Application Note 1397-1

Lowering Cost and Improving Interoperability by Predicting Residual BER: Theory, Measurements, and Applications

Introduction

This application note is intended for those who require a functioning knowledge of residual bit error rate (BER) in digital QAM (quadrature amplitude modulation) radio systems. Residual BER is critical to understanding and determining phase noise and linearity requirements necessary to uphold a given quality of service (QoS) in error rate limited radios. This note assumes a basic knowledge of radio frequency (RF) technology and measurement, but is addressed to a wide audience whose exact skills and experience may vary.

The demand to put more data through the wireless channel has driven the industry to complex modulations that require low-phase-noise oscillators and low-distortion power amplifiers (PAs). While design tools such as link power budgets are excellent for predicting performance at threshold, a different analysis is needed for 'normal' receive power levels. Prediction of residual BER allows a radio designer to determine both phase noise and linearity requirements for a given quality of service (QoS). Bit error rate performance is thus related to key analog metrics that drive an all-important parameter—cost!

The dramatic growth of the broadband market in recent years has created much interest in this little-understood topic—thus this application note.

Section 1 begins with an introduction to residual BER and a review of QAM systems for those unfamiliar with the technology. It presents a basic noise and error probability model that is commonly found in textbooks and will serve as a foundation for the phase noise and linear case used in the residual BER budget process. It offers a similar look at common non-linear distortion models. This is followed by a discussion of the two effects as a whole (composite), including a section on techniques for creating a system budget and optimizing its cost. This material is then tied together in an actual sample budget. A brief discussion of "golden unit" testing is given. Section 1 concludes with a brief concept summary.

Section 2 provides a closer look at phase noise measurement, examining its theoretical principles and practical applications. It covers the techniques for measuring phase noise: direct measurement, heterodyne frequency counter measurements, and carrier removal/demodulation. A selection of equipment for measuring phase noise is reviewed, followed by profiles of its target DUTs: oscillators, amplifiers, frequency multipliers, frequency synthesizers and converters, transmitters and receivers, and modulators and demodulators. Section 2 ends by discussing several unique applications of phase noise measurement with a vector signal analyzer (VSA).

Section 3 covers linearity distortion measurements in more detail, beginning with the use of the CCDF (complementary cumulative distribution function) for choosing proper vector and power test ranges. Linearity distortion is broken down into its constituents: AM to AM and AM to PM distortions. Measurement techniques are given for amplifiers, frequency converter receivers, and frequency converter transmitters. A special troubleshooting section is offered as a guide to making better measurements. High-power measurement setups using a vector network analyzer (VNA) are examined, followed by a profile of VNA use in component measurements. Section 3 ends with a discussion of complex stimulus-response techniques.

Section 4 is a short review of the entire document, briefly recapping the various measurements (and instruments) for phase noise, CCDF, and AM/AM and AM/PM linearity distortions.

The Appendices offer additional information, including Agilent phase noise and linearity measurement equipment (A), references and recommended reading (B), and symbols and acronyms (C).

Frequently, radio engineers find that they have been exposed to much of the material in this document, but have rarely seen it applied to the residual BER budget process and may be unfamiliar with the various measurement setups.

We trust that this application note will help you gain a solid understanding of residual BER and its critical role in helping you to both design and test of modern digital radio systems.

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1.1 Bit Error Rate (BER)

1.1.1 An Example

A large company was in the process of final-testing a high-capacity, broadband microwave point-to-point data link. It encountered a variety of problems in the final acceptance test phase.

The digital radio "dribbled" errors from the moment it began to transmit. In an attempt to find the cause, the project's systems engineer looped the modem back on itself and found it to be error-free. That possibility was eliminated. He reasoned that the RF itself might be bad, so he backed off (turned down) the transmit power. The dribbling errors stopped. The engineer suspected that the radio link's transmitter power amplifier might be the cause of the problem. He switched the amplifier to an error-free radio link and found that that link also began to dribble errors—at about the same rate as the original system. Therefore, he reasoned, the amplifier was probably the cause of the problem. He repeated the process using another error-free link and found that the result was virtually identical. He now concluded positively that the amplifier—a $3,000 component—was the cause of the problem and promptly scrapped it! After all, the problem followed the amplifier from link to link, so it had to be the cause, correct?

Not necessarily—the amplifier could be perfectly useable!

Let's assume now that the engineer had first created a BER budget specification for the radio system—in other words, he determined the "floor" of transmit-receive errors that could be expected when *all* components of the system were working properly, and then allocated a series of worst-case impairments that would be allowed for each key component in the system.

By analyzing this BER budget, the engineer might have seen that if the oscillator in the radio's demodulator had performed properly, the link would have worked perfectly—with the original power amplifier. In other words, a $3,000 PA was scrapped because of a $10 crystal elsewhere in the system! This exact scenario has occurred many times in the radio industry. Seemingly unrelated components interact in unexpected ways, creating vexing and seemingly untraceable problems.

Other, similar examples could be cited. For example, system engineers often destroy a product's economic viability by choosing improper allocations between phase noise and linearity. A noisy frequency conversion or modem oscillator can create the need for a very expensive power amplifier to compensate. Conversely, poor PA performance can necessitate very low phase noise and costly local oscillator (LO) frequency synthesizers. As was the case with our engineer, dribbling errors often misdirect one's attention to the power amplifier (PA), since it is usually the easiest component to change in performance by reducing transmit power. The PA thus *seems* to have the greatest effect on residual BER.

Interoperability is another frequently cited issue in predicting residual BER. With the explosion of high-quality Internet-related digital data links, ensuring that an RF interface will have low error rates, no matter which terminal is receiving a signal, is of paramount concern to many vendors.

With examples like this, you can see the critical importance of understanding bit error rate as a predictor of performance and as a diagnostic tool for precisely tracking down problems. Let us now explore the subject in detail.

1.1.2 What is Bit Error Rate (BER)?

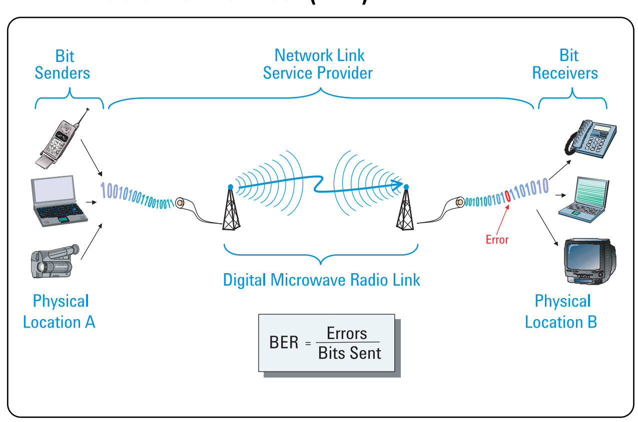


Figure 1: Digital Data Link and Bit Error Rate

Any digital communications link contains bit senders and bit receivers that are physically separated. These devices are linked together through a network link, often provided by a third-party service. The network links can be wire, coaxial cable, fiber optic, microwave, or some combination thereof.

One very important measure of the quality of service (QoS) of the network link provider is the ratio of bits sent correctly to bits received in error. This ratio is called the bit error rate, or BER.

Different levels of QoS are required depending on the type of network data being transported between locations. Voice traffic, for example, will tolerate much higher error rates than data traffic. Digitized voice can tolerate bit errors as high as 1 bit per thousand bits sent, or 10^{-3} BER. Computer data demands bit error rates of 1 per million to 1 per trillion, or BERs of 10^{-6} to 10^{-12}, depending on content. For example, Internet surfing does not demand the same quality of service as bank fund transfers, nuclear power plant control, or early warning systems for nuclear attack! BER is thus a very important part of the network operator's service offering, even more so for critical applications.

1.1.3 What is "Residual" BER?

What is "residual" BER? In a microwave data link, bit error rate is a function of the received signal level (RSL), sometimes called received signal strength (RSS).

Very weak signals create many bit errors. The transition from few errors to many errors at low power is called "threshold." A considerable body of knowledge has been developed on predicting threshold because it

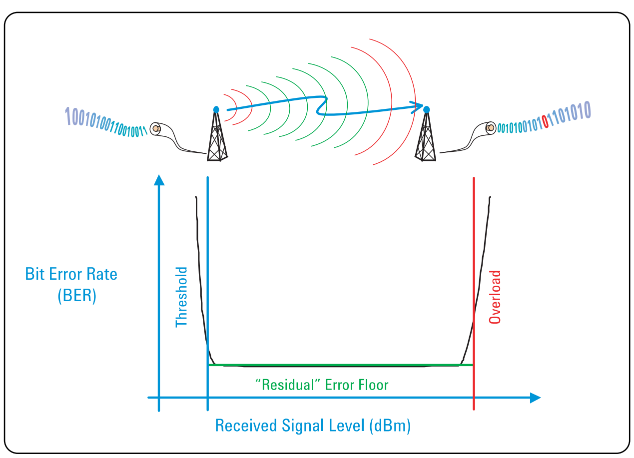


Figure 2: Residual BER & Receive Signal Level

is a key factor in determining the maximum physical distance of the data link and its availability to transport data due to meteorological conditions.

As received signal strength increases, the error rate will fall to a very low level, or "error floor." This error floor is called the "residual" bit error rate or "residual BER." It is the *normal* operating performance of the data link. On the other end of the curve, as received power increases, the receiver will ultimately reach an overload point where the error rate again increases quickly.

This application note examines the somewhat different approach for predicting residual BER than those traditionally used at threshold.

Why is residual BER so important? The network provider's customers demand a certain quality of service based on the type of data payload being carried. The better the quality of service, the bigger the potential market for the product. Thus the network provider uses residual BER as a measure of the quality of the equipment he has purchased.

Unlike threshold and dispersive fade margin—which are key availability metrics—residual BER characterizes the radio in its *normal* operating received signal strength range. This is the performance the network provider will experience most of the time.

Residual BER measures the combined effect of the digital radio's modulator, transmitter, receiver, and demodulator. It is thus a *composite* evaluation of the entire link. In essence, it is a single measure of a broadband radio link's quality of service. Residual BER is a metric defining the service performance, similar to a link budget's fade margin that defines the availability of service (AoS). The vast majority of the time, data links operate in the residual error floor region, with AoS specifications like 99.99% (i.e., signal strength is above threshold 99.99% of the time). AoS is often specified as a percentage of time the link is considered operable (in the residual error floor). A common notation is the number of 9s, 99.9% being three 9s, 99.999 being five 9s—or only 312 seconds of outage per year!

Residual BER is a key performance metric, but why is it so valuable to be able to predict it? *Residual BER prediction guarantees that modem and RF will integrate together to deliver consistent error floor performance.*

The measurements used to confirm residual BER prediction budgets allow the engineer or technician to separate modulator, transmitter, receiver, and demodulator issues, whereas other metrics such as loopback tests and error vector magnitude (EVM) cannot.

Many vendors now provide products with capacity upgrade paths by increasing the complexity of the radio's modulation. Sequential installation systems, such as the one illustrated in Figure 3, make the interoperability of subsequent installations essential for success. Residual BER budgets also ensure interoperability between different receivers or transmitters. Equally important, a residual BER budget is essential for assuring customer premises equipment (CPE) units will not dribble errors when deployed years after the base station is installed.

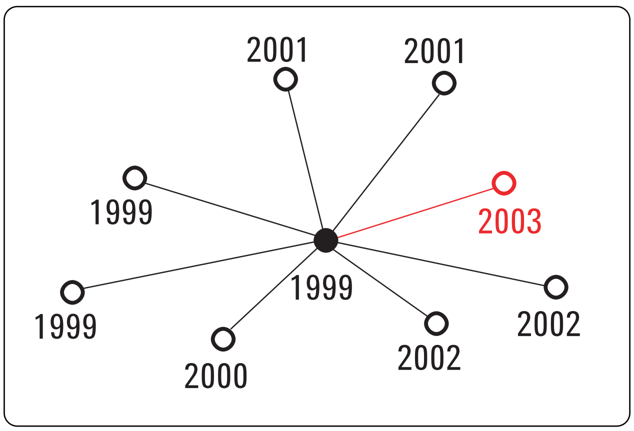


Figure 3: Sequential Installation and Interoperability

Residual BER prediction also gives manufacturers the ability to upgrade a modem with confidence that the currently-installed RF will support it.

Residual BER budgets are also an essential element for controlling cost of the sources and power amplifier —two of the most expensive pieces in any radio link.

The most important technical contribution of residual BER prediction is that it mathematically relates key analog metrics used to specify components to digital bit errors used to evaluate systems. This bridges the gap between the network provider's quality of service metric and the radio engineer's analog component metrics.

1.1.4 Digital Radio QAM Modulation

Most modern digital radios use quadrature amplitude modulation (QAM) or some form of it. QAM is a vector (or *phasor*) modulation. It is created by taking two vectors which are 90 degrees apart and amplitude-modulating them, then summing them together to from a resultant vector. The vector can be modulated in both amplitude and phase.

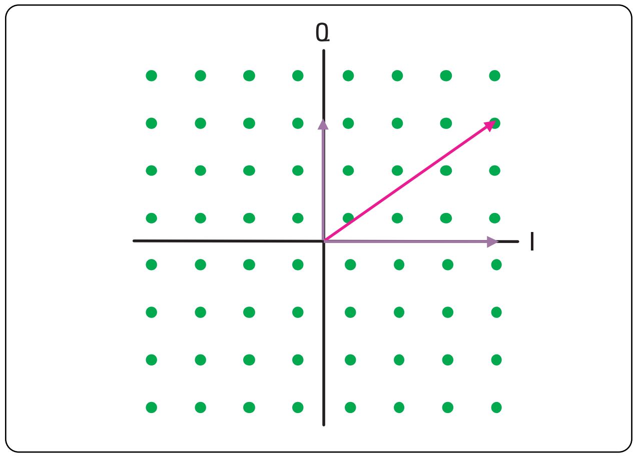


Figure 4: 64 QAM Constellation

This vector, or phasor, can be directed to any number of points, which together represent a symbol constellation. Typically, the number of points in the constellation is related to a power of 2 (2^n) to make digital processing easier.

In the remainder of this publication, we will use the 64 QAM symbol constellation for our examples (although the techniques are equally applicable to other QAM modulations). Often, for simplicity, only a single quadrant of the 64 QAM constellation is shown.

(NOTE: Many other digital modulations, such as orthogonal frequency division multiplexing (OFDM) and spread spectrum, can be analyzed similarly to single-carrier QAM. There are differences, however; for example, carrier recovery techniques and spread bandwidths must be accounted for. These differences require some modification to the basic assumptions of this application note and the reader is advised to account for them.)

1.1.5 QAM Modulation

Each symbol represents several bits, allowing more information to be sent with each sample of the vector's position.

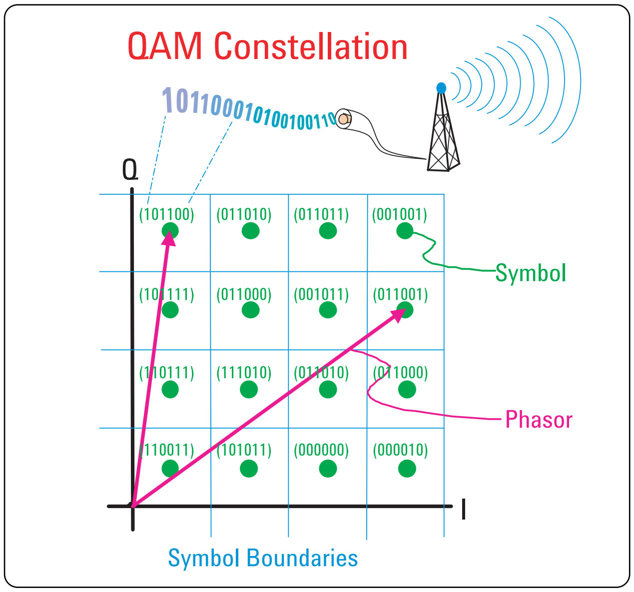


Figure 5: QAM Modulation Bits and Symbols

Sending several bits with each sample has the advantage of decreasing the rate at which the vector is modulated, thus decreasing the RF bandwidth required to transmit a given amount of information. Decreasing bandwidth provides high spectral efficiency, which is often a concern in broadband modulations.

On the receive side, the vector is matched up with the symbol it best fits and bit values are reassigned.

1.1.6 System Diagram

Let's review the process in Figure 6. In the typical digital radio, bits come to the modulator and are mapped to a symbol point. The vector is then driven to that symbol point.

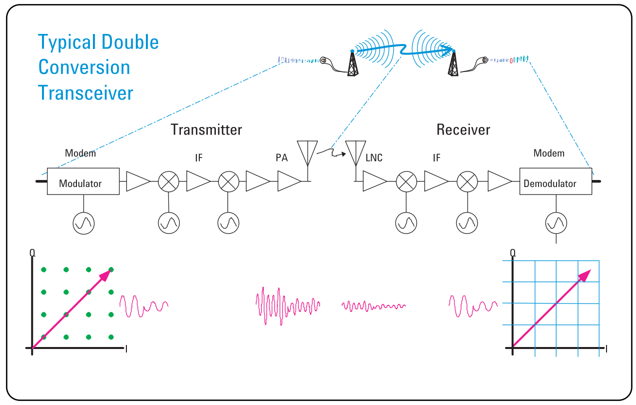


Figure 6: Radio Data Link System Diagram

The signal is then upconverted to a high frequency (filtering has been omitted), which radiates easily and with sufficient bandwidth to carry the required data rate. The upconverted signal's power is boosted by the power amplifier (PA) and directed out through the antenna towards the receiver. The signal travels to the receiver, suffering attenuation and distortion from the path.

Upon receiving the signal, the receiver amplifies it and downconverts it to a frequency at which signal processing is least costly. The demodulator then compares the phase and amplitude of the vector and makes a decision on which symbol best fits, assigning the appropriate representative bits.

Ideally, at the receive demodulator, a single infinitesimally small sample point will be found in the center of the symbol boundary area. Unfortunately, this ideal never occurs, because there is always noise, distortion, and interference components present.

1.1.7 Types of Receive Problems

Random noise has the effect of creating a distribution of sample points. Phase noise is similar to random noise but is only on the angular axis.

AM/AM (amplitude modulation to amplitude modulation) distortion causes the symbol point to fall short of the desired point on the radial axis based on vector length. AM/PM (amplitude modulation to phase modulation) distortion causes the symbol point to take on an angular error based on the vector length. Delay distortion (sometimes called inter symbol interference, or ISI) causes the symbol point to be distorted based on the previous symbol point.

Finally, spurious interference will cause the point to take on a variety of circular shapes.

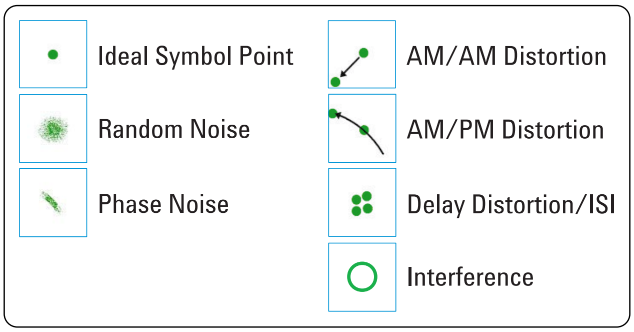


Figure 7: Types of Receive Signal Impairments

These are all *un*wanted signal impairments, which make the symbol decision process imprecise and result in bit errors.

Many of these receiver problems are dominated by key system elements. Generally, they fall into two broad categories: noise sources and distortion sources.

1.1.8 Primary Noise and Distortion Elements

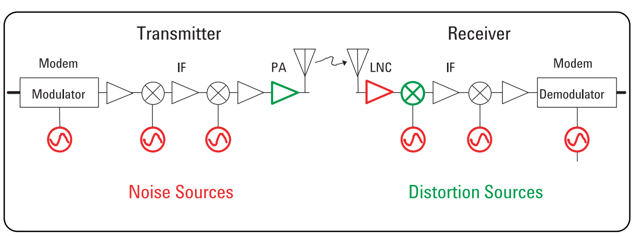


Figure 8: Primary Noise and Distortion Sources

Different components within the radio system contribute to different signal impairments.

Oscillators, which create and demodulate the modulated signal, as well as the local oscillators (LOs), which are responsible for the up- and downconversions, are the primary sources of phase noise.

The receiver noise figure is a primary cause of *random* noise. The transmitter's power amplifier and receiver's first mixer are the most common sources of *distortion,* such as AM/AM and AM/PM. It is interesting to note that these components often represent the majority of the cost of a broadband wireless link. Typically, *60% or more* of a radio's cost is found in the sources and power amplifier! (Thus, much of the interest in residual BER budgeting.)

Each of these sources of constellation problems has a unique impact on BER. At threshold, receiver noise dominates the BER error mechanisms. At overload, receiver distortion, primarily in the first mixer, dominates BER. The residual error floor is dominated by a combination of the phase noise from *all* sources as well as the PA's distortion.

1.1.9 Primary BER Influences

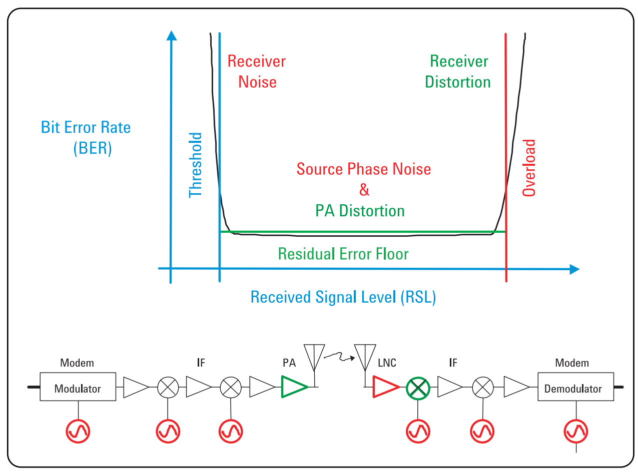


Figure 9: Primary BER Influences & Error Floor

The remainder of this document will focus on the residual BER floor. It is important to note, however, that the techniques that follow are equally applicable to overload and threshold. The dominant component(s) of error differ based only on the received signal level. Thus, the error vs. received signal level curve is really made up of three composite terms—threshold, residual, and overload.

Most systems engineers are familiar with the threshold calculation

$$\textbf{\textit{Threshold}} = C/N+NF+BW+KT.$$

For example, if C/N = 28dB, NF = 5dB, BW = 28MHz, and KT = -174.1dBm @ 290°K, then

Threshold = 28+5.0+74.4-174.1 = -66.7 dBm.

At first glance, this formula may not seem to be related to the techniques described in this publication However, carrier-to-noise ratio (C/N) for a given error rate does in fact embody the analogous statistical relationship presented in this tutorial. C/N is a ratio of carrier power to noise power, as we'll see in the next section. Random noise can add to the symbol vector, causing it to miss the intended symbol "box." Noise figure (NF), bandwidth (BW), Boltzmann's constant (K), and temperature (T) are terms used to scale the noise power level to traceable standards.

1.2. Noise and Error Probability

Now that we have had a brief review of QAM radio, let us examine a common noise and error probability model.

1.2.1 Error Probability

Errors occur when the received phasor sample falls outside the intended symbol boundary. The addition of Gaussian noise creates a distribution of sample points about the mean or "ideal" symbol point. If sliced on a single axis, the probability density function (PDF) is clearly visible. This distribution is similar to the "bell shaped curve" of test scores.

