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What are Embedded - System On Chip (SoC)?

System On Chip (SoC) integrates multiple functions of a computer or electronic system onto a single chip. Unlike traditional multi-chip solutions, SoCs combine a central

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Interestingly, it is not unusual to have a reported low downstream or upstream RxMER number, yet find the measured CNR and signal levels to be just fine. Why? Because one or more impairments that cannot be seen on a spectrum analyzer—poor in-channel frequency response, including group delay variation and micro-reflections, and even upstream data collisions—may be the cause of the low reported RxMER.

This paper discusses the terms listed in Table 1.

Table 1: Terminology for Various SNR Ratio Concepts

Abbreviation	Term	Definition	Units
CNR (or C/N)	Carrier-to-noise ratio	The ratio of carrier or signal power to the white-noise power in a specified bandwidth, as measured on an RF spectrum analyzer or similar equipment.	dB
C/N_0	Carrier-to-noise - density ratio	The ratio of carrier or signal power to white-noise spectral density.	dB-Hz
CNIR (or C/(N+I))	Carrier-to-noise-plus-interference ratio	The ratio of carrier or signal power to the total noise power (including white noise and interference) in a specified bandwidth, as measured on an RF spectrum analyzer or similar equipment.	dB
E_S/N_0	Energy-per-symbol to noise-density ratio	In digital modulation, the ratio of the average energy of a QAM symbol to the white-noise spectral density.	dB
EVM	Error vector magnitude	The ratio of RMS constellation error magnitude to peak constellation symbol magnitude.	Percent
MER	Modulation error ratio	The ratio of average signal constellation power to average constellation error power.	dB
RxMER	Receive modulation error ratio	The MER as measured in a digital receiver after demodulation, with or without adaptive equalization.	dB
SNR (or S/N)	Signal-to-noise ratio	(a) A general measurement of the ratio of signal power to noise power. (b) In a specific context, a measurement of the ratio of signal power to noise power made at baseband before modulation or after detection or demodulation.	dB
TxMER	Transmit modulation error ratio	The MER produced by a transmitter under test, as measured by an ideal test receiver.	dB

CNR and SNR from a Telecommunications Industry Perspective

Some of the confusion mentioned in the introduction arises from the fact that in the world of telecommunications outside the cable industry, the terms SNR and CNR are often used interchangeably. According to Roger L. Freeman's *Telecommunications Transmission Handbook*, "The signal-to-noise ratio expresses in decibels the amount by which a signal level exceeds its corresponding noise." Another reference, Tektronix's *Measuring Noise in Video Systems*, says "In the most general case, SNR is expressed as the ratio of RMS (root mean square) signal level, S_{RMS}, to the RMS noise, N_{RMS}, ($SNR = S_{RMS}/N_{RMS}$)." [Eq. 1]

Both of the previous SNR definitions can easily be applied to RF CNR measurements (after all, a carrier is a “signal”) as well as baseband SNR measurements (baseband video and audio are “signals,” too). If the specific measurement is not clearly defined, it is difficult to know whether SNR refers to a baseband or RF parameter. This paper distinguishes between SNR and CNR. In the subsequent sections, each term is defined and explained, and the distinction is illustrated following usage in the cable industry.

CNR and SNR from a Cable Industry Perspective

Modern Cable Television Technology, 2nd Ed., states, “Carrier-to-noise ratio (*C/N*) is defined as follows:

$$C/N(\mathrm{dB}) \equiv 10\log(c/n) \quad \text{[Eq. 2]}$$

where *c* and *n* are the scalar power levels of the carrier and noise, respectively.”[1]

When measuring CNR on a spectrum analyzer with thermal noise underlying the carrier, one actually is measuring not C/N but, more precisely, (C + N)/N = 1 + C/N. This distinction is not normally a concern unless the CNR is very low—say, single-digit decibel (dB) values, as we will see later.[2]

The cable industry has long used CNR and SNR to represent quite different measurement parameters, one in the RF domain (Figure 1) and the other in the baseband domain (Figure 2 on page 5). CNR is applied to the transmitted over-the-cable RF waveform, whereas SNR refers to the video and audio signal prior to modulation for broadcast, or after demodulation of the RF waveform at the receiver.

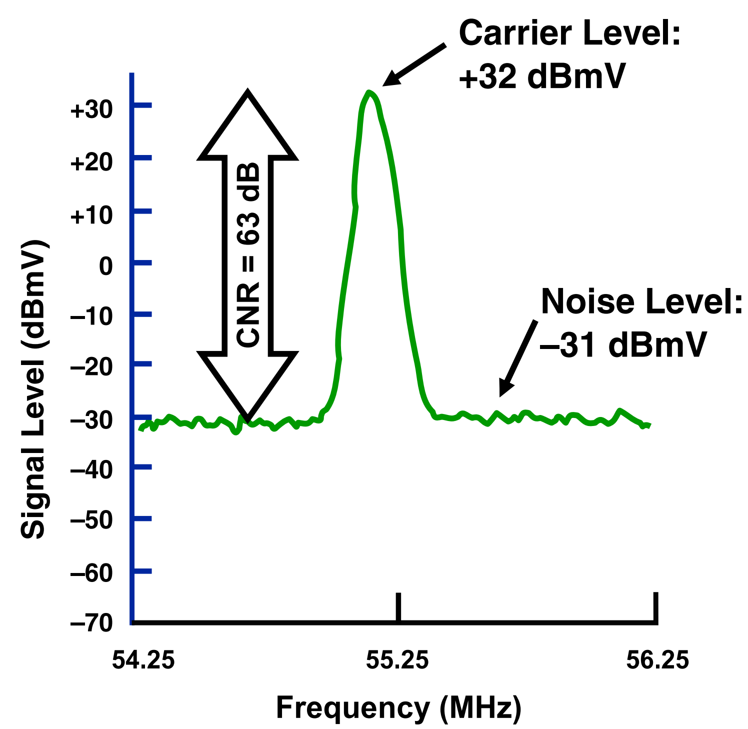


Figure 1: RF CNR Measurement (As Detected in the Test Equipment Resolution Bandwidth)

1. In this paper, all logarithms are base 10.
2. The expression 1 + C/N uses power quantities, not dB; that is, we are not adding 1 dB to C/N.

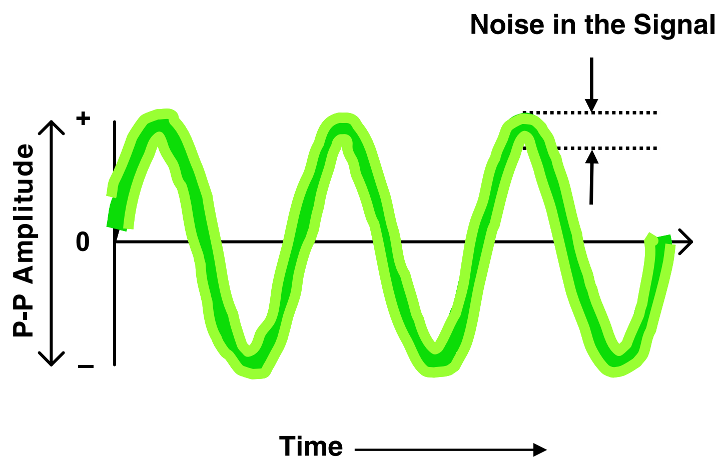


Figure 2: Baseband SNR Measurement

Analog Video CNR in Cable Networks

Consider CNR, which is generally accepted to be a predetection measurement—that is, one made at RF. When only analog TV channels were carried on cable networks, CNR was understood to be the difference, in decibels, between the amplitude of a TV channel visual carrier[3] and the root mean square (RMS) amplitude of system noise in a specified noise power bandwidth. In this application, noise power bandwidth is normally specified as the modulation bandwidth, which is approximately equal to the bandwidth of the baseband modulating signal. It is common practice to express power in terms of RMS voltage across a nominal resistance. For example, *NCTA Recommended Practices for Measurements on Cable Television Systems* defines 0 dBmV (decibels referenced to one millivolt) as the power of a signal of 1 millivolt RMS in 75 ohms, or 13.33 nanowatts = –48.75 dBm.

