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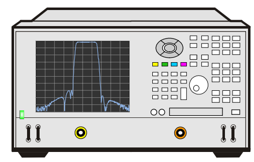
Getting the Greatest Dynamic Range from Your Vector Network Analyzer



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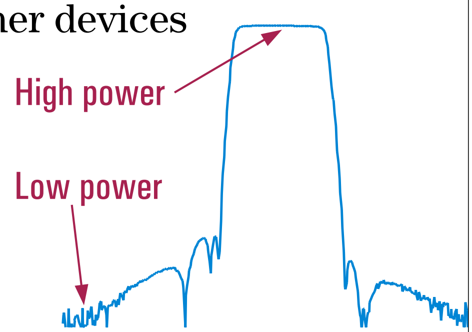
- Welcome to “Getting the Greatest Dynamic Range from Your Vector Network Analyzer!”
- Have you ever wondered if you are using your RF vector network analyzer or VNA to its fullest potential? Are you taking advantage of all the ways to improve dynamic range? Do you understand how the VNA's hardware can affect dynamic range? Are you making the best choices between speed and measurement accuracy?
- Over the course of today’s seminar, we’ll try to answer these questions and more.

VNA and DR Basics

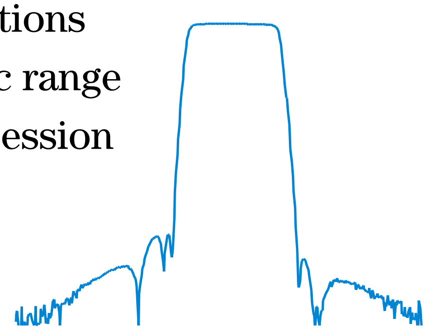
- Vector network analyzers are primarily used to characterize the **linear** performance of RF devices
  - filters (bandwidth, insertion loss, match …)
  - amplifiers (gain, match, isolation …)
  - plus lots and lots of other devices
- Dynamic range (DR) is the **range** of **power** levels that a network analyzer can measure



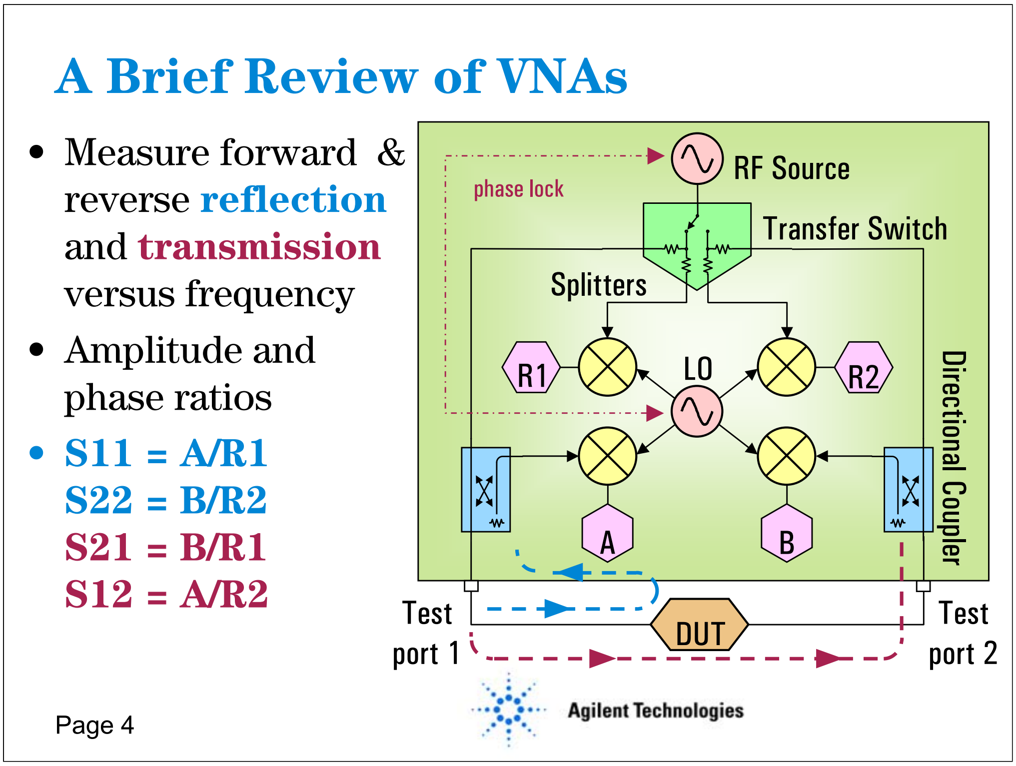
- Let’s start with some RF component-test basics. Vector network analyzers are primarily used to characterize the linear performance of RF devices versus frequency. Common network analyzer measurements for filters include bandwidth, insertion loss, input and output match, and for amplifiers, gain, input match and reverse isolation. Of course there are many, many other devices that are commonly tested on VNAs, such as cables, adapters, antennas, and so on.
- One of the most important performance features of the VNA is its ability to measure a broad range of power levels. The difference between the highest and lowest powers that can be measured is called “dynamic range”.

Agenda

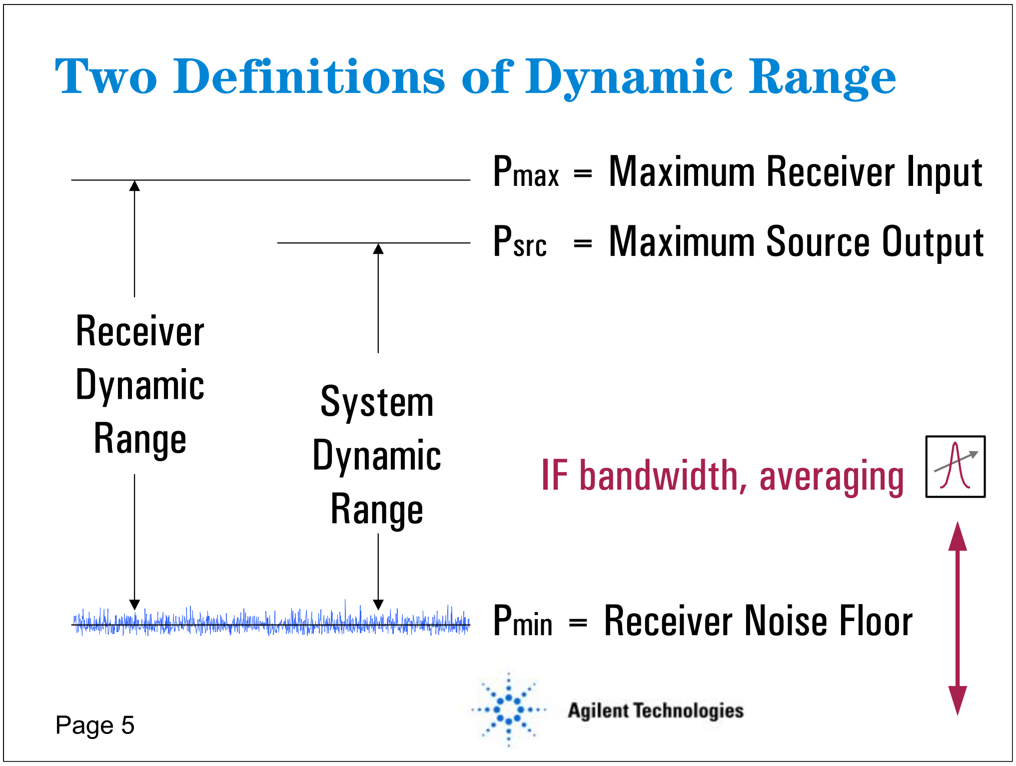
- Vector network analyzers: a quick review
- What is the strict definition of dynamic range and why does anyone care about it?
- VNA hardware considerations
- Ways to improve dynamic range
- Questions-and-Answers session



- Here's what we'll cover for the next 45 minutes or so:
- First we'll do a quick review of the block diagram of vector network analyzers. Then we'll relate that to how S-parameters are measured.
- Next, we'll offer a more rigorous definition of dynamic range and then spend some time exploring why RF designers and test engineers should even care about it.
- Next, we'll talk about various hardware considerations within the VNA itself that have an impact on dynamic range.
- In the following section, we will focus on various ways that users can optimize dynamic range, and explore the tradeoffs between dynamic range and measurement speed.
- Finally, you will have an opportunity at the end of today's seminar to ask questions about any of the material presented or about network analysis in general.



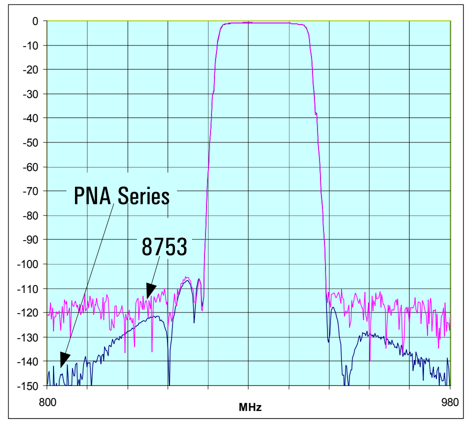
- Here is our one-slide review of vector network analyzers and S-parameter measurements. If this is new to you, at the end of the seminar we'll point you to some reference material that you can use to learn about network analyzers and RF measurements in more depth.
- VNAs are stimulus/response measurement systems. This means they have both an RF signal stimulus and several measurement receivers.
- VNAs measure the amount of signal that either reflects from the DUT or is transmitted through the DUT. Reflection results from mismatch between the DUT and the test system, and transmission is an indication of the device's loss or gain characteristics. The VNA can measure reflection and transmission in both the forward sense, where the stimulus comes out of test port one, and in the reverse sense, where the stimulus comes out of test port two.
- S-parameters are actually ratioed measurements. This means that reflection and transmission are referenced to the RF signal incident on the device. S-parameters are complex, requiring that both amplitude and phase ratios are measured.
- The block diagram shown is that of the new PNA Series of RF vector network analyzers, just introduced in September of this year. All of the RF signals are down-converted using either mixer- or sampler-based receivers. At the top, we show the RF source that provides the measurement stimulus. The RF source is phase-locked to the synthesized LO used for down conversion. The transfer switch directs the stimulus to the two test ports, one port at a time. The switch is shown in the forward mode. The splitters provide some signal for the reference receivers, R1 and R2, and some for the DUT. The directional couplers are used to separate the incident and reflected signals. The A receiver measures signals coming into port one, due to either forward reflection or reverse transmission, and the B receiver measures signals coming into port two, due to either reverse reflection or forward transmission.
- The slide also shows how the various reflection and transmission S-parameters are related to the different measurement receivers of the VNA.