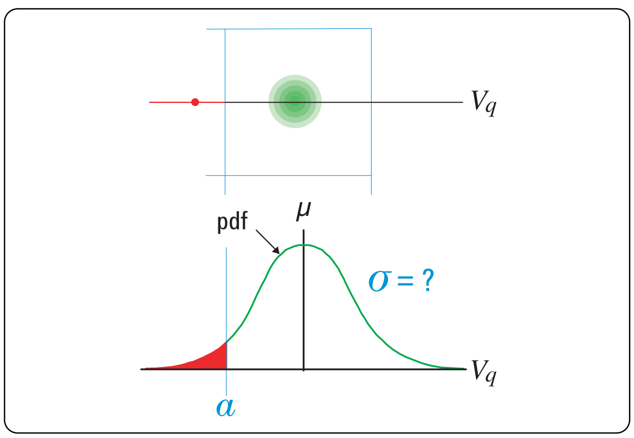


Figure 10: Symbol Errors and Gaussian Noise

The PDF area under the curve beyond the symbol boundary represents the probability of that type of error occurring. The probability can be calculated by integrating the area from the symbol boundary to minus infinity:

$$P\left(V_Q < a\right) = \int_{-\infty}^{a} \frac{1}{\sqrt{2\pi\sigma^2}} \exp\left[-\frac{\left(v_Q - \mu\right)^2}{2\sigma^2}\right] dv_Q$$

Figure 11: Probability Calculation from the Area Under PDF Curve

The central limit theorem can be used to normalize the curve to a Gaussian probability density function where the "standard deviation" (σ) is used to determine the probability of an error occurring. Error probability can then be expressed in terms of the standard deviation of samples. Thus, the primary question becomes; How many sigma (σ) are there to the symbol boundary?

Counting the number of standard deviations to the symbol boundary provides a means to determine the probability of a symbol error.

1.2.2 Symbol Error Rate

If there is only a single standard deviation to the symbol boundary (σ = 1), then the probability of that boundary error is 3.5 x 10^{-1} or 35%. If there are seven sigma to the boundary (σ = 7), then the probability of an error is 2.7 x 10^{-10}, which is quite small.

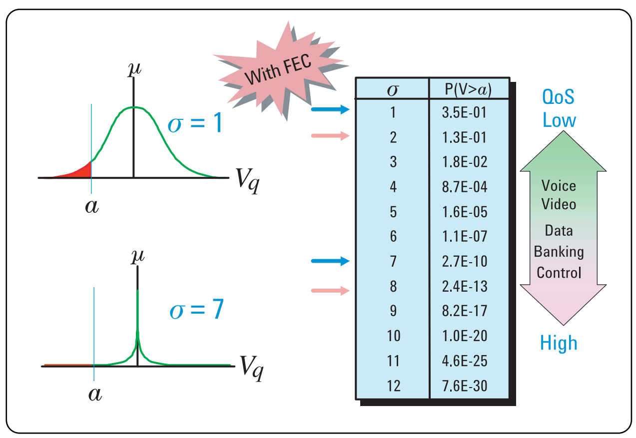


Figure 12: List of Error Probabilities Based on Sigma (σ)

In practice, many higher performance wireless broadband systems compete with the low end of fiber optic communications. Often, these broadband systems are held to the same QoS requirements as fiber, where residual BER rates of 10^{-12} are commonplace. Raw, uncorrected BER rates of 10^{-10} are usually sufficient to achieve 10^{-12} after forward error correction (FEC), thus providing a wireless service of the same quality as fiber. The typical number of sigma to the boundary for a high-quality wireless system is between 7 and 8.

(NOTE: The above paragraph is the only discussion of the effects of forward error correction in this publication. The improvement achieved by turning FEC on is predictable and fixed for low error rates; hence, its effects are easily factored into the raw uncorrected error rate. The remainder of our discussion deals with the raw, *un*corrected error rate. Incidentally, it is usually fastest to test systems with FEC turned off and apply a mathematical correction factor to determine performance with FEC on.)

1.2.3 Phase Noise Effects

We have briefly explored a simplified random noise model often found in textbooks for the analysis of threshold noise. This model is the basis for threshold effects, which are random in all directions and relevant only at low power.

This application note is concerned with "normal" operating power levels of the residual error floor, where oscillator phase noise is the dominant source of noise. Local oscillator phase noise is always present and, unlike threshold noise, is random on the angular axis.

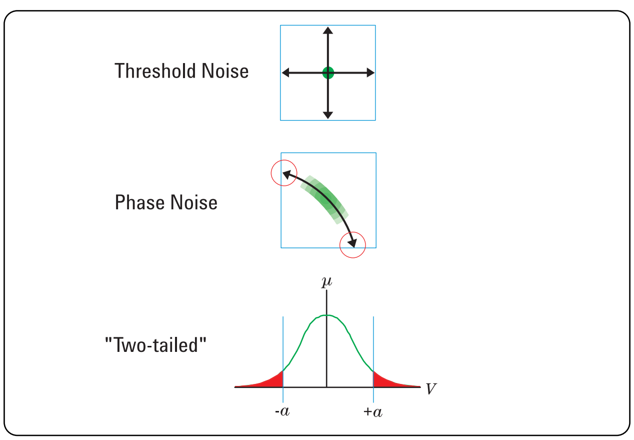


Figure 13: Types of Noise and "Two-tailed" Error Probability Model

The preceding analysis was simplified to examine only a single type of boundary error. In reality, errors can occur on both symbol boundaries, so a "two-tailed" probability model is required. The integration of both "tails" is sometimes seen as a 3 dB factor.

1.2.4 RMS Phase Noise

So how do we characterize and account for phase noise in the probability model? First, we must understand that phase noise is a measure of the source's or local oscillator's spectral purity, or how perfect the sine wave is.

Frequency (f) is the rate of change of phase with time. Phase noise is the deviation in phase from the mean rate of phase change (the center frequency).

Because sideband noise power is rarely seen as a flat Gaussian curve with frequency offset, it must be integrated to obtain phase noise.

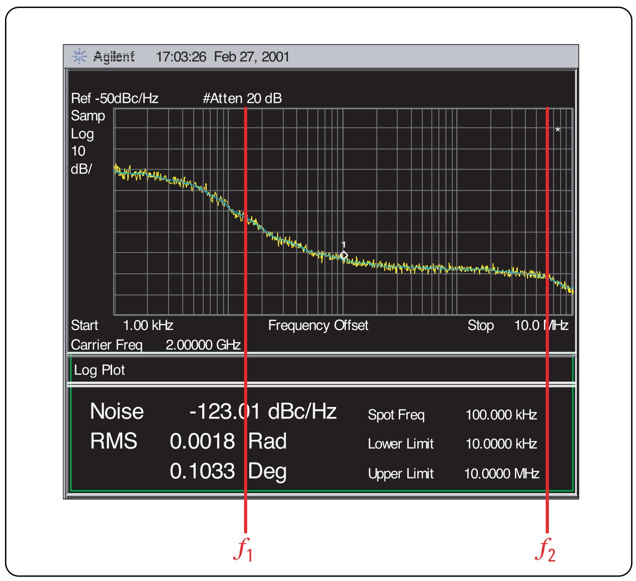


Figure 14: Phase Noise Measurement Showing RMS Calculation

The random nature of sideband noise necessitates a root-mean-squared (RMS) characterization, where 1Hz "buckets" are summed together using the central limit theorem to predict a composite deviation of many samples of varying amplitude. Hence, by integrating sideband noise (the difference between the mean power and the sideband power, dBc) in an RMS fashion, phase noise is expressed as an RMS angular error in degrees or radians.

$$\Delta\phi_{RMS} = \sqrt{\int_{f_1}^{f_2} 2 \cdot \mathscr{L}(f_o)^2 \cdot df_o}$$

Figure 15: Integrated Phase Noise

The limits of the integration should start just outside of the bandwidth of the carrier recovery tracking loop as the lower limit, and stop at the bandwidth of the symbol rate for the upper limit.

1.2.5 Standard Deviation and RMS Noise

A key relationship to understand is that the 1-sigma (σ=1) distance happens to be *identical* to the RMS error (with no dc term)! Often, statistics courses are taught without ever clearly pointing this out to the engineer.

$$\sigma = 1 \Leftrightarrow \mathrm{RMS}$$

Figure 16: Standard Deviation (σ) Relationship with RMS

If there is one point to remember from this publication, it is *this relationship between sigma (σ) and RMS.* As you will see, it makes it possible to calculate the probability of a BER from *analog metrics alone.*

1.2.6 Phase Noise and Error Probability

Integrating sideband noise power over the appropriate limits gives us the RMS angular phase noise in degrees ($\Delta\phi_{RMS}$) that affects the modulation. This RMS error ($\Delta\phi_{RMS}$) represents the angular degrees contained in 1 sigma's worth (σ=1) of standard deviation.

Knowing the angular magnitude of 1 sigma and the constellation geometry makes it possible to calculate the number of sigma to the symbol boundaries.

Given the number of sigma to the boundary, the normalized PDF yields the exact probability of that boundary error! Thus it is possible to calculate the probability of a symbol error for each boundary error type in the constellation. Hence the effect of oscillator phase noise on residual symbol errors can be analytically determined.

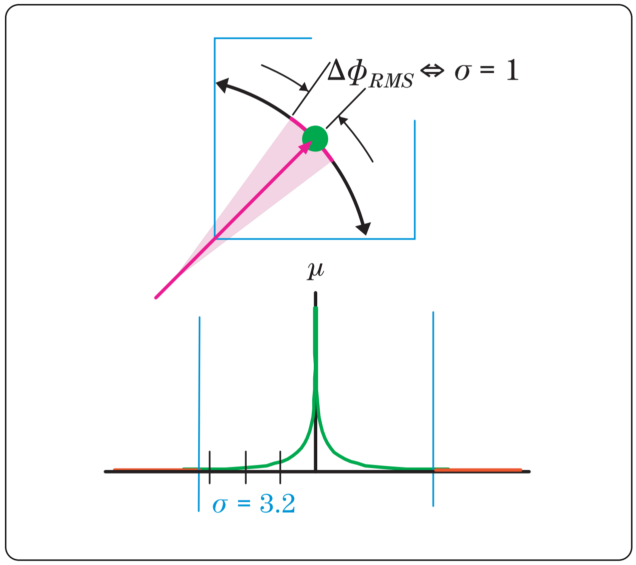


Figure 17: Phase Noise and Error Probability

1.3. Non-linearity Distortion

Having reviewed QAM digital radio concepts, a basic noise model, and the extension of that model to phase noise, let us now examine non-linear distortion as it applies to residual BER. As with the noise model, we will first review common non-linear models and then focus on those which are best for predicting residual BER.

1.3.1 Common Amplifier Linearity Metrics

Commonly used linearity metrics include two-tone intermodulation distortion (IMD) testing and amplitude modulation to phase modulation (AM/PM) conversion testing. Let's review each briefly.

Two-tone IMD testing is a scalar measurement based on the internal mixing of harmonics generated in the device at high power. It is indirectly related to BER and has long been used as a diagnostic tool. Unfortunately, studies of residual BER vs. IMD have shown correlations as low as 85%, making system residual BER performance uncertain. Despite the inaccuracies of IMD, historically it has been the traditional test for linearity characterization (mostly because of the perception of lower equipment costs).

Many vendors use intermodulation distortion (IMD) and phase noise to predict if a QAM radio will dribble errors. This approach is flawed, being purely empirical. IMD is a spot power characterization and does not account for the effects of changing compression characteristics as a function of power over the entire range of vectors that make up the modulation. Worse yet, IMD is often applied just below saturation, where its prediction utility is of little value (clipping begins to dominate over square law curves). While it may seem easy to measure error rates at different power levels and correlate them with IMD measurements, unfortunately the correlation coefficient can be as low as 85% for 128QAM. Said another way, 15 out of every 100 radios would fail their residual error count based on this metric! AM/AM and AM/PM (which is most accurate when clipping begins) and phase noise, which are mathematically related to BER, can calculate the error rate exactly.

AM/PM testing is a vector measurement that typically requires a vector network analyzer (VNA). It is based on measuring a relative microwave phase shift as power is increased. AM/PM is directly related to the BER mechanism and is an analytical part of the residual BER system budget. Recent decreases in the cost of vector network analyzers make AM/PM measurement less expensive than IMD measurement, so it is gaining popularity.

So which linearity test is best for QAM? AM/PM is clearly the best choice for high-QoS systems where residual BER is important. Let's examine why AM/PM is the measurement of choice.

At low power levels, amplifiers exhibit linear behavior, such that small signals are amplified by a fixed amount of gain. Given a specific power input, the amplifier is said to behave linearly if the power output is a constant ratio larger. As input signal power grows, it reaches a point at which the output signal will stop getting larger. The amplifier is then said to be "saturated" and the linear relationship between input and output no longer exists. Many measurements have been devised to characterize this phenomenon.

"Gain compression" is a term used to describe the difference between the saturating amplifier's performance and the theoretically ideal performance, or the difference between the ideal linear (constant) gain and the actual gain The so-called P_{1dB} is a measure of the output power at 1 dB of gain compression.

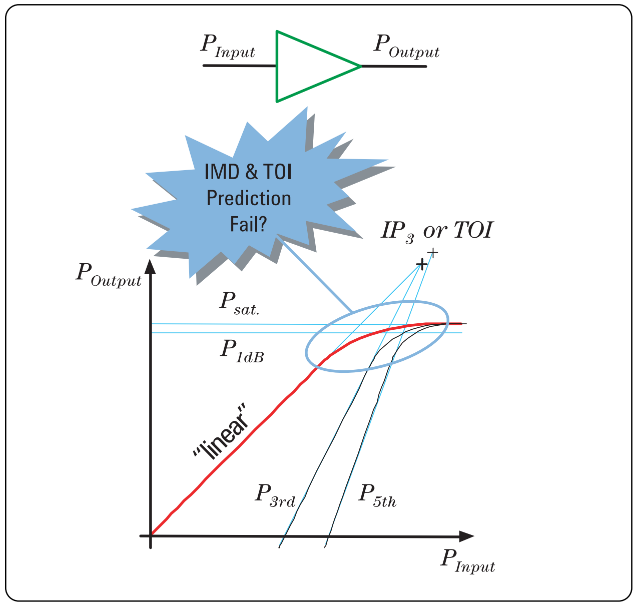


Figure 18: IMD and TOI Fail to Predict Residual BER Due to Non-linearity Effects

Two-tone intermodulation (IMD) is a measurement designed to predict the amount of unwanted modulation energy created by nonlinear saturation. It predicts 3^{rd}- or 5^{th}-order intermodulation products. A 3^{rd}-order intercept point can be calculated from the 3^{rd}-order products, for predicting the level of intermodulation distortion.

Though commonly used, none of these linearity metrics is ideally suited for predicting residual BER since PAs are operated in a region where an abrupt change in device linearity occurs and intermodulation distortion (IMD) is no longer predicted by third-order intercept point (TOI).

To predict residual BER, linearity metrics are needed which directly relate to the QAM vector in amplitude and phase. Gain compression—the difference between the ideal linear gain and the actual gain—is amplitude modulation due to amplitude modulation (AM/AM) conversion.

As power is increased, the phase delay through an amplifier begins to change as it nears saturation. This change in phase shift as the power is increased, or mod-

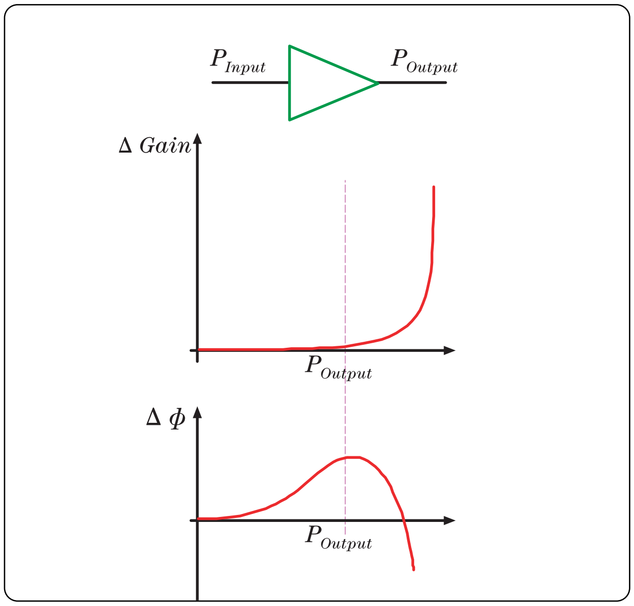


Figure 19: Gain Compression, AM/AM & AM/PM

ulated, is AM/PM modulation. As we will see later, this modulation is additive to the QAM vector. AM/AM and AM/PM are unwanted modulations that affect the accuracy of the symbol point position on the constellation.

Figure 19 shows the distortions in relation to each other (i.e., same power scale). Note that significant phase shifts occur before significant amplitude shifts occur. Typically, only a few tenths of a dB of AM/AM occur when several degrees of AM/PM have built up. The QAM modulation is much more "sensitive" to AM/PM distortion–a few degrees of distortion are quite significant, whereas 0.1 to 0.3 dB of AM/AM has little effect. This phenomenon yields a key simplifying assumption: AM/AM is a secondary effect and will be ignored in our worst-case analysis. (NOTE: If highly accurate BER prediction is sought, it may also be necessary to geometrically account for AM/AM effects on the symbol vectors.)

1.3.2 Amplifier Model

Sometimes it is helpful to review the mechanisms that generate these distortions in the power amplifier in order to provide insight into controlling factors and to understand how our designs affect system-level performance. Let's use a simple GaAs MESFET (gallium arsenide metal semiconductor field effect transistor) model to illustrate the principle. To keep it simple, we have omitted the effects of microwave matching circuits, which we will assume are linear in nature and have infinite bandwidth.

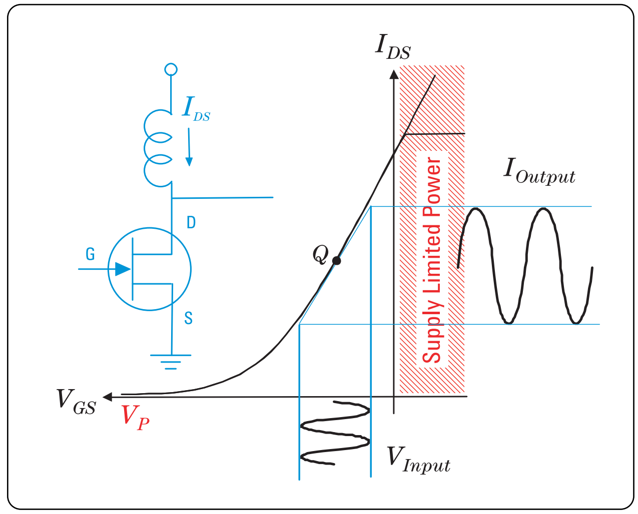


Figure 20: GaAs Voltage-to-Current Curve Showing Square Law Behavior

The GaAs MESFET voltage-to-current characteristic behaves as a square law device if driven with a small signal input voltage with current modulated about the Q point on the output. Typically, most QAM microwave power amplifiers are built as "Class A" designs for maximum linearity.

The raw dc power supplied to the FET to establish the Q point must be limited to constrain the device's operating temperature. Constraining the temperature slows the semiconductor's defect migration to assure a long operating life.

Next, let's see how IMD products, AM/AM and AM/PM, are created in the device.

1.3.3 Two-tone IMD

In a two-tone IMD test, identical amplitude tones slightly separated in frequency are injected into the input of the amplifier. The power supply cannot deliver additional current, so the superimposed sine waves are periodically "clipped." The Fourier series of the clipped signal gives rise to 2^{nd}, 3^{rd}, etc. harmonics, which mix together internal to the device to form the IMD products.

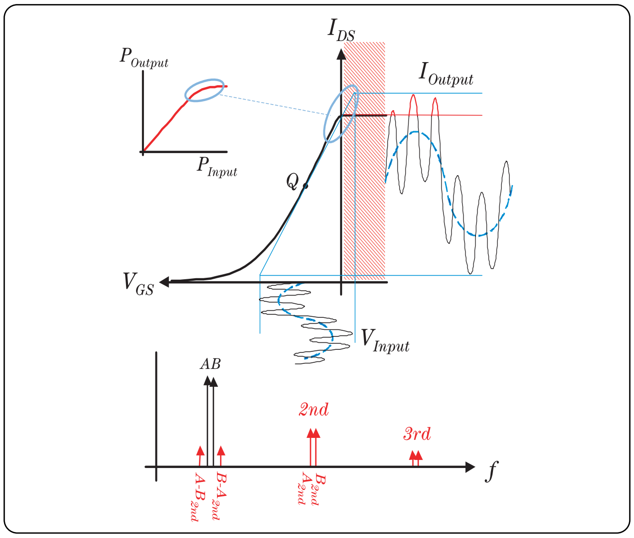


Figure 21: Two-tone IMD Mechanisms

This mixing action actually takes place in the junction of the device and the harmonics are often substantially attenuated by the finite band-limited output matching networks. Thus, it is often not possible to directly compute the intermodulation products using the measured harmonic power for microwave amplifiers. It is, however, a simple process to measure the relative attenuation of the IMD products with a spectrum analyzer.

Does saturation really produce such an abrupt clipping of the two sine waves? That depends very much on the particular FET and how it is biased. Forward gate rectification can be very abrupt, whereas pinch-off or the square law curve are much more subtle saturation characteristics. Hence the amount of harmonic energy available to be converted into intermodulation products is very dependent on the abruptness of the non-linearity. Here lies the major problem in using IMD to predict BER from correlation studies. The IMD measurement is only a single point on the P_{Output}/P_{Input} curve and predicts linearity only for small signal characteristics, while QAM modulation operates over a range of vector amplitudes near saturation that include abrupt changes in linearity.

(NOTE: Notice that at low power levels TOI accurately predicts IMD, but as the device approaches saturation, TOI no longer works. This is believed to be due to transition from the square law non-linearity of the FET to an abrupt power supply limitation.)

1.3.4 AM/AM and AM/PM Mechanisms

The AM/AM mechanism is very simple to examine via our amplifier model. If we put a single sine wave into the amplifier and the power source cannot supply the necessary current, the resulting clipping in the amplitude of the output is reduced. This reduction is gain compression or AM/AM.

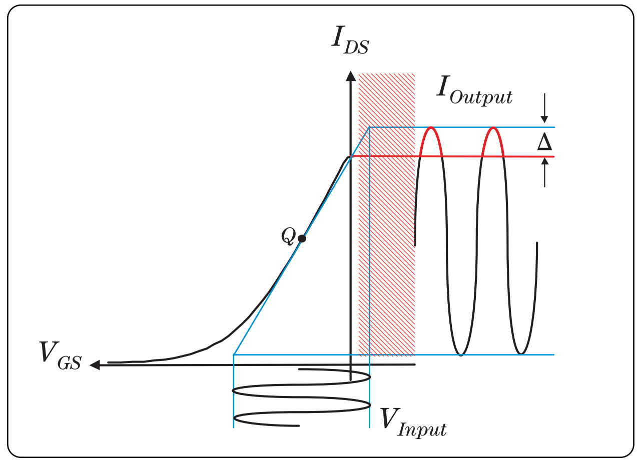


Figure 22: AM/AM Mechanism

AM/AM is easily measured with either a source and spectrum analyzer or a network analyzer.

The phenomenon that gives rise to AM/PM also begins when the power source limitation creates a clipping or "mushing" of the waveform. Since the top of the waveform is not correctly amplified, the average value or "zero crossing" is offset from its original position. This offset in "zero crossing" occurs where the sine wave has finite slope, creating a phase shift in the output signal.

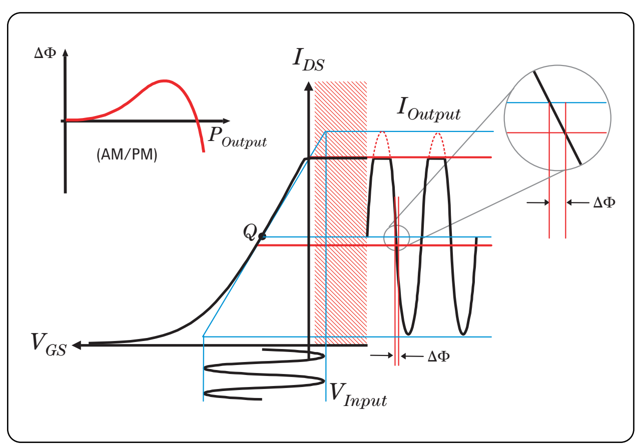


Figure 23: AM/PM Mechanism

It is interesting to note that if the input signal power is increased still further, clipping will ultimately occur on the bottom of the sine wave as the output begins to resemble a square wave. Clipping on both ends causes the average offset to migrate back toward that of the original sine wave—thus the characteristic rise of a few degrees of phase shift, then fall of the AM/PM curve.

