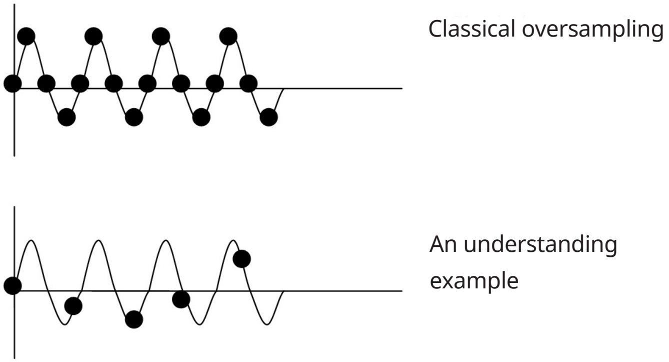


Figure 48. Pictorial diagrams of undersampling and oversampling are shown here.

Another method of detection - synchronous detection - works by performing a final down conversion to DC in an in-phase and in a quadrature sense. Since the A/D converters are operating at DC, the clocking structure can be simpler. Like homodyne systems, however, there are DC defects (e.g., offsets and channel skews) that must be correct for and/or minimized. The concept of synchronous detection is shown in Figure 49.

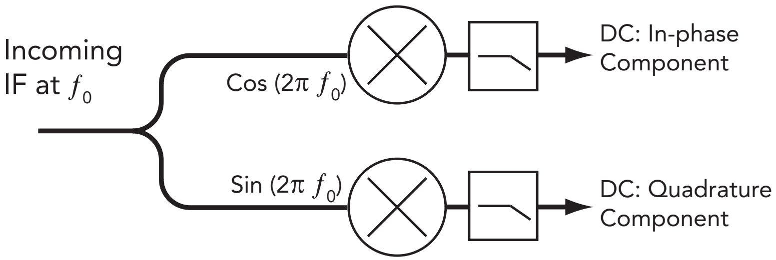


Figure 49. Depicted here is the concept of synchronous detection, where the final IF is down converted to DC for A/D sampling.

On a primary level, the dynamic range is set by the floor of the A/D converter, added noise and leakage. For conventionally-available A/D converters, this is normally not adequate. Other options include combining multiple A/D converters or adding variable gain to the IF chain.

The former is easier to implement from a state machine point-of-view, but calibrating the transition between converters can be challenging. The latter is relatively simple from a calibration point-of-view, but the measurement engine must be able to make decisions on-the-fly regarding how much gain to add. It must then also be able to force re-measurements, if necessary. Both schemes are shown in Figure 50.

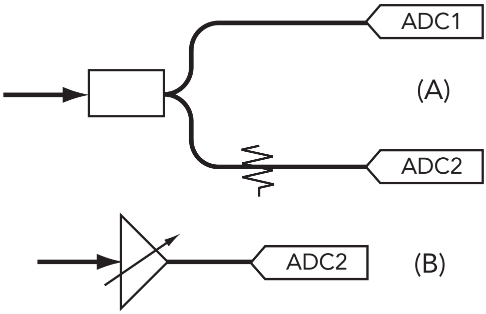


Figure 50. Two different methods of increasing the effective dynamic range of a digitizer are shown here. (A) illustrates the use of multiple A/D converters with a signal level offset, while (B) illustrates the use of variable gain prior to the A/D converter.

Implementation of filtering is another major topic. Usually some analog filtering is required to handle aliases, images and other known large interferers, to avoid overloading any IF amplifiers or A/D converters. Filtering is also required for noise reduction and sometimes reduction of close interferers. In VNAs, this is known as an IF bandwidth (IFBW), although an analogous concept of resolution bandwidth applies to spectrum analysis. Historically, IFBW was done with a collection of analog filters, but issues of stability and measurement accuracy arose when changing the setting between calibration and measurement. More recent instruments implement this filtering digitally. A common filtering implementation is shown in Figure 51.

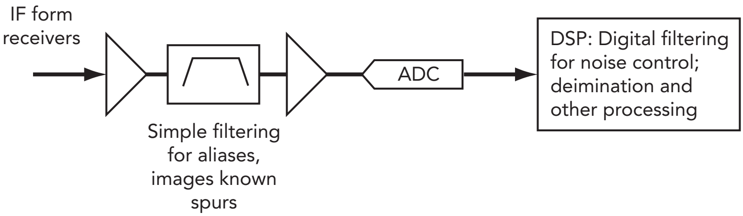


Figure 51. In this common IF-filtering scheme, some simple filtering is analog but most of the variable filtering (and narrow-bandwidth filtering) is done digitally.

System Performance Considerations

While interpretation of VNA specifications has been presented elsewhere, it is useful to examine how the performance of the blocks discussed thus far will impact those performance parameters.