- Now that we've reviewed vector network analyzers, let's look at some more rigorous definitions of what this seminar is all about: dynamic range. As was mentioned earlier, dynamic range is the range of power levels that the VNA can measure. The maximum power can be defined as either the maximum power that the receiver can measure, shown as Pmax on the slide, or the maximum power that the RF source can provide, shown as Psrc on the slide. The minimum power is set by the noise floor of the instrument's receivers, shown as Pmin.
- Although the slide makes it look like we subtract the maximum and minimum powers, we actually take the ratio between them. Since this ratio can be very large in a linear sense, we usually express dynamic range in logarithmic terms or dB. On a log-magnitude plot, dynamic range appears as the difference between two logarithmic values.
- Since there are two definitions for maximum power, we can define dynamic range in two ways. One way is to define dynamic range in terms of the measurement receivers. In this case, it is the ratio between Pmax and Pmin, which we call receiver dynamic range. This is the best possible dynamic range we could get from our VNA.
- If we think of the VNA as a stimulus/response system, then we must consider the available source power. This leads to the other definition, which is the ratio between Psrc and Pmin. We will call this system dynamic range, and it is generally the same or smaller than receiver dynamic range. At microwave frequencies, where RF source power is hard to come by, system dynamic range is usually smaller than receiver dynamic range.
- For testing passive devices using a standard VNA configuration, we are limited to the system dynamic range. However, if we are testing active devices, the gain of the DUT can compensate for loss in source power, so we can take advantage of the full receiver dynamic range of the network analyzer.
- The values that we choose for the maximum and minimum powers are somewhat arbitrary, depending on the definitions used. Pmax is limited by receiver compression, and the noise floor can be defined in several ways.
- Also note that receiver noise floor depends primarily on the bandwidth of the measurement system, which is set by the instrument's IF bandwidth. The narrower the bandwidth, the lower the noise floor, and hence the greater the dynamic range. For VNAs, noise floor is usually defined with a 10 Hz bandwidth, but sometimes it is defined in terms of 1 root-hertz. For many devices, it is desirable to trade off dynamic range for measurement speed, which we can do by using wider IF bandwidths. During this seminar, we will also explore effects other than IF bandwidth than can affect the instrument's noise floor.

Why Should I Care?

- DR is very important for:
  - measuring **high-DR** devices like base-station filters
  - measurement **speed**
  - measurement **accuracy**



Measurement Dynamic Range	8753ES Sweep Speed	PNA Sweep Speed	Improve-ment
100 dB	470 ms (1 kHz)	29 ms (35 kHz)	16x
120 dB	42.5 s (10 Hz)	1.2s (300 Hz)	35x

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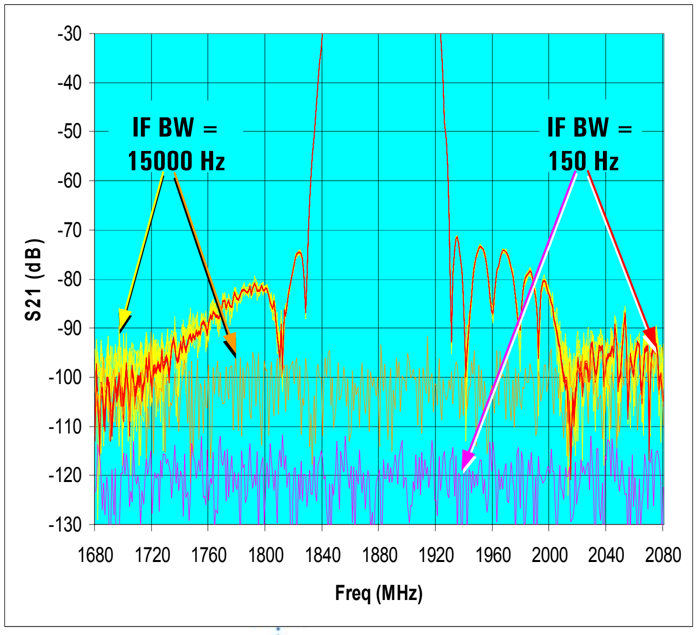
- Now that we have a good starting point to understand what dynamic range is, lets talk about why it is so important for testing RF components. The first reason is obvious: if the DUT has a lot of inherent dynamic range, then the network analyzer also needs to have a lot of dynamic range to measure the device properly. Many base-station filters fall into this category. The plot on the slide shows a 900 MHz base-station filter measured by both an 8753 and a PNA Series network analyzer, using a 10 Hz bandwidth. This results in a measurement taking around 30 to 60 seconds, depending on how many measurement points are used. As can be seen, the 8753's noise floor is too high to measure the close-in stopband performance of the filter. The PNA Series instrument on the other hand, has a much lower noise floor, which gives it the capability to measure around 150 dB of dynamic range at this frequency. This allows us to clearly see the transmission zeros or nulls in the stopband that are typical for high-reject, sharp-rolloff base-station filters like this one. We'll talk about how to set up the network analyzer for this type of measurement later on.
- Not all devices, however, require this much dynamic range. For many components, 100 dB is plenty. If the analyzer starts out with excellent dynamic range, then the measurement can be made in a wider, faster IF bandwidth, and still have sufficient dynamic range to accurately characterize mid-performance devices. The table shows measurement speed comparisons between the 8753 and the PNA Series, for a specific amount of instrument dynamic range. If 100 dB of dynamic range is needed, the PNA Series can sweep in a 35 kHz bandwidth, compared to 1 kHz for the 8753. This results in sweeps that are 16 times faster. If 120 dB of instrument dynamic range is needed, then the PNA Series can use its 300 Hz bandwidth, yielding sweeps that are 35 times faster. What takes about 43 seconds on an 8753, only takes about a second on a PNA Series analyzer. This is why it is desirable to have a lot of instrument dynamic range to begin with, so it can be traded off for measurement speed.
- The last reason why dynamic range is important is because it is needed for good measurement accuracy. The instrument's noise floor is usually the dominant contributor to measurement uncertainty for high-reject devices. It is often hard to see the underlying VNA noise floor because it can be masked by the device's response. However, we must still consider signal-to-noise ratios to understand measurement uncertainty. Greater signal-to-noise ratios result in lower measurement uncertainty and higher accuracy.

DR and Measurement Uncertainty



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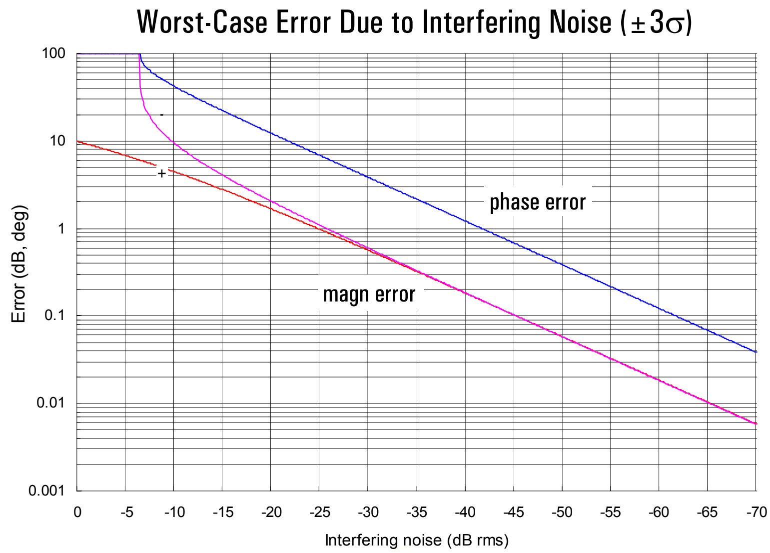
- Here is a graphical example of how signal-to-noise ratio can affect measurement uncertainty. Both the red and yellow plots of the filter's response are overlays of ten sweeps each, to visually show measurement repeatability. Repeatability is a good indication of measurement uncertainty for measurements near the analyzer's noise floor. The underlying noise of the measurement is also shown to help understand the concept. The measurement using a 15 kHz IF bandwidth shows a much wider variation in measured stopband response, as the signal-to-noise ratio is much lower than that which results from using a 150 Hz bandwidth. The next slide shows a more quantitative way to determine noise-induced measurement uncertainty based on signal-to-noise ratio.

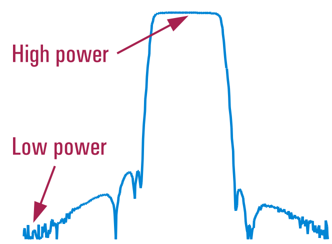
DR and Measurement Accuracy



- This plot shows both amplitude and phase error that results from adding noise to a coherent vector. Think of the vector as the desired measurement of the device's response, and the noise as resulting from the network analyzer. The horizontal axis essentially indicates signal-to-noise ratio in dB, with the level of noise determined from its root-mean-square, or rms value.
- In order to calculate the amount of error induced by the noise, we must choose a statistical-based value for the worst-case noise contribution. The rms value is not a very good choice, since much of the time, any given noise sample will exceed the rms value. And, in theory, if we wait long enough, noise can assume infinite proportions. Therefore, we must pick a value that is a reasonable trade-off between measurement uncertainty and the amount of time that the indicated worst-case error is true. For this example, the chart uses ± three-sigma or ± three standard deviations worth of noise. Thee sigma means that for each measurement point, 99.73% of the time the measurement uncertainty will be ≤ than the error shown on the plot. We could increase the percentage of measurements that fall within the worst-case error value by using more standard deviations, but the resulting error would also be higher. Three sigma is a reasonable tradeoff between measurement yield and worst-case error.
- Using ± three-sigma, we can see from the chart that -7 dB rms is the point where the worst-case noise signal is equal to the desired measurement vector, causing +6 dB and -infinity of amplitude error, and ±90° of phase error. Had we been adding two vectors, this point would have occurred at 0 dB on the x-axis.
- We can easily see that for less than ±1 dB of noise-induced error, we must have greater than 25 dB of signal-to-noise ratio. To get less than ±1 degree of noise-induced error, we must have greater than 42 dB of signal-to-noise ratio. It should be obvious that the farther the device's response is from the noise, the more accurate the measurement will be.