It should also be noted that the microwave matching circuits strongly affect the impedance, changing the voltage-current relationship in addition to clipping the output.

1.3.5 AM/PM and QAM Power Levels

As a function of signal amplitude or vector length, the AM/PM phase shift distorts the ideal symbol location of the QAM symbol constellation.

Outer symbols have the largest vector length and therefore suffer the most AM/PM distortion. AM/PM testing is typically done with a carrier wave (CW) sine wave (though it is possible and occasionally necessary to measure it with a modulated signal and vector signal analyzer, or VSA). This is important because it relates an analog CW parametric test to the actual distortion impairment of the QAM symbol constellation.

The ability to relate a parametric analog test to the actual error mechanism provides the means to predict digital error rates from traceable standards.

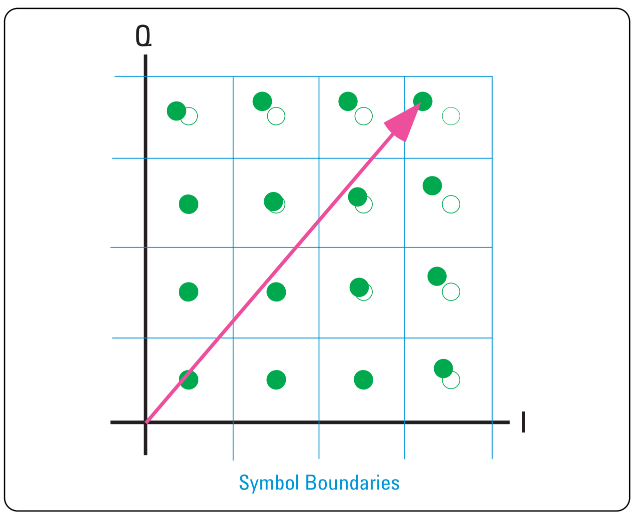


Figure 24: AM/PM Effect on QAM

So how do we relate the vector length to known traceable standards and over what range of vector lengths do we need to measure the distortion?

The power meters we used to set up our transmit power actually measure RMS CW or modulated power. Hence testing AM/PM with a CW source must relate the average modulated power to the average CW power.

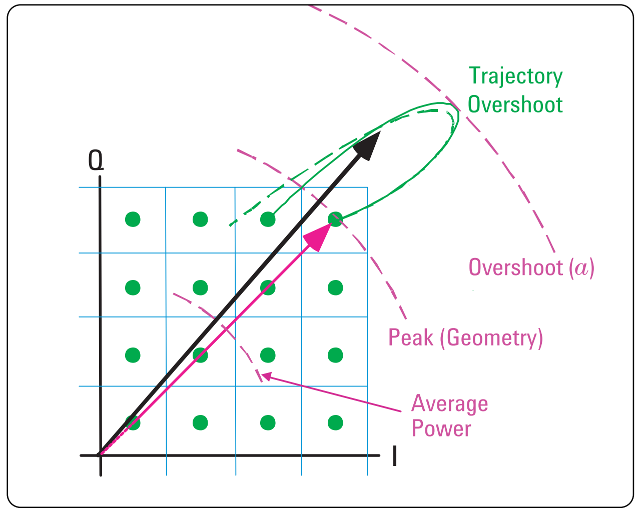


Figure 25: Modulated Power Levels

At first glance, we might think that using the constellation geometry to calculate the RMS vector length would do this. However, the worst-case AM/PM occurs at the longest vector length or peak power. Understanding the peak-to-average ratio of the constellation is an important step in determining what CW test signal power is needed to do AM/PM testing.

A geometric peak-to-average power correction is only one part of determining the highest power at which to test AM/PM. In between symbol states, the vector overshoots the boundaries of the constellation. This so-called 'overshoot power' or 'trajectory power' represents the longest vector length. Though well understood, the overshoot phenomenon is beyond the scope of this publication, so we'll simply say that it is a function of the baseband filtering α. (NOTE: The reader is cautioned that the term "peak to average" is often applied to both the geometric peak to average and the overshoot power to average.)

What test conditions are necessary to properly characterize the distortion of the power amplifier? The gain and phase distortions should be characterized over the range of modulation powers. This means that AM/PM should be tested from the smallest vector needed to produce the modulation to the largest vector (including overshoot)—this is the "vector range."

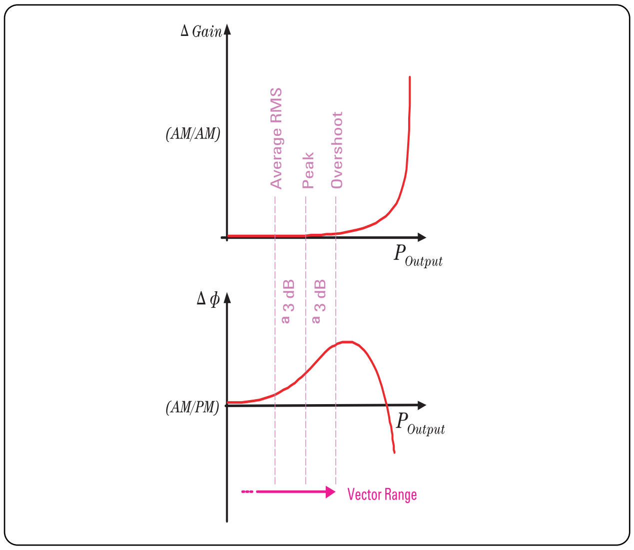


Figure 26: Vector Range and AM/PM

The important thing is to actually measure the amplifier's linearity at the overshoot power, where distortion will be the most significant. Testing at average power (what a power meter measures) would be very misleading.

1.3.6 Test Conditions Using CCDF

One approach to determining the appropriate vector range is to measure the complementary cumulative distribution function (CCDF) of the modulated signal. The vector signal analyzer (VSA) test in Figure 27 shows the overshoot or trajectory vector length vs. statistical frequency for a QAM signal (green = QAM, gray = random noise).

We will cover this measurement technique for determining the setup of the AM/PM test in depth in Section 3.

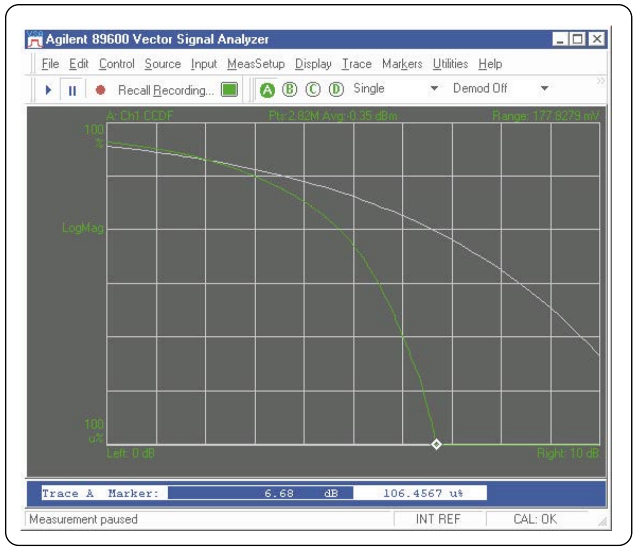


Figure 27: The CCDF Measurement

1.3.7 Spot vs. Swept Measurements

There are two types of AM/PM measurements: spot and swept. The classical diagram of AM/PM or $\Delta\phi$ vs. power level, often seen in linearizer work, is a spot frequency vs. swept power measurement. This measurement is best suited for fixed frequency operation since the matching, hence the AM/PM, is usually a function of frequency.

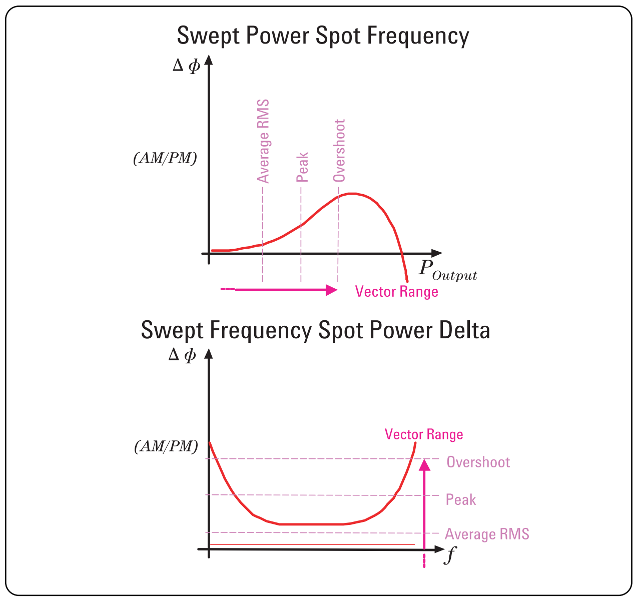


Figure 28: Swept Power vs. Frequency Measurements

Another approach is to use a swept frequency measurement at a spot power delta. This measurement is best suited for broadband devices that operate over a range of frequencies. It does assume that the AM/PM increases monotonically over the power range of interest, which is nearly always the case with QAM signal amplifiers.

Swept frequency AM/PM measurement requires PA output power to be calibrated across the band of interest at a low power (power cal). The phase is then calibrated to zero with the PA in place (since this is a relative measurement). Finally, the power is increased to the overshoot point and the AM/PM across the band is observed (this is explained in detail in Section 3).

Practically, AM/PM measurement must be done quickly to avoid junction-cooling effects that influence the accuracy of the measurement. If it is not possible to make the measurement quickly, the complex stimulus-response method is required.

The author has found that QAM amplifier designers often optimize their designs for maximum saturated power output or lowest IMD at high power. Power amplifiers used in high-QoS links should instead be optimized for minimum AM/PM over the vector range within the band of interest (swept frequency, spot delta power).

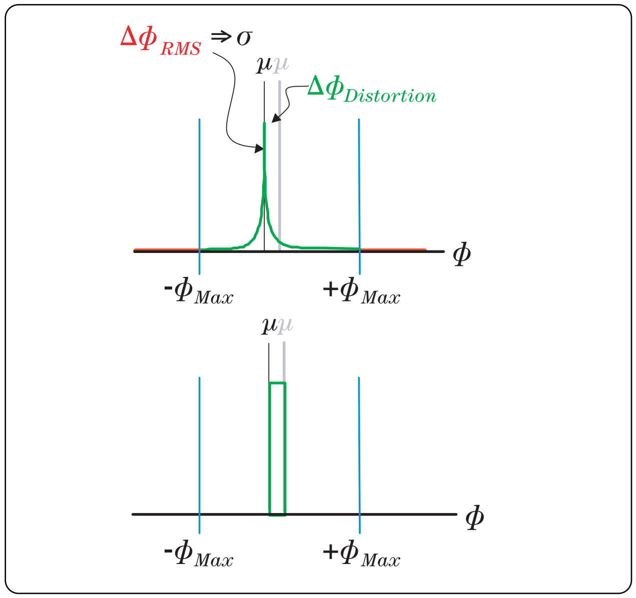


Figure 29: Noise and Distortion Effects

1.4 Composite System

Now that we have reviewed some of the basic principles of QAM digital radio, a phase noise probability model, and non-linear elements, we will review some of the mathematics necessary to combine them, as well as the assumptions and approximations involved.

1.4.1 Noise and Distortion

Distortion is directly additive to noise because it operates on the vector itself. This has the effect of offsetting the mean value of the probability density function, PDF.

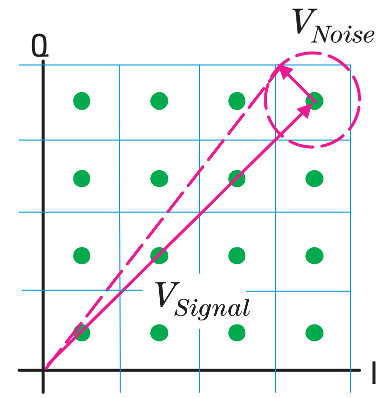
Distortion by itself has no error probability density function—it is purely deterministic! At first, this might seem counter-intuitive, but it requires the randomness of phase noise to create random dribbling errors. What if the distortion was so large that the sample point fell beyond the symbol boundary? It would make an error, but without noise it would always make the same error in a deterministic, periodic way. This doesn't happen in practice because there is always some randomness present.

Dribbling errors often misdirect the engineer's attention to the power amplifier because it is usually the easiest component to change in performance by reducing transmit power. The PA then "seems" to have the greatest effect on residual BER. However, the integrated phase noise contribution is multiplied by the number of sigma (σ) required for an acceptable error rate (usually between 7-8 for high-quality systems). Thus, changes in phase noise have a much greater impact than changes in distortion. Unfortunately, phase noise usually cannot be adjusted like PA output power.

Residual BER is thus a function of *both* the power amplifier linearity and the phase noise of *all* sources in the system. This is a very important point, for disagreement often occurs over whether the PA or one of the sources is the cause of dribbling errors. The reality is that they *both* influence the error floor and only a judicious allocation budget (usually based on implementation cost) can sort out which element is bringing the system down.

1.4.2 System Phase Noise and Geometric Effects

System phase noise can be represented by a noise vector, which adds geometrically to the desired modulation vector. Noise from more than one source can be added geometrically to obtain the total noise for the system.



$$\phi = \sqrt{(\phi_{Signal})^2 + (\Delta\phi_{Noise})^2}$$

$$\Delta\phi_{RMS_{Total}} = \sqrt{(\Delta\phi_{RMS_1})^2 + (\Delta\phi_{RMS_2})^2 + \ldots}$$

Figure 30: System Phase Noise Addition

The sources in the modem and the conversion process all contribute to overall system phase noise. Thus residual BER is a function of both the modem *and* RF sources. The residual BER is also affected by both transmitter *and* receiver sources. These are very important points. It is impossible to evaluate the system's residual BER without taking into account both the modem *and* the RF. Likewise, it is impossible to evaluate residual BER without taking into account both the modulator/transmitter *and* receiver/demodulator.

This means that loopback testing to exonerate the modem of dribbling errors is not a valid approach. Likewise, most golden transmitters and golden receivers are not capable of adequately testing for residual BER. Only golden units with sources and amplifiers having worst-case phase noise and distortion can be used for testing residual BER (a difficult proposition to arrange).

The QAM symbol constellation has some important geometric effects to consider. The outer symbol points can tolerate the least angular error. In the 64QAM constellation, the outermost point can tolerate a maximum of 7.7° of error before making a symbol decision error, whereas the innermost point can tolerate 45° of error. The outermost point also has the longest vector length and will thus suffer the highest corresponding angular AM/PM distortion.

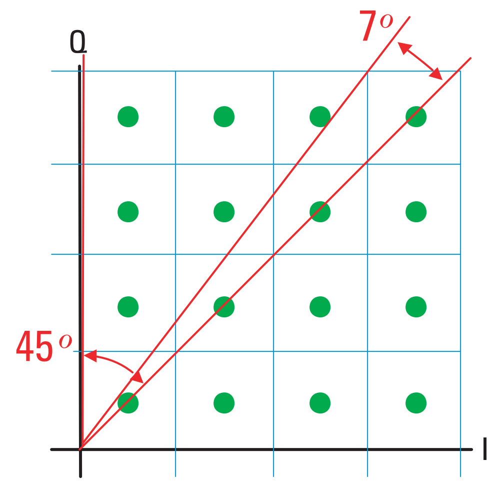


Figure 31: QAM Geometric Effects

The maximum angular error is symbol location dependent, but phase noise is a constant for all symbol points. Thus the number of sigma to the boundary is dependent on symbol location, with the outer symbols having the fewest sigma (σ) to the boundary.

Error probability of each symbol must be weighted based on the probability of symbol occurrence. Usually, symbols are equi-probable and symbol boundaries are set up on a simple grid pattern. Some QAM modulations have probabilities of occurrence and grid patterns that are not so simple (usually to take advantage of the fact most errors are made on the edge of the constellation). This must be taken into account in the system model.

In developing our model of residual BER, we have focused primarily on sources' phase noise and the power amplifier's AM/PM distortion. These key components represent the vast majority of the residual error budget and are often responsible for the largest cost of the radio. Other, secondary contributors can affect the error floor, such as group delay distortion or intersymbol interference (ISI).

1.4.3 Group Delay Distortion

Before we cover group delay distortion, you should be aware there are many other secondary contributors to the symbol error mechanisms. In our simple model, however, we will assume that they are insignificant, which in practice is usually the case, thanks to the excellent work of the modem designers.

In most cases, group delay distortion is not a significant factor because of the tremendous power of modern digital equalizers used in today's radios. However, in burst systems, where digital equalizer performance may be limited, you should take group delay distortion into account.

Filters introduce group delay into the modulated channel. The spectral energy associated with a change in phase is dependent on the magnitude of the phase modulation step. Small changes in phase occupy small bandwidths, and large changes occupy larger bandwidths.

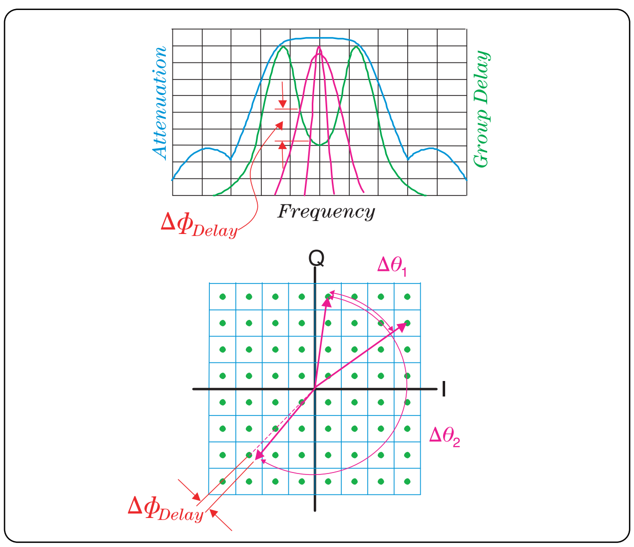


Figure 32: Group Delay Distortion

The delay difference between the two spectra produces a phase error in the modulation vector. This error is based on what the previous symbol sequence was and is another way of looking at intersymbol interference (ISI). Fortunately, for most modern digitally equalized radios, the equalizer reduces this error to an insignificant portion of the residual BER budget and it can be ignored.

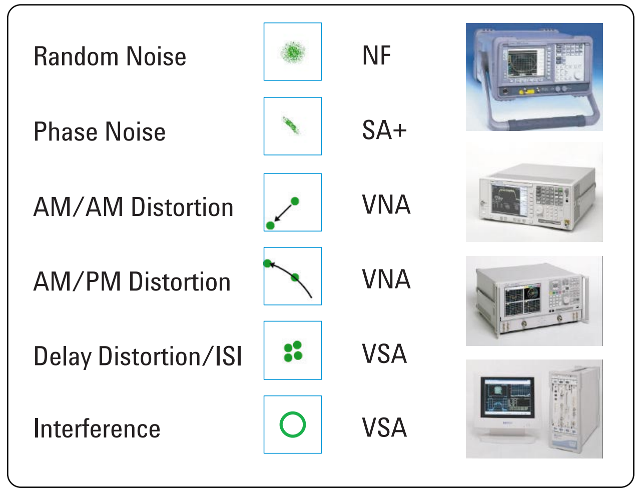


Figure 33: Measurement Equipment

There are many approaches to measuring the different types of error mechanisms. Among the most popular are noise figure analyzers (NFAs) for random noise characterization, spectrum analyzers (SAs) for phase noise integration (high-performance applications require phase noise test sets), and vector network analyzers (VNAs) for AM/AM and AM/PM measurement.

Delay distortion can be measured with either a vector network analyzer (VNA) or a vector signal analyzer (VSA). Vector signal analyzers also excel at identifying spurious interference with tools such as error vector spectrum.

1.4.4 Why Not Use EVM?

At this point you might ask: Why not use error vector magnitude (EVM) measurements to estimate BER vs. separate AM/PM and phase noise measurements? To answer that, let's examine two scenarios, one with a lot of phase noise and little distortion, the other with little phase noise and a lot of distortion. Which has the lower BER? Which has the lower EVM?

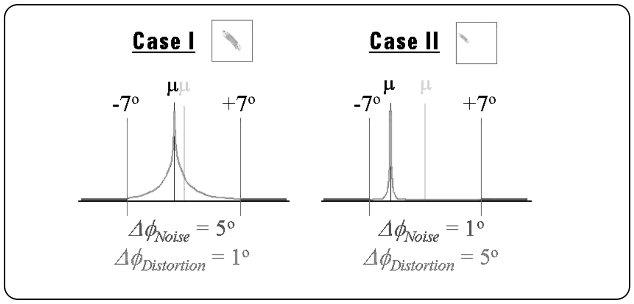


Figure 34: Two Phase Noise and Linearity Allocation Cases

The one with the least number of sigma to the boundary will have the highest BER; hence, Case 1 will have high BER (Case 1, $\sigma = 6/5 = 1.2$, vs. Case 2, $\sigma = 2/1 = 2.0$). If we add up the RMS phase noise and distortion components, both examples have 6° (5+1 = 6 and 1+5 = 6)! They both have the same EVM!

Error vector magnitude (EVM) is a summation of effects. It does not differentiate between random effects, which add geometrically and possess probability density functions (PDFs), and distortion effects, which add directly and are deterministic. EVM is useful because its characteristics can tell us something about the nature of the problem or where the signal was degraded, but it does not always directly relate to BER—the acid test for the network operator. EVM must be broken into different types of error mechanisms, so the appropriate mathematics (statistical or deterministic) can be applied to relate those components in order to predict residual BER. However, the temporal characteristics of the error vectors themselves can provide tremendous qualitative insight to the trained eye in diagnosing problems.

In the case study, we saw that EVM did not necessarily predict differing BER. What exactly happened? EVM is the summation of both deterministic effects like AM/PM and probabilistic effects like phase noise. EVM does not differentiate between the two types of impairments.

The mathematics of deterministic and probabilistic effects are, however, different. Deterministic effects such as AM/PM add to our angular distortion budget in arithmetic (1+1 = 2) fashion. AM/PM always rotates the vector in the same direction proportional to the vector magnitude. Probabilistic effects such as phase noise add in geometric ($1+1 = \sqrt{2} = 1.41$) fashion and rotate the vector randomly in direction. Thus for the residual error floor, where both deterministic and probabilistic effects play a role, it is necessary to break down EVM in order to make a BER prediction.

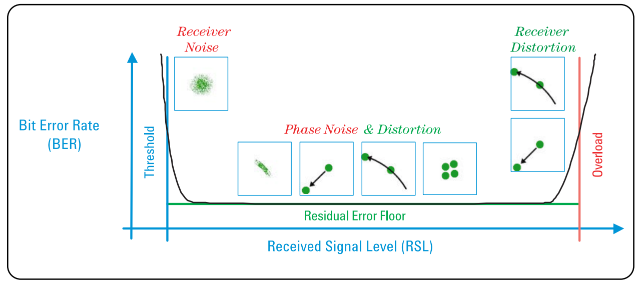


Figure 35: Dominant Symbol Impairments vs. RSL

EVM can predict BER only when a single type of impairment strongly dominates QoS performance. This is usually the case around threshold, where errors are caused primarily by Gaussian (probabilistic) noise. Likewise, at overload, where receiver distortion (deterministic) dominates, EVM will parallel BER. EVM, however, is not useful for predicting errors in the residual error floor.

1.5 Example System Budget

To pull all of these concepts together, let's review a sample residual BER system budget. Our goal is to predict the worst case residual BER of a 64QAM radio design by assigning phase noise and AM/PM performance.