1. **Dynamic Range** - There are two parts to dynamic range: noise floor and maximum power. The latter is addressed either by compression or port power. Noise floor is impacted by front-end loss (e.g., couplers and attenuators), mixer/sampler conversion loss and initial IF gain stages. A RF pre-amplifier can help at the potential expense of compression and stability.
2. **Trace Noise** - Usually trace noise is measured far away from the noise floor to avoid any impact. LO/source-phase noise folds over and converts to the IF. Trace noise also impacts IF system noise.
3. **Port Power** - Port power comes from the source power and the loss between source and port. Compression limits of switches and other test-set components may play a role.
4. **Power Accuracy** - Generally, power accuracy is limited by the structure of the ALC loop, temperature compensation employed, calibration procedure, and power ranges involved.
5. **Harmonics** - Usually harmonics come from the source and related components.
6. **Compression** - Usually the mixer/sampler linearity sets the compression limit, although the IF can sometimes contribute. While RF attenuators can help in some applications, front-end RF amplifiers can make the compression limit worse.
7. **Raw Port Parameters** - The front-end components (e.g., couplers, attenuators and transfer switches) tend to set these parameters.
8. **Residual Port Parameters** - Generally, the calibration kit and calibration algorithms set these limits. The first instrument parameters to affect the residuals are usually linearity-related.
9. **Stability** - Many factors, some of which are hard to measure (e.g., measurement dynamics), affect stability. For example, temperature stability of couplers, switches and cables affects the system, as does LO-power stability and the sensitivity of the mixer/sampler to its changes. The linearity of the system and stability of port power are two other factors affecting stability.

All of the blocks discussed play a role in how the VNA performs and how these specifications are created.

Measurement Fundamentals

The VNA represents a large suite of measurement capabilities. What follows are a few central concepts and vocabulary which will prove useful in any VNA discussion.

The Reference Plane

Perhaps one of the most important, and yet poorly understood, S-parameter measurement concepts is the reference plane. When performing a RF calibration with short-open-load-thru (SOLT), line-reflect-line (LRL) or some other algorithm, the engineer implicitly establishes a set of planes (one per port) that serve as the reference for the calibration. Imagine that two of these planes describe a pair of locations where, if a perfect thru could be connected between them, S_{21} and S_{12} would be precisely unity (0 phase) and S_{11} and S_{22} would be precisely 0 - excluding residual calibration errors. This can become an issue when things like adapters and fixtures are added or changed between calibration and measurement. The measured phases and amplitudes will be incorrect due to the distortions caused by those additions/changes. While the effects may be insignificant at low frequencies, they can be quite large above 30 or 40 GHz.

This concept is shown in Figure 52. Suppose the calibration is performed at planes 1A and 2A, while the DUT is connected to planes 1B and 2B (the horizontal lines between those planes may be adapters or fixture parts). Both reflection and transmission phases are altered by the electrical length between the A and B planes of the relevant port(s). Reflection and transmission magnitudes are altered by any loss (or change of loss) between these planes.

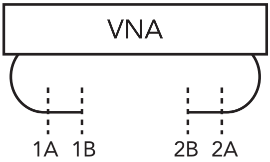


Figure 52. According to the reference plane concept, if adapters, fixtures and cables are added/ changed after a calibration, distortions can result.

If the characteristic impedances of any elements in this figure change, or if the desired reference impedance changes, the situation becomes more complicated. Since any line/ reference-impedance deviation makes the length change act like a transformer, large distortions are possible.

In some cases, it may be desirable to allow these plane distinctions to be more complex—to account for matching networks or because of complicated launching or fixturing that may be needed. A more flexible embedding/de-embedding engine can be used to allow for fairly complex reference-plane translations.

Introduction To Calibrations

Calibration is a central concept of many VNA measurements. Some fundamental concepts are provided here.

To begin with, assume that the VNA is physically perfect. The ports present exactly a 50-ohm (or some other reference) impedance, the couplers have infinite directivity, the receivers have a perfectly flat and known frequency response, and all of these characteristics are perfectly stable with time. Assuming this to be true, it is possible to make good S-parameter measurements without having to calibrate the VNA.

Unfortunately, these assumptions are not valid. The VNA's physical characteristics do change slightly with time. Additionally, the engineer may want to use different cables (which change with time) or adapters, change connector or media types, or use a test fixture—all of which radically change the test ports' electromagnetic properties. Resolving this semi-dynamic situation requires the engineer to periodically perform a calibration in the environment of choice. By connecting a combination of calibration standards (about which something is known), the properties of the VNA and test assembly can be deconvolved, to a certain degree, from DUT measurements to provide a more accurate depiction of the DUT's S-parameters.

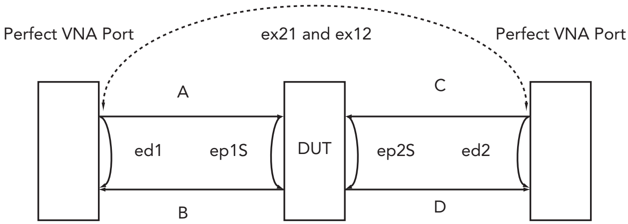
Some additional considerations to keep in mind are:

1. How many standards are needed, and what must be known about them, depends on the algorithm and calibration type selected. The choice of algorithm also determines how accurately the standards must be known. Some of the typically used standards are transmission lines (called thrus or lines), loads or matches (good terminations), opens, shorts, and reciprocal devices (usually lower-loss passive networks where S_{21}=S_{12}).

2. Calibrations are not perfect because the calibration standards are not perfectly known and the error models used to describe the measurement setup are not perfect. The errors remaining after the calibration are often termed residual errors and are expressed in the data sheet for a variety of calibration algorithms. These residual errors can be used to help compute measurement uncertainties and are distinguished from raw-port characteristics (sometimes specified), which describe things like match and directivity without correction applied.

3. The error models used to describe the setup cover the items listed below. Refer to Figure 53 for a diagram. The i and j terms refer to port numbers.