According to the Federal Communications Commission's (FCC's) cable regulations in §76.609 (e), system noise is the "total noise power present over a 4 MHz band centered within the cable television channel." This latter definition is applicable only to analog National Television System Committee (NTSC) TV channel CNR measurements, and defines the approximate bandwidth of the baseband video that modulates the channel visual carrier.

The FCC does not actually use the term CNR in the rules. §76.605 (a)(7) states "The ratio of RF visual signal level to system noise shall...not be less than 43 decibels." That definition is more or less in line with the general definition of SNR, although it is understood in this specific instance to mean CNR. Even though the FCC's cable rules mandate a minimum CNR of 43 dB, good engineering practice targets end-of-line analog TV channel CNR in the 46 to 49 dB range. More on this topic appears later in this paper.

3. Visual carrier amplitude is the RMS amplitude of the synchronizing peak or "sync." This amplitude corresponds to the visual carrier's peak envelope power (PEP).

Analog Video SNR in Cable Networks

What about SNR? SNR is, in cable industry vernacular, a premodulation (at the transmitter or modulator) or postdetection (at the receiver) measurement—one made on a baseband signal such as video or audio. The previously mentioned Tektronix application note says: "In video applications, however, it is the effective power of the noise relative to the nominal luminance level that is the greater concern." It goes on to define video SNR in decibels as:

$$SNR = 20\log(L_{NOMINAL}/N_{RMS}) \qquad \text{[Eq. 3]}$$

where $L_{NOMINAL}$ has a value of 714 millivolts peak-to-peak (100 IRE units) for NTSC or 700 mV peak to peak for PAL. These luminance values do not include sync.

Equation 3 simply states that baseband video SNR is the ratio of the peak-to-peak video signal, excluding sync, to the noise within that video signal. The noise is measured in a bandwidth defined by a combination of low-pass, high-pass, and weighting filters. These filters limit the measured noise to a bandwidth that is roughly the same as the video signal and may be used to remove certain low-frequency noise from the measurement. Weighting filters are used to simulate the eye's response to noise in the TV picture. Various standards such as RS-170A, RS-250B, and NTC-7 specify the characteristics of filters that are used in baseband video SNR measurements.

To recap: CNR is a predetection measurement performed on RF signals. It is the difference, in decibels, between carrier power and noise power in the RF transport path only—for instance, a coaxial cable distribution network or a stand-alone device such as an upconverter or headend signal processor. As such, CNR is ideal for characterizing network or individual device impairments. SNR, when applied to analog video or audio signals, is a premodulation or postdetection measurement performed at baseband. It is equal to the ratio of the peak-to-peak baseband signal to the noise within that signal (refer to Figure 1 on page 4). SNR includes noise in the original signal—say, noise in the video from a TV studio camera—as well as noise contributions from the transmitter or modulator, transport path, receiver, and demodulator. It is ideal for characterizing end-to-end performance—the overall picture quality seen by the end user, in the case of baseband video SNR.

Discrete Versus Modulated Signals and Carrier-to-Noise Density Ratio

A measurement that is closely related to CNR is carrier-to-noise-density ratio (C/N_0), defined as the ratio of carrier or signal power (in watts) to the underlying white-noise power spectral density (in watts/Hz). Noise power spectral density N_0 is the noise power in a 1 Hz bandwidth—that is, watts per Hz. Because of the impracticality of making a 1 Hz bandwidth noise power measurement, noise power spectral density is usually measured in a larger, more convenient bandwidth—the test equipment resolution bandwidth (RBW)—or, to be more precise, the equivalent noise bandwidth of the RBW filter. The measured value in watts is then divided by the test equipment resolution bandwidth in Hz, which yields the power (in watts) in a 1 Hz bandwidth. If the noise power measurement is in dBmV, subtract 10log(RBW in Hz) from the measured value to get the 1 Hz bandwidth equivalent, also in dBmV.

Taking the ratio of units shows that C/N_0 has units of Hz:

$$\frac{C\ \text{(watts)}}{N_0\ \text{(watts/Hz)}} = C/N_0(Hz) \qquad \text{[Eq. 4]}$$

In decibels, C/N_0 is expressed in dB-Hz, which means "decibels referenced to one Hz." Because C/N_0 is not unitless like other SNR and CNR metrics, care must be taken to reference the noise measurement to a 1 Hz noise bandwidth.

C/N_0 is especially useful for measuring the CNR of a narrowband signal such as an unmodulated or continuous wave (CW) carrier. Consider a spectrum analyzer capture of a CW signal in a white-noise background, with the analyzer RBW set to 100 kHz. Assume that placing the analyzer marker on the CW signal indicates an amplitude of –10 dBmV, and moving the marker to the displayed noise floor shows –40 dBmV. Be careful—this noise reading is *not* in dBmV/Hz, but represents the noise power in the analyzer RBW as mentioned previously, giving a spectral density of –40 dBmV/(100 kHz). To convert to a 1 Hz bandwidth, subtract 10log(100,000 Hz) = 50 dB-Hz. So the noise density is actually –40 dBmV – 50 dB-Hz = –90 dBmV/Hz. Some spectrum analyzers have a marker noise function that provides automatic readings in a 1 Hz bandwidth, eliminating the need for this conversion. The true C/N_0 is then (using dB quantities):

$$\begin{aligned} C/N_0 &= \text{Signal} - \text{Noise} + \text{RBW} \\ &= -10\ \text{dBmV} - (-40\ \text{dBmV}) + 50\ \text{dB-Hz} \\ &= 80\ \text{dB-Hz} \end{aligned} \qquad \text{[Eq. 5]}$$

To convert C/N_0 to CNR in a given bandwidth B, we use

$$\text{CNR} = C/N = C/(N_0 B) \qquad \text{[Eq. 6]}$$

So in decibels, to convert C/N_0 to CNR, subtract 10log(B). The CW signal in this example would have a CNR in a 6 MHz bandwidth (using decibel quantities) of:

$$\begin{aligned} \text{CNR} &= C/N_0 - 10\log(B) \\ &= 80\ \text{dB-Hz} - 10\log(6\ \text{MHz}) \\ &= 80\ \text{dB-Hz} - 67.8\ \text{dB-Hz} \\ &= 12.2\ \text{dB} \end{aligned} \qquad \text{[Eq. 7]}$$

This example illustrates an important principle when measuring a mix of discrete signals (CW or any signals that are much narrower than the RBW of the analyzer) and spread signals (such as noise or modulated signals that are much wider than the analyzer RBW). The spectrum analyzer marker simply measures power in the RBW. For a narrowband signal, this measurement equals the carrier power. For noise, it gives the density referenced to the RBW. Scale to a 1 Hz bandwidth to get C/N_0, and scale to a desired bandwidth B to get CNR.

Digitally Modulated Signal CNR

What about CNR measurement of digitally modulated signals on a cable plant? The DOCSIS *Radio Frequency Interface Specification* states an assumed minimum 35 dB CNR for downstream digitally modulated signals. If the network analog TV channel CNR is maintained in the 46 dB or higher range, in most cases, there will be little or no problem complying with the DOCSIS assumed minimum for downstream digitally modulated signals. The DOCSIS assumed minimum upstream CNR for digitally modulated signals is 25 dB. Carrier power—the "C" in CNR—is the average power level of the digitally modulated signal, often called digital channel power. It is measured in the full occupied bandwidth of the signal; for example, 6 MHz for a North American DOCSIS downstream signal.

Is a Digitally Modulated Signal a "Carrier"?

Quadrature amplitude modulation results in a double-sideband, suppressed-carrier RF signal. Before modulation is applied, the unmodulated signal is certainly a carrier—a CW carrier. But when modulation is applied, the carrier is suppressed. From one perspective, it could be correctly argued that the "haystack" of a QAM signal is not a carrier—it is just an RF signal—because the carrier is suppressed. From another perspective, it is legitimate to argue that a QAM signal "carries" information because its modulation was impressed on a carrier frequency. The underlying carrier can be recovered and tracked by a digital receiver. In fact, a QAM signal is *cyclostationary*—its carrier can be regenerated and observed on a spectrum analyzer by first passing the QAM signal through a nonlinearity such as 4th power. To minimize confusion with baseband signals and to follow common industry parlance, this paper refers to digitally modulated RF signals as "modulated carriers," or simply "carriers."