The first step is to calculate symbol vector lengths for every point in the constellation. Using symmetry simplifies the work by requiring only the computations of a single quadrant. Second, the maximum possible phase error for each symbol is calculated. Third, the phase noise and distortion components are allocated on a trial basis. Fourth, the symbol error probability and BER is calculated from the normalized probability density function.

If the results are unacceptable, reallocation of the phase noise and distortion components must be repeated until the correct results are achieved. Finally, once the desired residual BER has been established, the phase noise and distortion allocations must be subdivided across the system.

With this procedure, let's see how it works using an example constellation and numbers.

First, using the Pythagorean theorem, we calculate the magnitude of each symbol vector in the constellation, using a single quadrant for simplicity:

$$A_{iq} = \sqrt{I_i^2 + Q_q^2}$$

We then determine the peak symbol magnitude, or in some cases, magnitudes (32 or 128 QAM). Summing up each vector and dividing by the total number of vectors gives us the average vector length.

$$\left|A_{iq}\right| = \sqrt{I^2 + Q^2}$$

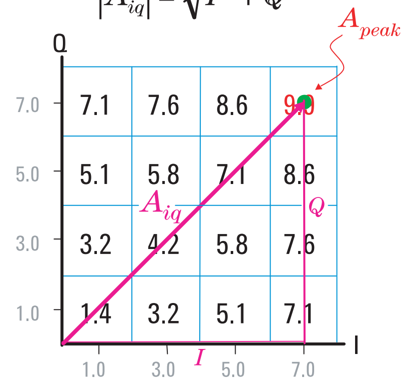
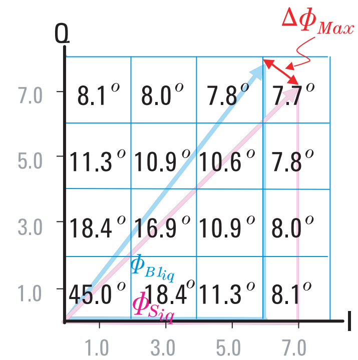


Figure 36: Symbol Vector Lengths

1.5.1 Maximum Phase Error

Using the constellation geometry, we then calculate the maximum permissible phase error to the symbol boundary. We begin by calculating the angle to the symbol point (ϕ_{iq}) using an arccosine relationship, the previously calculated symbol vector length being the hypotenuse of the triangle.

Similarly, we can calculate the angle to the symbol boundaries (ϕ_{B1iq} and ϕ_{B2iq}) by adding or subtracting the distance from the symbol point to the boundary and use the arccosine relationship again. Two equations are necessary for this, depending on where the symbol point is located in the constellation. Symbol boundaries are intersected either on the horizontal axis, vertical axis, or both (diagonal) by angular rotation of the vector.



$$\Delta\phi_{Max_{iq}} = \mathrm{Cos}^{-1}\left(\frac{Q \pm \Delta Q_B}{A_{iq}}\right) - \mathrm{Cos}^{-1}\left(\frac{Q}{A_{iq}}\right)\Bigg|_{i \geq q}$$

$$\Delta\phi_{Max_{iq}} = \mathrm{Cos}^{-1}\left(\frac{I \pm \Delta I_B}{A_{iq}}\right) - \mathrm{Cos}^{-1}\left(\frac{I}{A_{iq}}\right)\Bigg|_{i \leq q}$$

Figure 36: Symbol Vector Lengths

Subtracting the symbol point angle from the boundary angle gives us the maximum angular error permissible before an incorrect symbol is detected.

We can now make an important simplification by assuming that the clockwise and counterclockwise maximum angular errors ($\Delta\phi_{Max}$) are the same. Though they differ slightly, in practice the difference is small enough to be neglected for most worst-case models.

The geometry of the constellation locks down the maximum angular error ($\Delta\phi_{Max}$).

1.5.2 Phase and Distortion Allocation

The angular error is comprised of a distortion component and some number of sigma (σ) times the phase noise component (neglecting other possible error sources, which are usually negligible). Next, we have to allocate the distortion component and phase noise components of the error.

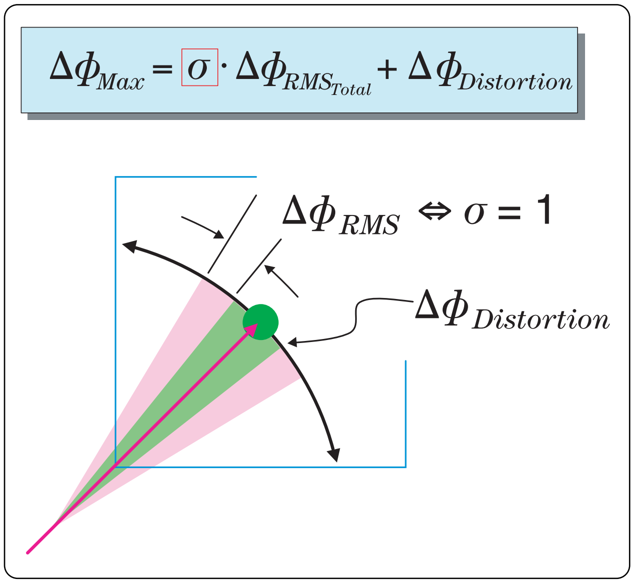


Figure 38: Phase and Distortion Allocation

Properly allocating between distortion and phase noise can have a major impact on the overall cost of the system. This cannot be overstated, because the allocation affects the performance requirements of the sources and power amplifier, the most costly parts of the radio (typically 60% or more of the entire radio cost)! This is where systems expertise and a sound budget pay off.

It is not within the scope of this publication to discuss the economics of allocation in detail—this decision is usually assigned to the experienced system engineer. However, a couple of guidelines may be helpful. High-QoS systems that require very low "fiber-like" error rates usually call for 7 to 8 sigma to the symbol boundary, favoring a total phase noise of around 10% of the maximum angular error. The distortion allocation favoring the lowest cost is usually just below 50% of the total maximum angular error. As transmit frequency increases, 20LogN multiplication of the phase noise allows more of the budget to be allocated to phase noise.

Let's assume a typical allocation in our 64 QAM example of 3.000° of AM/PM distortion and 0.600° of total RMS phase noise and from this calculate the probability of symbol errors.

1.5.3 Standard Deviation

The previous equation can be solved for sigma for each of the constellation points. We see that the number of sigma to the boundary (σ_{iq}) is much lower for the outer states than for the ones close to the origin. This means that virtually all of the residual errors occur in the outermost symbol points.

$$\sigma_{iq} = \frac{\left(\Delta\phi_{Max_{iq}} - \Delta\phi_{Distortion}\right)}{\Delta\phi_{RMS_{Total}}}$$

(σ_{iq})

Q \ I	1.0	3.0	5.0	7.0
7.0	8.6	8.3	8.0	7.8
5.0	13.8	13.2	12.6	8.0
3.0	25.7	23.1	13.2	8.3
1.0	70.0	25.7	13.8	8.6

Figure 39: Number of Sigma to Boundary

In Figure 40, we have made an important simplifying assumption —distortion is small, leaving phase noise essentially symmetric about center of the symbol point. In doing this, we have neglected to account for the differing number of sigma to the boundaries due to the fact that AM/PM has the effect of offsetting the mean in a single direction. At first glance, this may seem like a gross approximation, but in practice it results in reasonable errors, for several reasons. In the case of virtually no distortion, true symmetry does exist. In the case of large distortion, a single tail of the PDF dominates, and the model is off by a factor of two. A factor of two would seem to be large, but when evaluating BER, our concern is usually the *exponent,* not the number preceding it (e.g., rarely do we care if the BER is 1.0 x 10^{-12} or 2.0 x 10^{-12}, but we do care if it is 10^{-12} vs. 10^{-8} !). If your application requires the extra precision, you will need to create two separate tables—one to account for clockwise boundary errors and the other to account for counterclockwise boundary errors.

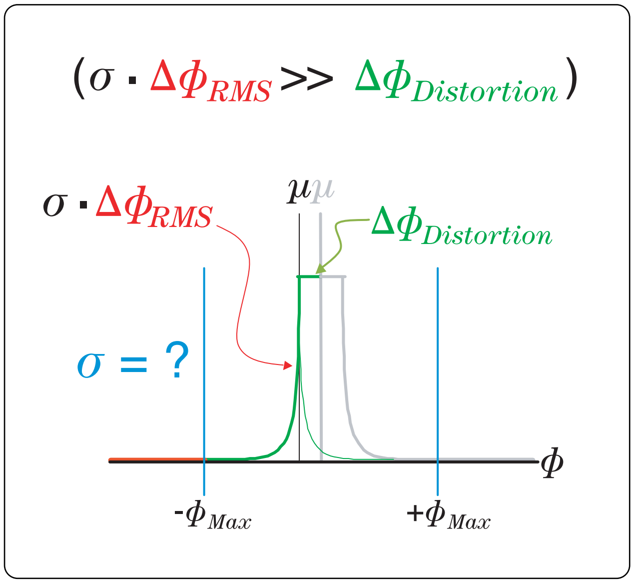


Figure 40: Key Assumptions

Q \ I	1.0	3.0	5.0	7.0
7.0	7E-15	5E-14	3E-13	2E-12
5.0	2E-39	1E-35	2E-32	3E-13
3.0	5E-139	6E-112	1E-35	5E-14
1.0	0E+00	5E-139	2E-39	7E-15

$$P_S = \frac{\sum_{i=1}^{4}\sum_{q=1}^{4} P_{iq}}{m} = 1.7\text{x}10^{-13}$$

Figure 42: Symbol Error Probability

The number of sigma to the boundary allows us to use the normalized probability density function (PDF) to calculate the probability of each symbol error (P_{iq}). The integral of the PDF has no closed-form solution, but fortunately there are tables and spreadsheet functions such as NORMDIST (X, μ, σ, C) which make it very easy to numerically evaluate.

$$P_{iq}(\Phi > \Delta\phi_{Max}) \approx 2 \cdot \int_{\Delta\phi_{Max} - \Delta\phi_{Distortion}}^{-\infty} \frac{1}{\sqrt{2\pi\sigma_{iq}^2}} \exp\left[-\frac{(\phi-\mu)^2}{2\sigma_{iq}^2}\right] d\phi$$

Figure 41: Approximate Symbol Error Probability Equation

1.5.4 Symbol Error Probability

Once we have calculated the individual symbol probabilities, we can average them to yield the probability of a symbol error. In this example, the probability of a symbol error is approximately 1.7×10^{-13}. This is quite good and on a par with fiber optic performance. Moreover, this is before forward error correction (FEC), which can gain us an additional one or two orders of magnitude improvement.

1.5.5 Bit Error Rate (Probability)

The symbol error probability must now be converted to the bit error rate (BER), or probability of a bit error. The conversion of symbol errors to bit errors is dependent on bit mapping (i.e., the arbitrary assignment of six different bits to each symbol in the constellation).

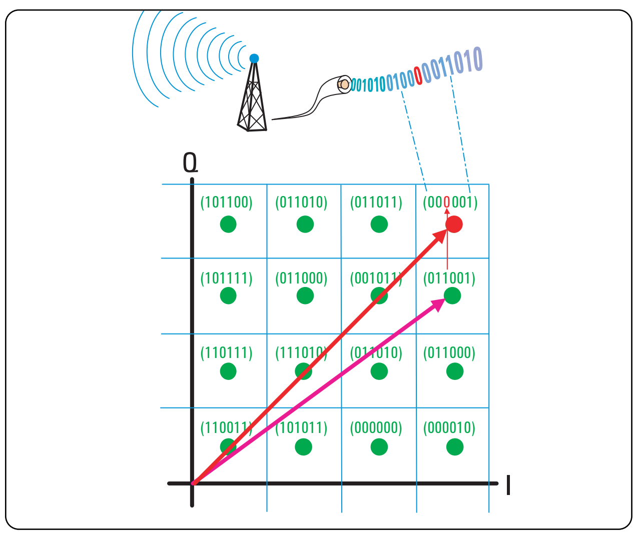


Figure 43: Symbol Error Rate to Bit Error Rate Mapping

Generally speaking, in mapping bits to symbols it is important to make adjacent symbols differ by as few bits as possible. Ideally, only a single bit will differ in the adjacent symbols and a symbol error will create only a single bit error (no error multiplication).

The conversion factor between symbol errors and bit errors is typically a low number. Again, the usual focus in high-QoS systems is on orders of magnitude rather than a specific number. In practice, with most bit mappings, it is usually a close approximation to assume that a symbol error results in a single bit error. *The calculated symbol error probability is usually a close approximation of the BER.*

Once the calculated BER is found to be acceptable, no reallocation of distortion or phase noise is necessary, and we can proceed to subdividing the top level allocation to its individual components.

1.5.6 Subdivided Allocation Budget

The final step is to subdivide the total allocation into the components of the system which generate it.

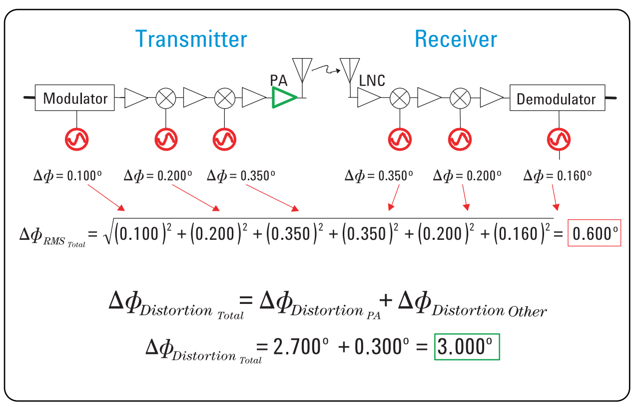


Figure 44: Subdividing the Allocation Budget

In the case of phase noise, our example system has six different sources. Since noise adds in RMS fashion, each of the sources of integrated phase noise is geometrically summed to add up to our total of 0.600°. Again, the systems engineer must choose the phase noise requirement of each source to minimize cost. The key consideration is the 20LogN relationship of multiplied noise (i.e., high-frequency sources tend to have more noise than low-frequency sources).

A similar subdivision of the distortion budget can be made but to a lesser extent. It is generally wise to reserve approximately 10% of the distortion budget for secondary effects such as incompletely compensated group delay, PA power leveling, etc.

The skill of the systems engineer in making the right allocations and subdividing the budget into practical, realizable values affects the cost and the competitiveness of the radio system in a *major* way, as we pointed out earlier.

A tool that is useful in subdividing the phase noise budget across a system is a simple pair of pie charts. One displays phase noise while the other displays the cost of the phase noise.

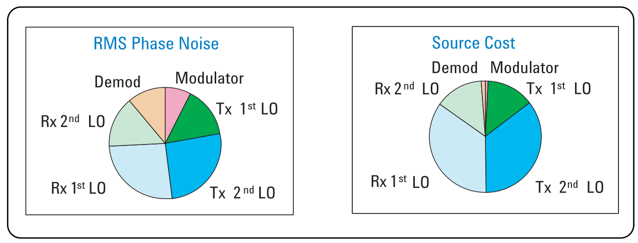


Figure 45: System Comparison of Phase Noise vs. Cost

It is often helpful to compare the phase noise contribution of a component of the system against its cost. A large phase noise contribution made by a low-cost source may warrant tightening that source's performance criteria. Modem oscillators often fall into this category —inexpensive surface-mount crystal oscillators can save a few pennies, but can require expensive microwave synthesizers to tighten their requirements. The result—the overall system cost skyrockets!

Another useful tool to minimize cost is to calculate the cost-per-degree of phase noise. The lowest cost-per-degree may be the best place to start in searching for savings.

A final approach to looking at the systems residual BER budget is to examine the error budget for the outermost constellation point (the one that usually suffers the worst distortion).

Let's look at a realistic example that parallels our discussion in the introduction.

Figure 46 shows the outer symbol angular data for three example links.

	Spec.		Case #1		Case #2		Case #3	
Modulator Phase Noise	?			0.10		0.10		0.10
1st Tx LO Phase Noise				0.20		0.20		0.20
2nd Tx LO Phase Noise				0.30		0.30		0.30
PA Distortion			2.69		1.50		2.00	
Other Distortion			0.30		0.30		0.30	
1st Rx LO Phase Noise				0.35		0.35		0.35
2nd Rx LO Phase Noise				0.20		0.20		0.20
Demodulator Phase Noise				0.30		0.32		0.35
			2.99	0.63	1.80	0.64	2.30	0.65
Total Angle Error	7.69		7.88		6.77		7.39	

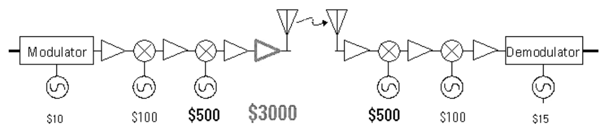


Figure 46: Total Angular Error Without A System Budget

In this example, the transmit/receive link in Case 1 exceeds the maximum angular error, causing it to dribble errors at an unacceptable rate. Again, the test technicians suspect the power amplifier in situations like this. Indeed, if the PA of Case #1 is swapped with the PA of Case #2—a radio which is working properly—the link in Case #2 will also begin to dribble errors (note that most of the numbers are the same). Similarly, the link in Case #3, which starts out well, will also begin to dribble errors when the technician swaps the PA from Case #1 for the PA of Case #3. In situations like this, technicians are usually convinced that because the problem followed the "bad" PA, the PA must be the cause. As a result, a $3000 PA is put aside as scrap!

But are all those PA's really unusable? Actually, they may all be fine. Let's see why.

We have now added a system residual BER budget specification (left-hand column, "Spec"). Notice that if the crystal oscillator from the demodulator in Case #1 had met its spec, the link in Case #1 would perform fine with the original PA #1.

	Spec.		Case #1		Case #2		Case #3	
Modulator Phase Noise		0.10		0.10		0.10		0.10
1st Tx LO Phase Noise		0.20		0.20		0.20		0.20
2nd Tx LO Phase Noise		0.35		0.30		0.30		0.30
PA Distortion	2.70		2.69		1.50		2.00	
Other Distortion	0.30		0.30		0.30		0.30	
1st Rx LO Phase Noise		0.35		0.35		0.35		0.35
2nd Rx LO Phase Noise		0.20		0.20		0.20		0.20
Demodulator Phase Noise		0.16		0.30		0.32		0.35
	3.00	0.60	2.99	0.63	1.80	0.64	2.30	0.65
Total Angle Error	7.69		7.88		6.77		7.39	

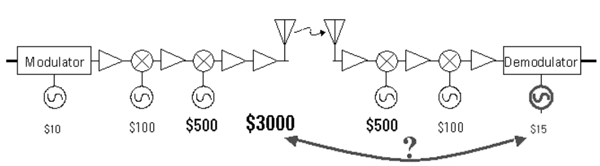


Figure 47: Total Angular Error With A System Budget

The author has encountered this exact case in two different companies where the cost of the modem source was ~$10, compared to a ~$3000 cost for the PA. Put another way, improvements in the PA would cost hundreds or thousands of dollars to make, whereas improvements in the demodulator source would cost pennies! Likewise, system engineers can often make or break a product's economic viability when choosing the allocation between phase noise and linearity. A noisy frequency conversion or modem oscillator can make the power amplifier's cost unacceptably high, just as poor PA performance can drive up the cost of an LO frequency synthesizer.

Budgeting tools like these provide badly-needed insight which, along with the systems engineer's wisdom in trading off phase noise against linearity, can radically alter the radio system's overall cost—a very important consideration in high-QoS products.

The reader who wishes to evaluate the cost-effectiveness of a system may want to take advantage of a free spreadsheet from Agilent. It serves as a general-purpose start to the residual BER budgeting process that can easily be modified for your specific needs. Simply go to **www.agilent.com/find/rxtx.**

1.6 Golden Unit Testing

One last important point in characterizing residual BER: Do *not* use "golden" units for testing. "Golden" units are product assemblies used to test other products. The idea is to test other assemblies against a fixed, non-varying mating assembly. Residual BER is a function of many factors distributed across the entire system, all of which add together in a complex way. If the total exceeds the maximum allowable angular error, BER will be too high. The problem with golden units is finding one with sources and amplifiers having worst-case phase noise and distortion, a rare possibility. In fact, a "good" unit is often chosen on the basis of having lower-than-average phase noise and distortion. Figure 48 (pg. 26) illustrates this by analyzing the error budget's complex summation for the outer constellation point.

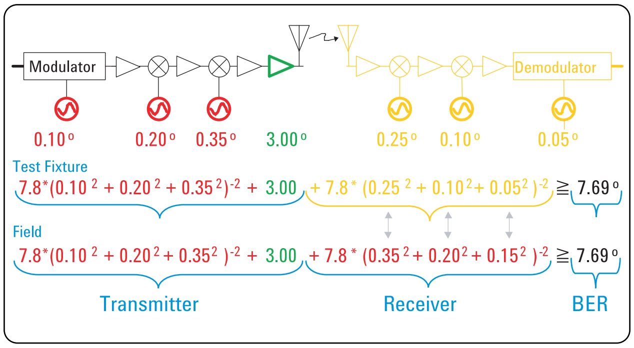


Figure 48: Golden Unit Testing Issues

Another key issue in golden unit testing is that no quantitative parametric data is generated. This makes it impossible to predict compatibility with other units or vendors without retesting. In addition, the "go/no go" nature of golden unit testing provides little in the way of statistical process control (SPC), due to its poor resolution (pass/fail only) and inability to separate different impairments. Many of the components which determine phase noise and linearity are difficult to control and can vary from one production lot to another. Hence, statistical process control (SPC) techniques are warranted in order to minimize final integration interruptions—thus high-resolution, traceable parametric measurements are essential.

We repeat, do *not* use golden modems as part of your production process in measuring residual BER. With their better-than-normal performance, golden modems can easily mask other problems in your radio chain, causing a unit to pass in production but dribble errors in the field! The author has seen cases in which entire product lines had to be shut down (at tremendous loss) because the manufacturer relied for interoperability on golden unit testing which failed, giving an otherwise good design an unreliable reputation.

1.7 Concept Summary

We have covered the theoretical foundations of bit error rate (BER) in some detail. Hopefully by this time you have developed a better understanding of this important subject, which is so vital to producing wireless data links that meet both the quality-of-service and cost-effectiveness requirements of the network provider. The system engineer who judiciously allocates a residual BER budget creates a significant competitive advantage in product cost as well as the interchangeability necessary for today's highly-complex radio products.

We outlined the process for predicting the worst-case residual BER from the analog metrics which influence it, contrasting that to empirical approaches, many of which don't work. Using a combination of phase noise and AM/PM distortion measurements, we showed how the various sources and PA can be characterized to deliver consistent combined residual BER performance.

In the following sections, we will cover the two most significant contributors to BER—phase noise and nonlinearity—in more detail, concentrating on the many different measurements, techniques, and equipment options available to characterize them.

2.1 Understanding Phase Noise Applications

From our analysis of residual BER so far, we have seen that phase noise plays an important role in system performance and can affect the linearity requirements and cost of the PA. In fact, phase noise is one of the *most* expensive parameters to design for in high-performance radio systems. It is critical to not only have the optimum balance between phase noise and linearity, but also to measure both characteristics accurately.

Most radio links which demand low BER require constant, vigilant monitoring of component performance. Why is this so? The cost of achieving the low phase noise suitable for data links using spectrally efficient modulations is so expensive that engineering excess margin for it may be neither practical nor cost-effective. Instead, careful monitoring of marginal-performance components is usually the least-costly approach.

Several techniques can be used to measure phase noise. A clear understanding of each is essential to minimizing the cost of test and ensuring proper equipment selection in order to get the most out of every piece of equipment purchased. Before reviewing these methods, let's take a moment to review the definition and sources of phase noise.

2.2 Defining Phase Noise and Measurement Principles

Many different units can be used to quantify phase noise. Let's examine the most useful ones to for predicting residual BER, how they are derived, and how they relate to each other.

2.2.1 Frequency Stability

Frequency stability is defined as the degree to which an oscillating source produces the same frequency throughout a specified period of time. Every RF and microwave source exhibits some frequency instability. Frequency stability can be broken down into two components—long-term and short-term.