- **Source Match (epiS)** - Corrects for the imperfect match of a measurement reference plane when that port is driving.
- **Load Match (epiL)** - Same as above, but when the port acts as a termination. Some algorithms do not use this term. Instead they use a pre-correction step to account for any differences between source match and load match.
- **Directivity (edi)** - Corrects for imperfections in the couplers, which lead to a measured reflection even when a perfect termination is attached.
- **Reflection Tracking (etii)** - Frequency response of the system when measuring a reflection.
- **Transmission Tracking (etij)** - Same as above, except when making a transmission measurement. This term is not entirely independent of reflection tracking. Some algorithms use this information.
- **Isolation (exij)** - Corrects for some leakages between the system's receivers and source. While most algorithms support some aspect of isolation, it is not commonly used due to relatively small improvements.



et11 = A * B
et22 = C * D
et21 = A * D
et12 = C * B

Figure 53. In this diagram of the error model, the idea is to replace the physical VNA ports with ideal VNA ports connected to simple networks containing all of the imperfections (called error boxes).

4. Sometimes it is not convenient to calibrate all the way up to the reference planes at the DUT. In such cases, techniques such as de-embedding and reference-plane extension are used to help correct for any discrepancies. In some sense, these techniques are like calibration (and may use much of the same math), but rely more on models (circuit-based or file-based descriptions of the networks involved) rather than measurements of standards through the networks in question. Some combination of these approaches may be appropriate depending on the measurement setup involved.

5. Calibrations do not last forever. As temperature changes, cable behaviors change. With many cable-bending cycles, the phase shift through those cables may change. Replacements of cables or adapters change the calibration. Calibrations can last anywhere from a few hours to a few days, but it depends greatly on the environment and the desired stability.

Linearity

A fundamental assumption in most VNA measurements is that the system is linear. Commonly, this is expressed either in terms of compression or TOI, as follows:

Compression example: 0.1 dB compression occurs at a port power of 10 dBm.

TOI example: Given a tone spacing of 1 MHz and a port power of -10 dBm, the TOI is 35 dBm.

In the context of linearity, compression means that if a DUT (e.g., amplifier) is measured such that the power hitting the port is 10 dBm, as in the example, then the result will be 0.1 dB off (usually lower) from the expected value at a lower port power. Compression is usually due to the VNA's down-converting element but can come from other sources. Its effect can be minimized by using pads after the DUT or using receive-side attenuators. Receive-side attenuators are attractive since they are in the coupled arm, as opposed to in the test port path, and therefore do not degrade raw directivity and stability (Figure 54). Note that compression can also be caused by high source power and the DUT need not have gain for it to occur. Compression can affect reflection measurements as well.

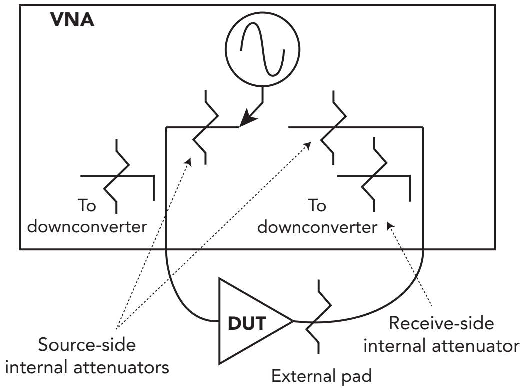


Figure 54. A partial VNA block diagram is shown here to illustrate the placement of internal step attenuators. The source-side versions can be used to reduce drive power, while the receive-side versions can help reduce compression/linearity effects. For simplicity, reference couplers are not shown.

TOI is an analogous way of expressing the receiver's linearity, although the quantity is somewhat different. This specification is important when using the VNA to measure intermodulation products or a DUT's TOIs.

While less common, linearity can sometimes be an issue at lower-signal levels. Here, the nonlinearities of A/D converters tend to appear. Usually though, enough gain is used in the VNA to prevent this from becoming an issue. Other IF and RF chain nonlinearities can be present as well, but are usually not explicitly specified. They are typically accounted for in uncertainty calculations.

Data Formats

A large number of file formats and data formats are used by the VNA. The two main classes pertaining to users are DATA and Text. The DATA class contains a number of export formats which are available for use in external applications or for archiving. In contrast, the Text (.txt extension) class is a tab-delimited format, with an optional descriptive header, in which every trace's data in the active channel is saved to a desired location. Each trace's data is saved as an X and Y column (e.g., to accommodate mixed frequency and time domain). Subsequent traces are added as additional columns.

Some of the common formats include:

- Common-Separated (.csv)—A more spreadsheet-friendly version of the Text class.
- Bitmaps (.bmp)—A graphics capture of the data and menu areas.
- .snp (.s1p, .s2p)—The standard microwave simulator text-data format. This is similar to .txt except a controlled header is used and only one or four S-parameters are saved. If a .s2p file is requested, but all S-parameters are not being measured, a "0" is entered for the missing S-parameters. If a full 2-port calibration is applied, all S-parameters are measured but need not be displayed. In this case the .s2p file is complete. Note that, unlike the previous file types, .snp files can be recalled and displayed as trace memory. They are loaded into the active channel.
- Trace Memory Formatted (.tdf)—This saves the active trace's memory following post processing and is useful for comparing against a 'golden device' to see if a DUT's performance has changed.
- Trace Memory Unformatted (.tdu)—Same as above, but occurs before most trace post-processing (e.g., time domain, smoothing and group-delay calculations).