To measure the white noise—the "N" in CNR—underlying a digitally modulated signal, a noise power bandwidth (also called equivalent noise bandwidth or effective bandwidth) equal to the symbol rate should be used. The noise measurement should be performed when the signal is not present, or in an empty band near the signal. Table 2 and Table 3 summarize noise power bandwidth values for DOCSIS downstream and upstream channels.

Table 2: Noise Power Bandwidth for DOCSIS Downstream Channels

Channel RF (Occupied) Bandwidth	Symbol Rate R_S	Noise Power Bandwidth (approximate)
6 MHz	5.056941 Msym/sec	5.06 MHz
6 MHz	5.360537 Msym/sec	5.36 MHz
8 MHz (Euro-DOCSIS)	6.952 Msym/sec	6.95 MHz

Table 3: Noise Power Bandwidth for DOCSIS Upstream Channels

Channel RF (Occupied) Bandwidth	Symbol Rate R_S [note a]	Noise Power Bandwidth
200 kHz	160 ksym/sec	160 kHz
400 kHz	320 ksym/sec	320 kHz
800 kHz	640 ksym/sec	640 kHz
1.6 MHz	1.280 Msym/sec	1.28 MHz
3.2 MHz	2.560 Msym/sec	2.56 MHz
6.4 MHz	5.120 Msym/sec	5.12 MHz

a. In the DOCSIS 2.0 Physical Media Dependent Sublayer Specification, the term "modulation rate in kHz" is used instead of symbol rate in kilosymbols per second. This definition encompasses both TDMA and S-CDMA transmission.

Downstream Digitally Modulated Signal CNR — An Example

Assume that a cable network has been designed to provide a downstream end-of-line CNR of 46 dB with +15 dBmV subscriber tap levels for analog TV channels, and a downstream 64-QAM DOCSIS digitally modulated signal is carried at –10 dBc. Thus the digital channel power of the 64-QAM signal at the tap spigot will be +5 dBmV, or 10 dB lower than the +15 dBmV analog TV channel levels. What is the 64-QAM signal CNR? It is *not* 36 dB, as one might first assume.

Because analog TV channel CNR is 46 dB at the tap spigot, the noise-floor amplitude N_{NTSC} for the analog channels is +15 dBmV – 46 dB = –31 dBmV (4 MHz noise power bandwidth for analog NTSC TV channels). To determine the 64-QAM signal CNR, we have to first calculate what the noise-floor amplitude is for that QAM signal, based on a noise power bandwidth equal to its symbol rate. For DOCSIS 64-QAM signals, the symbol rate is 5.056941 Msym/sec, so the noise power bandwidth is 5.06 MHz (refer to Table 2). From this, we can calculate the noise-floor amplitude $N_{64\text{-}QAM}$ for the QAM signal with the equation:

$$N_{64\text{-}QAM} = N_{NTSC} + [10\log(5.06/4)] = -29.98 \text{ dBmV} \qquad \text{[Eq. 8]}$$

The 64-QAM signal CNR is +5 dBmV – (–29.98 dBmV) = 34.98 dB.

The CNR of the digitally modulated signal is degraded by more than the 10-dB reduction in signal level because the wider noise bandwidth in the digital signal case allows more AWGN through to the demodulator.

E_S/N_0 and CNR of a Digitally Modulated Signal

E_S/N_0 is the most prevalent parameter used in digital communications to represent the SNR of a signal. It is defined as the ratio of the average energy E_S per QAM symbol to the noise power spectral density N_0 with the noise assumed white. It is a unitless ratio and is normally expressed in dB. Energy per symbol is equal to the average power—that is, the energy in 1 second—of the signal divided by the number of symbols in 1 second.

How can we reconcile E_S/N_0 with CNR? If we multiply the numerator and denominator by the symbol rate R_S, we get:

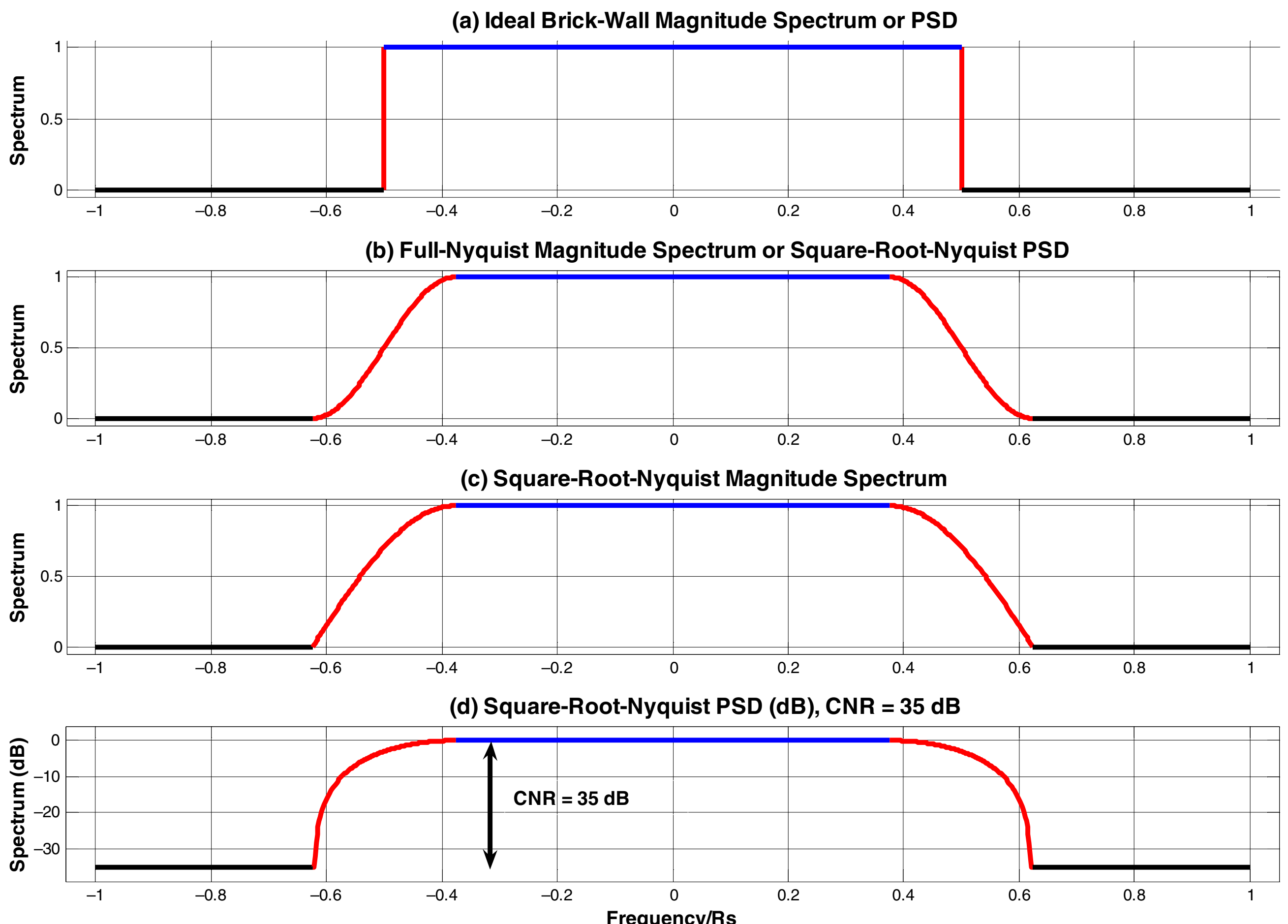
$$\frac{Es}{N_0} = \frac{E_s R_s}{N_0 R_s} = \frac{\text{Signal Power}}{\text{Noise Power in Symbol Rate Bandwidth}} = CNR \qquad \text{[Eq. 9]}$$

(assuming no synchronous code-division multiple access [S-CDMA] spreading, which we discuss later). This equation tells us why the measurement of CNR for digitally modulated signals typically uses the noise power bandwidth equal to the symbol rate: It results in CNR being equal to E_S/N_0.

from cascading the same filter twice; in the latter case, the squaring results from converting the magnitude spectrum to a power spectrum. In a real system, using a spectrum analyzer, we are used to observing the power spectrum of the Tx signal in decibels, which is $20\log|H_I(f)|$ or $10\log|H_I(f)|^2$.