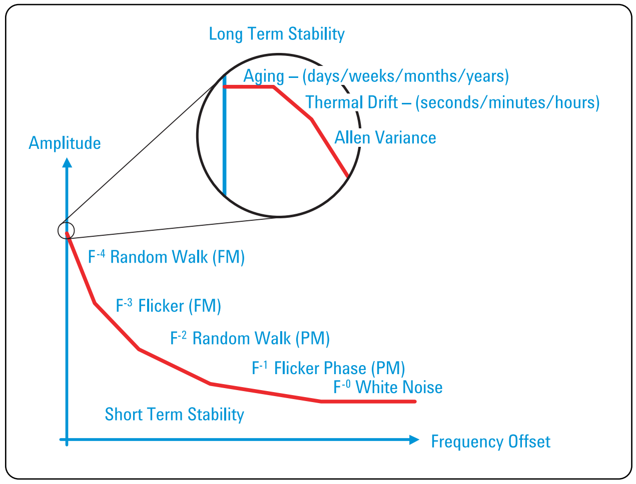


Figure 49: Frequency Stability, Long-term and Short-term

Long-term stability is related to the frequency variations occurring over long time periods, expressed in parts per million per hour, per day, per month, or per year. Short-term frequency relates to those changes occurring in less than a few seconds' duration. Mathematically, an ideal sinewave can be described by:

$$V(t) = V_o \sin(2\pi f_c t)$$

where V_o = nominal amplitude

$2\pi f_c t$ = linear growing phase component

f_c = nominal center frequency

(NOTE: There is often confusion over the terms "frequency" and "frequency offset" as they relate to phase noise. The term f_o can mean frequency "not" or null—that is, center frequency—or it can mean frequency offset from the carrier. In this application note, f_o will mean frequency offset from the carrier, while f_c will mean the frequency of the carrier.)

An actual signal is better modeled by:

$$V(t) = |V_o + \mathcal{E}(t)| \sin[2\pi f_c t + \Delta\phi(t)]$$

where $\mathcal{E}(t)$ = amplitude fluctuations

$\Delta\phi(t)$ = randomly fluctuating phase or phase noise

Fluctuating phase could be observed on an ideal spectrum analyzer (one which has no sideband noise of its own), as shown in Figure 50. There are two types of fluctuating phase terms. The first is deterministic, in which discrete signals appear as distinct components in the spectral density plot. These signals, commonly called *spurious*, can be related to known phenomena in the signal source, such as power line frequency, vibration frequencies, or mixer products.

2.2.2 Random Fluctuation (Phase Noise)

The second type of phase fluctuation is *random*, or probabilistic, in nature, and is commonly called phase noise. Sources of phase noise in an oscillator include thermal white noise, flicker noise, shot noise, popcorn noise, and random walk.

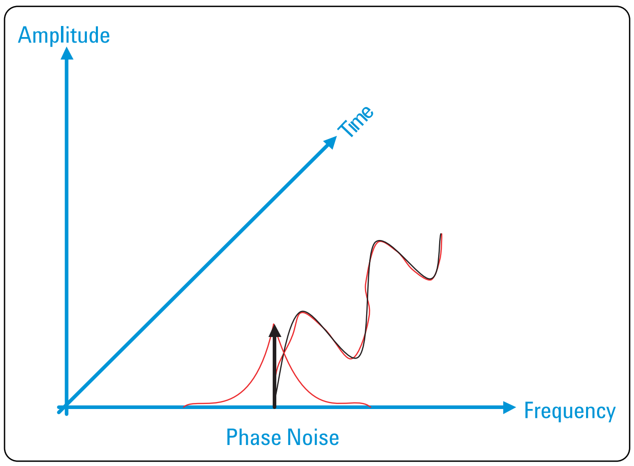


Figure 50: Frequency Stability and Phase Fluctuations

Many terms can be used to quantify the randomness of phase noise. Essentially, all of them are based on measuring the frequency, or phase deviations, of the source in either the frequency or time domains. Since frequency and phase are related, these terms are also related.

One fundamental description of phase noise is the one-sided spectral density of phase fluctuation on a per-Hertz basis. The term "spectral density" describes the energy distribution as a continuous function, expressed in units of energy within a specified bandwidth. Thus, $S\Delta\phi(f_0)$ is defined as:

$$S_{\Delta\phi}(f_o) = \frac{\Delta\phi^2_{RMS}}{BW \text{used to measure } \Delta\phi_{RMS}} \quad \frac{rad^2}{Hz}$$

If the modulation sidebands are such that the total phase deviations are << 1 radian, another useful measure of the noise energy is $\mathscr{L}(f_o)$, which is then directly related to $S\Delta\Phi(f_o)$ by $\mathscr{L}(f_o)$.

$\mathscr{L}(f_o)$ is an indirect measure of noise energy that is easily related to the RF power spectrum observed on a spectrum analyzer. As shown in Figure 51, $\mathscr{L}(f_o)$ is defined by the U.S. National Institute of Standards Technology (NIST) as the ratio of the power in one phase modulation sideband, on a per-1 Hertz of bandwidth power spectral density basis, to the total signal power, at an offset f_o Hertz away from the carrier.

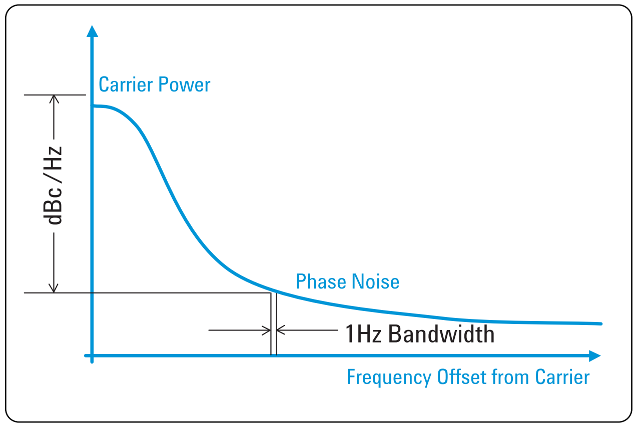


Figure 51: Definition of Single Side Band Phase Noise

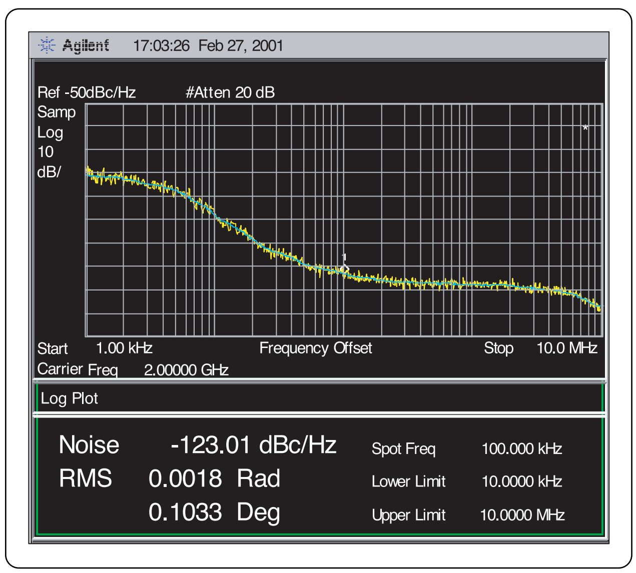


Figure 52: Single Sideband Phase Noise Measurement

Now that we have a definition of single sideband phase noise, let's look at the principal approaches to measuring it.

2.2.3 Direct Measurement Method

The most straightforward method of measuring phase noise is to input the test signal into a spectrum analyzer or vector signal analyzer and directly measure the power spectral density of the oscillator with a 1Hz resolution bandwidth. Called the direct measurement method, this is probably the most widespread technique for measuring phase noise for most residual BER applications. Its advantages are that it is fast and easy to set up and that it captures data across a wide range of carrier offsets. The direct method does have a few limitations, however. It assumes that all carrier sideband energy is due to phase modulation, since it cannot distinguish between that and sideband energy from amplitude modulation sources. It requires a spectrum analyzer with lower internal phase noise than the source being measured. Finally, it requires a source with adequate long-term stability to make the measurement given the resolution bandwidth of the spectrum analyzer. This means that if the source being measured is drifting so fast the resolution bandwidth filter cannot adequately capture the signal, significant measurement errors can occur. This is often the case with voltage controlled oscillators (VCOs).

2.2.4 Heterodyne Frequency Measurement

Heterodyne frequency measurement is a time domain technique that begins by downconverting the signal under test to an intermediate frequency (IF) in the range of a frequency counter. Then a high-resolution frequency counter is used to make repeated counts of the signal, with the time period for the counts held constant. This allows for several calculations to be made of the fractional frequency difference, in phase (y), over the time period. From these values for y, the Allan variance, $\sigma_y(\tau)$, can be computed. The term $\sigma_y(\tau)$ in the time domain corresponds to $\mathscr{L}(f)$ in the frequency domain.

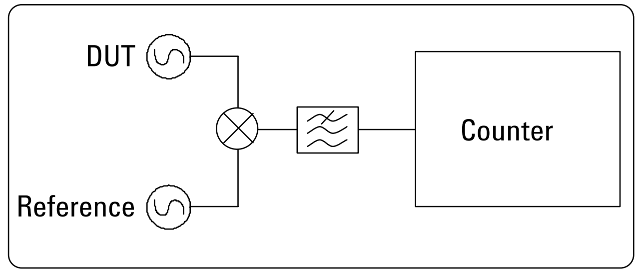


Figure 53: Counter Method

This method is particularly useful for short-term frequency instabilities occurring over periods of time greater than 10 ms (less than 100 Hz offsets in the frequency domain), where single sideband phase noise is falling rapidly. It is ideally suited for very close-in measurements. Typically, most residual BER issues occur on modulations that are far wider than those for which the counter method is best suited. The counter method is useful for making measurements in the region where carrier tracking problems occur–a subject related to residual BER prediction. Key disadvantages are: Like the direct method, it requires a conversion (heterodyning) source with known noise less than the source being measured, it is not applicable to wide frequency offset measurements, and it does not work well when measuring noise floors which are flat.

2.2.5 Carrier Removal/Demodulation

Most of the techniques for dedicated phase noise measurement equipment fall into this class. The principle is similar to that of an actual radio demodulator. Increased sensitivity can be obtained by nulling or demodulating the carrier, then measuring the noise of the resulting baseband signal. The most common methods are measuring the noise with a frequency discriminator or with a phase detector.

Frequency discriminator methods often use delay lines/mixer, cavities, and bridge configurations connected to measure the frequency noise at specific offsets from the carrier. The method has the unique advantage of not requiring a separate reference source for downconversion. The primary disadvantage is poor close-to-carrier, or close-in, sensitivity.

Phase detector implementations use a quadrature relationship between the measured signal and the reference signal to create a resulting signal that is proportional to the phase fluctuations. This method provides the best sensitivity and is used by most high-performance phase noise test sets. The primary disadvantages are that it requires a reference source better than the one being measured, is more complex to set up, and does not work well with unlocked VCO sources.

2.3 Phase Noise Measurement Equipment Considerations

As noted in the previous section, different methods of phase noise measurement equipment operation are available for different applications. Choosing the best equipment for a particular application depends very much on the nature of the application, as we shall soon see.

The first thing to consider when testing phase noise for residual BER applications is: What type of component connection is available—an individual crystal oscillator, a frequency converter with integrated synthesizer, or an entire transmitter? A crystal oscillator may have a convenient SMA connector and a high-level signal that is easily measured, or it could be integrated onto a PC board where a trace probe is necessary. Similarly, a frequency converter with an integrated synthesizer could be a single-component with convenient access to each LO via SMA connectors, or it might have only an IF input and RF output with no direct access to the LOs. This type of component requires a low phase noise test source to stimulate the converter so that the degradation can be measured at the output. Complete transmitter chains can be characterized, but this requires the ability to turn off the modulation so that only the CW carrier can be characterized.

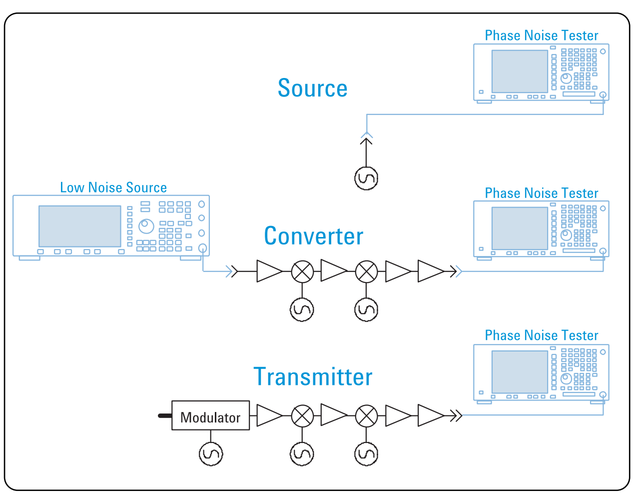


Figure 54: Characterizing Sources, Converters and Transmitters

A device's physical technology and manufacturing source often determine the level of integration and types of connections available to characterize its phase noise contribution. For new designs, it is generally recommended that a manufacturer add provisions to gain access to the source signals. A small probe point or an SMA connector tapping into a signal with adequate power for direct measurement can greatly simplify the test engineer's job, as well as lowering the cost of test.

The next consideration in setting up a phase noise test system for residual BER prediction is the long-term stability of the signal source to be measured. In particular, many voltage controlled oscillators (VCOs) have very poor long-term stability (seconds, minutes, hours, etc.) and cannot be measured directly using the spectrum analyzer method. The VCOs must either be locked up to their reference or must be measured via the frequency discriminator method. Thus, when choosing a phase noise test system to measure VCOs, you should either lock the VCO to its reference, use a test system that will lock the VCO, or use the frequency discriminator method.

The dynamic range required for the measurement is also an important factor to consider in choosing test equipment and connection points. For example, if you measured a crystal oscillator with phase noise of –170 dBc/Hz at 1MHz offset from the carrier and carrier power of +10 dBm, an inexpensive spectrum analyzer with a noise floor of –113 dBm/Hz would give you inaccurate results. Why? Because the analyzer with a limited dynamic range can measure a phase noise of only –123 dBc/Hz (+10 dBm - [-113] dBm/Hz), not the –170 dBc/Hz of the source's "actual" phase noise.

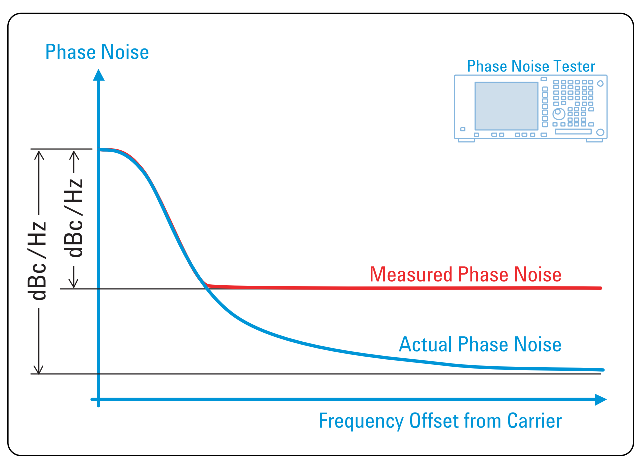


Figure 55: Phase Noise Dynamic Range Considerations

Dynamic range can also be a factor in using the direct measurement technique. You should choose a spectrum analyzer with adequate dynamic range for the measurement. It is also a good idea consider boosting the signal level with a pre-amplifier before using the spectrum analyzer, overcoming the high thermal noise floor in the test equipment, which then allows you to make full use of its dynamic range. Phase noise is a relative measurement and should not change with linear amplification of the source's output. If it does change as amplification is increased, it is not the source's phase noise which is being measured, but rather the dynamic range of the instrument.

Similarly, another important consideration is the phase noise performance of the instrument, which must typically exceed that of the source being tested. Historically, phase noise performance had to exceed the performance of the DUT by about 10 dB. Modern high-performance phase noise measurement equipment now allows the capability to "subtract out" some of the analyzer's phase noise via complex digital algorithms and careful calibration of the noise at a particular frequency. This process is called near noise correction and helps to reduce the cost of the measurement equipment while extending its useful range.

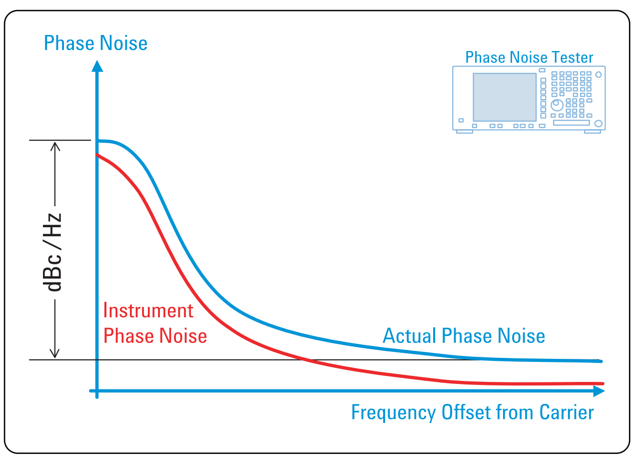


Figure 56: Instrument Phase Noise Considerations

Though near noise correction can be very helpful in stretching the performance limits of the test equipment, it is still best to purchase development equipment with phase noise performance that is lower then that of the DUT by 10 dB. This eliminates the time and effort required to calibrate for near noise correction. In the production environment, calibration issues can often be automated and are not a significant problem.

One final factor that often plays an important role in selecting measurement equipment is the "20LogN relationship" of multiplied signals. When a frequency source is multiplied, the phase noise grows by the following factor:

Sideband noise after multiplication = 20 log(N)+sideband noise before multiplication

where N = number of times of multiplication.

For example, if the phase noise of a 2GHz signal before a multiplier is –125 dBc/Hz at 100 kHz offset from center frequency and the signal is multiplied by 2 to 4 GHz, the phase noise will grow to –119 dBc/Hz at 100 kHz offset. This relationship can often be used to advantage when testing a source. Performance levels which demand high-cost measurement equipment at low frequencies (MHz) can easily be measured after multiplication using lower-cost spectrum analyzers. A classic example is a crystal source being multiplied with a step recovery diode (SRD). After multiplication by a factor of 100 or 40 dB, sideband noise power which could have been measured only with an expensive phase noise test set can now easily be measured with a lower-cost spectrum analyzer.

2.4 Common Phase Noise Measurement Applications

Let's now look at how these measurement principles and considerations apply in practice to specific component types.

2.4.1 Crystal Reference Oscillators

We'll look first at crystal reference oscillators, which are often the starting point for complex modulation generation or complex LO synthesizers.

Crystal reference oscillators are used in most radios to ensure precise frequency control. The high quality factor (Q) of the crystal resonator and excellent long-term stability are ideal for this application. Crystal oscillators typically have very low phase noise, requiring the dynamic range and sensitivity of a dedicated phase noise test set, using the demodulation carrier removal method of the phase detector method.

Though these systems are costly, they can provide invaluable insight into the actual source of the phase noise. Why is this so? Most crystal sources in modern radios are followed by elaborate frequency synthesizers. If the crystal source is "assumed" to be good but is in fact not, considerable effort may be expended in vain trying to fix the wrong components. In high-performance radios with complex modulations, achieving the lowest possible phase noise begins with the crystal source, so precise characterization is essential.

2.4.2 Voltage Controlled Oscillators (VCOs)

The next component to consider is the voltage controlled oscillator (VCO). VCOs come in many types, usually based on the resonant structure used– crystal VCOs (commonly called VCXOs), voltage controlled dielectric resonant oscillators (DROs), transmission line VCOs, and YIG oscillators (actually a current controlled oscillator), to name a few. All of these oscillators share a common trait–their long-term stability (seconds, minutes, hours, etc.) is typically very poor. Thus they either need to be phase-locked to a stable reference oscillator in order to measure their phase noise, or the frequency discriminator demodulation method should be used.

Free-running VCOs usually exhibit significant frequency variation and must be measured by either the frequency discriminator method, with a dedicated VCO test set, or phase-locked and measured by the direct method. Most VCOs are best characterized using a dedicated VCO test set designed to phase-lock the VCO, then measure the phase noise. Dedicated VCO testers also measure other important parameters such as tune sensitivity, which is essential to the proper use of a VCO inside a control loop, such as a phase-lock loop.

Of all components requiring phase noise testing, VCOs tend to need the most constant monitoring during production. Why is this? The voltage control aspect of the VCO naturally gives it a higher sensitivity to phase modulation from noise on the tune control line. This higher sensitivity is compounded by variations in the characteristics of the "varactors" commonly used to tune the VCO using voltage variation. Varactor characteristics are dependent on semiconductor processing, which in turn relies on photo-lithography resolution and controlled addition of impurities–both of which vary from wafer to wafer. The difficulty in controlling these processes so that each diode exhibits identical tune sensitivity, resistance, and noise characteristics is immense. Consequently, the phase noise performance of these components tends to change over time (from wafer to wafer) much more than other components. Many companies go so far as purchasing an entire wafer lot at one time to obtain better consistency in performance.

Adding to VCO performance variations from one unit to another is the fact that VCOs are generally used in phase-locked synthesizers. This brings into the phase noise equation all of the loop parameters which set crossover points between reference phase noise characteristics and VCO characteristics and damping response, some of which are based on the varactor's properties. Consequently, VCOs are often among the highest initial sources of phase noise in many systems, *as well as* varying in performance over the lifetime of a product line! Hence they should be continuously monitored to avoid expensive over-engineering of other components. The cost of continuous monitoring of marginal VCO parts and vigilant statistical process control (SPC) is usually very small when compared to extremely tight part tolerance or highly complex designs.

2.4.3 Amplifiers and Frequency Multipliers

Amplifiers can add phase noise to the desired signal from bias modulation, filament modulation, poor common mode rejection, high thermal noise floor (from poor noise figure), and a variety of other sources. Frequency multipliers suffer from similar problems, while also increasing the phase noise by 20LogN, the number of times the signal is multiplied.

Amplifiers and multipliers present several unique challenges in characterizing their phase noise. First, they require a low phase noise source to stimulate the DUT, since amplifiers and multipliers do not actually produce a sinewave to measure. The key to the source of stimulation is that its phase noise/noise floor must be lower than the phase noise to be measured.

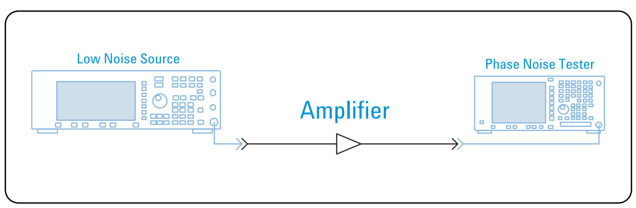


Figure 57: Testing Amplifier Phase Noise

Often, the preferred method of testing amplifiers and multipliers is the direct measurement method, since it is done with a spectrum analyzer. The spectrum analyzer is also capable of searching for spurious emissions and harmonic levels that can be related to the same mechanisms that cause high phase noise. Thus, the direct method usually provides a degree of convenience and a lower cost that other approaches cannot.

Unfortunately, for some applications the phase noise performance of even the best spectrum analyzers is inadequate to measure the miniscule phase noise added by very low noise solid-state amplifiers. For these applications, the carrier removal/demodulation approach (i.e., frequency discriminator or phase detector method) of a dedicated phase noise test set is best.

Frequency multipliers may also require a phase noise test set rather than a spectrum analyzer because of their very low phase noise contribution. The 20LogN relationship substantially lessens this requirement over simple amplifiers and can even be used to lower test equipment costs for some devices. For example, if you tried to measure the phase noise on the 10^{th} harmonic of a 10 MHz crystal reference using a step recovery diode (SRD) multiplier, it might be well below the measurement capability of even a high-performance spectrum analyzer. However, if you measured the phase noise on the 100^{th} harmonic of 10 MHz (1GHz), the 40 dB increase in phase noise from the 20LogN relationship would enable you to measure it using a spectrum analyzer, providing insight into what performance might be like at the 10^{th} harmonic. Given the significant cost difference between a spectrum analyzer and a phase noise test set, the 20LogN relationship should be used to approximate actual performance whenever possible.