Note that these last two formats can be recalled into the active trace, but the VNA must normally be in hold or sweeping very slowly for this to be useful. They can also be recalled into the active trace's memory. The trace can later be recalled into the same part of the chain where it was saved. Formatted trace memory is recalled after post-processing steps, while unformatted trace memory is recalled before those steps.

Other Terms of Interest

A few other common terms of interest are defined below. More detail can be found throughout this book.

Power - Power delivered to the port when that port is driving.

Scale - The dB/div or units/div of a given graph. It also defines the magnification of the Smith chart graph.

Limit Lines - Test values on a graph that can be used to mark a DUT as pass or fail. Upper and lower values can be defined for a point, or a range of points, that may indicate specifications for the DUT.

Interpolation - A means of acquiring information between calibration or measurement points. This term is used in a number of contexts within the VNA.

- **Calibrations -** Normally measurements can only be made at frequency points where a calibration was performed. By enabling interpolation, the system attempts to calculate error coefficients at other points in between. Some error may result, but it can be very convenient.
- **Embedding/De-Embedding -** When a file is loaded that describes a network to be embedded/de-embedded, its frequency list may not exactly match the frequencies being swept. In this case, the system will attempt to interpolate the data to the frequencies being used.
- **Markers -** Normally, in discrete mode, two markers hop between frequency points being measured. In continuous mode, the markers go to other frequencies and display data values based on interpolation between measurement points.

System Architecture And Modes Of Operation - The VNA's flexibility provides unique opportunities for specialized configuration. It is helpful to understand the differences in system architecture that can be found in different VNAs. Understanding the strengths and limitations of different VNA system architecture helps in choosing the best VNA for a particular application and in configuring the measurement for optimum results. The different modes of operation and types of system architecture include:

- **Receiver Offset Mode -** This mode allows for independent source and receive functions for mixer, harmonics, IMD, and other measurements where the source and receive frequencies are offset.
- **Source Modes -** Some non-Anritsu VNAs provide synthesized (step) mode and non-synthesized (open-loop) modes of operation. Verify that the VNA is properly configured for synthesized mode to ensure accurate phase measurements. Verify also that the VNA has the best trace noise to ensure accurate transmission tracking.
- **IF Chain Methodology -** Microwave VNAs convert the RF signal to IF before analysis. There are different methods of down conversion, but they usually fall into two categories: harmonic sampling and harmonic mixing. VNAs utilizing harmonic mixing often have a lower

compression point so care must be taken to ensure that accuracy is not lost due to compression errors.

- **Receiver Calibration -** Unlike conventional VNA RF calibrations (e.g., SOLT and LRL/LRM) that are used to calibrate the VNA for S-parameter measurements, the receiver calibration is more of an absolute power calibration to help with measurements like:
  - Harmonics
  - IMD, IP3 and other multi-tone distortion measurements
  - Mixer-conversion loss, in simpler scalar cases
  - Other times when the VNA is used just as a channelized receiver

The idea behind receiver calibration is to take a known source power, at some source-reference plane, and transfer that knowledge to the receiver at a desired receiver-reference plane. If it is convenient to use the test port as the source-reference place, the built-in factory ALC calibration can be used to establish the 'power knowledge.' If this is not convenient—because of frequency translation or the requirement for some other network or greater accuracy—then a power calibration can be performed with the help of a GPIB-controlled power meter to establish that power knowledge.

Specifications and Measurement Accuracy

Specifications help the user determine the level of measurement accuracy available from an instrument under a given set of conditions. When considering the VNA, specifications such as dynamic range, noise floor and available power provide information on the DUT conditions that can be measured and are not intrinsically "accuracy specifications."

Measurement accuracy is determined by factors such as directivity, source and load match, and isolation. These factors are often presented in the form of raw uncorrected performance and/or residual performance after calibration. In addition, the S-parameter performance of the DUT also contributes to the overall measurement accuracy as it interacts with the residual performance of the VNA. When considering measurement accuracy, it is important to take into account all parameters of the VNA's residual performance, combined with the S-parameter performance of the DUT at each frequency point-of-interest.

Figure 55 provides an example of the transmission uncertainty for different S_{21} DUT characteristics over the frequency span from 20 to 70 GHz. The plot provides an indication of the complexity of calculating measurement accuracy for each device of interest. In order to help generate uncertainty curves for individual devices, a software tool was developed and is available on the Anritsu website (www.anritsu.com). This program takes into account all salient performance parameters of the VNA and plots uncertainty curves for specific levels of DUT performance.

Magnitude Measurement Uncertainty

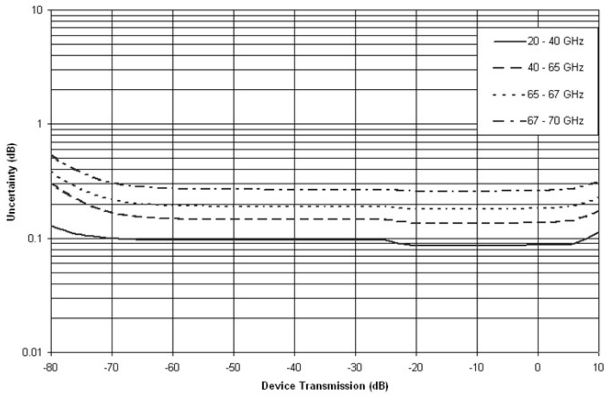


Figure 55. An uncertainty plot provides the overall measurement uncertainty, taking into account all system residuals and DUT performance.