Ideal QAM Spectra

Now we return to the discussion of QAM spectra. Figure 4 shows four ideal QAM spectra, all with the same symbol rate R_S, which is normalized to 1 in these plots. In part (a) of the figure, a perfect brick-wall rectangular spectrum is shown. As the magnitude response $|H_I(f)|$ of a matched filter, this signal is unrealizable in practice because the pulse-shaping filter $h_I(t)$ would have to be infinitely long in time duration. Part (a) of the figure also represents the power spectrum $|H_I(f)|^2$ of the waveform transmitted using such an ideal filter. Despite its impracticability, it is useful as an illustration of an ideal world in which the occupied bandwidth equals the symbol rate and there is no excess bandwidth.



(a) Ideal Brick-Wall Spectrum
(b) Full Nyquist Spectrum
(c) Square-Root Nyquist Spectrum
(d) Square-Root Nyquist Spectrum in dB, with Noise Floor Giving CNR = 35 dB
The passband is shown in blue, transition bands in red, and stop bands in black.

Figure 4: Ideal QAM Spectrum Plots

Part (b) of Figure 4 shows the full-response magnitude spectrum $|H(f)|$ used in the DOCSIS upstream, representing the cascade of the transmit and receive filters. In order to make the filters realizable, an excess bandwidth of 25 percent (alpha = 0.25) is used, resulting in the S-shaped "raised cosine" roll-off regions shown in red, while the passband ideally remains flat. This spectrum possesses the Nyquist property in the frequency domain: If the frequency response $H(f)$ is replicated many times shifted by multiples of the symbol rate, and the copies are overlaid and added as illustrated in Figure 5, the result is a flat spectrum, which results in zero intersymbol interference (ISI).[5] Because of the cascading property mentioned previously, part (b) of Figure 4 also represents the power spectrum $|H_1(f)|^2$ of the transmitted signal and of the square-root Nyquist filter, described next.

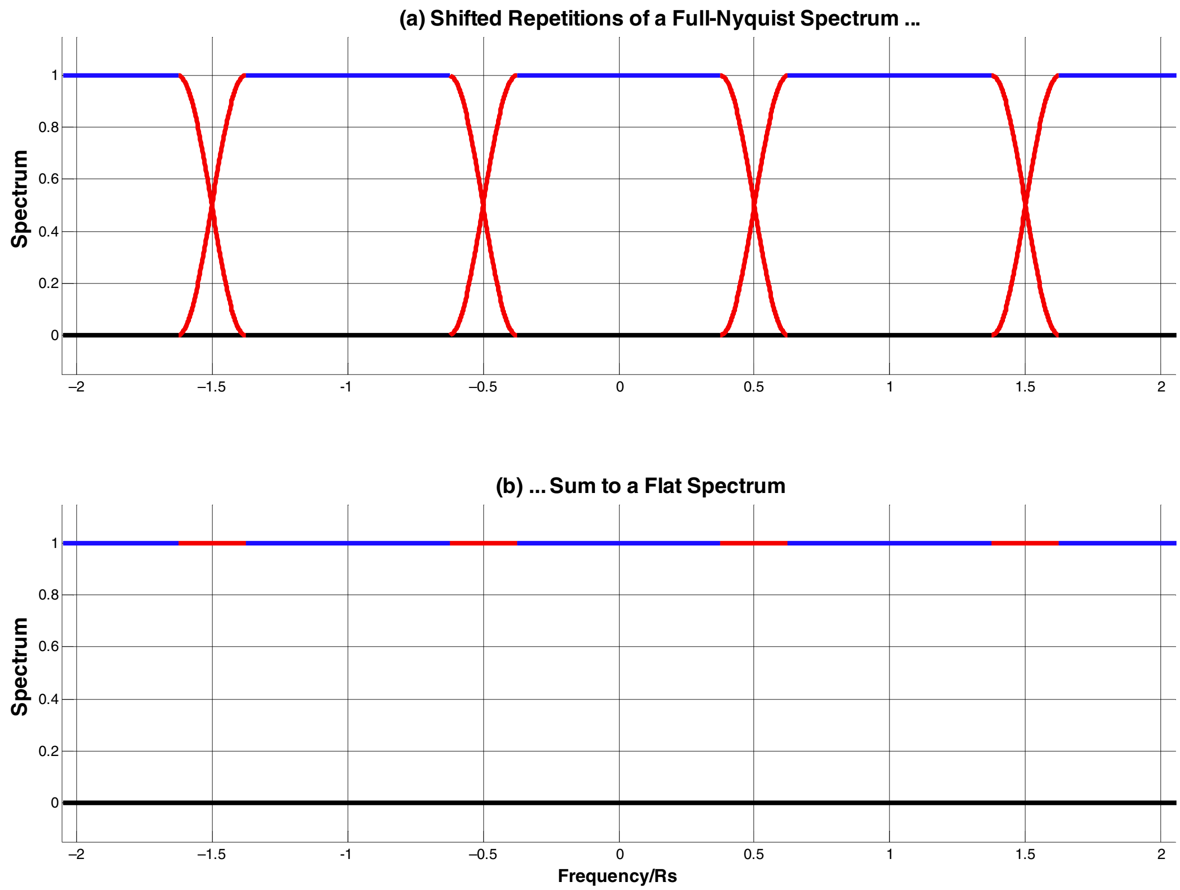


Figure 5: The Nyquist Property States That When Copies of the Spectrum Are Shifted by Multiples of the Symbol Rate and Added, the Result Is a Flat Spectrum, Which Results in Zero ISI.

In practice, the full Nyquist spectrum $H(f)$ of part (b) in Figure 4 is divided into two identical cascaded "square-root Nyquist" filters $H_1(f)$, one in the cable modem upstream transmitter and one in the CMTS burst receiver, using the matched filtering concept discussed earlier. The square-root Nyquist magnitude response $|H_1(f)|$ is shown in part (c) of the figure. Again, because of the cascading property, the power spectrum $|H_1(f)|^2$ of the square-root Nyquist filter and of the transmitted signal, is given in part (b) of the figure.

5. Historically, H. Nyquist discovered this property and published it in 1928, in the context of telegraphy.

For practical use, we can rewrite the previous equations using dB quantities:

$$\text{signal\_pwr\_offset\_dB} = 10\log\left(1 + \frac{1}{10^{\text{haystack\_height\_dB}/10} - 1}\right) \qquad \text{[Eq. 14]}$$

$$\text{true\_signal\_pwr\_dBmV} = 10\log\left(\frac{10^{\text{haystack\_top\_dBmV}/10}}{1 + \frac{1}{10^{\text{haystack\_height\_dB}/10} - 1}}\right) \qquad \text{[Eq. 15]}$$

$$\text{true\_CNR\_dB} = 10^{\text{haystack\_height\_dB}/10} - 1 \qquad \text{[Eq. 16]}$$

where:

- *haystack_height_dB* is the height of the signal haystack (including the noise-floor contribution) above the displayed noise floor, in dB.
- *haystack_top_dBmV* is the power reading on the spectrum analyzer at the top of the signal haystack (including the noise-floor contribution), in dBmV in the analyzer RBW.
- *signal_pwr_offset_dB* is the offset to the signal power measurement caused by the noise floor, in dB.
- *true_signal_pwr_dBmV* is the true signal power reading with the noise-floor contribution backed out, in dBmV in the analyzer RBW.
- *true_CNR_dB* is the true CNR with the noise-floor contribution backed out, in dB.

The formula for signal power offset or error as a function of height above the noise floor (Eq. 13) is graphed in Figure 8. As a general rule, if the signal is at least 10 dB above the noise, the measurement offset will be less than about 0.5 dB. If the signal is at least 15 or 16 dB above the noise, the measurement offset will be less than about 0.1 dB and can be neglected for all practical purposes.

the nearest ideal symbol, or hard decision, from the QAM constellation. The decisions are passed to the Trellis decoder, descrambler, deinterleaver, Reed-Solomon (RS) FEC decoder and MPEG deframer, and on to the MAC layer, which assembles and outputs received packets to the user.

