# NEW MEGABITS, SAME MEGAHERTZ: PLANT EVOLUTION DIVIDENDS

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*Abstract*

*The use of 256-QAM in the downstream path has nearly completely replaced 64-QAM as the modulation of choice for MSO's for two simple reasons:*

> *(1) 256-QAM is more bandwidth efficient, providing a 33% increase in spectrum efficiency compared to 64-QAM*
> *(2) 256-QAM has been proven to work reliably.*

*A natural question to ask, given the relatively smooth transition from 64-QAM to 256-QAM, is whether there is a convenient next step in terms of improved bandwidth efficiency. The industry has yet to make a significant move towards 1024-QAM. Prior papers, including by this author, have pointed out some of the potential hurdles. However, as the HFC network has evolved in support of new service demands, and the downstream multiplex has done the same, variables that affect the ability to reliably implement 1024-QAM are beginning to work in favor of this more bandwidth efficient approach.*

*It is now time to understand and quantify the practical performance and the potential limitations in order that it may be deployed properly. This paper will examine the required specification of impairments for successful transmission of 1024-QAM. The discussion will summarize the effect of HFC-specific impairments on 1024-QAM and compare them to 256-QAM and 64-QAM. Finally, we will present some conclusions, along with supporting measured data, on the proper architecture limitations and system thresholds to ensure high-performance delivery of 1024-QAM. These guidelines can be used to enable MSOs to reliably extract another 25% more bandwidth from their digital multiplex.*

## INTRODUCTION

Appetite for more bandwidth has continued to increase, and there is every expectation that it will continue. In particular, as MSOs migrate from broadcast-like models with a VOD component, to delivery models trending towards all-switched unicast services, the Mbps required per service group climbs dramatically. Existing tools (analog reclamation, MPEG-4, SDV, etc.) can and will continue to make the bandwidth explosion manageable. Nonetheless, continued bandwidth growth demands expansion of the HFC toolkit. The requirements are driven by a combination of high-consumption trends – personalized streams, the content itself evolving to HD and beyond, the growth in multiple and non-traditional consumption venues and devices, and the desire to continue to increase data tiers for high-speed Internet service.

The cable plant has kept up with the bandwidth consumption by adding RF bandwidth and using efficient digital modulations to mine the capacity effectively, and with robustness. What started as 64-QAM digital signals became yet more bandwidth efficient with the deployment of 256-QAM downstream, which is the dominant

QAM approach today. The ability to successfully deploy such schemes is due to the fact that the downstream channel in a cable plant has very high SNR, and a very low distortion. This is because it was designed to ensure proper conditions for supporting much less robust analog video, which had historically dominated the downstream payload. In addition to high linearity and low noise, the downstream channel has a flat frequency response on a per-channel basis, minimizing both amplitude and phase distortion, although it can be prone to reflection energy.

What has powered digital modulations historically is the ability to deliver robust link performance and services over varying degrees of link quality by boiling the required receiver function down to choosing between 1's and 0's, instead of replicating and infinitely-valued analog waveform with high fidelity. And, when channel conditions are of high quality – such as the downstream cable plant – the situation is ripe for exploiting bandwidth using very efficient digital techniques.

As a simple example of the possibilities, the theoretical capacity of a 6 MHz channel with a 40 dB SNR is approximately 80 Mbps. Yet, for 256-QAM, the transmission rate is only about 40 Mbps. When accounting for overhead, there is even less throughput. The next higher order, square-constellation, modulation is 1024-QAM. This technique achieves an efficiency of 10 bits/symbol, or another 25% efficiency over 256-QAM, and an impressive 67% improvement relative to 64-QAM. Practically speaking, three MPEG-2 HD's per QAM fit comfortably, as compared to the threat of visual artifacts of jamming a third HD channel into a 256-QAM payload. Alternatively, it would represent at least two more SD streams per QAM. By also considering statistical multiplexing efficiencies and implementing wider channels, this could possibly be increased to four HDs per QAM [2].

Of course, there is no free lunch when it comes to modulation efficiency. To support 1024-QAM, a more stringent set of specifications must be met. The goal here is to identify how pristine the plant must be, or must become, in order to migrate to this modulation profile effectively in terms of the likely candidates for disruptions to robust transmission: SNR, Beat distortion interference, and to a lesser extent, phase noise. It is worthwhile to point out that an intermediate step using 512-QAM, and its additional 12.5% efficiency, has merit in moving up the modulation complexity chain. All of the tools developed here apply to 512-QAM as well, with of course a different set of numbers drawn from them. For this analysis, however, we will focus our attention on 1024-QAM, as these modems exist for the cable space, and the downstream channel capacity favors the likelihood of adding increasingly more bandwidth efficiency to the plant.