2.4.4 Frequency Synthesizer Subassemblies

In the last decade, frequency synthesizers have become pervasive in most radio applications. The ability to tune easily to any channel is a highly desirable feature of most radio systems. Even fixed point-to-point licensed backhaul systems provide this feature. Frequency synthesizers generally have higher phase noise than many of their component building blocks, permitting characterization via the direct measurement method.

There are many advantages in using a spectrum analyzer for testing frequency synthesizer subassemblies and super-components. The ability of a spectrum analyzer to make related measurements such as spurious emissions, harmonics, power, time-gated signals, and long-term frequency drift, in addition to phase noise, is a significant factor in equipment selection. In addition, the spectrum analyzer is among the lowest-cost items of equipment for measuring phase noise.

In rare cases, a synthesizer DUT's performance is so good that it necessitates a dedicated phase noise test set.

2.4.5 Frequency Converter Subassemblies

There are two approaches to characterizing entire frequency converter subassemblies.

First, each local oscillator (LO) can be measured independently and its phase noise root-mean-square added to determine the overall phase noise. The advantage of this approach is that each LO contribution can be viewed separately, so if a problem occurs, its source is known. The disadvantage is that each LO must be connected in turn, and the resolution and dynamic range of the measuring instrument must be better, since each LO is less than the RMS total. Depending on the system's performance, this may mean that low-cost direct measurements with a spectrum analyzer may not be possible for all LOs, substantially raising the cost of test.

The second approach is to inject a low-noise stimulus into the converter and measure the phase noise added to the output of the converter. This again requires an LO which has lower phase noise than the noise of the converter to be measured. The advantage of characterizing the entire converter with an injected carrier wave tone is that it does not require access to individual LOs and thus saves on connection costs (particularly if some LOs are in the mm-wave frequencies). The disadvantage is that for multiple-frequency converters, it provides little insight into where a problem lies. Injecting a low-noise stimulus and measuring its phase noise with a spectrum analyzer is a simple way to validate that an outdoor unit (ODU) meets the required portion of a phase noise budget.

2.4.6 Transmitters and Receivers

Testing the phase noise of an entire transmitter can be done easily–if the modulation of the carrier can be turned off. The output of the transmitter can then usually be measured directly with a spectrum analyzer and an appropriate attenuator (to protect the analyzer from excessive power). Typically, the strong signal levels of the transmitter, which are far from the thermal noise floor of the instrument, and the combined phase noise of all the sources in the system, make for an easy measurement. While this is an excellent way to characterize high-level system performance and compatibility with numerous receivers, it lacks the resolution for diagnostics within the transmitter.

Testing the phase noise of an entire receiver is more complex than that of a transmitter, because a key element in the receiver is the carrier tracking loop, which contains an oscillator that adds to the total receiver phase noise. The carrier tracking loop oscillator is used to downconvert the received signal to baseband, often baseband I and Q signals. This presents some unique measurement challenges and requires a different test interface.

Vector signal analyzers equipped with baseband I and Q inputs can be used to analyze the phase noise in a receiver. Complete receiver testing requires the injection of a carrier signal (unmodulated CW sine wave) with a lower phase noise than that of the receiver. The receiver will then downconvert the signal and phase-lock it to the carrier. At this point, the baseband I and Q signals can be connected to a vector signal analyzer equipped with baseband inputs. The vector signal analyzer can then be used to look at the phase noise by choosing the power spectral density (PSD) display (the direct method). The vector signal analyzer reconstructs the I and Q inputs, enabling analysis of the phase noise.

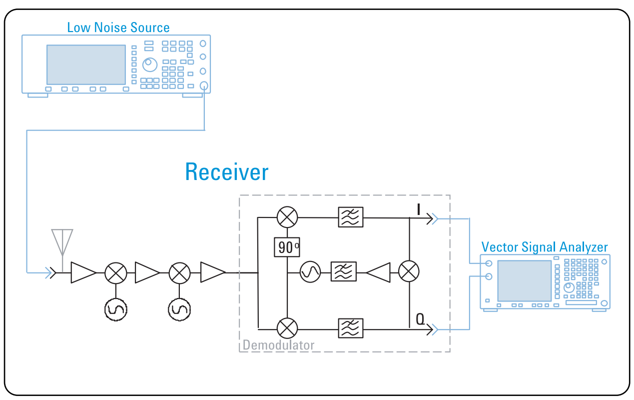


Figure 58: Testing Receiver Phase Noise

Thus the IQ reconstruction allows us to measure the entire receiver's performance, including the VCO embedded in the carrier recovery circuit, as shown in Figure 58. This can be a very insightful measurement, because quite often the embedded VCO's noise performance is impaired, as we will see in the next section.

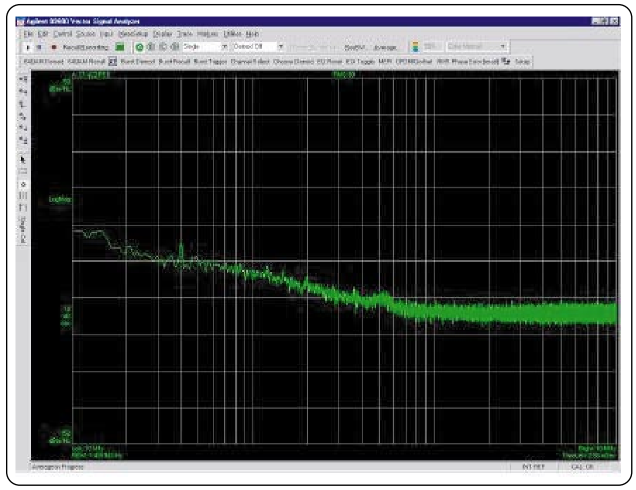


Figure 59: Reconstructed I and Q Phase Noise

2.4.7 Modulators and Demodulators

Modulators and demodulators present unique challenges in characterizing their phase noise. The low frequency sources used in these devices, usually crystal resonators, are often neglected as contributors to phase noise. The belief is that the low-frequency and high-Q crystal oscillator used in modems contribute little compared to the very high-frequency microwave sources, some of which have low-Q varactor-controlled oscillators which tend to have much higher phase noise. In practice, this is often *not* true! Why is this? Although the low-frequency crystal sources tend to be lower in phase noise, the applications inside modulators and demodulators can easily degrade their performance. Let's see why.

Sources used in modulators and demodulators are typically located in a very noisy digital environment. It is common practice to place crystal oscillators in close proximity to large digital application-specific integrated circuits (ASICs) or field-programmable gate arrays (FPGAs) that produce noise-like spectra which can be conducted via power supply lines to the oscillators or simply radiated to sensitive oscillator points–for example, a voltage controlled oscillator that is being used for the carrier recovery in a demodulator. Even very low levels of digital noise radiated onto the tune line will increase the phase noise of the VCO over that tested outside the demodulator. Carrier tracking loops in a demodulator with incorrect damping can also elevate phase noise to higher levels than expected for a low-frequency crystal oscillator.

Amazingly, many designs attempt to cut the cost or size of a modem by using inexpensive crystal oscillators with poorly-controlled phase noise performance. It is critical to realize that these oscillators contribute to the overall system budget so that *a small savings in the modem can translate into a major cost or size penalty in the RF portion of the system!* Thus, it is highly advisable to characterize the phase noise performance of the modem and, in many cases, it is financially advantageous to do so on an ongoing basis, when tolerances of degrading factors are difficult to control (PC board etching, power filters, isolation shields, varactor impurities, rise times and f_t of ASICS, etc.).

Modulators, like transmitters, are relatively straightforward to characterize if the modulation can be turned off and only the carrier signal comes through. The carrier signal can then be measured directly with any technique having adequate sensitivity to produce an accurate reading. The direct method (spectrum analyzer) has the advantage of convenience—other measurements, such as spectral emission masks, can be done using the same analyzer and connection. The vector signal analyzer adds even more convenience, since it is also capable of doing constellation analysis, EVM, and a host of other modulation quality measurements. High-performance systems may require a dedicated phase noise test set to measure very low levels of phase noise.

Demodulators, like receivers, can have the added complication that the carrier recovery loop must lock to a stimulus signal with lower noise than the one to be measured. Ideally, the carrier recovery VCO has the capability to connect directly, making it easy to measure. Depending on the technology used to implement the VCO, it may be necessary to measure the phase noise using the I and Q technique discussed in the receiver measurement section.

In digital applications, it may also be necessary to measure the phase noise using the I and Q technique. For applications in which demodulator carrier tracking loops are completely implemented with digital technology (digitally-controlled oscillators or DCOs), it may be necessary to extract I and Q digitized bit streams to measure the phase noise. Simply measuring the DCO reference oscillator will usually *not* account for the phase variations created by most DCO implementations.

2.5 Phase Noise Measurement with a Vector Signal Analyzer (VSA)

Because of the vector signal analyzer's flexibility in making so many kinds of measurements, including phase noise, we should examine it in more depth.

As we have seen from our earlier discussion of RMS, the phase deviation in degrees, radians, or seconds of jitter can be very useful in predicting residual BER. Vector signal analyzers greatly facilitate these measurements if they include demodulation and time-domain features.

The vector signal analyzer is particularly useful for doing modem and top-level (transmitter/receiver characterization) system work. VSAs are often overlooked in measuring phase noise when engineers focus on error vector magnitude (EVM) measurements. However, the advantages of VSAs in making phase noise measurements are many. For example, if the signal-under-test is unlocked and drifting too much to make a satisfactory averaged measurement, a VSA with demodulation and auto-carrier functions can frequently track these phase perturbations, providing the necessary stability to make VCO-like measurements.

Likewise, when modulation quality measurements are made, it is possible to measure the carrier phase noise using the same vector signal analyzer to determine the portion of EVM impairment due to phase noise. This can provide valuable insight into the probable causes of EVM degradation, as well as assurance of transmitter/receiver interoperability.

Unlike small signal measurements, which are essentially independent of power level, linearity measurements are by their nature power level dependent. As we saw in Section 1, we should characterize the linearity of a device over the *expected* range of modulation vectors. Thus, before we can relate AM/PM performance to BER, we must characterize the range of vectors that the power amplifier will likely produce. Because of this, we'll begin our linearity section with complementary cumulative distribution function (CCDF), using this measurement to understand the vector range above average power over which the AM/PM measurement should be made.

3.1 Choosing the Correct Vector Range/Power Levels with CCDF

Although instruments like a power meter can measure average RMS power, what is often needed for today's digital radios is *peak* power statistics. CCDF is a valuable technique for examining the power statistics of digitally-modulated signals.

Why must we resort to statistics to measure the power of a signal? Most modern digital modulations have noise-*like* signals whose peak vector length is a function of several factors (baseband shaping/filter alpha, previous symbol states, etc.). This noise-like characteristic forces us to choose peak vector length/peak power based on statistical probability.

So what exactly is CCDF? Complementary cumulative distribution function is a tongue-twister to say—but conceptually easy to understand!

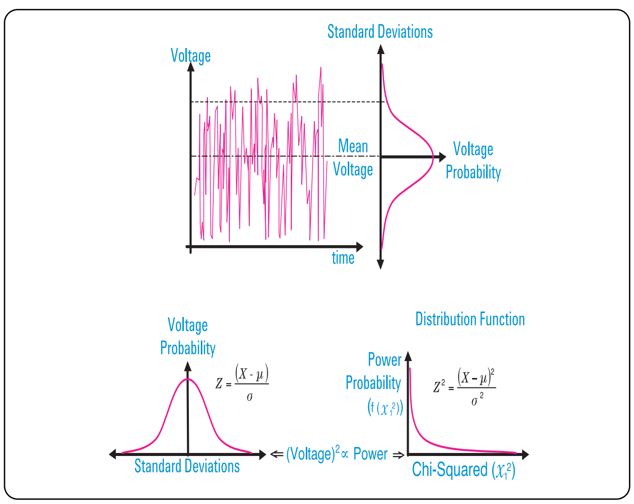


Figure 60: Modulated Signal Voltage and Power Probability

Figure 60 shows the voltage vs. time plot of a signal. How would this information be useful for design purposes? In fact, in this form, the information is *not* very useful.

The CCDF curve is generated for the normal distribution starting with the familiar normal probability density function (PDF). Note that this distribution function is a characteristic of the signal's modulation. (Be sure not to confuse this noise-*like* distribution of the power of a time domain modulated signal with the much smaller, unwanted phase noise. They are different matters altogether—amplitude vs. phase.)

To construct a power CCDF curve, we must square the normal PDF of the signal. Squaring generates a new type of function, the chi-squared PDF. This curve can be integrated based on the power level of interest. The question is normally: What percentage of time will the signal be less than some value? However, for amplifier design we must ask: What percentage of the time will the signal be *greater* than a specified power level? Hence it is necessary to subtract the integrated (or cumulative) chi-squared PDF from 1, or 100% probability. This gives the complementary cumulative distribution function, CCDF.

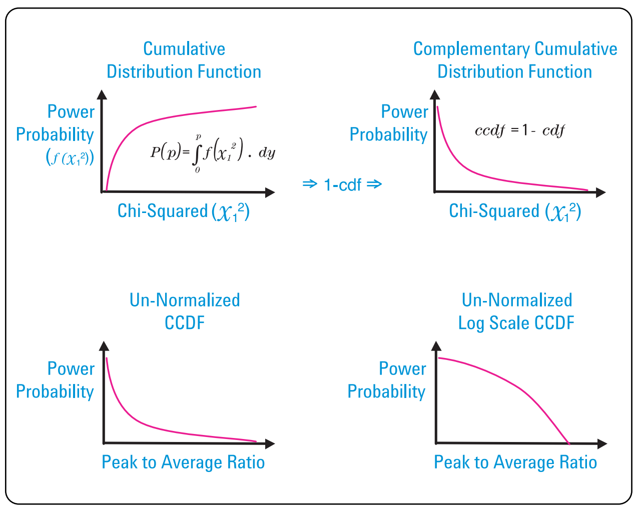


Figure 61: Steps to Log CCDF

The CCDF curve X-axis is then re-scaled to average power and both axes displayed on a log scale. The power CCDF curves depict only information from average power on up. These axis transformations yield better resolution of low-probability power events and are directly related to the RMS average power as measured on a power meter. They also allow CCDF power curve comparisons of signals having different average power values.

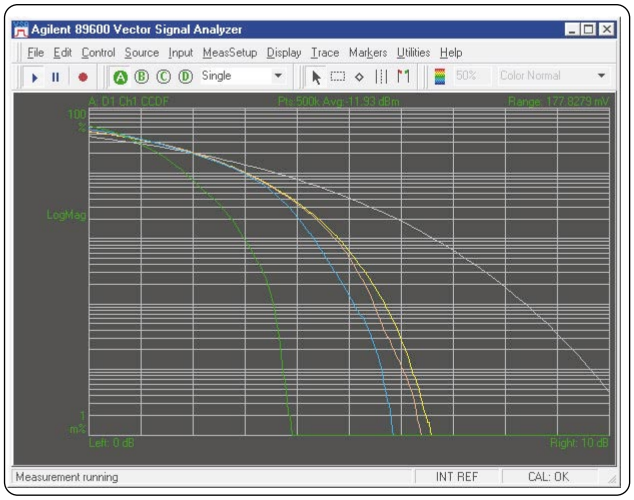


Figure 62: QPSK, 16, 64, and 256 QAM CCDFs

In Figure 62, the CCDF curve of the 16QAM signal is higher than that of the QPSK (quadrature phase shift keying) signal, which is to be expected since the 16QAM signal shows wider power excursions than the QPSK signal.

The amplifiers and components of a 16QAM system will have different design requirements than lower order modulations because of the higher peak-to-average statistics.

Factors other than modulation format can affect the CCDF curve. For example, the same modulation format with different filter coefficients can create a different CCDF because, once again, the power excursions for the two signals are different.

Now to apply the CCDF curve.

The power CCDF curve is scaled on the abscissa (x) axis to dB above signal average power. The ordinate (y) axis is also shown in a logarithmic scale to gain better resolution of the low-probability events.

We can find a percentage of time on the ordinate, then read across the graph and determine the signal's corresponding number of dB above average power for

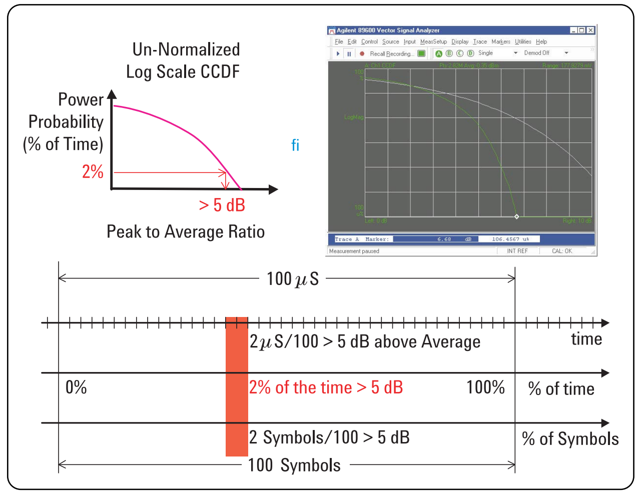


Figure 63: Relating CCDF Peak Powers to Symbols Affected

that time. For example, if we wanted to know the power above average which the modulated signal would exceed 2% of the time, we would read across the graph and see that it is 4.1 dB.

Sometimes, the complexity of the CCDF display can become overwhelming, causing confusion over how to apply it. It is helpful to remember that "percentage of time" can be directly related to the percentage of symbols transmitted. For example, if for 2% of the time the power level exceeds 4.1 dB, we can also say that for 2% of the symbols the power level exceeds 4.1 dB. This could mean that 2% of the symbols may reach a level at which linearity will become a problem. The CCDF curve answers the question: What is the vector range of interest? Most communications systems demand far *lower* levels of symbols potentially being impacted by distortion. For example, high-quality systems are at risk of excessive power in less than 1 symbol per every million symbols sent, or only 0.00001% of the time!

Measuring the CCDF of a distortion-free (out of the modulator) signal gives us a curve that relates average power—what the power meter measures—to peak power for a given level of symbols at risk.

Thus, if we determine that under the worst case, 1 symbol in every 166,666 sent (1 bit error per symbol error, 10^{-6} BER, 6 bits/symbol) is of concern, we can convert this to percentage of the time (0.6 m%), then go to the CCDF distribution and find that this represents a power about 6.5 dB above the average.

Once a vector range of interest is established, it is possible to go to the AM/PM characteristics of our power amplifier and find the *worst-case* AM/PM allocated in our system budget, count back from the peak, and determine the average power output (what the power meter measures) that will support that BER.

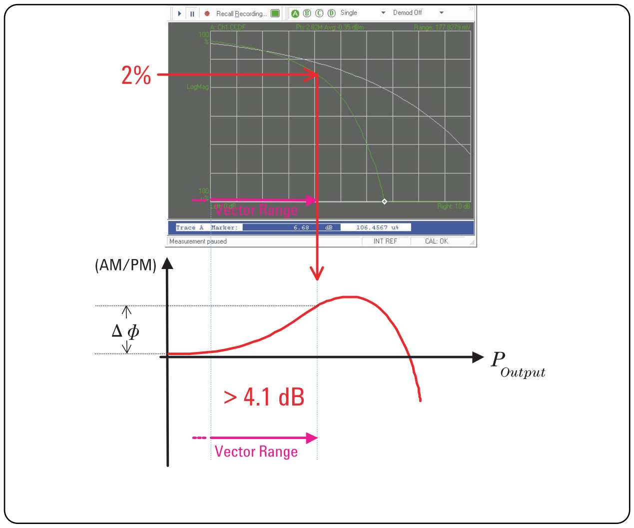


Figure 64: Relating the CCDF Vector Range to AM/PM

Remember: To know if your AM/PM performance is within the budgeted error allowance, you need to know its value *over the vector range*. You can determine both the vector range and the AM/PM phase error directly from the CCDF and the AM/PM curves.

3.2 AM/AM and AM/PM Measurement Techniques

AM to PM distortion is a measure of the undesired phase shifts that occur as a result of increasing signal amplitude. It is expressed in degrees per dB at a specified level. It is important to specify the level at which the distortion occurs, since AM to PM is a non-linear distortion.

As we have seen, CCDF allows us to relate the average power, or that measured by the power meter, to the peak power significant for our application. Thus, it is possible to define the range of power over which AM/AM and AM/PM must be measured. Measurements of AM to PM distortion and system phase noise can be then used to estimate the residual BER rate, as described in Section 1.

To characterize linearity of components, a variety of test configurations are available to suit almost any component, subassembly, assembly, or system. Let's review some of the more common approaches.

3.2.1 Amplifiers

Simple amplifiers or amplifier chains can be connected to a vector network analyzer (VNA). A compatible power meter is essential to properly calibrate the network analyzer. A hidden benefit to using a VNA for device linearity characterization is that it can also characterize S-parameter performance using the same setup, often without even changing connections. This simple configuration is used to test many low-power amplifiers for both linear S-parameter and non-linear AM/PM and AM/AM performance.

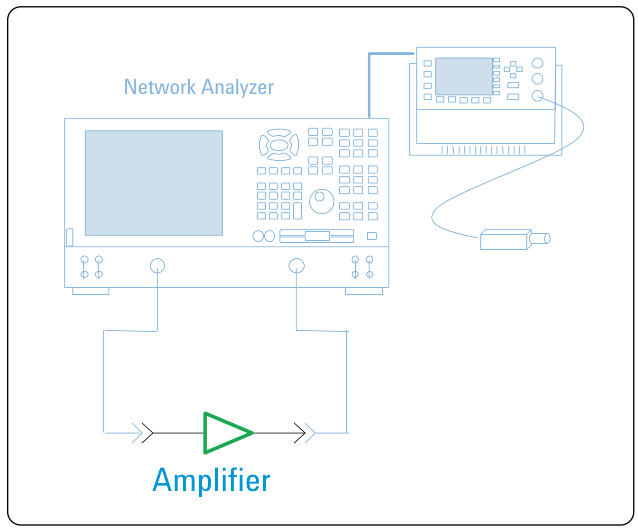


Figure 65: Amplifier Linearity Measurement Setup

3.2.2 Frequency Converter Receivers

The expense and difficulty of high-frequency interconnects often does not allow test connections to be made at intermediate signal points, so testing requirements can demand linearity testing on frequency converters. These devices are different from simple amplifiers in that their input and output frequencies are not identical. This creates a major issue—a vector network analyzer (VNA) must keep track of phase differences between their input and output. To do this, you must have a reference phase path that undergoes a frequency conversion identical to that of the DUT. Nothing changes in the basic measurement, but we must accommodate the frequency conversion.

The receiver downconverter assembly shown in Figure 66 can be characterized by adding an external frequency reference path comprised of a mixer, sideband select filter, and amplifier. Linearity measurements of AM/AM and AM/PM can then be performed to determine overload points, as well as conversion gain and flatness.

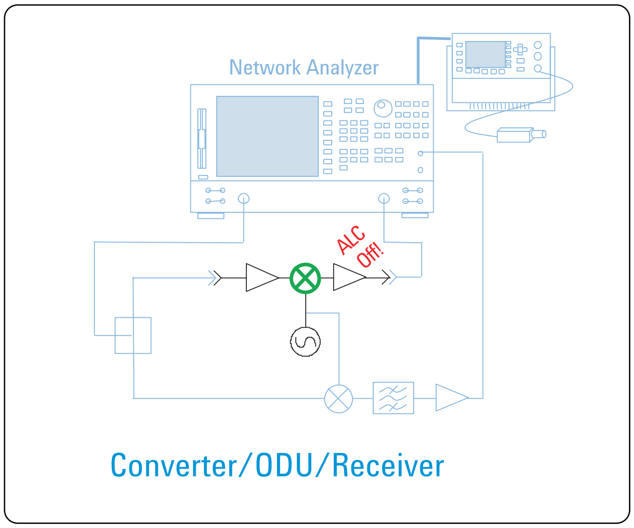


Figure 66: Receiver Linearity Measurement Setup

Note that the reference path does *not* undergo the same change in amplitude as the DUT path. Obviously, the signal level in the reference path must be adjusted so that no AM to PM distortion occurs.