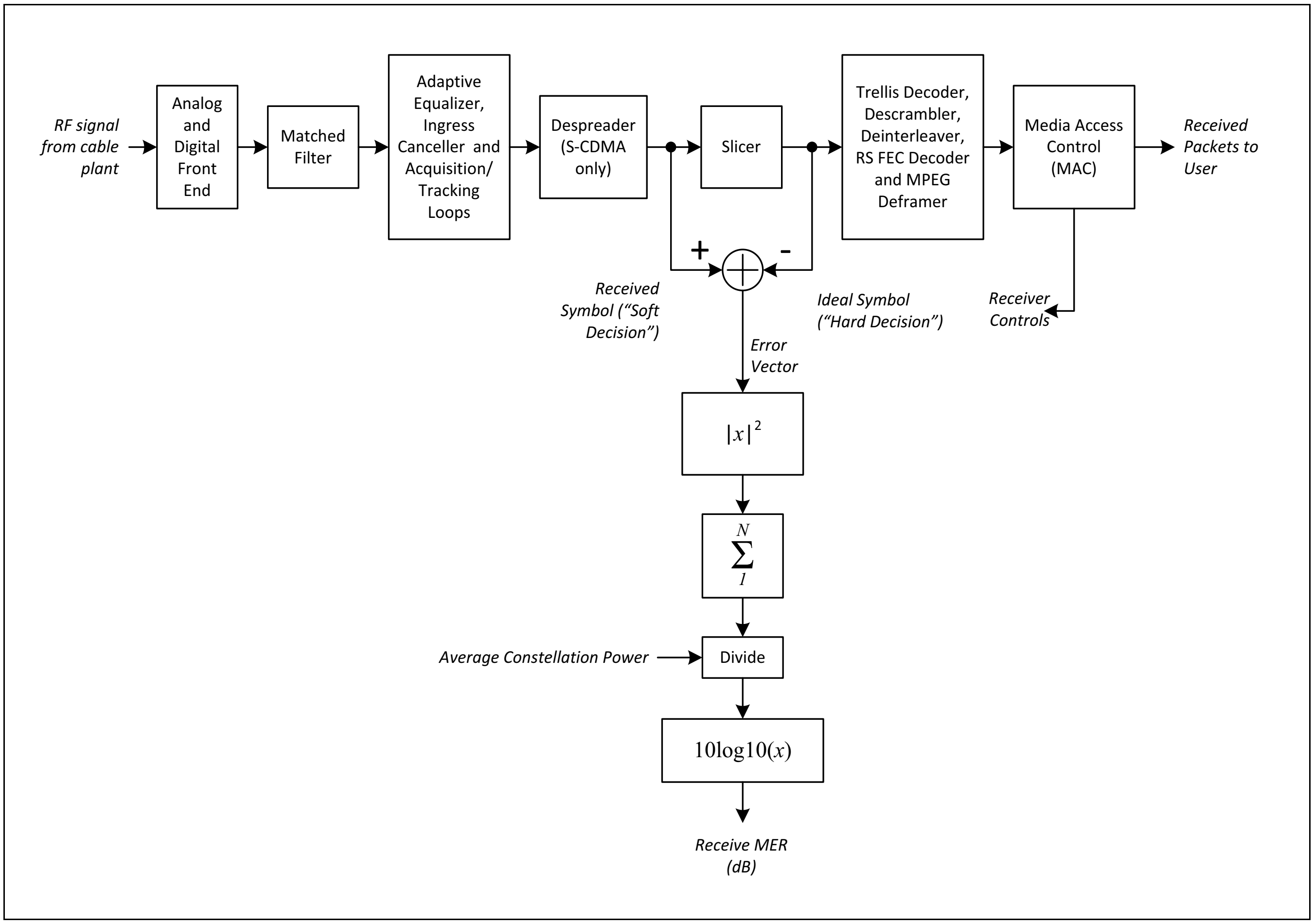


Figure 11: Block Diagram of Generalized Digital QAM Receiver, Showing Computation of Receive MER

What Is Inside the Blocks in a Digital QAM Receiver?

Analog and digital front end: Analog and digital front-end components perform tuning, automatic gain control, channel selection, analog-to-digital conversion, and related functions. Their purpose is to preprocess the signal so that the individual QAM RF channels are available for further digital processing.

Matched filter: The square-root Nyquist filter has a response matched to the symbol or S-CDMA chip. An identical filter is located in the transmitter; this "matched-filter" arrangement gives optimal receive SNR in white noise. The cascade of the transmit and receive square-root filters gives a response with the "Nyquist property." This property, expressed in the time domain, ideally results in zero ISI, even when symbols are transmitted so close together in time that their responses significantly overlap.

Adaptive equalizer: This element compensates for channel effects, including group delay variation, amplitude slope or tilt, and microreflections. It adapts its filter coefficients to dynamically varying channel responses so as to maximize the receive MER. In effect, an adaptive equalizer creates a digital filter with the opposite response of the impaired channel.

Ingress canceller: An ingress canceller is normally included in a CMTS burst receiver to remove narrowband interference (including CB, ham and shortwave radios, etc.). It operates by dynamically detecting and measuring the interference, and adapting its coefficients to cancel it.

Acquisition and tracking loops: Tracking loops provide estimates of frequency, phase, and symbol timing, allowing the receiver to lock to the incoming signal. Acquisition refers to the initialization and pull-in process that occurs when the receiver is first powered on or changes channels.

Despreader: (S-CDMA upstream only) Despreading consists of multiplying the composite received signal by a given code sequence, and summing over all 128 chips in the code. There are 128 despreaders, one for each code. The output of the despreader is a soft symbol decision.

Slicer: The slicer selects the nearest ideal symbol, or hard decision, from the QAM constellation.

Trellis decoder: (Downstream and some S-CDMA upstream modes) The trellis decoder uses the Viterbi algorithm to choose the most likely sequence of symbols and thereby reject noise.

Descrambler: The descrambler adds a pseudorandom bit sequence to the received data bits, reversing the scrambling operation performed at the transmitter. The purpose of scrambling is to randomize the transmitted data in order to provide an even distribution of QAM symbols across the constellation.

Deinterleaver: The deinterleaver pseudorandomly reorders groups of received bits, reversing the interleaving operation performed at the transmitter. The purpose of deinterleaving is to break up long bursts of noise so that the errored bits can be corrected by the Reed-Solomon decoder.

Reed-Solomon (RS) FEC decoder: This device processes groups of bits (7- or 8-bit symbols) arranged in codeword blocks, in terms of an algebraic code using Galois field arithmetic. By processing the received code words, which include redundant parity symbols, receive symbol errors can be found and corrected, up to one corrected RS symbol for each two redundant RS parity symbols.

MPEG deframer: The downstream DOCSIS signal is grouped into 188-byte MPEG transport packets, permitting the multiplexing of video and data over the common physical layer. The MPEG deframer removes the MPEG transport overhead to recover the bytes that are delivered to the MAC layer.

MAC: The MAC layer controls the physical (PHY) layer and is the source and sink of PHY data. The MAC layer processes data frames delineated by DOCSIS headers. In the upstream, the MAC layer governs how cable modems share the channel through a request or grant mechanism.

The input and output of the slicer are complex numbers or vectors, each represented by two components: magnitude and phase, or equivalently, real (in-phase or "I") and imaginary (quadrature or "Q") parts, as shown in Figure 12. In an ideal zero-noise, zero-ISI condition, the soft decision would lie exactly on one of the constellation points, and the magnitude of the error between them would be zero. In a real-world receiver, subtracting the hard-decision vector from the soft-decision vector gives the error or noise vector at each symbol time. The implicit assumption is that a low symbol error rate exists – that is, very few decisions are incorrect, ensuring that the "decision-directed" error vector from the nearest symbol nearly always equals the true error vector from the correct reference symbol.