## PRIOR ANALYSIS, ESTIMATES, AND TESTING

An early analysis of 1024-QAM was presented at the 2002 SCTE Cable-Tec Expo [1]. While early modem ICs existed that could support this modulation mode [5], little attention had been given to it in the cable world by operators, and subsequently little was understood about how well it would perform in a typical plant. The paper used communication theory and what was learned during the implementation phases of 64-QAM and 256-QAM to draw conclusions about the expectations for 1024-QAM, which was thought to be coming around the corner. However, because of the potential complexities of taking on this advanced scheme and because of other more pressing priorities, this modulation mode has not been exercised, nor has there been any significant trial activity.

The conclusions of the 2002 paper covered a range of items, but ultimately focused on the major items likely to be challenging to 1024-QAM under typical HFC performance delivered: SNR, analog beat distortions, and phase noise. In each case, the conclusion was that current RF characteristics represented performance that would lead to minimal margin. And, for "below average" performing HFC plants, the result could be operating in a bit-error prone region that would add to the FEC's work of delivering error-free data. The concerns regarding these three parameters are quickly summarized below.

SNR

The difference in SNR requirements between 64-QAM and 256-QAM is 6 dB, and the difference between 256-QAM and 1024-QAM is an additional 6 dB. This is nearly exact in the non-FEC case, and is close also in the case of added coding (of the same type), although post-FEC error rate curves are much steeper. So, while 64-QAM, which delivers a 1E-8 BER with no error correction at a 28 dB SNR, zips along comfortably on a cable plant, each modulation order increase gets more difficult. At 28 dB, this allows the QAM load to be implemented with up to 10 dB of signal power back-off relative to the analog carriers and yet still support a large link margin. For example, 46 dB delivered for analog, means 36 dB delivered for digital, resulting in more than 8 dB to spare without even including FEC. This amount of back-off is important, because it allows the digital load to become much less consequential to the total power load, adding only about 1 dB to the total RF load on an 870 MHz system, assuming a 12-14 dB tilt.

Going from 64-QAM to 256-QAM means 6 dB of lost margin in the example above, which sounds perilous – 36 dB of SNR against an uncorrected 1E-8 performance threshold of 34 dB. However, the relative digital power in this case is typically -6 dB, so four of those lost dBs are recovered, at the expense of a larger impact on the total RF load of about 2.4 dB. Because that could mean 5 dB of third order distortion degradation (such as CTB), development of hybrids around digital loads had already followed suit in the expectation of larger loads. In fact, the expansion of analog bandwidth in the plant has continued to spur the development of new actives, and these are often designed with the ability to have the extended bandwidth still filled with analog channels. So, as the needs of more digital SNR for 256-QAM payloads started to matter from the RF load perspective, actives had been keeping pace so as not to increase distortion and degrade analog quality.

Now consider 1024-QAM. Figures 1 and 2 (see end of paper) show constellation diagrams of 256-QAM @ 34 dB SNR and 1024-QAM @ 40 dB SNR. These are equivalent uncorrected BER cases supporting a 1E-8 performance. The similar relative relationship of QAM symbol cloud to hard decision boundary is apparent. The congested look of the 1024-QAM diagram, emphasized by the small symbol decision regions, signals the sensitivity this scheme has to disturbance, and is illustrative of the battle ahead. It takes only small impairments to move an otherwise good symbol across a boundary. It is this 40 dB SNR and the sensitivity to analog beat distortions that led to the suggestion that 1024-QAM represents the first digital modulation choice that needs to be treated more like analog modulation than digital.

Consider what 40 dB means in terms of use on the plant. In light of the above example for digital loading, 46 dB of plant analog CNR becomes 40 dB of digital SNR. By adjusting the modulation order upward, and without any other steps taken, we have instantly evaporated virtually the entire link margin, and are now into a region of measurable bit errors, relying on FEC to finish the job under even the most benign circumstance of thermal noise only. Clearly, for plants or areas of plant that deliver poorer SNR, the situation becomes that much more challenging.