Elements of Dynamic Range

- Things that determine maximum receiver input
  - Damage level
  - Compression
- Things that affect the inherent noise floor
  - Conversion efficiency / noise figure
  - Crosstalk
  - Mixer/Sampler “bounce”
  - Spurious receiver residuals



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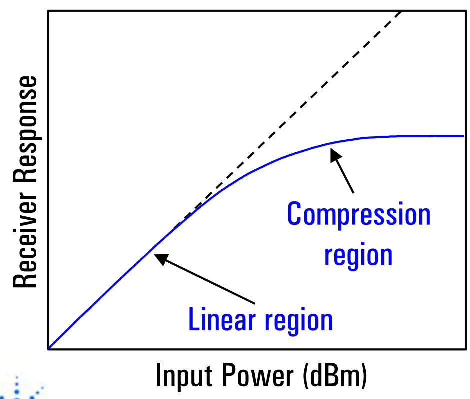
- Now that we know why dynamic range is so important, let's take a closer look at what things determine the receiver's maximum input power level and its noise floor. For maximum input, we could use the damage point. Obviously we want to keep our power below the damage point, but even if we did, we could still have a large amount of measurement uncertainty due to receiver compression. We'll cover this topic on the next slide.
- Looking at the minimum portion of dynamic range, there are four effects that can affect or degrade instrument noise floor, thereby decreasing dynamic range. The biggest contributor to instrument noise floor is the effective receiver noise figure, largely determined by what type of down-conversion topology is used. Crosstalk is the term for undesired signal leakage. Mixer or sampler bounce is the result of non-linear performance of the network analyzer, and it only affects certain measurements. Receiver residuals are the last receiver impairment we'll look at, and for most vector network analyzers, they are the smallest contributor. We will go into more detail for each of these topics in later slides.

Receiver Compression

- Power level where receiver is no longer **linear**
- No **standard** value defined for compression point
- Effect of compression **cannot** be removed by vector error correction
- Refer to **dynamic accuracy** or system **uncertainty** curves on network analyzer data sheets



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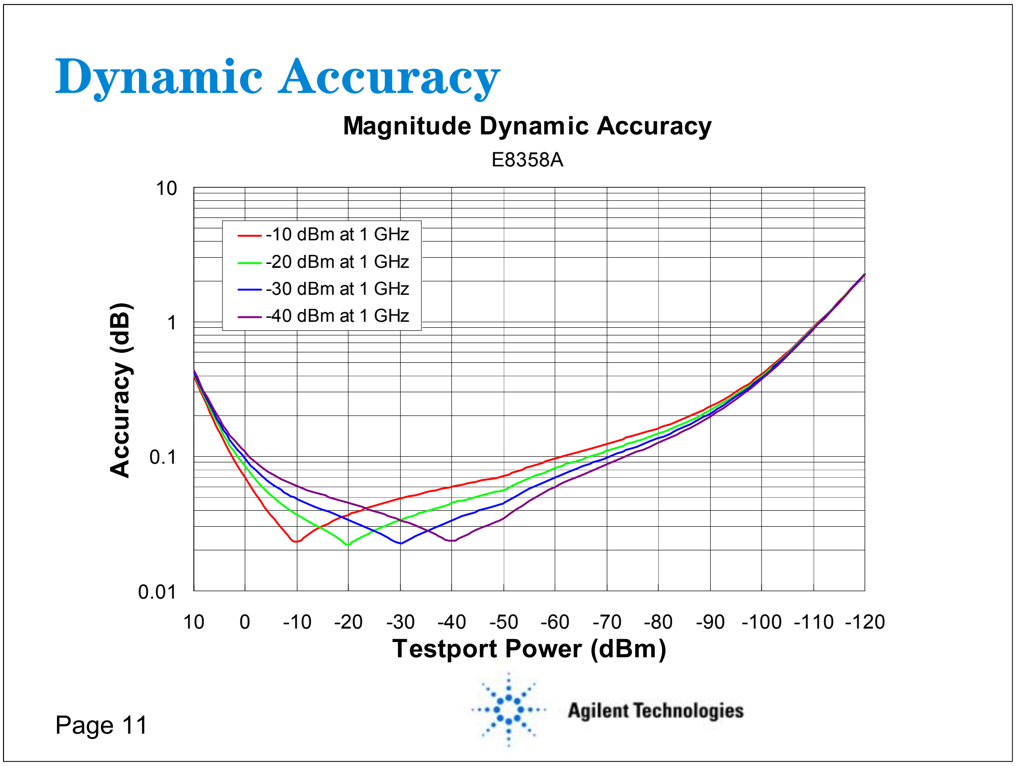
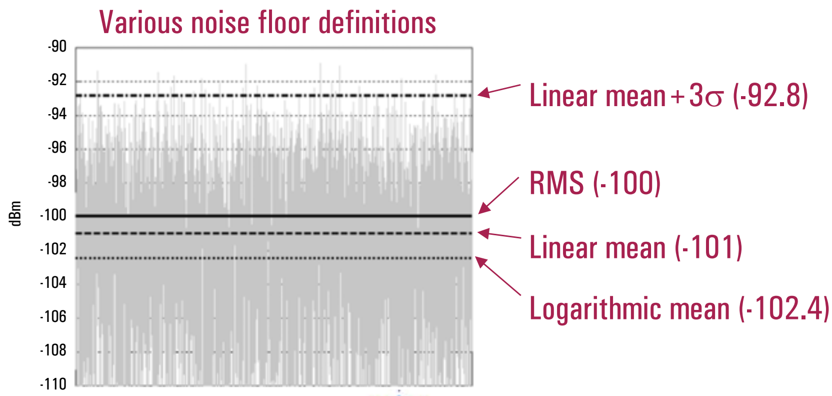
- Receiver linearity is a very important attribute of vector network analyzers for achieving high accuracy. For most of the range of the instrument, the receivers are very linear. However, as we increase input power, at some point the receiver no longer responds linearly. This means as power is increased, the indicated receive power gets lower and lower compared to the actual input. When this occurs, the receiver is said to be in compression. The power level beyond which we consider the measurement error to be unacceptable is the receiver's compression point, typically between 0.1 and 1 dB. This is the value used for maximum receiver input, from which we calculate receiver dynamic range.
- There is no standard value for receiver compression point, either between manufacturers, or even between network analyzers from a given manufacturer.
- We can easily determine the amount of receiver compression that a particular network analyzer uses for its dynamic range calculations by looking at the dynamic accuracy or system uncertainty curves in the instrument's published data sheet.



- This slide shows an example of a dynamic-accuracy chart. It shows for various calibration power levels, how much error is introduced by deviating from that reference power. No matter what power level is used during calibration, we can clearly see that when we input +10 dBm of power, the receiver is compressing by about 0.4 dB. At 0 dBm, where we have given up 10 dB of dynamic range, there is between 0.07 dB and 0.1 dB of compression, depending on the power we used during calibration. A chart such as this is very useful for determining how much accuracy we can trade off for say filter passband measurements in order to increase dynamic range and accuracy for stopband measurements.

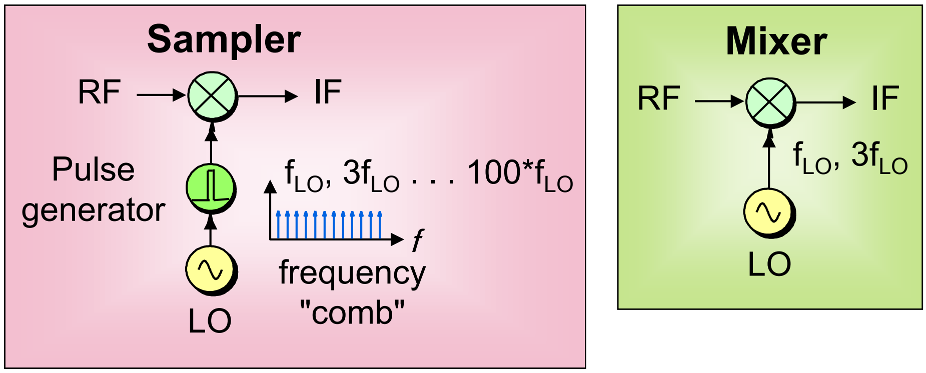
Receiver Noise Floor

- No **standard** definition for noise floor
- PNA Series uses **RMS** (equivalent input noise power)



- Now let's cover in more depth the things related to the receiver's noise floor, which sets the minimum power that the network analyzer can measure. First of all, let's look at the ways that noise can be defined. Depending on the definition used, our dynamic range specification can differ by around 10 dB.
- The slide shows Gaussian noise with an rms value of -100 dBm, calculated by computing 10*log of the rms of the linear values. Using this noise, various other definitions were also calculated and indicated on the slide. One calculation is to simply take the mean of all of the linear values, and then convert the mean to dBm. This results in a value of -101 dBm, which gives a 1 dB dynamic range advantage compared to using the rms value.
- Another approach is to calculate the mean of the logarithmic values, which results in -102.4 dBm. This approach gives an even larger increase in dynamic range of 2.4 dB, compared to using the rms value. This is also the value that we would measure by applying lots of smoothing, or if we used video averaging or a small video-bandwidth filter on a spectrum analyzer.
- The most conservative approach is to take into account some decrease in dynamic range due to noise peaks. Using the linear mean plus three standard deviations ($+3\sigma$), converted to dBm, we see the noise floor is degraded to -92.8 dBm, which is about 7 dB higher than the rms value. The problem with this approach is that the manufacturer may not choose the right amount of standard deviations to add to the mean. For this reason, Agilent chose to define the noise floor of its new PNA Series of vector network analyzers using the rms value. This is probably the best understood and most versatile definition of noise level. With this definition, it is easy to calculate the noise peaks using any desired number of standard deviations. The rms value of noise is also equal to the equivalent input noise power of the measurement receivers.