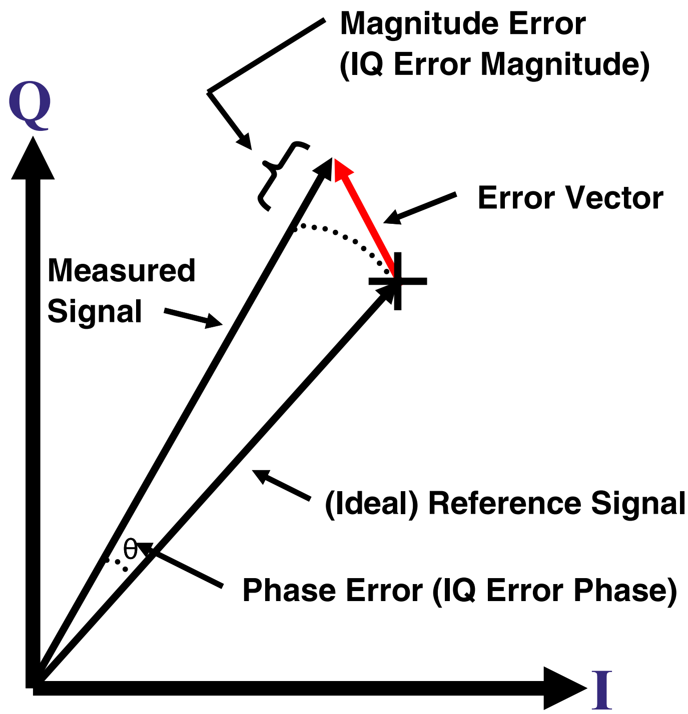


Figure 12: The Error Vector Is the Difference Between the Measured Signal (Soft Decision) and the Reference or Target Signal (Hard Decision). (Source: Hewlett-Packard)

For RxMER, we are concerned with the average power of the error vector, which is computed, as shown previously in Figure 11, by taking the squared magnitude of the complex error vector and accumulating or averaging it over a given number of symbols *N*. This process gives the error vector power (or noise power) at the slicer. Because we want the ratio of signal to noise, we divide the average signal power (a known constant for each constellation, such as 64-QAM or 256-QAM) by the average error vector power. We then take the logarithm to convert to decibels, giving RxMER in dB. To summarize: RxMER is simply the ratio of average symbol power to average slicer error power, expressed in dB.

More About Modulation Error Ratio

Modulation error ratio is digital complex baseband SNR—in fact, in the data world, the terms "SNR" and "MER" are often used interchangeably, adding to the confusion about SNR, especially considering that, as mentioned previously, in the telecommunications world, the terms "CNR" and "SNR" are often used interchangeably.

Why use MER to characterize a data signal? It is a direct measure of modulation quality and has linkage to bit error rate. Modulation error ratio is normally expressed in decibels, so it is a measurement that is

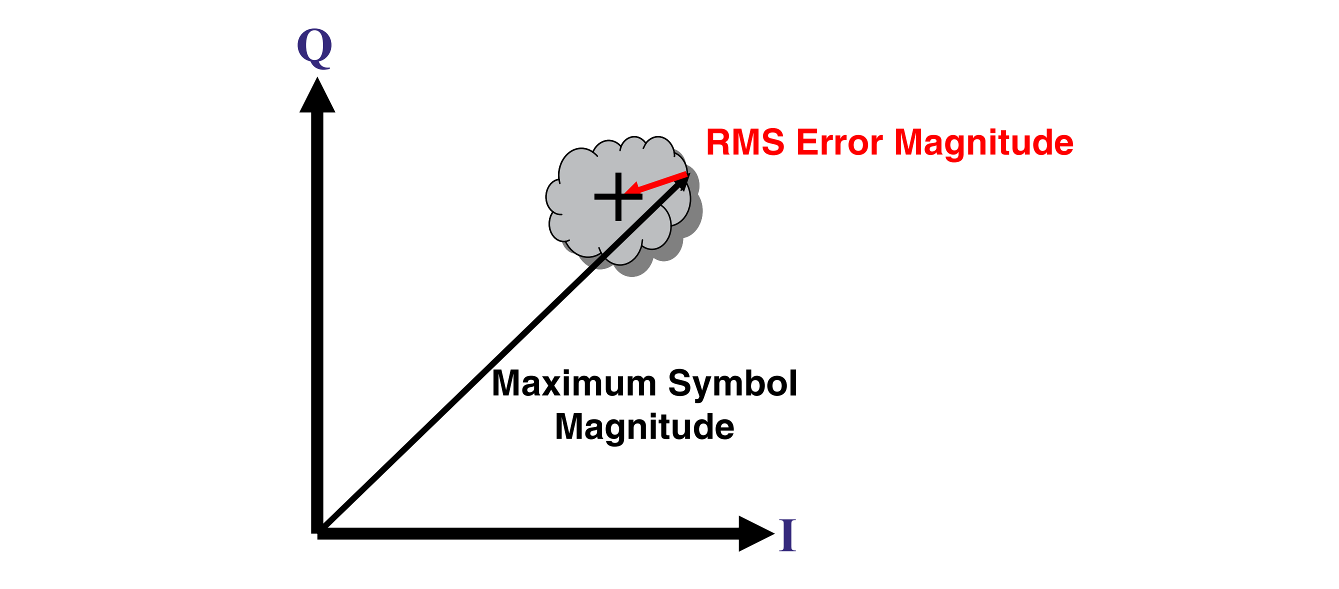


Figure 15: Error Vector Magnitude Is the Ratio (in Percent) of RMS Error Magnitude to Maximum Symbol Magnitude. (Source: Hewlett-Packard)

DOCSIS MIB Definition of Upstream RxMER

In DOCSIS, the upstream RxMER MIB measurement is defined as an estimate, provided by the CMTS demodulator, of the ratio:

$$\frac{\text{average constellation energy with equally likely symbols}}{\text{average squared magnitude of error vector}} \qquad \text{[Eq. 23]}$$

The CMTS RxMER is averaged over a given number of bursts at the burst receiver, which may correspond to transmissions from multiple users. The MIB does not specify whether receive equalization is enabled; this is implementation-dependent.

EVM Versus MER

Another measurement metric that is closely related to MER is error vector magnitude (EVM). As shown previously in Figure 13, EVM is the magnitude of the vector drawn between the ideal (reference or target) symbol position of the constellation, or hard decision, and the measured symbol position, or soft decision. By convention, EVM is reported as a percentage of peak signal level, usually defined by the constellation corner states. The mathematical definition of EVM follows:

$$EVM = (E_{RMS}/S_{max}) \times 100\% \qquad \text{[Eq. 24]}$$

where E_{RMS} is the RMS error magnitude and S_{max} is the maximum symbol magnitude. EVM is illustrated in Figure 15. From this, it is clear that the lower the EVM, the better. Contrast EVM with MER, where the higher the MER, the better.

Error vector magnitude is normally expressed as a linear measurement in percent, and MER is normally expressed as a logarithmic measurement in decibels. Why use EVM instead of MER to characterize a data signal? Many data engineers are familiar with EVM, and for some, linear measurements are easier to work with than logarithmic measurements. Error vector magnitude links directly with the constellation display, and there is a linear relationship between EVM and a constellation symbol point "cloud size" or "fuzziness."

Maximum-to-Average Constellation Power Ratio and EVM/MER Conversion

Because EVM and MER are referenced differently, in order to relate EVM to MER, we must first compute the ratio of the peak constellation symbol power to the average constellation power. The peak constellation power is the squared magnitude of the outermost (corner) QAM symbol. Its formula for a square QAM constellation on an integer grid follows:

$$P_{peak} = 2(\sqrt{M} - 1)^2 \quad \text{[Eq. 25]}$$

where M is the number of points in the constellation (M = 4, 16, 64, 256, etc.) and the points are spaced by 2 on each axis. For example, for 16-QAM, the I and Q coordinates take on values from the set {–3, –1, 1, 3} and the peak power is $2(4 - 1)^2 = 3^2 + 3^2 = 18$. (Use of the integer grid is for illustration purposes only and does not imply any particular power normalization.)

The average constellation power (averaged equally over all symbols in the constellation) follows:[12]

$$P_{av} = \frac{2}{3}(M - 1) \quad \text{[Eq. 26]}$$

For example, for 16-QAM, the average constellation power is (2/3) (16 – 1) = 10. Note that this result happens to equal the power of one of the constellation points; the point (3,1) also has power $3^2 + 1^2 = 10$.

The maximum-to-average constellation power ratio (MTA) is, therefore, the unitless ratio:

$$MTA = \frac{P_{peak}}{P_{av}} = 3\frac{\sqrt{M} - 1}{\sqrt{M} + 1} \quad \text{[Eq. 27]}$$

which approaches 3, or in decibels, 10log(3) = 4.77 dB, for very-high-order QAM. MTA (converted to dB) is tabulated in Table 5, which contains entries for the standard square constellations as well double-square constellations. A double-square constellation is a subset consisting of half the points of the next-higher square constellation, arranged like the black squares on a checkerboard, and contains the same peak and average values as the next-higher square constellation. DOCSIS uses 64-QAM and 256-QAM square constellations for downstream transmission, and specifies both square and double-square constellations from QPSK to 128-QAM for upstream transmission.