Another important consideration in making receiver measurements on complex modulation systems is ensuring that any automatic leveling control (ALC) is turned off. ALC can alter the power levels, gain, AM/AM , and AM/PM of the DUT. (NOTE: It is possible to measure the ALC range with a vector network analyzer, but this is beyond the scope of this paper.)

3.2.3 Frequency Converter Transmitters

Similar to a receiver, converter measurements can be applied to transmitter or ODU applications. As with receiver applications where frequency conversion is involved, you must create an external phase reference path. This is done using a power splitter, external mixer, filter, and amplifier.

There are two main differences between receiver measurements and transmitter measurements. First, let's take receiver measurements. As with most transmitter measurements, the power levels required to properly drive the upconverter assembly are much higher. An external amplifier must be added when the drive requirements exceed the maximum output power of the network analyzer stimulus port. Note that error correction is still possible if an external coupler is used to sample the reference signal. Care must be taken to assure that the reference phase path remains within its linear range, well away from the point at which AM/PM effects begin.

With transmitter testing, output power levels are often quite high, so the vector network analyzer must be configured for high-power operation by adding isolators.

As in the receiver case, ALC leveling loops should be turned off on the transmitter power amplifier.

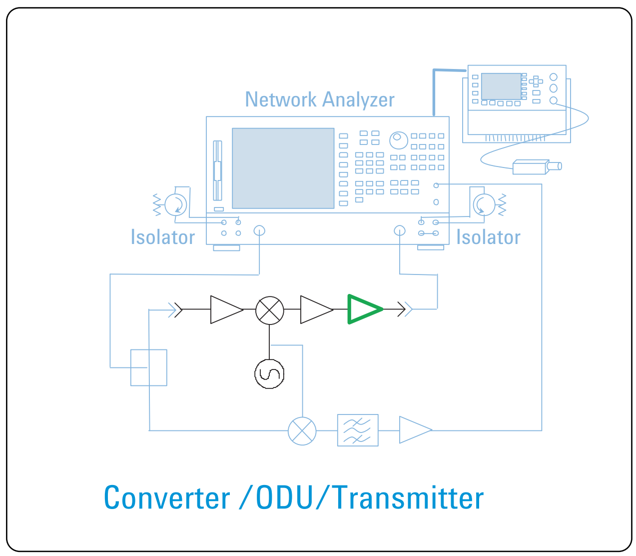


Figure 67: Transmitter Linearity Measurement Setup

At higher frequencies, it can be beneficial to measure the transmitter assembly with the duplexing filter in place. The interaction between this filter and the power amplifier can be very significant at the band edges, even with an isolator in between them! The VNA is the ideal tool for displaying both linear and non-linear metrics, since both can rob performance.

3.3 AM/AM and AM/PM Swept Power Measurements

In Section 1, we mentioned two ways of making AM/PM measurements, the first being swept power measurements. The swept power measurement plots AM/PM vs. increasing PA output power. This measurement is done at a single frequency and is popular with those who manufacture pre-distortion linearizers. The advantage of this type of measurement is that it can characterize a DUT over a wide range of powers. Characterizing a DUT over a wide power range can provide invaluable insight into both the range of possible applications and diagnostic information about bias conditions.

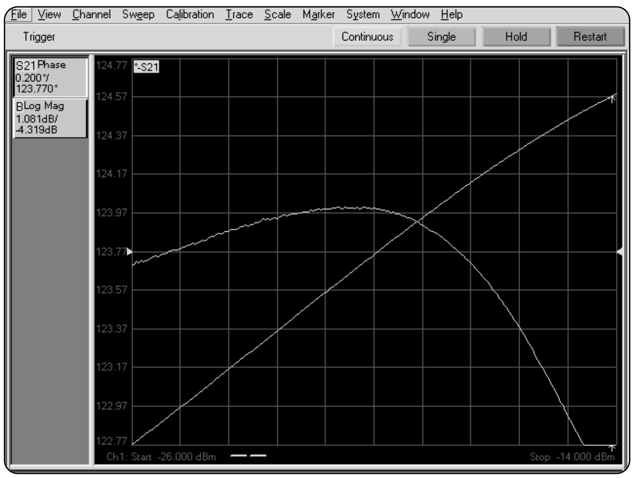


Figure 68: Swept Power AM/PM Measurement

Swept power is often the measurement of choice at the systems level, where entire transmitters or receivers are characterized. The bandwidth limitations of IF filtering eliminates the need to characterize devices over a wide frequency range at the system level.

In top-level systems characterization, system gain (transmitter power – threshold) is paramount and insight can be gained into the maximum possible power. The single-frequency swept power measurement can be done well beyond the intended operating point, offering insight not only into the specified performance, but also the margin available before errors begin. This information is invaluable in choosing a producible operating point for the PA.

The first consideration in making swept power measurements of AM/AM and AM/PM is choosing the appropriate channels and displays for the analyzer. When making AM/PM measurements, the analyzer compares the phase of the reference vs. phase at a received port. Phase is always measured relative to a reference signal—thus the requirement for a vector network analyzer capable of keeping track of phase. Typically, this is identical to an S_{21} measurement, where the analyzer is set to B/R (B channel divided by reference channel). The display format is then set to Phase.

Gain compression, or AM/AM, can be observed in several ways. One popular approach is to use only the B channel receiver and observe the power increasing until saturation is reached. This provides insight into the highest level of output power of which the amplifier is capable. This is an indirect measurement of AM/AM, since it does not directly show the amount of amplitude modulation. To show AM/AM directly, set the analyzer to measure forward gain (S_{21}) during the power sweep and normalize the small signal gain to zero. This method will directly indicate how much the vector length is reduced due to AM/AM. You can choose either log or linear display, depending on the application.

Some important final points to bear in mind when setting up swept power measurements. First, choose the power sweep ranges very carefully. An important consideration is: What power is the DUT capable of handling before damage occurs? Also, avoid using ranges which are close to the analyzer's capability, where power leveling loops may be slow to acquire, since this can create an anomaly at the start of the power sweep that looks like a small step in the AM/PM curve. This can often be eliminated by narrowing the IF bandwidth, which slows the sweep speed, but this also reduces measurement throughput. (NOTE: It is usually best to narrow the IF bandwidth to effect a slower sweep speed, because it reduces the amount of phase noise at the analyzer's detector and thus improves measurement accuracy by reducing trace noise.)

3.4 AM/AM and AM/PM Swept Frequency Converter Measurements

When predicting residual BER, we must consider the *worst-case* operating scenario. Normally, AM to PM distortion is worse in channels near the upper and lower operating frequencies of the DUT, but we must measure each channel to be sure. The best way to do this is by using the swept frequency, single power step method.

The basic technique is to determine AM to PM distortion by comparing the phase response of the DUT at a *low* output power level, to its phase response at a *high* output power level. Since we are often looking for worst-case degradation of signal, this should be done across the DUT's entire operating band. This can include multiple channels if the DUT, for example, is a wide-band PA. The swept power method can be applied at each channel and the results pieced together to provide a total picture of AM/PM response over a wide range of frequencies. However, this will be quite time-consuming if we are looking only for the worst-case AM/PM in the band of interest. Instead, we should consider using the swept frequency, single power step method.

Practical DUTs, such as power amplifiers, typically experience increased AM/PM in the channels nearest their high and low operating frequencies. This is usually because it is difficult to optimally match to the low junction impedance of power devices over a broad band.

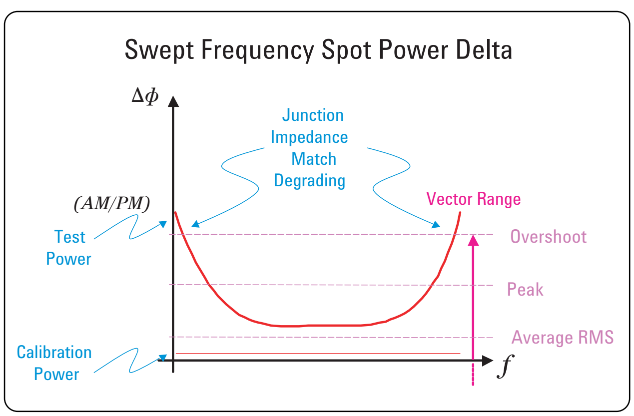


Figure 69: Swept Frequency AM/PM Measurement

When looking for the worst-case AM to PM across the band, we must step the amplitude from some low value with virtually no AM to PM distortion to some high value, usually the approximate phase shift we think we can tolerate. Doing a simple swept power measurement at a mid-range frequency gives us an idea of where to start. The equipment connections for swept frequency AM/PM measurements are essentially the same as those for the swept power technique.

To set up a swept frequency AM/PM measurement, begin by inserting the DUT and attaching a power meter to the vector network analyzer for a power calibration. Adjust the output power of the amplifier to a level where AM/PM is negligible (referring to the swept power AM/PM measurement can be helpful in determining this level).

Next, because the gain of most power amplifiers varies across a wide band and we are interested in AM/PM relative to the DUT's output power, we must do a power calibration. During the power calibration, the analyzer takes control over the power meter and will then adjust the input power level of the DUT to yield a constant output power across the desired frequency band. This can be checked following power calibration by slowing the frequency sweep speed of the vector network analyzer and observing the power level on the output of the DUT—it should be constant.

Like swept power measurements, the swept frequency, stepped power measurement requires the channels to be set to B/R (the same as S_{21}) and the trace format set to Phase for AM/PM and Gain for AM/AM.

Before making the measurement, you should also calibrate the reference phase and gain. This can be done after the DUT is calibrated for power and reconnected to the B port of the analyzer. With the DUT in place, a thorough calibration can be done, "flattening" both the phase and gain curves to zero.

Now the output power can be stepped up to the overshoot trajectory peak power, and the AM/PM and AM/AM observed.

This swept frequency AM/PM curve provides a wealth of information about the DUT's capability relative to frequency. The worst-case AM/PM at any point in the band of interest can be read directly from the curve for a given overshoot power. Because the impedance match of the DUT's output directly affects the current and voltage requirements needed to prevent non-linear clipping, we can determine how well the broadband matching circuits are working relative to AM/PM. We can also determine how well the center frequency of matching and combining networks has been implemented by the center frequency of the minimum AM/PM. In other words, many key design and production parameters that relate directly to residual BER can be evaluated from a single graph!

The associated AM/AM swept frequency, stepped power plot is also very useful. It indicates the amount of AM/AM experienced by the signal vector across the band of interest. Depending how much linearity degradation you decide to accept, it may be necessary to account for the AM/AM effects on the symbol constellation as well.

Another important consideration in AM/AM measurement is accounting for the reduced output power on the AM/PM measurement. If, for example, the DUT is driven hard to saturation, there could be, say, 0.5 dB of AM/AM associated with 5 degrees of AM/PM. Since the network analyzer is stepping the input power to the DUT, if it is 0.5 dB compressed, the output power is actually 0.5 dB less than indicated by the power step in the analyzer. This can be verified by doing a slow sweep with the power meter and compensated for by driving the analyzer an extra 0.5 dB in its power step.

3.5 Troubleshooting Tips

Here are some helpful, time-proven tips for making successful measurements:

1. Calibration: Depending on how an amplifier is integrated into the test system, calibration may be limited. Choose ranges and calibration options carefully from the menus. In any case, calibration becomes especially challenging when the DUT includes frequency conversion.

2. Heating Effects: Ideally, the temperature of the device junction(s) when measured would be identical to their temperature under normal modulated operating conditions. This is not always possible, but acceptable results are often obtained in spite of this limitation.

3. Sourcing and Measuring High Power: Depending on the DUT, you may have to amplify input signals and/or attenuate output signals to fall within the acceptable power ranges of the analyzer.

4. Narrow the Resolution Bandwidth: Typically the AM to PM distortion will be less the 6 degrees. Narrowing the bandwidth will result in less noisy measurements.

5. Verify the Input Range: Power calibration on most network analyzers can operate only over a single power range. Make sure you are in a range that can output both the low and high input power required. Note that source power ranges normally overlap considerably; see your network analyzer user's guide for more information.

6. Verify Output Power Flatness: If the output power in the B receiver isn't flat due to AM/AM, the resulting AM to PM will be correspondingly off. Adjust the input power as needed to flatten the output power, either through repeating a power cal or through manual correction of the leveling.

7. Verify Levels: It is wise to make sure the output signal will not cause AM to PM distortion in the B receiver of the network analyzer. An attenuator may be required on the output of the amplifier to lower the signal to an acceptable level.

8. Automation: You may be able to mitigate some of the heating problems by running the device at the average output power and then quickly switching the level for each measurement (see following sections).

9. User's Guide: Refer to the user's guide of your vector network analyzer for detailed measurement procedures. Capabilities and options of instruments can vary significantly and a user's guide can help keep you from wasting time in setting up measurements.

3.6 High Power Measurement Setup

Modern VNAs can be configured with the addition of a driver amplifier, coupler, and isolators to handle high-power signals directly, while still retaining the ability to make very accurate S-parameter measurements, particularly S_{22}. Again, this provides both linear and nonlinear characterization on a single test instrument.

Inside the network analyzer are a pair of directional couplers, as well as a set of detectors, which actually measure the reflected and transmitted power on each test port. It is possible to boost and absorb the signal before and after (respectively) the couplers. This allows you to stimulate the DUT at higher power levels than the basic VNA can handle, or to absorb stronger signals than the basic VNA can handle, while maintaining the S-parameter accuracy of the instrument. Placing the driver amplifier before the stimulus coupler allows the VNA to measure high-power amplifier input parameters. Likewise, the addition of isolators can absorb high power levels while still allowing direct analyzer connections to the high-power output and retaining S_{22} measurement accuracy. (Adding a high-power attenuator directly on the output of the PA would prevent measurement of S_{22}.)

The addition of an external coupler after the driver amplifier allows the internal reference signal to measure variations in flatness and drive level of the external driver amplifier.

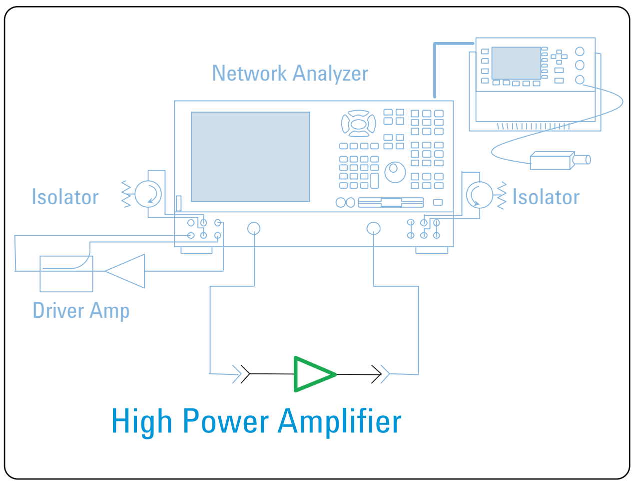


Figure 70: High Power Amplifier Linearity Measurement Setup

The ports necessary to gain instrument access for these high-power measurements are typically available as an option to the basic instrument.

3.7 Advantages of a VNA for Component Linearity Characterization

A vector network analyzer (VNA) has several advantages over other instruments for making linearity measurements. These include simultaneous capture of S-parameters, characterization over a band of frequencies, fast data capture, the ability to measure linearity with frequency conversion, multi-term error correction for accuracy, and low cost relative to other instrument approaches.

In a test environment, it is sometimes necessary to control the impedance match as well as the linearity of power amplifiers and their components. Transferring maximum power from the PA to the antenna is critical to being able to transmit the highest power with the least amount of AM/PM distortion. Amplifier impedance and linearity may be measured simultaneously on a VNA, eliminating the need to switch out test gear. This allows you to characterize the PA quickly and completely, with fewer items of capital equipment.

A VNA is also suited to measuring many broadband and multi-channel devices, using its high sweep rate to characterize many MHz per millisecond. Complete S-parameter and linearity characterization can be done quickly, and with a minimum number of test stations, in medium- to high-volume production environments.

A VNA also has the best error correction and calibration for S-parameter measurement accuracy. Its sophisticated, multi-term error correction and wide range of precision calibration kits are helpful in characterizing systems for which antenna match is critical. VNAs are *not* well suited for applications in which substantial temperature change takes place in a solid-state junction or for characterizing average AM/PM across a frequency band.

In a Class A amplifier, dc power is converted to heat in the absence of an RF signal. When an RF signal is applied, more RF power is transmitted from the output of the amplifier (i.e., the PA has gain). The energy of the output signal is derived from the dc input to the amplifier. Thus, in a Class A amplifier, some of the heat that would normally be dissipated in the junction leaves the device as RF energy. This means that there is actually *less* dissipation in the junction when an RF signal is applied, which results in a cooler junction temperature!

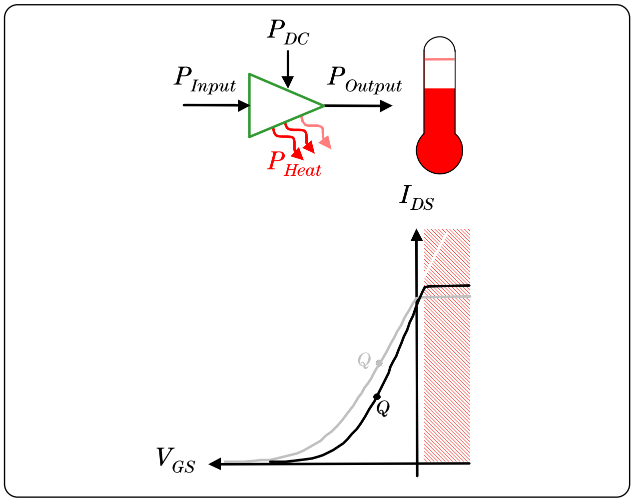


Figure 71: RF Output Power and FET Thermal Effects

So exactly how does temperature affect the linearity of an amplifier? In the case of FET amplifiers, lower channel temperature increases the forward small signal transconductance (g_m), which generally means higher gain for the part. Temperature affects the AM/AM, AM/PM, and group delay distortion of an amplifier, so it is essential to test these linearity metrics using accurate thermal conditions.

Inaccuracies arise with a VNA because it uses a carrier wave (CW) signal that does not have a power statistic matching the CCDF of the real modulated signal. The thermal effect can often be mitigated by setting the VNA to a CW power slightly above the average power before the measurement is taken. In effect, instead of blanking the signal between sweeps, an "idling power" is chosen to minimize temperature changes in the junction. Then the power sweep, or power step, can be made quickly to avoid further temperature change in the junction.

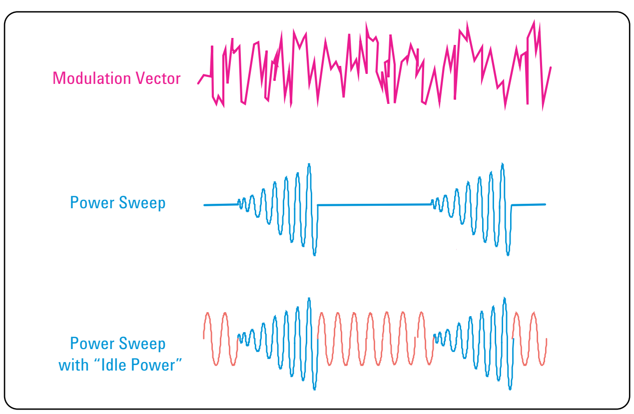


Figure 72: "Idling Power" Approximation of the Modulated CCDF

For example, imagine a Class A RF amplifier consuming up to 50 watts of dc power and transmitting 10 watts of RF energy. In a swept frequency, stepped power measurement, the junction temperature can change significantly, since 10 watts of power is removed from an area of only a few square microns! It is easy to detect this type of problem because the measurement stability is poor and AM/PM distortion begins to "creep" with successive sweeps of the analyzer as the junction temperature changes. This can be mitigated by setting the amplifier to its normal operating power, allowing the junction temperature to stabilize, then making the AM/PM measurement as quickly as possible so the junction temperature does not change appreciably.

Always consider the efficiency of the power amplifier when choosing your measurement technique. If it is a single-stage PA with most of the energy dissipated in the device (i.e., 2% efficient), significant temperature changes will probably *not* occur between modulated and unmodulated signals. On the other hand, if the amplifier stage is 30% efficient, considerable energy will leave the device as a modulated signal is applied. Thermal changes in the device will then cause the AM/PM response to change as the temperature of the junction changes. If you see the measurement continuing to change after a power step, junction temperature is probably the cause. Consider using an "idling" power level to pre-cool the junction before making a measurement sweep, then making the sweep as quickly as possible, as discussed above. In most cases, quick measurements can mitigate this effect because the junction's thermal time constants are sufficiently long. If they cannot be mitigated, consider using the complex stimulus-response technique (see next section).

Another area in which the VNA's continuous wave limitations cause measurement problems is characterization of average AM/PM across a band (used in pre-distortion tests). In some cases (feed-forward linearizers in particular), you will need to measure the average AM/PM across a frequency band to derive pre-distortion seed coefficients. Unlike residual BER analysis, in which *worst*-case AM/PM information is usually sought, applications such as linearizers in which average AM/PM corrections are needed across a wide band of interest require a different method of characterization, currently not possible with a vector network analyzer. For these applications, or those in which your goal is to accurately determine the precise residual BER (*not* worst-case), consider using the complex stimulus-response approach for characterizing AM/AM and AM/PM.

3.8 Complex Stimulus-Response

As we have seen, most VNAs have a relatively slow power sweep (or step), which can be significant in high-power devices. At times, it is impossible to avoid the thermal effects of modulation at the junction and testing linearity with an accurate modulation becomes a requirement. Using a vector signal generator (VSG) to measure AM/PM by creating a signal with the identical CCDF distribution as the actual QAM signal can improve measurement accuracy for some devices and applications. This is done in conjunction with a vector signal analyzer (VSA) that compares the amplifier's input and output signals to the reference signal. The vector signal generator and vector signal analyzer both have the advantage of substantial measurement bandwidth and the ability to provide a signal of virtually any CCDF.

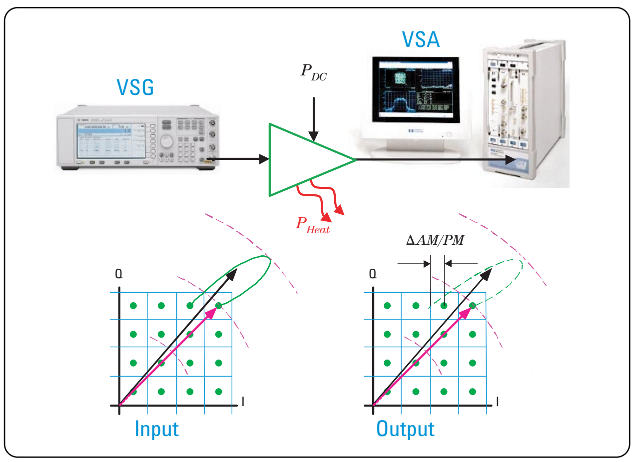


Figure 73: Complex Stimulus-Response AM/PM Testing

An identical CCDF allows the junction to be at a precise operating temperature continuously, eliminating thermally-related inaccuracy in vector network analyzer measurements. Likewise, the wideband nature of the modulated signal with an identical CCDF enables an accurate average of the AM/PM results over a modulated band—a key step in moving from worst-case to actual BER prediction.

The complex stimulus-response approach to measuring AM/PM distortion is still relatively young. At present, the majority of applications are not designed for residual BER prediction, but rather linearizer applications of mask-limited, multi-carrier PAs.

Using complex stimulus from the VSG and analyzing the response for linearity in a VSA has some significant drawbacks that should be considered carefully before taking this approach. First, it is by nature slower. Second, it lacks the inherent resolution of a vector network analyzer (VNA).