Mixers versus Samplers: Hardware



- Mixer-based front ends yield **lower** noise floors than sampler-based front ends
- Lower NF gives **dB-per-dB** improvement in DR

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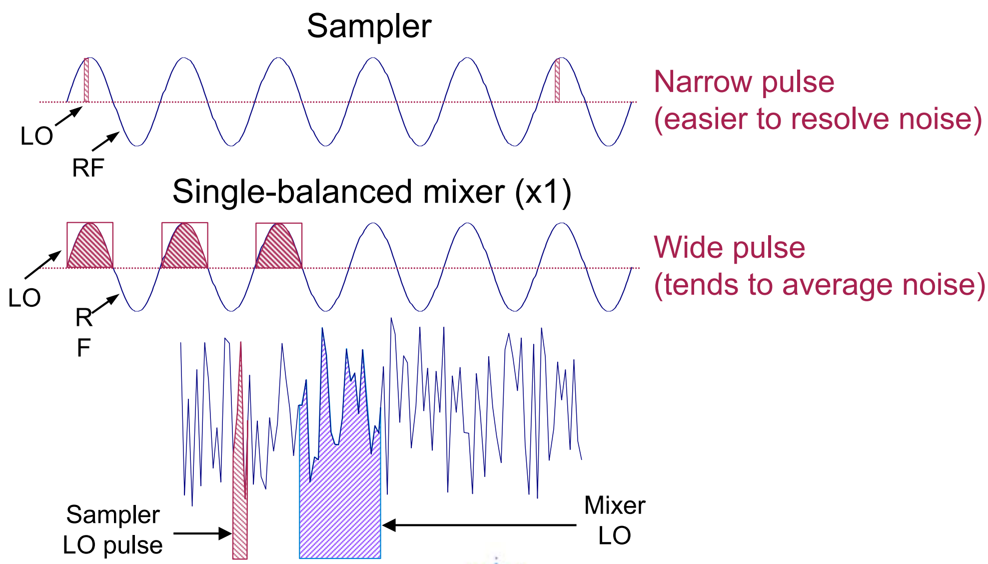
- The biggest contributor to the noise floor of the network analyzer is the receiver front-end topology chosen for the down-conversion process. All vector network analyzers use either sampler-based or mixer-based front ends. Mixer-based front ends yield lower noise floors than sampler-based front ends. Another way to say this is that mixer-based front ends have a better noise figure. Every dB reduction in receiver noise figure yields a dB increase in dynamic range. The 8753 and 8720 families of vector network analyzers use samplers, while the new PNA Series uses mixers.
- The sampling approach, while more complicated conceptually, is actually easier and thus cheaper to achieve in hardware, especially at microwave frequencies. A sampler can be thought of as a mixer driven with a narrow pulse that is rich in harmonics. The LO for a sampler-based front end does not need to sweep as high as one used for a mixer-based front end, since we are using the LO's harmonics for mixing. The RF stimulus present at the input to the sampler mixes with one of the harmonics, which are often called "comb teeth", since the frequency-domain spectrum of the LO pulse looks like an upside-down comb. The result of this mixing is an IF signal. With this approach, we can easily need 100 or more harmonics to cover the desired frequency range of the analyzer.
- The mixing approach is conceptually very simple. The RF signal mixes with either the fundamental of the LO, or as is common for upper frequency bands, with the third harmonic of the LO. The next two slides will attempt to explain why samplers have higher noise figures, using both a time domain and a frequency domain approach.

Mixers Versus Samplers: Time Domain



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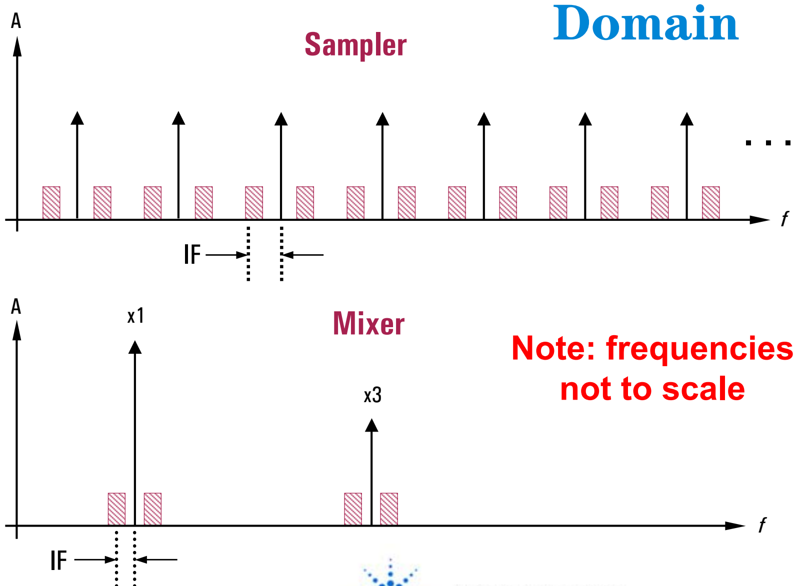
- Let's look at the difference between samplers and mixers in the time domain first. Samplers use very narrow pulses to sample the RF input, compared to fundamental or third-order mixing. The narrow pulse is what makes a harmonic-rich LO in the frequency domain. This narrow pulse also gives more time-domain resolution, making it easier to follow the peaks and valleys of the noise. The result is that there is more noise on the IF signal.
- In contrast, the mixer's LO is on for roughly half of the RF cycle, assuming a single-balanced mixer, which is typically the case for RF front ends. This longer period provides much more noise averaging. The result is less noise on the IF signal.

Mixers Versus Samplers: Frequency Domain



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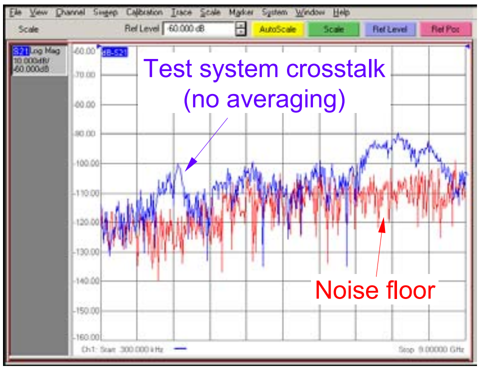
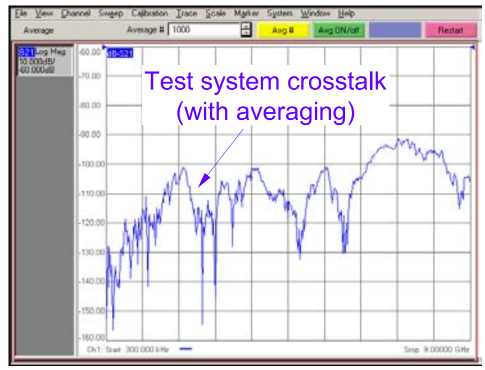
- Now let's use a frequency-domain approach to explain why there is more noise conversion using samplers.
- As was mentioned earlier, there are many harmonics of the LO in the frequency domain when using a sampler. Any noise present one IF away from every comb tooth, on either side, will be down-converted and detected in the IF. Since there are so many more harmonics, much more noise conversion takes place compared to using mixers, where noise is converted only around the fundamental and third harmonic of the LO. The noise multiplication effect from all of the sampler LO harmonics result in the sampler having a worse noise figure than the mixer. Typically, the difference is around 20 to 30 dB, depending on the frequencies involved.
- Both the time domain and frequency domain approaches are valid ways at looking at the down-conversion process. They are just two different ways of explaining the same phenomenon.

Crosstalk

- Crosstalk is signal leakage that **bypasses** the DUT during a transmission measurement
- Usually very **close** to the noise floor