12. Simon, Hinedi, and Lindsey, *Digital Communication Techniques*, equation 10.25, page 628.

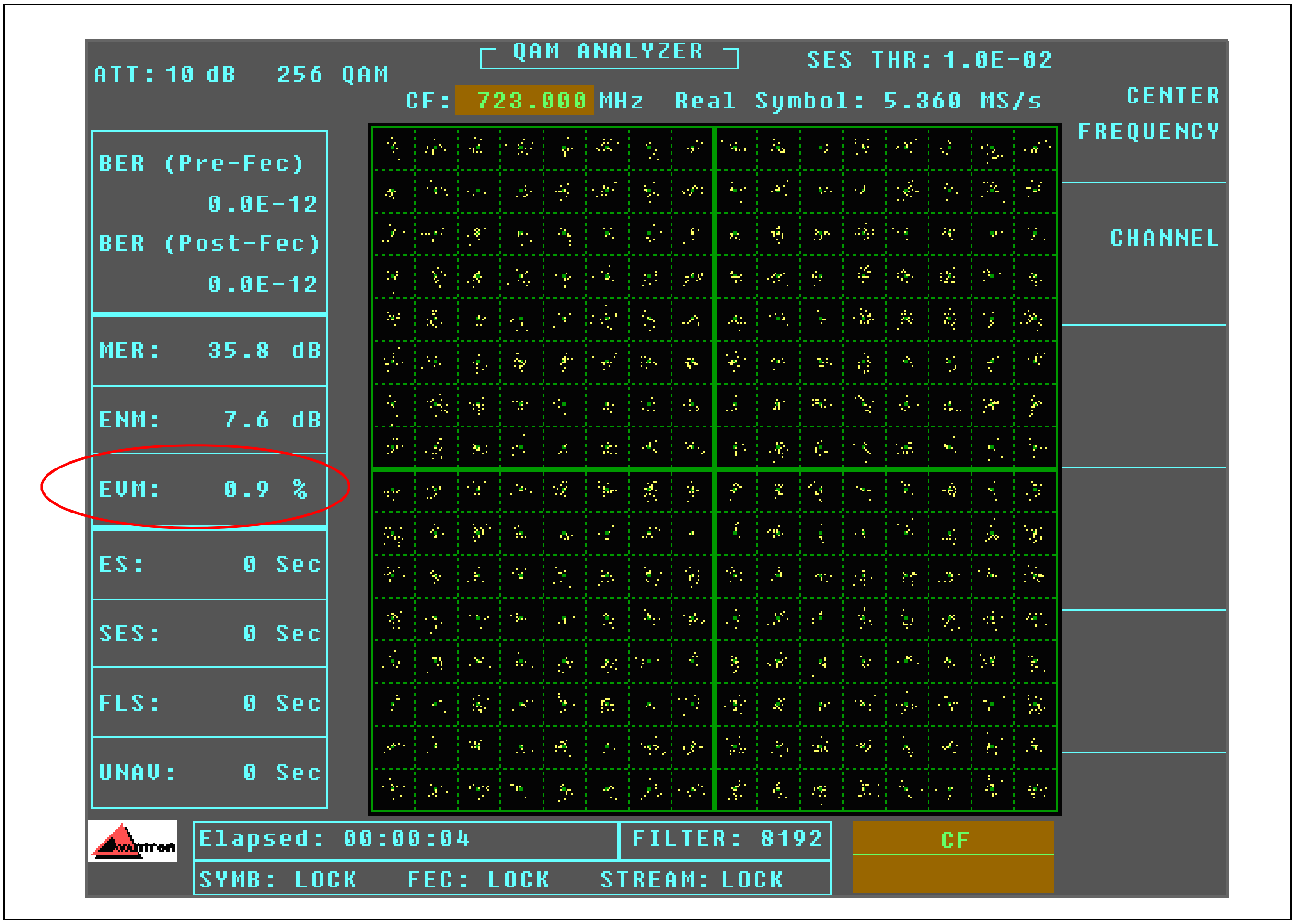


Figure 16: The EVM of This Downstream 256-QAM Digitally Modulated Signal Is 0.9 Percent. (Courtesy of Sunrise Telecom)

Figure 17 and Figure 18 on page 37 show two examples of a 16-QAM upstream digitally modulated signal. The constellation in Figure 17 illustrates a relatively unimpaired signal[14] whose unequalized MER is 27.5 dB. Figure 18 shows an impaired signal, where the unequalized MER is only 19 dB. This level is close to the failure threshold for unequalized, uncoded 16-QAM for some demodulators. In fact, in Figure 18, we can see that some soft decisions are close to the decision boundaries. Note that there is really no way to tell for sure *what* is causing the low MER in Figure 18 by simply observing the display—it could be because of low CNR or perhaps because of one or more linear or nonlinear distortions.

14. The signal has a slight amount of phase noise, observable in the angular spread of the corner constellation points, but is otherwise what could be considered a "clean" signal.

The spectrum analyzer screen shot in Figure 19 shows the upstream spectrum of a cable plant in the band from 0 to 100 MHz. The region from 5 MHz to 22 MHz contains relatively low-level ingress. Clean upstream spectrum is seen in the 22 MHz to 42 MHz range. The upstream DOCSIS carrier is located at 32 MHz under the rightmost red cursor. The diplexer roll-off is seen in the 42 MHz to 54 MHz range. This upstream is relatively clean: The CNR of the carrier is excellent at about 36 dB; there is no visible common path distortion, ingress, or strong impulse noise anywhere near the signal; and we see only a modest amount of ingress well below the signal frequency. Yet although QPSK worked fine in this upstream, unequalized 16-QAM was found to be unusable. The reason? Linear distortions. A severe impedance mismatch about 1100 feet from the node caused a micro-reflection, resulting in amplitude and group delay ripple. These distortions were not visible on the spectrum analyzer display, yet were significant enough to degrade the upstream RxMER and impair 16-QAM transmission. The CMTS reported low SNR —in reality, unequalized RxMER—of 21 dB, which agreed with third-party test equipment measurement of unequalized MER (refer to Figure 20 on page 39). To support 16-QAM transmission on this plant, the operator could enable the preequalization function that is available in DOCSIS 1.1 and higher modems, or increase the level of FEC coding.