Since modulated signals rarely reach their peak amplitudes, a huge number of vector sample points may be needed to cover the signal's full dynamic range of interest. This can take a very long time for high-QoS signals if you have to characterize the 1-in-1-millionth symbol! To get a statistically significant AM/PM measurement, many millions of symbols may need to be compared in phase before and after the DUT. Depending on the symbol rate, test time could be quite long!

Complex stimulus-response also provides no information beyond measurement power—unlike swept-power VNA measurements. To determine how much more power a PA is capable of, another test must be run at the next higher power.

The measurement equipment costs for complex stimulus-response are still considerably more than for the vector network analyzer approach.

In some cases (e.g., a multi-channel linearizer), the ability of complex stimulus-response to characterize a device under actual use conditions may outweigh these drawbacks.

SECTION 4

Conclusion & Review

In concluding, let us review some of the more salient points in this application note. First and most importantly, remember that the overall cost of a radio is a strong function of the residual BER budget. Hence residual BER should have a prominent place in your overall system design. Otherwise, it is simply not possible to properly and completely diagnose sources of error—too many misallocated or unknown causes may result, along with loss of product cost control.

Residual BER can be predicted using the following measurements:

1. Integrated Phase Noise Measurements

- Using a signal or spectrum analyzer and the direct method of measurement, local oscillators (LOs), transmitters, and frequency converters can be characterized
- Using VCO test sets or vector signal analyzers (VSAs) and the demodulation method, voltage controlled oscillators (VCOs) can be characterized
- Using a phase noise test set and the demodulation method, measurements of crystals and very high-performance sources can be made

2. Signal Analyzer Measurements of CCDF

- Using a signal analyzer to determine the appropriate range of modulation vectors to consider for complex, noise-like modulations

3. AM/AM and AM/PM Distortion Measurements

- Using a vector signal analyzer configured for a wide range of devices, including amplifiers, converters, ODUs, and high-power devices
- Using complex stimulus-response measurement setups in special cases where thermal effects or AM/PM characteristics are sought

We hope that this application note has helped you to become more aware of and interested in the key role played by residual BER in radio system design and testing. For additional information, see Appendix B, References and Recommended Reading

Appendix A

Agilent Phase Noise and Linearity Measurement Equipment

Agilent Technologies offers a complete line of phase noise and linearity measurement equipment designed to help determine residual BER. Included below are typical products, detailed tables, and brief descriptions of features and benefits (pg. 50). (NOTE: There may be overlap between various products and functions.)

A.1 Agilent Phase Noise Measurement Equipment

Figure 74 is an overview of Agilent instruments for measuring phase noise, followed by a detailed breakout in Table 1.

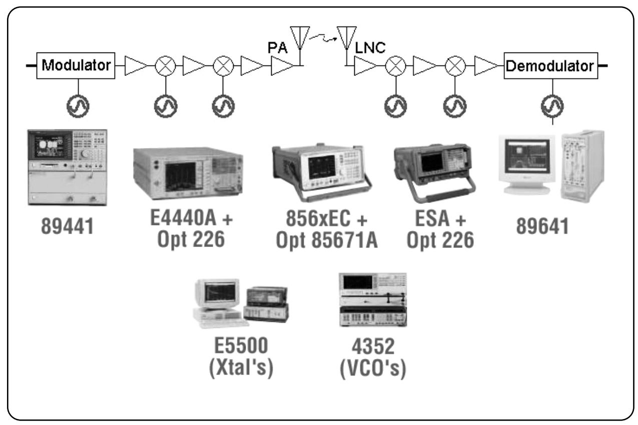


Figure 74: Popular Agilent Equipment for Measuring Phase Noise

Table 1:
Agilent Phase Noise Measurement Equipment and Selection Criteria

	High Performance			Mid Performance			Economy
							
Product Families	**Dedicated Phase Noise Testers** **E550xA/B Series**	**Spectrum Analyzers** **PSA Series***	**Vector Signal Analyzers** **896xxA Series***	**VCO Test Set**	**Rugged Portable Spectrum Analyzers** **856xEC Series***	**Vector Signal Analyzers** **894xx Series***	**Spectrum Analyzers** **ESA-E Series***
Model Numbers Frequency Ranges	**E5501A/B** 50 kHz – 1.6 GHz **E5502A/B** 50 kHz – 6.0 GHz **E5503A/B** 50 kHz – 18.0 GHz **E5504A/B** 50 kHz – 26.5 GHz	**E4440A** 3 Hz–26.5 GHz **E4443A** 3 Hz–6.7 GHz **E4445A** 3 Hz–13.2 GHz **E4446A** 3 Hz–44 GHz **E4448A** 3 Hz–50 GHz	**89610A** dc–40 MHz **89640A** dc–2.7 GHz **89641A** dc–6.0 GHz	**4352S** 10 MHz–3.0 GHz **+ 43521A** 10 MHz–12.6 GHz	**8560EC** 30 Hz–2.9 GHz **8561EC** 30 Hz–6.5 GHz **8562EC** 30 Hz–13.2 GHz **8563EC** 30 Hz–26.5 GHz **8564EC** 30 Hz–40 GHz **8565EC** 30 Hz–50 GHz	**89410A** dc–10 MHz **89441A** dc–2.65 GHz	**E4401B** 30 Hz–1.5 GHz **E4402B** 30 Hz–3.0 GHz **E4404B** 30 Hz–6.7 GHz **E4405B** 30 Hz–13.2 GHz **E4407B** 30 Hz–26.5 GHz
Key Selection Criteria							
Technique	Carrier Removal/ Demodulation -160 dBc/Hz @ >32 kHz (Refer to E5500 Cofiguration Guide)	Direct Spectrum -114 dBc/Hz @ 10 kHz, 1 GHz -120 dBc/Hz @ 100 kHz, 1 GHz -144 dBc/Hz @ 1 MHz, 1 GHz	Direct Spectrum Phase Detector -110 dBc/Hz @ 100 kHz, 1 GHz	Carrier Removal/ Demodulation -130 dBc/Hz @ 10 kHz, 1 GHz -140 dBc/Hz @ 100 kHz, 1 GHz -150 dBc/Hz @ 1 MHz, 1 GHz	Direct Spectrum -113 dBc/Hz @ >10 kHz, 1 GHz -117 dBc/Hz @ 100 kHz, 1 GHz	Carrier Removal/ Demodulation -116 dBc/Hz @ 10 kHz, 1 GHz	Direct Spectrum -120 dBc/Hz @ 10 kHz, 1 GHz -110 dBc/Hz @ 100 kHz, 1 GHz
Measurements							
Crystal Reference	●						
Voltage Controlled Oscillators	●		●	●		●	
Amplifiers & Frequency Multipliers	●	●			●		●
Frequency Synthesizers	●	●			●		●
Frequency Converter Subassemblies	●	●			●		●
Transmitters	●	●	●		●	●	●
Receivers			●			●	
Modulators	●	●	●		●	●	●
Demodulators			●			●	

** These analyzers are capable of measuring CCDF within their bandwidth capability.*

A. 2
Agilent Linearity Measurement Equipment

Figure 75 is an overview of Agilent instruments for measuring linearity distortion, followed by a detailed breakout in Table 2.

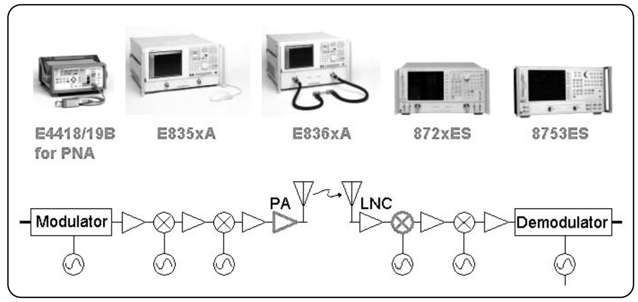
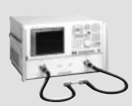


Figure 75: Popular Agilent Equipment for Measuring Linearity

Table 2: Agilent Linearity Measurement Equipment and Selection Criteria

	High Performance		Mid Performance		Compatible Power Meters	Complex Stimulus	Complex Response
							
Product Families	**Vector Network Analyzers** **PNA Series**	**Vector Network Analyzers** **PNA Series**	**Vector Network Analyzer**	**Vector Network Analyzer**	**Power Meters Meters** **EPM, EPM-P Series**	**Vector Signal Source** **E4438C**	**Vector Signal Analyzers** **896xxA Series**
Model Numbers Frequency Ranges	**E8362A** 45 MHz–20 GHz **E8363A** 45 MHz–40 GHz **E8364A** 45 MHz–50 GHz	**E8356A** 300 kHz–3 GHz **E8357A** 300 kHz–6 GHz **E8358A** 300 kHz–9 GHz	**8719ES/ET** 50 MHz–13.5 GHz **8720ES/ET** 50 MHz–20 GHz **8722ES/ET** 50 MHz–40 GHz	**8753ES/ET** 30 kHz–3.0 or 6.0 GHz	**E4418B** 9 kHz–110 GHz **E4419B** 9 kHz–110 GHz	**E4438C-501** 250kHz–1 GHz **E4438C-502** 250kHz–2 GHz **E4438C-503** 250kHz–3 GHz **E4438C-504** 250kHz–4 GHz **E4438C-506** 250kHz–6 GHz	**89610A** dc–40 MHz **89640A** dc–2.7 GHz **89641A** dc–6.0 GHz
Key Selection Criteria							
Technique	Sinewave Stimulus	Sinewave Stimulus	Sinewave Stimulus	Sinewave Stimulus	Sinewave Stimulus	Modulated Stimulus	Modulated Stimulus
Typical Sweep Speed	<26µs/point	<35µs/point	264 µS/point	264 µS/point	20 Msa/s	NA	NA
Measurements							
Amplifiers	●	●	●	●	●	●	●
Converters			●	●	●	●	●

A. 3
Agilent CCDF Measurement Equipment

Table 3 details Agilent equipment for measuring complementary cumulative distribution function (CCDF):

Table 3: Agilent CCDF Measurement Equipment and Selection Criteria

	High Performance			Mid Performance		Economy
						
Product Families	**Spectrum Analyzer** **E444xA Series**	**Infiniium 54800 Series + 89601A Vector Signal Analyzer Software**	**Vector Signal Analyzer** **896xxA Series**	**E4406A VSA Series Transmitter Tester**	**894xxA Vector Signal Analyzer**	**Spectrum Analyzer** **E440xB Series**
Model Numbers Frequency Ranges	**E4440A** 3 Hz–26.5 GHz **E4443A** 3 Hz–6.7 GHz **E4445A** 3 Hz–13.2 GHz **E4446A** 3 Hz–44 GHz **E4448A** 3 Hz–50 GHz	**54830B** 600 MHz (BW) **54831B** 600 MHz (BW) **54832B** 1.0 GHz (BW) **54845B** 1.5 GHz (BW) **54846B** 2.25 GHz (BW)	**89610A** dc–40 MHz **89640A** dc–2.7 GHz **89641A** dc–6.0 GHz	**E4406A** 7 MHz–4.0 GHz	**89410A** dc–10 MHz **89441A** dc–2.65 GHz	**E4401B** 30 Hz–1.5 GHz **E4402B** 30 Hz–3.0 GHz **E4404B** 30 Hz–6.7 GHz **E4405B** 30 Hz–13.2 GHz **E4407B** 30 Hz–26.5 GHz
Key Selection Criteria						
Technique	Scalar	Vector	Vector	Vector	Vector	Scalar
Bandwidth	8 MHz	500+MHz	36 MHz	8 MHz	10 MHz	8 MHz

A.4
Product Descriptions

NOTE: Products appearing in the above tables are described below in alphanumeric order. For more information, go to our website at **www.agilent.com**, select "Test and Measurement Instruments," and search for the appropriate product.

**E4401/2/4/5/7B ESA-E Spectrum Analyzers,
30 Hz to 26.5 GHz**
Fast, accurate, portable spectrum analyzers covering a wide frequency range, with 4 ms sweeps and 30 measurement/sec updates. Software link allows operation with 89601A Vector Signal Analysis Software (see below).

**E4406A VSA Series Transmitter Tester,
7 MHz to 4 GHz**
Full-featured vector signal analyzer designed to meet the test needs of wireless communications for R&D and manufacturing. Leads the wireless test market in demodulation for standards-based, one-button measurements. Supports multiple wireless standards, providing a flexible, expandable platform.

**E4418B/9B EPM, EPM-P Power Meters ,
9 kHz to 110 GHz**
Low-cost, high-performance programmable power meters with single and dual channels, respectively. Measures from –70 dBm to +44 dBm and 9 kHz to 110 GHz, taking up to 100 (dual-channel) or 200 (single-channel) readings per second.

Agilent E4438C ESG Vector Signal Source
Offers a broad array of capabilities for evaluating performance of communication systems to meet the requirements of nearly all popular wireless interface standards. Test signals can also be customized to cover proprietary and other non-standard modulations. Available in 1, 2, 3, 4, and 6 GHz models.

**E4440/3/5/6/8A PSA Performance Spectrum Analyzers ,
3 Hz to 50 GHz**
Modern, high-performance spectrum analyzers with one-button measurements and several measurement personalities (including phase noise) with digital demodulation. A market leader in speed, accuracy, and coverage.

**E5501/2/3/4A/B Phase Noise Measurement Solutions,
50 kHz to 26.5 GHz**

Phase noise solutions designed to minimize test times for one-port VCOs, DROs, crystal oscillators, and synthesizers. Fastest phase noise measurements available plus direct measurement of AM noise.

**E8362/3/4A PNA Microwave Network Analyzers,
300 kHz to 50 GHz**
Combines speed and precision with wide coverage and low trace noise. Features 128 dB dynamic range at test point (143 dB with direct receiver access), 35µs/point sweep speed, and ample 256Mb RAM.

**4352S VCO/PLL Signal Test System,
10 MHz-3.0 GHz**
Evaluates characteristics of VCOs and PLLs, using both signal analyzer and VCO tester modes. Measures RF power, frequency, phase noise, spectrum, frequency transient, DC consumption current, and FM deviation at up to 3.0 GHz and, with 43521A Downconverter Unit (see below), up to 12.6 GHz.

**Agilent 43521A Downconverter Unit,
10 MHz to 12.6 GHz**
Designed to operate with the 4352S VCO/PLL signal test system for test applications at frequencies above 3 GHz. The enhanced 4352S is a complete system with wider range for measuring phase noise, RF power, transients, settling time, and many other parameters required for VCO/PLL evaluations.

**8356/7/8 PNA Series RF Vector Network Analyzers ,
300 kHz to 9 GHz**
Exceptional coverage, dynamic range (123 dB at test point, 138 dB with direct receiver access), high sweep speed (35µs/point) and ample RAM (256Mb). Powerful automation and self-help tools serve both R&D and manufacturing uses.

**8560/1/2/3/4/5EC Spectrum Analyzers,
30 Hz to 50 GHz**
Portable SAs with the color display, measurement capability, and performance of larger, more expensive benchtop analyzers, with digital resolution bandwidths (1-100 Hz) and low phase noise.

8719/20/22ES/ET Vector Network Analyzers, 50 MHz to 40 GHz
Complete, full-featured characterization of RF and microwave components, including magnitude and phase measurements. IntuiLink software connects VNAs to PC applications for fast, easy data analysis and display.

8753ES/ET Vector Network Analyzers, 30 kHz to 6 GHz
Complete characterization of RF components, including magnitude and phase measurements. IntuiLink software connects VNAs to PC applications for fast, easy data analysis and display. Available in 50Ω and 75Ω models.

Infiniium 54800 Series Oscilloscopes, 600 MHz-2.25 GHz
Combine high performance of digital technology with simple look and feel of analog scopes. Either 2 to 4 channels and up to 4 or 8 GSa/sec sample rates, respectively. Voice-control option (English only) for hands-free operation. Superior probing with bandwidth up to 4 GHz. Use with 89401A Vector Signal Analysis Software (see below).

89601A Vector Signal Analysis Software
Wide range of precision troubleshooting tools for analyzing digital satellite video, LMDS, 802.11b wireless LAN, 802.11a and HiperLAN/2 OFDM wireless LAN, *Bluetooth*™ systems, and more. Works with 89600, PSA, and ESA-E series spectrum analyzers, Infiniium 54800 Series oscilloscopes, and E4406A VSA transmitter tester.

89410A Vector Signal Analyzer, dc-10 MHz
Flexible demodulation for troubleshooting designs, ranging from dc to 10 MHz. Performs 8 MHz signal analysis in RF, or 10 MHz at baseband. Integrated, coherent source allows translated frequency response readings.

89441A Vector Signal Analyzer, dc-2.65 GHz
Companion to the 89410A, for flexible demodulation in troubleshooting designs, but with a wider range. Integrated, coherent source allows translated frequency response readings.

89610A Vector Signal Analyzer, dc-40 MHz
Signal, modulation, and error analysis to identify impairments in digital radios. Wide bandwidth (38 MHz at RF, 39 MHz baseband) and coverage (dc-40 MHz) make it ideal for RF/satellite applications. Custom units also available.

89640A Vector Signal Analyzer, dc-2.7 GHz
Flexible demodulation with adjustable parameters, 36 MHz bandwidth. Links to Agilent's Advanced Design System (ADS) for powerful design, simulation, and analysis tools. Analyzes measurements from 54800 Series Infiniium oscilloscopes, plus ESA-E Series, PSA Series, and E4406A signal analyzers.

89641A Vector Signal Analyzer, dc-6.0 GHz
Companion to the 89610A with a wider range. Links to Agilent's Advanced Design System (ADS) for powerful design, simulation, and analysis tools. Analyzes measurements from 54800 Series Infiniium oscilloscopes, plus ESA-E Series, PSA Series, and E4406A signal analyzers.

Appendix B

References and Recommended Reading

B.1 Agilent Application Notes and Technical Papers

1. Agilent Application Note AN 57-1, *Fundamentals of RF and Microwave Noise Figure Measurement* (lit. no. 5952-8255E)

2. Agilent Application Note AN 57-3, *10 Hints for Making Successful Noise Figure Measurements* (lit. no. 5980-0288EN)

3. Agilent Application Note AN 150, *Agilent Spectrum Analysis Basics* (lit. no. 5952-0292)

4. Agilent Application Note AN 154, *S-parameter Design* (lit. no. 5952-1087)

5. Agilent Application Note AN 355-1, *Tools for Digital Microwave Radio Installation and Maintenance* (lit. no. 5962-9920E)

6. Agilent Application Note AN 1287-7, *Improving Network Analyzer Measurements of Frequency-translating Devices* (lit. no. 5966-3318E)

7. Agilent Application Note AN 1298, *Digital Modulation in Communication Systems–An Introduction* (lit. no. 5965-7160E)

8. Agilent Application Note AN 1303, *Spectrum Analyzer Measurements and Noise* (lit. no. 5965-7160E)

9. Agilent Application Note AN 1313, *Testing and Troubleshooting Digital RF Communications Transmitter Designs* (lit. no. 5968-3578E)

10. Agilent Application Note AN 1314, *Testing and Troubleshooting Digital RF Communications Receiver Designs* (lit. no. 5968-3579E)

11. Agilent Application Note AN 1354, *Practical Noise-figure Measurements and Analysis for Low-noise Amplifier Designs* (lit. no. 5980-1916E)

12. Agilent Product Note PN E5500-1, *Pulsed Carrier Phase Noise Measurements* (lit. no. 5968-5662E)

13. Agilent Product Note PN 89400-14A, *10 Steps to a Perfect Digital Demodulation Measurement* (lit. no. 5966-0444E)

14. Agilent Product Note PN 89400-2, *Measuring Phase Noise with the Agilent 89400 Series Vector Signal Analyzers* (lit. no. 5091-7193E)

15. Kent K. Johnson, RF Engineer, Agilent Technologies, "Optimizing Link Performance, Cost, and Interchangeability by Predicting Residual BER, Parts I and II," *Microwave Journal,* July-August 2002

16. Agilent Color Poster, *Essential Tests & Tips for Better Transmitter/Receiver Performance,* (lit. no. 5988-6041EN)

For more information or to download free copies of the above, go to **www.agilent.com/find/rxtx.** To receive a free copy of the color poster (item #16) by mail, call **1-800-452-4844.**

B.2 Webcasts

Agilent offers a variety of live and archived webcasts that provide valuable information on BER-sensitive applications:

1. Predicting Residual BER (Theory and Calculations)

2. Predicting Residual BER II (Theory and Applications)

3. Learn RF Signal Source Basics

4. Learn RF Spectrum Analysis Basics

For more information and scheduling, go to **www.agilent.com/find/events.**

B.3 Training

Agilent's training classes use state-of-the-art Agilent equipment and techniques designed to help you make the most of your test systems and applications. The following classes involving BER measurements are currently available:

1. RF and Microwave Fundamentals

2. RF Measurement Basics

3. Introduction to Digital RF Communications

4. Digital Microwave Radio Basics

5. Spectrum Analyzer Measurements

For information and scheduling, go to **www.agilent.com/find/rxtx.**

Appendix C

Symbols and Acronyms

ACP	Adjacent Channel Power
ADC	Analog-to-Digital Converter
ALC	Automatic Leveling Control
AM/AM	Amplitude Modulation to Amplitude Modulation
AM/PM	Amplitude Modulation to Phase Modulation
AoS	Availability of Service
ASIC	Application Specific Integrated Circuit
B/R	B Channel Divided by Reference channel
BER	Bit Error Rate
BW	Bandwidth
BWA	Broadband Wireless Access
CCDF	Complementary Cumulative Distribution Function
CDU	Combiner/Divider Unit
C/N	Carrier-to-Noise Ratio
CPE	Customer Premises Equipment
CW	Carrier Wave
dc	Direct Current
$\Delta\Phi$	Angular Error
$\Delta\Phi_{Max}$	Maximum Angular Error
$\Delta\Phi_{RMS}$	Angular Error in One Sigma (s)
dB	Decibels
dBc	Decibels Below the Carrier
dBm	Decibels Relative to .001 Watt
dBrad	Decibels of Angular Rotation in Radians
DRO	Dielectric Resonant Oscillator
DUT	Device Under Test
EVM	Error Vector Magnitude
Φ_{iq}	Angle to Symbol Point
Φ_{B1iq}/Φ_{B2iq}	Angle to Symbol Boundaries
f	Frequency
FEC	Forward Error Correction
FET	Field Effect Transistor
FTD	Frequency Translation Device
FPGA	Field Programmable Gate Array
g_m	Forward Small Signal Transconductance
I	Current/In Phase with Carrier
IMD	Intermodulation Distortion
ISI	Intersymbol Interference
K	Boltzmann's Constant
$\mathscr{L}(f_o)$	Frequency Offset
LO	Local Oscillator
GaAs MESFET	Gallium Arsenide Metal Semiconductor Field Effect Transistor
NF	Noise Figure
NFA	Noise Figure Analyzer
ODU	Outdoor Unit
OFDM	Orthogonal Frequency Division Multiplexing
P_{iq}	Probability of Symbol Error
P_{Input}	Input Power
P_{Output}	Output Power
PA	Power Amplifier
PDF	Probability Density Function
PLL	Phase Locked Loop
PNA	Portable Network Analyzer
RF	Radio Frequency
RMS	Root-Mean-Square
RS	Received Signal
RSS/RSL	Received Signal Strength/ Received Signal Level
Q	Quiescent Bias Point or Quadrature Phase
QPSK	Quadrature Phase Shift Keying
QoS	Quality of Service
QAM	Quadrature Amplitude Modulation
σ	Standard Deviation
σ_{iq}	Standard Deviation to Symbol Boundary
SA	Spectrum Analyzer
SPC	Statistical Process Control
SRD	Step Recovery Diode
T	Temperature
TOI	Third Order Interference
VNA	Vector Network Analyzer
VSA	Vector Signal Analyzer
VSG	Vector Signal Generator

Agilent Technologies' Test and Measurement Support, Services, and Assistance

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