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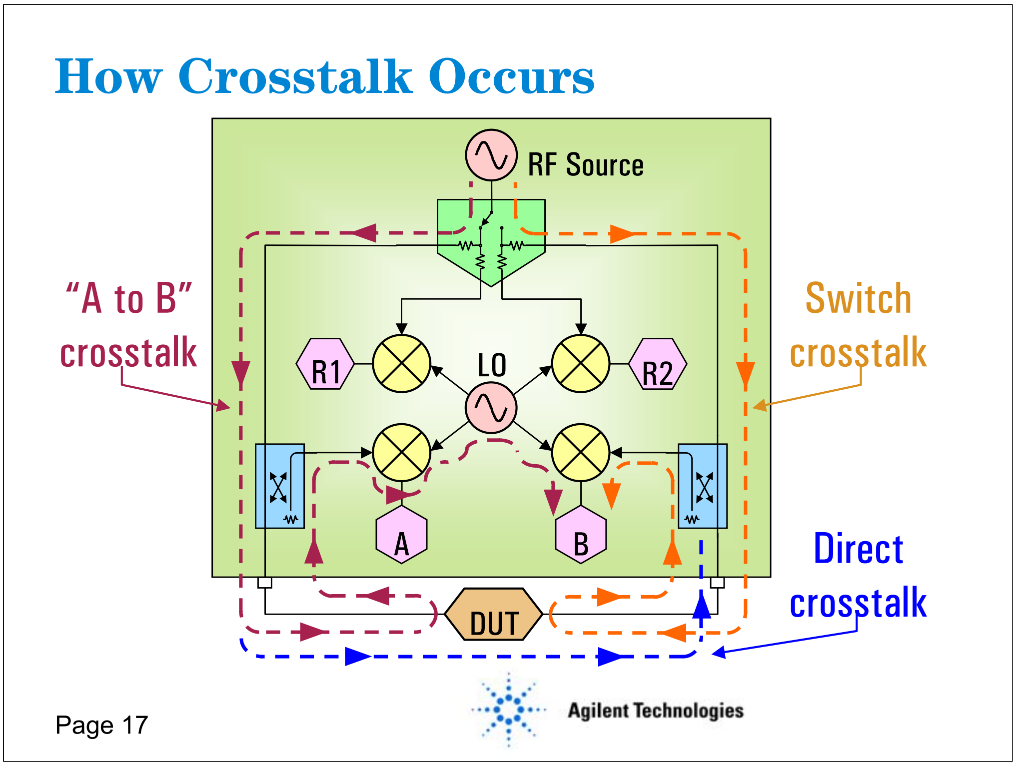
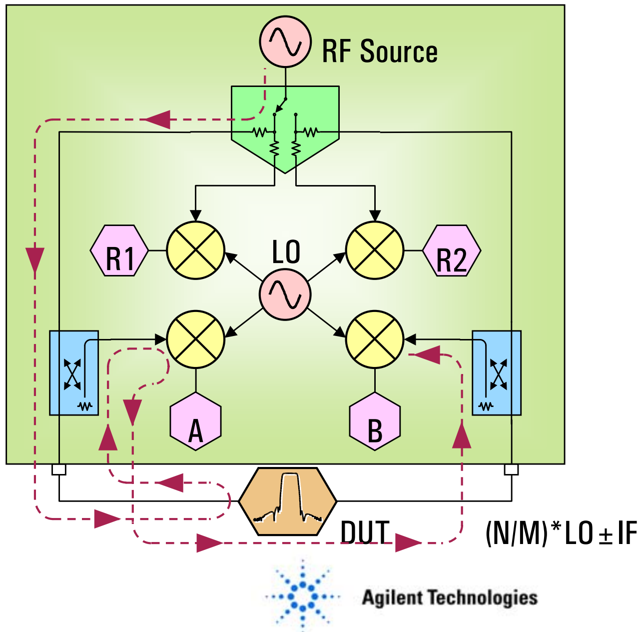
- Now that we have seen how the down-conversion architecture affects noise floor, let's look at some effects that can be higher than the noise floor, causing dynamic range degradation. The first impairment we'll talk about is crosstalk. Crosstalk is the signal leakage that bypasses the DUT during a transmission measurement. The leakage can either be caused by mechanisms within the network analyzer itself, which we'll cover in the next slide, or it can result from elements of the test system outside the network analyzer, such as test cables, fixtures or probes.
- Crosstalk is usually very close to the noise floor of the test system. We hope that it remains below the analyzer's noise floor, but often it will pop up out of the noise at certain frequencies, where it limits dynamic range. The plot on the left shows two traces: one shows test system crosstalk and one shows the analyzer's noise floor without any crosstalk. To simulate test system crosstalk for this example, two test cables were attached to the two test ports of the network analyzer, while the free ends had loosely-attached 50-ohm terminations.
- Because of the low signal-to-noise ratio typical of crosstalk, we must generally use lots of averaging to clearly see it. The plot on the right shows the test system crosstalk with about 100 averages. The effective noise floor of the analyzer has been reduced considerably, resulting in a better measurement of the crosstalk.



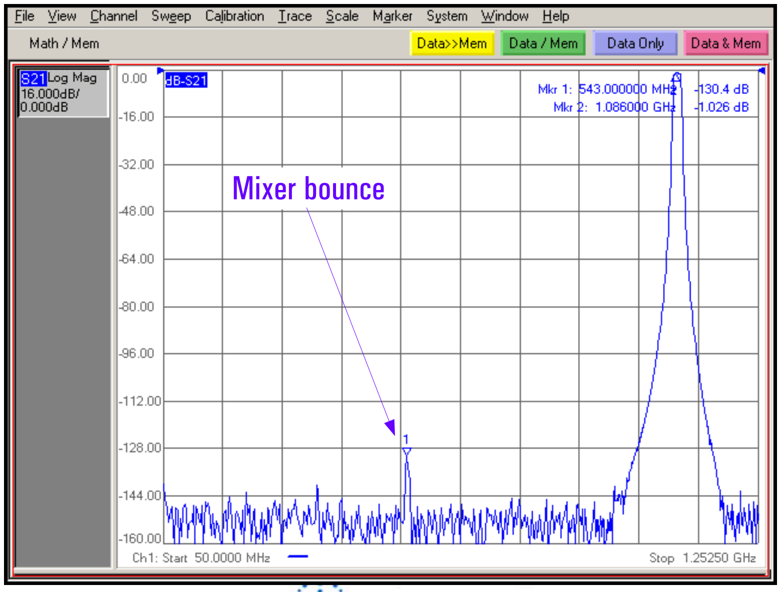
- There are actually three mechanisms that can cause crosstalk. Direct crosstalk is the most obvious: it's the signal that leaks around the DUT outside of the network analyzer. It can be greatly minimized or effectively removed by using high-quality test cables, adapters, fixtures, and so on. Good connector techniques are also important to reducing direct crosstalk. This means cleaning the connectors properly, using the correct amount of torque when tightening the connectors, and replacing damaged or worn connectors.
- There are two types of crosstalk that can occur within the network analyzer itself. The first type is called "A to B" crosstalk because it is the crosstalk that results from signal leakage between the A and B measurement receivers. As shown on the slide, A to B crosstalk results from RF signals that reflect from the input of the DUT and get back into the analyzer, where they are coupled into the A receiver. Since we only need the B (and a reference) receiver for a forward transmission measurement, any signal present in the A receiver is ignored (for transmission only -- the signal is not ignored for measuring forward reflection, or S11). However, since there is only finite isolation through the LO chain, a very small amount of the RF signal can leak from the A to the B receiver, where it adds to the direct crosstalk term. A to B crosstalk is very dependent on the match of the DUT. The better the match, the smaller the signal which reflects back into the network analyzer.
- The other type of crosstalk that can occur within the network analyzer is due to the RF transfer switch, which also has finite isolation. Any RF signal that leaks across the open side of the switch comes out of the test port used for measuring transmission, which, in the example shown on the slide, is port two. This signal can then reflect off the output match of the device, and then re-enter the network analyzer where it is coupled to the B receiver. This crosstalk term is also dependent on the match of the DUT, just as A to B crosstalk was.
- Crosstalk can be removed as part of vector-error correction. We'll explain how to do this near the end of the presentation. Much work went into the design of the new PNA Series to ensure that the crosstalk terms caused by the network analyzer are below the analyzer's noise floor in a 10 Hz bandwidth. This gives these analyzers lots of usable dynamic range and largely eliminates the need to attempt removal of the crosstalk term during calibration.

Mixer/Sampler Bounce



- The next type of dynamic range impairment we'll study is called mixer or sampler bounce, depending on which approach is used for down-conversion. Mixer bounce is caused by non-linear performance of the network analyzer. It is the result of signals within the network analyzer converting to different frequencies, as opposed to crosstalk where the leakage terms are always at the same frequency as the RF stimulus.
- Mixer bounce is only noticeable for devices that have both a high-reflect region and a low-loss region, such as a filter. The example we'll use is a bandpass filter. As the slide shows, just like with crosstalk, some of the RF stimulus will reflect off the input match of the DUT during a transmission measurement, and this reflected signal will then couple into the A receiver. What makes this different than crosstalk, is that this RF signal mixes with the LO present within the mixer or sampler, and new signals come out of the receiver that are at frequencies equal to (N/M)*LO $\pm$ IF. If one of these signals happens to fall within the passband region of the filter, it passes through the DUT, couples into the B receiver, and then mixes down to the IF, where it is detected as a response. Any of the mixing products that fall outside of the filter's passband are attenuated by the filter's stopband response, and will not appear on the network analyzer's display. There can be multiple points during a network analyzer sweep where a mixer bounce product can appear on screen. Since there are usually buffer amplifiers before the receivers, any mixer or sampler bounce signals that come out of the receivers are attenuated by the buffer's reverse isolation. For this reason, mixer or sampler bounce responses are usually very small. Most are below the instrument's noise floor, but occasionally one can be seen poking out of the noise. The next slide shows an example of this, using a 1.09 GHz bandpass filter.

Example of Mixer Bounce

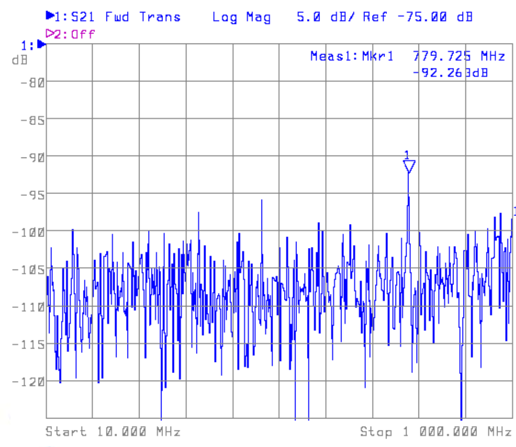


- In this example, using a PNA Series analyzer, the mixer bounce occurs when the RF signal is at 543 MHz. This happens to be a sub-harmonic of the filter's passband. The 543 MHz signal gets multiplied by two inside the A receiver to 1.086 MHz, where it then comes out of test port one and travels through the passband of the filter, causing a false response.
- There are ways to minimize or remove the mixer bounce term, which we'll look at in the next section of this seminar.

Receiver Residual Responses

- Response shows even when **no** RF signal present
- Caused by LO **self-mixing** with its spurious sidebands to produce an IF signal

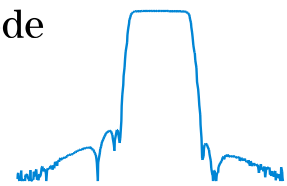
- Result is very **sharp** responses -- look like spikes poking out of noise floor
- Usually not a limiting factor for high performance VNAs



- The last dynamic range impairment we'll cover is due to receiver residual responses. These are responses that show up on the screen even when there is no RF stimulus signal present in the receivers. Receiver residuals are the result of spurious sidebands on the LO self-mixing to produce an IF signal. These signals are generally at a very low level, and are usually hard to even see. They typically produce very sharp responses, looking like spikes poking out of the noise floor. From sweep to sweep, they tend to be very noisy.
- Receiver residuals are usually not a limiting factor for high performance VNAs. The example on the slide shows a residual on an instrument from the 8712 family of low cost vector network analyzers.