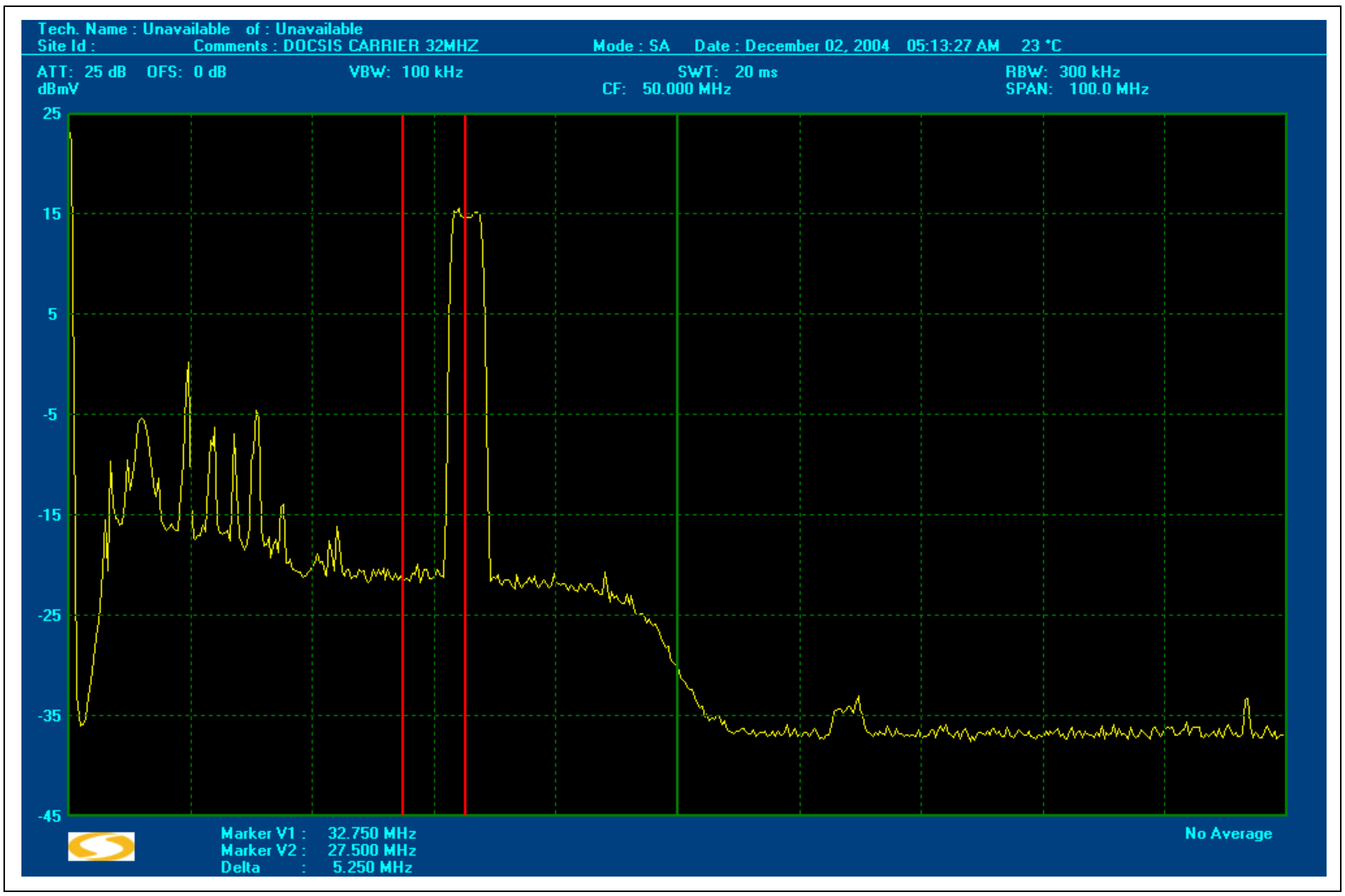


Figure 19: A Spectrum Analyzer Display Shows What Appears to Be a Relatively Clean Upstream, But Unequalized 16-QAM Would Not Work. The CNR Is About 36 dB. (Courtesy Sunrise Telecom)

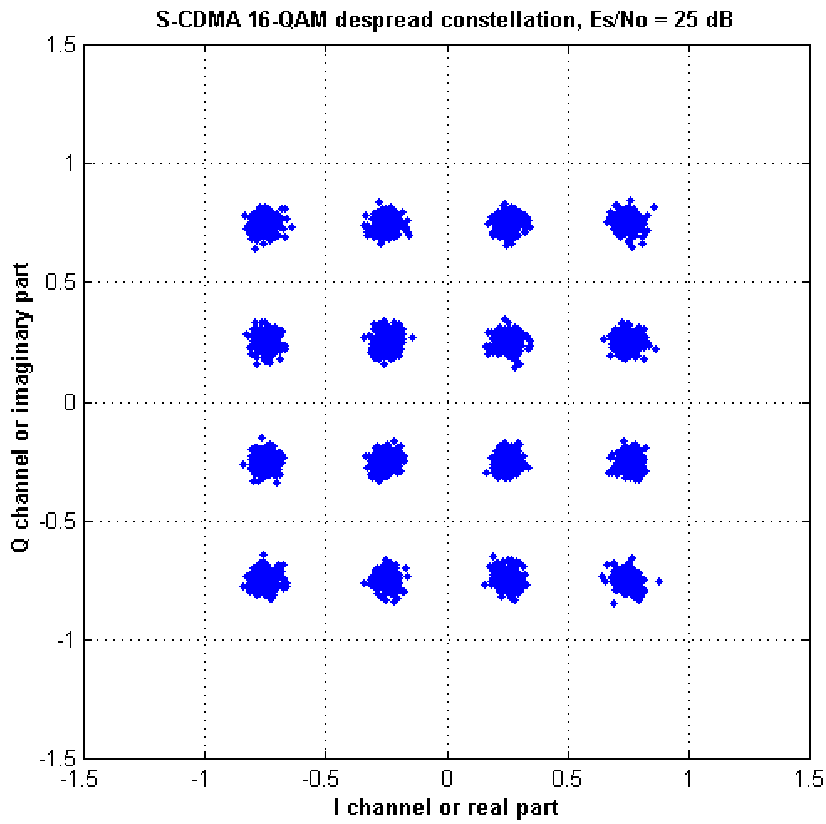


Figure 23: S-CDMA 16-QAM Constellation with 25 dB RxMER

Divergence of CNR and MER in S-CDMA

Because the codes are effectively independent, as shown previously in Figure 22, turning some codes off reduces the total signal power in the channel but does not affect the E_S/N_0 on the other codes, meaning the CNR seen on a spectrum analyzer will appear to fluctuate as some codes are transmitted and others are not, but the E_S/N_0 per code will remain constant. This effect is seen in Figure 24. In the upper trace, all 128 active codes are transmitted. The CNR, E_S/N_0 per code, and RxMER per code are all approximately equal to 25 dB. In the lower trace, all but 32 of the codes have been turned off, while keeping the received power per code unchanged. The CNR is reduced by 6 dB, but the E_S/N_0 per code and received RxMER remain approximately unchanged at 25 dB. Thus, in S-CDMA, the CNR measured on a spectrum analyzer can vary dynamically, and is a valid indication of the E_S/N_0 or RxMER per code only when all active codes are being transmitted in a given frame.

Summary of CNR, SNR, and MER

SNR is a general signal-to-noise-ratio measurement, and can refer to measurements performed at RF or baseband. The cable industry has long used SNR to refer to a baseband measurement; for example, baseband video or audio. CNR is the RF carrier-to-noise ratio seen on a spectrum analyzer. E_S/N_0 is the most common measurement of the quality of a digitally modulated signal. RxMER (often called receive SNR estimate) is the modulation error ratio of the demodulated digital constellation, a "bottom-line" measurement that includes transmitter imperfections, plant distortions, thermal noise, and receiver imperfections. CINR is the carrier-to-interference-plus-noise ratio, which includes narrowband ingress in addition to white noise.

SNR, CNR, E_S/N_0, RxMER, and CINR are equal in an ideal system with no impairments other than AWGN, and with full traffic loading. When impairments occur in a real system, the differences among these measurements provide clues about what the problem is.

Downstream RxMER is measured by the cable modem after the adaptive equalizer, which cleans up most linear distortions. Upstream RxMER is reported by the CMTS burst receiver, usually also after equalization, although some implementations may report an unequalized value. Most QAM analyzers provide equalized MER measurements.

If upstream RxMER is low but the spectrum analyzer shows high CNR, there is most likely a plant or equipment problem to be addressed. Likely candidates are micro-reflections from impedance mismatches or high group delay variation from the roll-off of diplex filters. If DOCSIS preequalization does not cure the problem, then nonlinear impairments, such as phase noise, spurious, burst or impulse noise, etc., may be involved.

TDMA CNR measurements are complicated by the fact that a signal is not continuously present, but the noise is. The maximum hold function on the spectrum analyzer may help. However, maximum hold will display the signal power of the highest-power cable modem burst and highest noise, if the noise is fluctuating in time, typically resulting in errors in the average CNR measurement of 1 to 2 dB or more.

S-CDMA presents challenges for accurate E_S/N_0 measurement because the power of an S-CDMA burst depends on the number of codes used in the burst. If a small amount of data is transmitted, only a small number of codes will be required and the signal power will be low, and low CNR, which may erroneously be perceived as implying low E_S/N_0, will result. An accurate measurement of CNR and E_S/N_0 from a spectrum analyzer can best be obtained when all possible codes are transmitted in a burst. Maximum hold on a spectrum analyzer can provide a reasonably accurate measurement as long as at least one burst is transmitted with all codes used during the measurement period.

CNR measurements accurate to within tenths of a dB are difficult and require knowledge of proper measurement bandwidths, conversion formulas, and compensation for instrument imperfections. If measurements are required only to within 1 to 2 dB or so, straightforward delta marker measurements will suffice as long as proper bandwidth compensations are made.

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