What follows is a summary of some of the important characteristics that determine the VNA's measurement-performance level. In some cases a brief description is provided on how to measure the corrected residual performance.

1. **Dynamic Range -** This is a calculated value based on noise floor and either the compression level (for receiver dynamic range) or maximum port power (for system dynamic range). For dynamic-range values referenced to a direct-receiver access port, the compression level is referred to that input. For system dynamic range, the maximum available port-power level from the test set—well within the 0.1 dB compression level—is used.
2. **Compression Level -** This is the receiver power level at which the reported value is 0.1 dB below what is expected based on lower power results. Due to the accuracy required, this almost always is computed as a ratioed measurement where a second receiver gets a padded version of the signal fed to the target receiver. The second receiver is well away from compression and therefore, no inadvertent cancellation of distortion occurs. Padding is typically used to minimize match interactions. In general, compression is referred to test ports but can also be referred to direct receiver-access ports. A diagram is shown in Figure 56. For the direct receiver-access ports, the setup is relatively easy to implement. For the main

test-port-based test receivers, a thru line and the internal source are used to provide drive. The driving-port test receiver is used as the norming receiver. Since this receiver sees a load-match value, it has the desired effect of reducing its drive level well below compression.

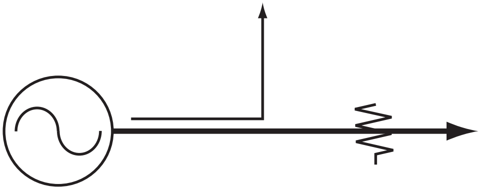


Figure 56. A schematic diagram of a compression-test setup is shown here.

As previously discussed, a ratioed measurement is normally required. In the two test-port case, this ratio is normally |b2/b1| or |b1/b2|. This parameter is swept at low power (-15 dBm) in a modest IFBW (e.g., 100 Hz) so that trace noise is much smaller than the desired compression level to be observed. This data is stored to memory and used for normalization. The drive power is then increased until the indicated level of compression occurs (typically 0.1 dB). The drive power at the subject port when this occurs forms the result of the measurement. Although the term compression is used, expansion may sometimes be observed due to match interaction. Deviation in either direction is counted in this measurement.

3. **Noise Floor -** This is a measure in absolute power (dBm) of the noise floor of the system referenced to the test port. Typically this is calculated by measuring S_{21} and S_{12} with a short-thru line connected and the port power set to some value X. The cable loss is normally subtracted separately or, alternatively, a flat-power calibration can be performed at the end of the cable. The traces are normalized with this thru in place. The ports are then terminated with loads (typically), and S_{21} and S_{12} measured in a 10 Hz bandwidth with no averaging. A minimum of 10 sweeps are acquired in linear magnitude mode and the root-mean-square (RMS) value is computed at each frequency point individually. This result is then normally converted back to log magnitude.

4. **Trace Noise (High Level Noise) -** This is a measure of the scatter of data when measuring a high-level signal (full reflect or transmission through a short transmission line). The measurement is specified at default power with a 1 kHz IFBW and default auto-reduction for frequencies below 3 MHz. The measurement

is performed by acquiring a minimum of 10 sweeps of data from the desired parameter, in linear magnitude and phase, after a trace-math normalization. At each frequency point, the population-based standard deviation is computed. This forms the RMS high-level noise number. For magnitude, this is normally converted back to log magnitude.

5. **Power Range -** Maximum port power is determined by the power at which the system is still leveled. The same applies to minimum power when step attenuators are not involved. For those systems with attenuator options, the maximum attenuator value for the model is subtracted from the minimum ALC power to get the minimum system power.

6. **ALC Power Accuracy And Linearity -** Power accuracy and linearity are both determined with an external power meter. Accuracy is evaluated at default power across frequency. Linearity is evaluated over both power and frequency with the reference being the value at default.

7. **Frequency Accuracy And Stability -** Frequency stability is a pass-through specification of the 10 MHz time base and is not tested separately at instrument level. Accuracy is measured using a frequency counter synchronized to a GPS-based time base after the 10 MHz calibration is complete.

8. **Harmonics -** This is also port-referred and based on default power. It can be calculated as follows, using a properly-calibrated spectrum analyzer or the VNA itself, using multiple source control:

- Choose b2/1 as the measurement variable.
- Perform a flat-power calibration at the end of a cable from port 1 over the entire range of the instrument.
- Perform a receiver calibration on b2.
- With multiple source equations all set to 1/1(f+0), acquire data and normalize.
- Change the receiver and receiver source equation to look for the desired harmonic (e.g., 2/1(f+0) for the 2nd harmonic).

9. **Raw Directivity -** This expresses the raw directivity of the test coupler assemblies as affected by internal match conditions. While it could be measured directly using direct-access loop options, it is normally done through the calibration engine. Perform a reasonably high-quality calibration and compute the term edx/etxx, where 'x' refers to the port in question. The magnitude of this term represents the raw directivity. Use the form -20log10|edx/etxx| to get the value in the form of a specification.

10. **Raw Source Match -** Much like raw directivity, this can be extracted from the

calibration system. The appropriate value is -20log10|epxS|, where 'x' is the port in question.