Agenda

- What is DR and why does anyone care about it?
- VNA hardware considerations
- Ways to improve dynamic range
  - Maximize **input** power
  - IF bandwidth **reduction** and **averaging**
  - **Direct** receiver access/reversed coupler
  - Adding a low-noise amplifier (**LNA**)
  - **Alternate** versus “**chop**” mode
  - Performing an **isolation** cal
  - Minimizing **residuals**



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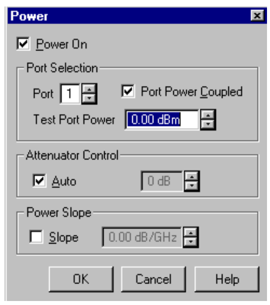
- Now that we fully understand all of the ways in which dynamic range can be affected, it's time to talk about the various techniques that you can use to improve the dynamic range of your network analyzer measurements.
- We'll start off with the basics of test power, and then explain some subtle differences between using IF bandwidth reduction and averaging to lower the instrument's noise floor.
- As was previously promised, we'll then cover how to configure the network analyzer to make extended dynamic range measurements of up to 150 dB, like we saw earlier. We'll also see how adding an LNA to the measurement can improve the dynamic range of most network analyzers.
- Next, we'll cover the difference between "chop mode” and alternate sweeps, which affects mixer or sampler bounce.
- Then, we'll talk about removing system crosstalk with an isolation calibration, and finish up with a brief mention on minimizing receiver residuals.

Maximizing Input Power

- Always use as **much** source power as possible
- Passive devices: limited by maximum **source power** available or receiver **compression** (note: default power level is often not maximum)
- Active devices: **lower** source power to account for gain of device (requires source attenuator)



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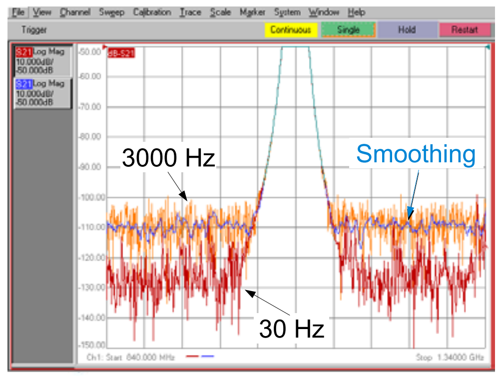
- The first step to improving dynamic range is to always use as much source power as possible. While this sounds obvious, it is very easy to skip this step when setting up a network analyzer measurement.
- For passive devices, maximizing input power to the DUT is either limited by the maximum source power available from the network analyzer, or by receiver compression considerations as was covered earlier. If you are trying to make an extremely accurate measurement of a low-loss device, it is often necessary to back off from the maximum source power condition.
- It should be noted that the default value for test port power is often not the maximum amount available. If you want to maximize dynamic range, even at the expense of some receiver compression at the high-power end, then you usually have to consciously set the source power to its maximum value. This is the step that is often forgotten.
- When testing active devices like power amplifiers or transistors, the source power level must be lowered to account for the gain of the DUT. For example, if we want to keep the output of a 25 dB amplifier to less than +10 dBm to avoid excessive receiver compression, we must set the source power to -15 dBm. This requires a source attenuator, which is generally available either as a standard feature or as an option. Usually control of the source attenuator is automatic so that it sets itself to the correct attenuation value based on the desired test port power. Many vector network analyzers allow different test port powers for the forward and reverse sweeps.

IF Bandwidth Reduction & Averaging

- **Reducing** IF **bandwidth** by a factor of ten **lowers** the noise floor (and increases DR) by 10 dB
- Vector **averaging** has a similar effect: **increasing** the amount of averaging by a factor of ten **lowers** the noise floor (and increases DR) by 10 dB
- **Smoothing** cleans up trace, but does **NOT** improve DR



- Let's now take a look at using IF bandwidth reduction and averaging to reduce the noise floor of the measurement receivers. When we first talked about dynamic range in the beginning of this seminar, we mentioned that the receiver's noise floor is a direct function of measurement bandwidth. Reducing the IF bandwidth by a factor of ten lowers the noise floor by 10 dB and correspondingly increases dynamic range by 10 dB. This also causes the measurement to be slowed by about a factor of 10.
- The example on the slide shows about a 20 dB reduction in the noise floor resulting from a 100 times decrease in IF bandwidth, from 3 kHz down to 30 Hz.
- Vector averaging has a similar effect: increasing the amount of averaging by a factor of ten again lowers the noise floor by 10 dB and again increases dynamic range by 10 dB. That averaging and IF-bandwidth reduction have the same effect on the noise floor is not surprising since IF-bandwidth reduction is accomplished by averaging more samples on a point-by-point basis. Increasing the amount of averaging by a factor of ten also causes the measurement to be slowed by a factor of 10.
- As a side note, smoothing is sometimes confused with vector averaging. Smoothing is a mathematical technique that can be used to clean up noise on a trace by using a sliding multi-point average. But, this technique does not fundamentally improve dynamic range. It simply reduces the peak-to-peak excursions of the noise, as can be seen on the slide.

Speed Tradeoffs (Using PNA Series)

Measurement time impact with wide IF bandwidths

IF bandwidth	Number of averages	Noise floor reduction (dB)	Measurement time increase factor
10 kHz	0	0	1
10 kHz	10	10	10
1 kHz	0	10	7.75
10 kHz	100	20	100
100 Hz	0	20	74.8

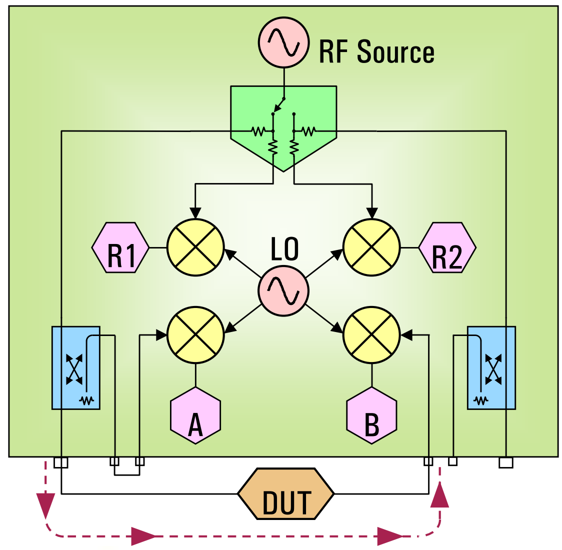
Measurement time impact with narrow IF bandwidths

IF bandwidth	Number of averages	Noise floor reduction (dB)	Measurement time increase factor
100 Hz	0	0	1
100 Hz	10	10	10
10 Hz	0	10	9.9
100 Hz	100	20	100
1 Hz	0	20	99.5

- This slide shows the speed tradeoffs between using IF-bandwidth reduction and averaging to reduce receiver noise. In the upper table, where we start out with a relatively wide IF bandwidth, we see a much larger difference between the two techniques than we see in the lower table, where we start with a relatively narrow IF bandwidth.
- Let's take a closer look at this phenomenon. In the upper table, the 10 kHz bandwidth with no averaging is our reference point. If we perform 10 averages, we see measurement time increase by a factor of 10, which makes sense since we are taking exactly 10 times as many sweeps. However, when we reduce the IF bandwidth by a factor of 10, down to 1 kHz, without using averaging, we see the measurement time increase only by a factor of 7.75, resulting in a 22.5% improvement in sweep speed compared to using averaging. This is because we don't multiply the dead time between sweeps by a factor of 10, as occurs when we perform 10 averages. When we decrease the IF bandwidth by a factor of 100, we get a 25.2% advantage. So, for speed reasons, it is better to reduce the IF bandwidth rather than use averaging, when starting with a wide IF bandwidth.
- In the example in the lower table, 100 Hz is used as the reference bandwidth. In this case, there is not nearly as much speed advantage to using IF-bandwidth reduction compared to using averaging. We see only a 1% and 0.5% advantage with a reduction factor of 10 and 100 respectively. This is because the sweep times are very long to begin with when using narrow IF bandwidths, so the dead time between sweeps becomes much less of a factor. These slow sweep times also give a reason why it might be advantageous to use averaging instead of bandwidth reduction. With averaging, you can see one sweep in one-tenth the time it would take to see one sweep with a bandwidth 10 times narrower. Although we wouldn't see a reduced noise floor after only one average, it is 10 times quicker to see if the measurement is set up properly. This doesn't waste as much time if we wish to change something as compared to waiting for the longer sweep.

Direct Receiver Access & Segmented Sweeps

- Requires access to receivers using **jumpers**
- Must use **segmented sweep** to increase DR
- **Response** calibration only



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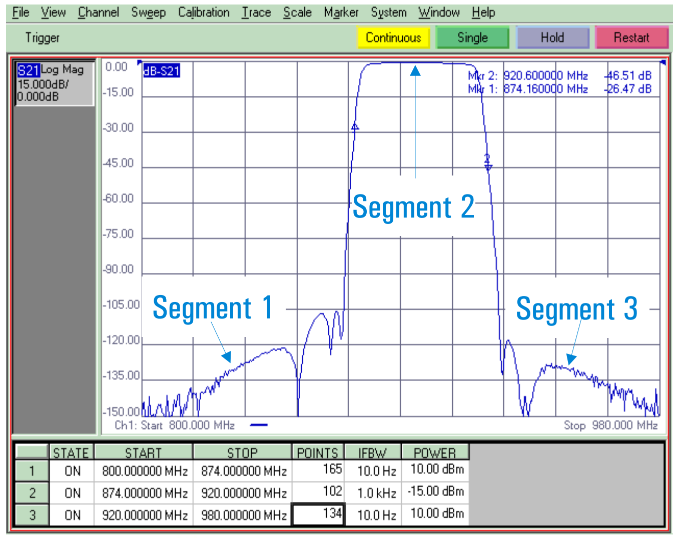
- This slide shows how we can re-configure the network analyzer's hardware to significantly enhance dynamic range in the forward direction. If we have direct access to the instrument's measurement receivers, we can bypass the directional coupler at test port two, which can increase our dynamic range by an amount equal to the coupling factor of the directional coupler. Using a PNA Series analyzer, this increase is about 15 dB. The jumpers required to gain access to the measurement receivers are standard with these analyzers. While it is necessary to bypass the directional coupler to achieve extended dynamic range, it is not sufficient.
- In order to realize a gain in dynamic range, we must also use a segmented sweep. This sweep type is called a list sweep in the 8753. When measuring a device like a base-station filter, where this technique is often used, a segmented or list sweep allows us to keep the power level high in regions where the filter has high rejection, and lower the power in regions where the filter exhibits low loss. This is necessary so we don't cause excessive receiver compression, since the maximum power level we can input to the receiver is lowered by the coupling factor. If we used a normal linear sweep with a fixed power level, we would only shift the range of measured powers downward, without increasing dynamic range. The next slide shows an example of a segmented sweep to measure a bandpass filter.
- When we bypass the instrument's directional coupler, only a response calibration can be performed. Often, this is adequate for filter stopband measurements, where accuracy is usually limited by noise, not by vector error correction. The next technique we'll explore will also achieve extended dynamic range, but it allows full two-port calibration for better measurement accuracy.

Setting Up a Segmented Sweep

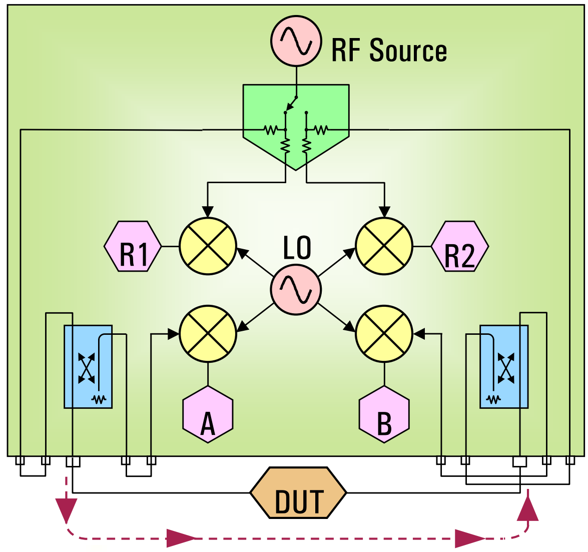
- **Maximum** power and **narrow** BW in stopbands
- **Lower** power and **wider** BW in passband
- **Skip** bands where no data is needed



- Setting up a segmented sweep is pretty straightforward, especially using a PNA Series analyzer as shown on the slide. We define a series of segments, where each segment has its own start and stop frequencies, number of trace points, IF bandwidth, and, most important for dynamic range improvements, its own power level. In the example shown on the slide, we used only three segments. Segments 1 and 3 were used to measure the close-in stopbands of the filter, and segment 2 was used to measure the filter's passband response.
- In the stopbands, we can use maximum power and narrow IF bandwidths for good dynamic range. In the passband, we must lower the power to avoid receiver compression. We can also use a wider IF bandwidth, which has the added benefit of increasing measurement time in that segment. Since we usually don't need a lot of dynamic range to measure a low-loss region, there is no need to use a narrow IF bandwidth in this segment. However, if we are trying to simultaneously achieve very accurate passband measurements by minimizing high-level trace noise, at the same time as wanting extended dynamic range, we might choose a narrow IF bandwidth even in the passband segment.
- Segmented sweeps can also save measurement time be skipping frequency bands where data is not needed. This works well when limit lines are used for certain regions of the device's response -- you only need to take data where a limit test will be performed. As is readily apparent, segmented sweeps are a great technique to optimize both dynamic range and measurement speed.

Reversing the Transmission Coupler

- Requires necessary **jumpers** for access
- Must use **segmented sweep** to increase DR
- Can use **two-port** calibration

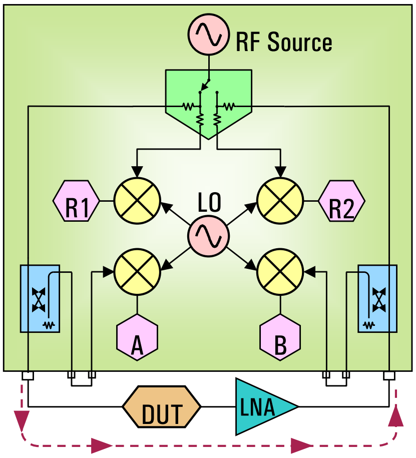


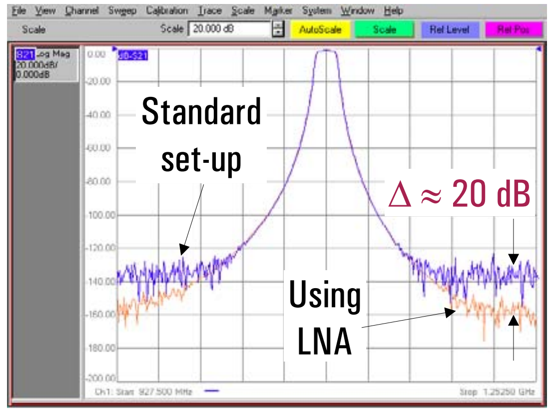
Note: Option 015 for PNA Series adds extra jumper for access to RF source

Option 014 for 8753 family adds necessary receiver and source jumpers

- There is another slightly more complicated technique that also can be used to extend dynamic range. Instead of bypassing the directional coupler, we can reverse the coupler by swapping the main and coupled arms as shown on the slide. This can be done by reconfiguring front panel jumpers or by installing switches inside the network analyzer.
- Following the transmission signal on the slide, we see that it enters port two of the network analyzer, where it travels through the main arm of the coupler to the B receiver. With the network analyzer configured normally, the signal would travel through the coupled arm, resulting in around 15 dB of loss. This technique provides extended dynamic range for one direction only, which in this case, is the forward direction. For reverse measurements, the stimulus travels through the coupled arm, resulting in 15 dB less port power than normal. However, since S21 = S12 for passive devices like filters, this is a good tradeoff since we only need to characterize transmission in one direction. The 15 dB loss in output power is not significant when measuring S22, the device's output match, since we don't need as much dynamic range for reflection measurements.
- This configuration also requires the use of segmented sweeps, just as was needed for the bypassed-coupler configuration.
- Reversing the directional coupler allows us to use two-port error correction on our filter measurements for the best accuracy. Since the coupler is still present in the signal path, instead of being completely bypassed as before, it still provides a signal-separation function at port two. Although we may not want to display the reverse S-parameters, we can still measure them. The reverse S-parameters can then be used for the 2-port error-correction calculations.

Adding a Low-Noise Amplifier





- LNA improves noise figure of VNA
- Must use segmented sweep
- Response calibration only

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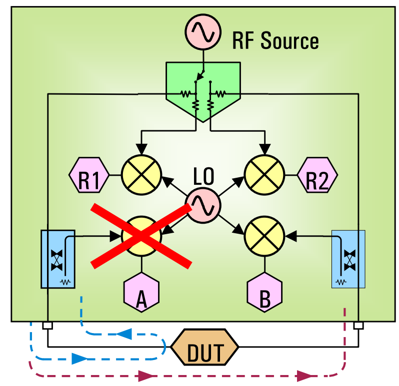
- Adding a low-noise amplifier or LNA to the measurement configuration can also extend the instrument's dynamic range. In this case, we are effectively improving the noise figure of the B receiver inside the network analyzer. This is similar to using an LNA in front of a spectrum analyzer to improve measurement sensitivity.
- Just as before, we must use a segmented or list sweep to extend the dynamic range. If we don't, we are just shifting the range of powers the network analyzer can measure, without increasing dynamic range.
- The slide shows the LNA inserted between the output of the DUT and port two of the network analyzer. Notice that this configuration does not require the use of front-panel jumpers for direct-receiver access or for reversing the directional coupler at port two. This technique can be used with any two-port network analyzer that can do segmented or list sweeps.
- The plot shows that a 20 dB improvement in dynamic range was obtained with this particular filter.

Alternate versus "Chop" Sweeps

- Chop-sweep mode (default) measures **two** S-parameters at a time (S11 & S21, or S22 & S12)
- Alternate-sweep mode measures **one** S-parameter at a time and **shuts off** the unused A or B receiver
- Alternate sweeps **improve** mixer/sampler bounce (and A to B crosstalk), but it takes **twice** as long to measure all four S-parameters (four versus two sweeps)



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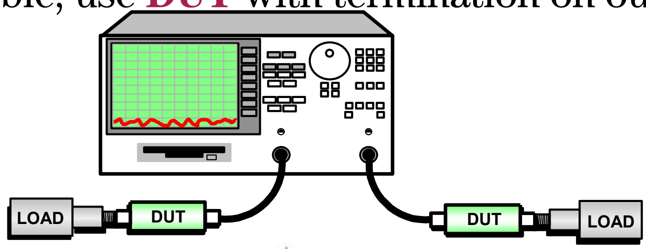
- At the end of the explanation of mixer and sampler bounce, mention was made of a measurement setup that can greatly reduce or completely eliminate the problem. This setup uses the alternate-sweep mode of operation. The default mode of sweeping for the network analyzer is what Agilent calls "chop mode". In this mode, at each measurement point of the sweep, the network analyzer multiplexes or "chops" between both the A and B receivers. This allow us to measure two S-parameters at a time. All four S-parameters requires only one forward and one reverse sweep.
- In alternate-sweep mode, only one S-parameter at a time is measured, and the receiver not needed for a particular S-parameter is shut off. For example, when measuring forward transmission, where mixer bounce can occur in chop mode, the A receiver is turned off in alternate-sweep mode. This prevents any signals from mixing within the A receiver and coming back out of test port one. This can also improve A-to-B crosstalk, which is caused by leakage between the A and B receivers via the LO path.
- The tradeoff with alternate sweeps is that it takes twice as long to gather all four S-parameters. In other words, it takes four sweeps to measure all four S-parameters, instead of just two sweeps as is the case using chop mode. If only forward transmission is needed, without two-port error correction, then there is no difference in sweep speed between using alternate or chop sweeps.