11. **Raw Load Match -** This can also be extracted from the calibration system. Perform the calibration so that the reference plane is at the test port of the port under study. The value in this case is -20log10|epxL|. Alternatively, perform a 1-port calibration at the end of a cable and then connect this cable to the port under study. The direct measurement in this case yields raw load match.

12. **Residual Directivity -** Residuals are all based on after-calibration results and, as such, reflect imperfections in the calibration. The residuals are determined primarily by the calibration method/algorithm, the calibration components involved and the skill of the person performing the calibration and verification. The residual measurements are generally traceable to airline standards which form the impedance reference for VNA measurements. Directivity is measured using an airline attached to the test port and terminated in either a load (preferred for broadband) or an offset (acceptable for lower frequencies). A ripple extraction method is applied to the resulting Sii data to determine the directivity. In an automated setup, circle-fitting and polynomial-fitting routines are commonly used, but each has its weaknesses. An adaptive sorting algorithm is generally advised.

13. **Residual Source Match -** This is also an airline ripple measurement but is performed with the airline shorted.

14. **Residual Load Match -** This is a direct reflection measurement with a thru or airline connected after calibration.

15. **Residual Reflection Tracking -** This is a direct-reflection measurement of a short, but not the one used during calibration for the case of a defined standards calibration such as SOLT, short-short-load-thru (SSLT) or short-short-short-thru (SSST). The peak deviation from a linear fit is used as the starting point to remove the effect of loss of the short standard. The contribution due to source match and trace noise is then removed from that result to get reflection tracking.

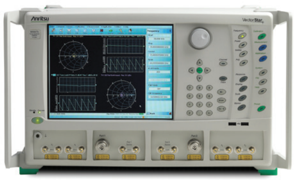
16. **Residual Transmission Tracking -** This is a direct-transmission measurement of a thru or airline, depending on the calibration. The peak deviation from a linear fit is again used as the starting point. The contribution due to the load-match source match interaction and that due to trace noise is removed from the result to get transmission tracking.

Vector Network Analyzers - VectorStar

Anritsu VNAs encompass a wide range of high performance solutions for R&D, manufacturing, and field applications in the wireless, satellite, defense, broadband communication, and optoelectronic components markets. Choose between the VectorStar, Shockline or VNA Master family for the perfect solution of advanced performance, accuracy, and reliability. Ideal for measuring any RF and Microwave component or system - from design and manufacturing to measurements anytime, anywhere.

MS4640B Series VectorStar Vector Network Analyzer

- Broadest frequency range from 70 kHz to 145 GHz single sweep, and up to 1.1 THz in waveguide bands.
- Superior dynamic range - up to 140 dB.



The world's widest frequency range in a single instrument – 70 kHz up to 70 GHz. Massive 106 dB dynamic range at 70 GHz. Sweep from 70 kHz to 70 GHz in less than 4 ms, locked and levelled. And innovative Precision AutoCal calibration makes test routine configuration easy. Whether you are testing performance characteristics of on-wafer devices or developing communications systems for aerospace, military or security use, the MS4640B is the best that money can buy.

	Features	Benefits	Applications
VectorStar VectorStar Family Microwave and mmwave VNA MS4642A Series 70 kHz to 20 GHz MS4644A Series 70 kHz to 40 GHz MS4645A Series 70 kHz to 50 GHz MS4647A Series 70 kHz to 70 GHz	Broadest frequency span from a single coaxial test port covering 70 kHz to 70 GHz in a single instrument and 70 kHz to 110 GHz in the broadband configuration.	• Obtain the most thorough and accurate broadband device characterization. • Eliminate time consuming concatenation process across the RF, microwave/mm-wave bands. • Decrease test instrument expenses by eliminating the need for a 2nd RF VNA. • Reduce the risk of DC extrapolation errors in your device modelling.	Device characterization Microwave and millimeter wave component test On-wafer Waveguide S-parameter R&D and production environments Mixer measurements including automatic de-embedded measurements with absolute Phase and group delay Embed/De-embed applications Amplified testing Broadband characterization Parameter extraction Device modeling
	Fastest swept synthesized measurement speed <20 usec per point	• Increase manufacturing revenue by increasing throughput. • Quickly and easily spot the most hard to find failures and reduce the risk of shipping defective products.	
	Superior Dynamic Range - up to 140 dB	• Accurately measure medium and high loss devices • Catch all potential feedthroughs in out-of-band regions	
	High compression point - up to 15 dBm at 70 GHz	• Eliminate the need for additional attenuators • Improve calibration and measurement accuracy	
	Best test port characteristic - up to 50 dB Directivity, Source Match, Load Match.	• Reduce measurement uncertainty • Reduce measurement guard bands • Improve productivity • Optimum precision in R&D	
	Highest point resolution - 100,000 points	• Zoom in on narrow band responses without re-calibration	
	Best device modeling data	• Accurate design cycle • Accurate DC modelling • Eliminate the need for 2nd VNA	
	Best time domain analysis	• 100,000 points and 700 kHz frequency step size provide the most accurate, highest resolved, low pass mode measurements. • Measure long transmission lines with the best non-aliasing range.	
	Most convenient automatic calibration system with best accuracy.	• Use Precision AutoCal for an easy, one-button method of VNA calibration and better accuracy than traditional SOLT calibration. • Spend less time setting up the VNA for the next production run.	