Performing an Isolation Calibration

- Adds **noise** to error-corrected traces
  (isolation term is often near system noise floor)
- Should only be done when **crosstalk** limits DR
- May need **narrow** IF bandwidth and **averaging**
- If possible, use **DUT** with termination on output



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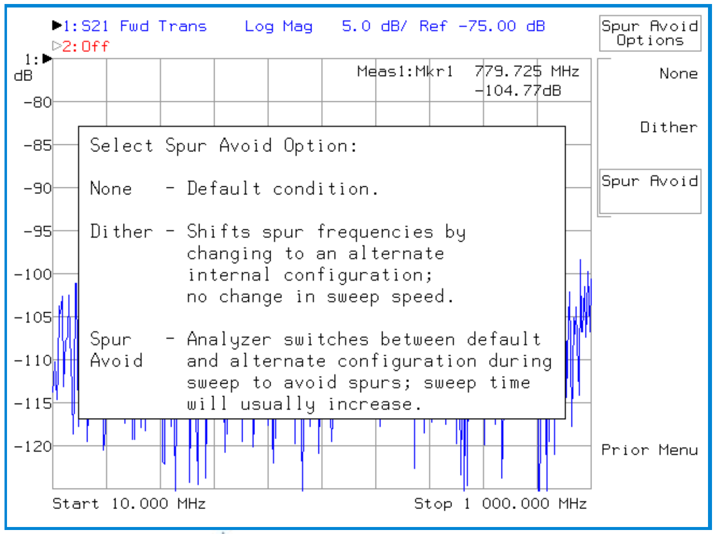
- When performing a two-port calibration, the user has the option of omitting the part of the calibration that characterizes crosstalk or isolation. The isolation calibration adds noise to the error model since we usually are measuring near the noise floor of the system. For this reason, one should only perform the isolation calibration if crosstalk would otherwise limit dynamic range. If the isolation portion of the calibration is done, a narrow IF bandwidth and trace averaging should be used to ensure that the system crosstalk is not obscured by noise.

- The best way to perform an isolation calibration is by placing the devices that will be measured on each test port of the network analyzer, with terminations on the other two device ports. Using this technique, the network analyzer sees the same impedances versus frequency during the isolation calibration as it will during subsequent measurements of the DUT. If this method is impractical (for example, if only one DUT is available), then placing a terminated DUT on the source port and a termination on the load port of the network analyzer is the next best alternative (the DUT and termination must be swapped for the reverse measurement). If no DUT is available, or the DUT is in a test fixture, or if the DUT will be tuned (which will change its port matches), then the user can either use a 50-ohm termination or a high-reflection standard like a short or open. The proper choice can be found empirically.

Receiver Residual Responses

- **Changing** start or stop frequencies slightly will often avoid problem
- Use **spur-avoidance** sweep mode if available



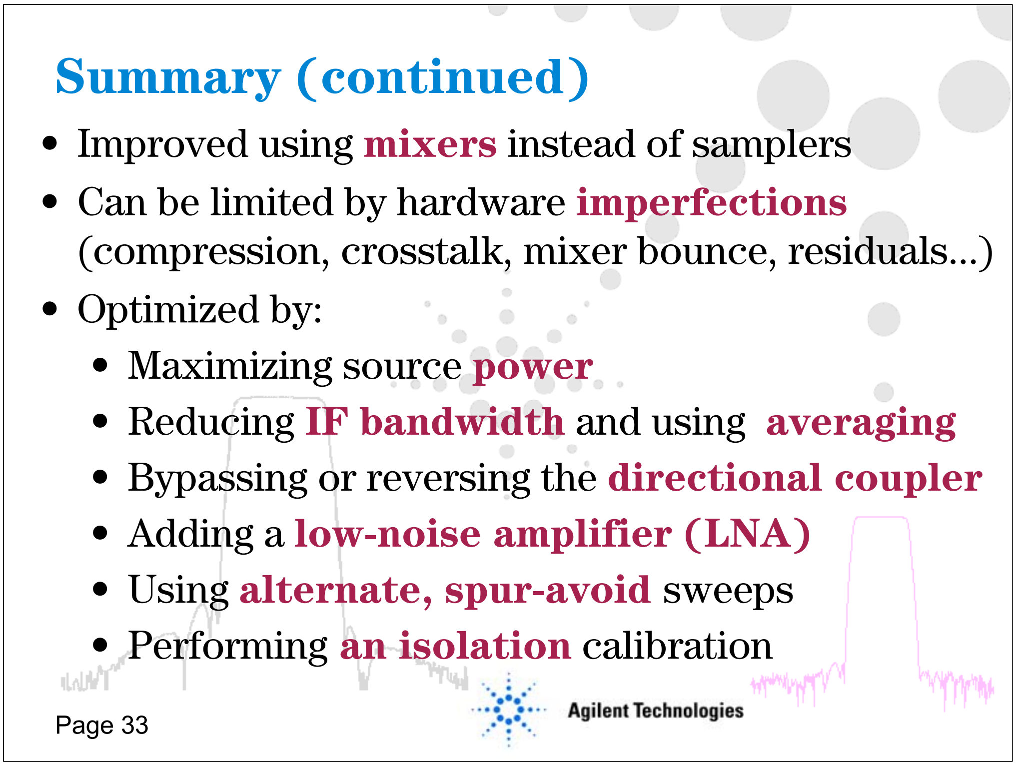
- The last technique we'll cover to improve dynamic range is that of spur-avoidance sweep mode. Spur avoidance is a technique in which one or more of the internal LOs used for the down-conversion process is shifted slightly to avoid situations where mixing products might be produced at the IF. Although our simplified block diagram has shown only one LO, there are actually more since network analyzers typically use double or triple down-conversion schemes.
- Many network analyzers, like the new PNA Series, perform spur-avoidance sweeps automatically, since there is no significant time penalty to do so. However, some instruments, like the 8712 family of low cost network analyzers, allow the user to control whether or not spur-avoidance techniques are used during the sweep. The example on the slide is from one of these analyzers. This allows users to make yet another tradeoff between sweep speed and dynamic range.

Summary

Dynamic range ...

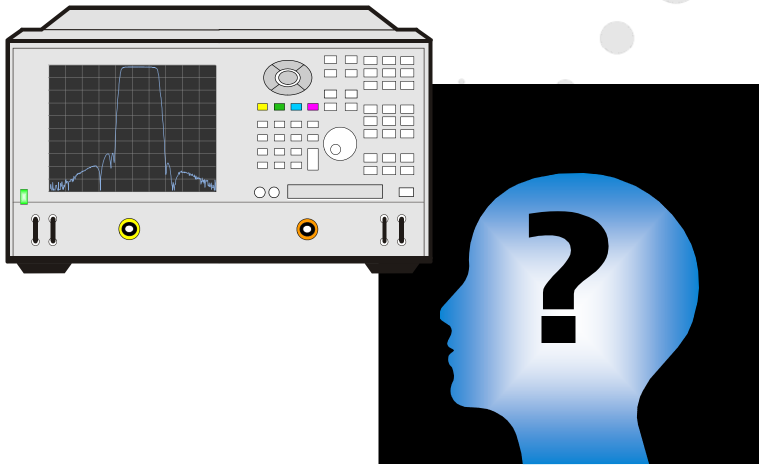
- Determined primarily by maximum source or receiver **power** and receiver **noise floor**
- A very **important** aspect of VNA performance
  - Testing **high-dynamic-range** components
  - Trade-off DR for measurement **speed**
  - Essential for good measurement **accuracy**

- Let's now briefly review what we have covered in today's seminar.
- Dynamic range is defined as the ratio of the maximum power the network analyzer can measure, set by its RF source or by receiver compression, and the minimum power it can measure, primarily set by the noise floor of the receivers.
- Dynamic range is a very important aspect of vector network analyzer performance, for several reasons. Number one is that the network analyzer must have high dynamic range in order to test high-dynamic-range components like base-station filters. Two, having good instrument dynamic range to begin with allows us to trade off some dynamic range for measurement speed, by utilizing wider IF bandwidths. And finally, the instrument must have significantly more dynamic range than the DUT to ensure good measurement accuracy.



- Dynamic range is generally greater on network analyzers that use mixers for down-conversion instead of samplers.
- We also explained several hardware imperfections that can limit dynamic range, such as receiver compression, signal leakage that can cause crosstalk and mixer or sampler bounce, and receiver residuals.
- We also covered several measurement techniques that can be used to optimize dynamic range. First, you should always use as much test port power as you can, taking into consideration receiver compression. The two easiest ways to lower the noise floor of the receivers and thus increase dynamic range, is to reduce the IF bandwidth and/or use averaging.
- We demonstrated three alternate configurations of the network analyzer to extend dynamic range, namely bypassing or reversing one of the directional couplers, and by using a low-noise amplifier to improve noise figure of the receivers. All of these configurations require segmented or list sweeps, where we can vary the source power segment by segment.
- We talked about two sweep choices to improve dynamic range. Alternate sweeps disable whichever receiver is not needed for a particular S-parameter, reducing or eliminating mixer or sampler bounce. Spur-avoidance sweeps can be manually selected on some network analyzers to reduce receiver residuals.
- And we described how performing an isolation calibration can be useful for lowering crosstalk.

Questions and Answers



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- Now it's time for a question and answer session, where you have an opportunity to ask us questions relating to any aspect of dynamic range or network analysis in general.