Vector Network Analyzers - ShockLine™

- Patented ShockLine™ VNA-on-a-chip technology.
- A radical departure from conventional VNA design.
- High accuracy, sensitivity and sweep speeds.
- Considerably more cost efficient than comparable bench top VNAs.

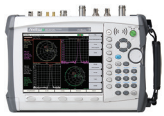
ShockLine VNAs are a family of 1-, 2-, and 4-port VNAs ideal for testing passive devices in production, design and education. Utilizing patented VNA-on-a-chip technology these units offer significant performance benefits in a very cost efficient format.

Model	Range	Dynamic Range	Sweep Speed	Corrected Directivity
MS46121A ShockLine 1 port USB VNA	40 MHz to 4 GHz 150 kHz to 6 GHz	N/A	120 us/point	42 dB typical
MS46122A ShockLine Compact USB Series VNA	1 MHz to 8 GHz 1 MHz to 20 GHz 1 MHz to 43.5 GHz	>100 dB, typ	220 us/point	≥42 dB, 1 MHz to 10 GHz ≥36 dB, 10 GHz to 30 GHz ≥30 dB, 30 GHz to 40 GHz
MS46322A ShockLine Economy Series 2-port VNA	1 MHz to 4 GHz 1 MHz to 8 GHz 1 MHz to 14 GHz 1 MHz to 20 GHz 1 MHz to 30 GHz 1 MHz to 43.5 GHz	>100 dB, typ	220 us/point	≥42 dB, 1 MHz to 10 GHz ≥36 dB, 10 GHz to 30 GHz ≥30 dB, 30 GHz to 40 GHz
MS46522B ShockLine Series 2-port VNA	50 kHz to 8.5 GHz 50 kHz to 20 GHz 50 kHz to 43.5 GHz 55 GHz to 92 GHz (Ext-ended E-band)	>120 db, typ 50 MHz to 43.5 GHz, >115 dB, typical E-band	30 us/point	>42 dB, 300 kHz to 10 GHz >=36 dB, >10 GHz to 20 GHz >=32 dB, >20 GHz to 30 GHz >=30 dB, >30 GHz to 40 GHz
MS46524B ShockLine Series 4-port VNA	50 kHz to 20 GHz 50 kHz to 43.5 GHz 55 GHz to 92 GHz (Ext-ended E-band)	>120 db, typ 50 MHz to 43.5 GHz, >115 dB typical E-band	30 us/point	>42 dB, 300 kHz to 10 GHz >=36 dB, >10 GHz to 20 GHz >=32 dB, >20 GHz to 30 GHz >=30 dB, >30 GHz to 40 GHz

Vector Network Analyzers - Handheld

MS202xC/3xC Handheld VNA Master™

- Active 2-port, 2-path fully-reversing VNA; measures and displays all S-parameters with a single connection.
- 350 µsec per data point, ideal for filter tuning.
- 100 dB transmission dynamic range.
- Vector Voltmeter option, ideal for cable phase matching.



The industry's first fully-reversing handheld VNA with true 12-term error correction for S-parameter measurements in the field anytime, anywhere. The VNA Master spans the widest frequency range to support a wide variety of RF and microwave systems, both coax and waveguide.

Model	VNA Frequency	SPA Frequency	Measurements
MS2026C	5 kHz to 6 GHz	-	Measures all 4 S-parameters, quad-display, Return loss, cable loss, VSWR, Smith Chart, Polar Display, gain, and optional Time Domain, includes LP Processing and Gated Time Domain, DTF, Secure Data
MS2027C	5 kHz to 15 GHz	-	
MS2028C	5 kHz to 20 GHz	-	
MS2036C	5 kHz to 6 GHz	9 kHz to 9 GHz	Capabilities of MS2028C PLUS High performance spectrum analysis, channel scanner, interference analysis, spectrograms
MS2037C	5 kHz to 15 GHz	9 kHz to 15 GHz	
MS2038C	5 kHz to 20 GHz	9 kHz to 20 GHz	

Summary

The major functional analog blocks in a RF/microwave VNA have been presented and examined in detail. While there are many more details regarding these blocks which could have been covered, the intention is to point out the possible trade-offs in the technologies chosen and the resulting performance attributes which can be achieved based on those choices. General information pertaining to the fundamentals of VNA measurements, specifications and performance was also covered, including a significant number of definitions which will prove useful in understanding the more advanced concepts.

References

R. A. Hackborn, "An automatic network analyzer system," Microwave Journal, vol. 11, pp. 45-52, 1968.

Microsemi, "The PIN diode circuit designer's handbook," 1992.

M. Kahrs, "50 years of RF and microwave sampling," IEEE Trans. Microwave Theory and Tech., vol. 51, pp.1787-1805, June 2003.

S. A. Maas, Microwave Mixers, Artech House, 1993.

R. S. Pengelly, Microwave Field Effect Transistors – Theory, Design and Applications, Research Studies Press, 1986.

G. D. Vendelin, A. M. Pavio, U. L. Rohde, Microwave Circuit Design Using Linear and Nonlinear Techniques, Wiley, 2005, chp. 12.

T. Van den Broeck and J. Verspecht, "Calibrated vectorial nonlinear network analyzer," Proc. Int. Micr. Symp., San Diego, 1994, pp. 1069–1072.

W. Van Moer and Y. Rolain, "A large signal network analyzer: why is it needed?" Microwave Magazine, vol. 7, issue 6, pp. 46-62, Dec. 2006.

J. A. Crawford, Frequency Synthesizer Design Handbook, Artech House, 1994.

Anritsu envision : ensure

Specifications are subject to change without notice.

- **United States**

Anritsu Company
1155 East Collins Blvd., Suite 100, Richardson,
TX 75081, U.S.A.
Toll Free: 1-800-267-4878
Phone: +1-972-644-1777
Fax: +1-972-671-1877

- **Canada**

Anritsu Electronics Ltd.
700 Silver Seven Road, Suite 120, Kanata,
Ontario K2V 1C3, Canada
Phone: +1-613-591-2003
Fax: +1-613-591-1006

- **Brazil**

Anritsu Eletrônica Ltda.
Praça Amadeu Amaral, 27 - 1 Andar
01327-010 - Bela Vista - São Paulo - SP - Brazil
Phone: +55-11-3283-2511
Fax: +55-11-3288-6940

- **Mexico**

Anritsu Company, S.A. de C.V.
Av. Ejército Nacional No. 579 Piso 9, Col. Granada
11520 México, D.F., México
Phone: +52-55-1101-2370
Fax: +52-55-5254-3147

- **United Kingdom**

Anritsu EMEA Ltd.
200 Capability Green, Luton, Bedfordshire, LU1 3LU, U.K.
Phone: +44-1582-433200
Fax: +44-1582-731303

- **France**

Anritsu S.A.
12 avenue du Québec, Bâtiment Iris 1- Silic 612,
91140 VILLEBON SUR YVETTE, France
Phone: +33-1-60-92-15-50
Fax: +33-1-64-46-10-65

- **Germany**

Anritsu GmbH
Nemetschek Haus, Konrad-Zuse-Platz 1
81829 München, Germany
Phone: +49-89-442308-0
Fax: +49-89-442308-55

- **Italy**

Anritsu S.r.l.
Via Elio Vittorini 129, 00144 Roma, Italy
Phone: +39-6-509-9711
Fax: +39-6-502-2425

- **Sweden**

Anritsu AB
Kistagången 20B, 164 40 KISTA, Sweden
Phone: +46-8-534-707-00
Fax: +46-8-534-707-30

- **Finland**

Anritsu AB
Teknobulevardi 3-5, FI-01530 VANTAA, Finland
Phone: +358-20-741-8100
Fax: +358-20-741-8111

- **Denmark**

Anritsu A/S
Kay Fiskers Plads 9, 2300 Copenhagen S, Denmark
Phone: +45-7211-2200
Fax: +45-7211-2210

- **Russia**

Anritsu EMEA Ltd.
Representation Office in Russia
Tverskaya str. 16/2, bld. 1, 7th floor.
Moscow, 125009, Russia
Phone: +7-495-363-1694
Fax: +7-495-935-8962

- **Spain**

Anritsu EMEA Ltd.
Representation Office in Spain
Edificio Cuzco IV, Po. de la Castellana, 141, Pta. 8
28046, Madrid, Spain
Phone: +34-915-726-761
Fax: +34-915-726-621

- **United Arab Emirates**

Anritsu EMEA Ltd.
Dubai Liaison Office
P O Box 500413 - Dubai Internet City
Al Thuraya Building, Tower 1, Suit 701, 7th Floor
Dubai, United Arab Emirates
Phone: +971-4-3670352
Fax: +971-4-3688460

- **India**

Anritsu India Private Limited
2nd & 3rd Floor, #837/1, Binnamangla 1st Stage,
Indiranagar, 100ft Road, Bangalore - 560038, India
Phone: +91-80-4058-1300
Fax: +91-80-4058-1301

- **Singapore**

Anritsu Pte. Ltd.
11 Chang Charn Road, #04-01, Shriro House
Singapore 159640
Phone: +65-6282-2400
Fax: +65-6282-2533

- **P.R. China (Shanghai)**

Anritsu (China) Co., Ltd.
Room 2701-2705, Tower A,
New Caohejing International Business Center
No. 391 Gui Ping Road Shanghai, 200233, P.R. China
Phone: +86-21-6237-0898
Fax: +86-21-6237-0899

- **P.R. China (Hong Kong)**

Anritsu Company Ltd.
Unit 1006-7, 10/F., Greenfield Tower, Concordia Plaza,
No. 1 Science Museum Road, Tsim Sha Tsui East,
Kowloon, Hong Kong, P.R. China
Phone: +852-2301-4980
Fax: +852-2301-3545

- **Japan**

Anritsu Corporation
8-5, Tamura-cho, Atsugi-shi, Kanagawa, 243-0016 Japan
Phone: +81-46-296-6509
Fax: +81-46-225-8359

- **Korea**

Anritsu Corporation, Ltd.
5FL, 235 Pangyoyeok-ro, Bundang-gu, Seongnam-si,
Gyeonggi-do, 463-400 Korea
Phone: +82-31-696-7750
Fax: +82-31-696-7751

- **Australia**

Anritsu Pty. Ltd.
Unit 20, 21-35 Ricketts Road,
Mount Waverley, Victoria 3149, Australia
Phone: +61-3-9558-8177
Fax: +61-3-9558-8255

- **Taiwan**

Anritsu Company Inc.
7F, No. 316, Sec. 1, NeiHu Rd., Taipei 114, Taiwan
Phone: +886-2-8751-1816
Fax: +886-2-8751-1817

Part No 11410-00724, Rev B. 04/2016