Additionally, on the STB side, there is similar margin-limited mathematics. For STB noise figures in the 10-14 dB range, and for QAM signals arriving at the STB at the low end of the power range, some simple math shows the following:

*Residual Thermal Noise Floor*:
-58 dBmV/5 MHz
*Add STB Noise Figure (middle of range):*
-46 dBmV/5 MHz
*Analog Level into STB*: 0 dBmV
*Digital Level into STB*: - 6 dBmV
(increased for 1024-QAM)
*STB SNR contribution*: -6 -(-46) = 40 dB

The SNR delivered *to* the STB and the SNR created *by* the STB combine to further aggravate the situation of being on the edge, in this case by 3 dB, delivering a 37 dB SNR. This led to the conclusion that existing conditions and typical deployment scenarios place the ability to ensure a smooth 1024-QAM roll-out at risk. It also reveals the necessity of the 4 dB or so of coding gain – some coding gain reduction occurs between 256-QAM and 1024-QAM for the same scheme – just to ensure link closure at a reasonable error rate.

Field tests on live plant [2] bear out the fact that running 1024-QAM means dealing with imperfections of transmission and limited margin. At the original digital transmit levels used for 256-QAM in this testing example, the authors noted errors accumulating. Increasing the digital power by 6 dB removed most, *but not all*, of these errors. It is not clearer whether the original digital level was relative to analog. A 6 dB increase of the single test channel on the multiplex would have no significant effect on loading whether the level began at -10 dB or -6 dB. However, the authors did note that the resulting SNR measured at the receiver was 36 dB, which is consistent with what is expected based on the sample calculations above. The authors additionally note that the system was tested on a relatively short cascade (N+3), which we will later see to be advantageous.

We will subsequently discuss how changing HFC variables are improving this scenario and helping change it for the better.

<u>Distortion</u>

Prior analysis [3], [4] had investigated the effects of analog beat distortions on 256-QAM. Furthermore, laboratory characterization had developed relationships for the comparative performance of 64-QAM and 256-QAM cable links against narrowband interference, which analog beat distortion represents. It was noted that in the modem receiver technology at the time, there was a 10-12 dB difference in susceptibility to a single, static, in-band narrowband interferer at the main CTB offset frequency of interest. It seemed logical to extend this relationship when discussing the difference between 256-QAM and 1024-QAM for this same situation.

Furthermore, analysis supported by test results indicated that the receiver would begin to count errors in a very high SNR environment for 256-QAM when C/I reached about 36 dB, then slowly degrade over the next 5-7 dB or so before becoming unacceptably error prone. This seemed to agree with field performance, as it would take poor CTB to achieve peaks of distortion of this magnitude, even given the noise-like quality to its amplitude. Minimum FCC requirements of 53 dB along with a noise-like peak-average, digital back-off, and the addition of a thermal noise component correlated with field results that generally showed good deployment results with occasional troubling installations, with a mathematical explanation as follows:

53 dB CTBmin – 6 dB (back-off) – 12 dB (pk-avg) = 35 dB

It was postulated that, relative to static CW RF interference (RFI), the interference cancellation mechanism in the receive equalizer would struggle considerably more with an interferer such as CTB because it has some finite bandwidth and a randomly varying amplitude. This expectation turned out to be the case. The concern echoed at the time for 1024-QAM was that, under the measured performance of that generation of receiver on 256-QAM, another 10-12 dB of interference sensitivity may not be tolerable in many more cases, and in cases where the CTB may be average or better.

Consider Figure 3 in the context of the above discussion about interference effects for 1024-QAM. A 1024-QAM constellation diagram with a 35 dB S/I is shown in the figure. This simulation shows clearly how such a tone, without mitigation, would cause errors with noise added in a 1024-QAM system. In fact, Figure 4 shows the 38 dB S/I case with a 40 dB SNR, and it is clear that hard decision errors are occurring that would require FEC support to correct. Analog distortion events also tend to be slow in duration relative to symbol times – a function of the frequency tolerance of the tone contributors, which results in an effective noise power bandwidth and associated time constant [3]. This then taxes the interleaver as well as the error correction mechanism, possibly requiring these receiver functions to be adapted for the increased relevance of this disturbance.

Again, we will discuss how changing variables in HFC evolution are supporting more robustness in this area as well.

<u>Phase Noise</u>

Untracked phase error leads to angular symbol spreading of the constellation diagram as shown in Figure 5 for 1024-QAM with .25 deg rms of Gaussian-distributed untracked phase error imposed. It was observed in [1] that this represents a reasonable limit to ensure that implementation loss due to phase noise is about 1 dB, assuming low uncorrected BER conditions, and with no practical phase noise-induced BER floor. A floor in the 1E-8 or 1E-9 region will be induced at roughly 50% more jitter, or .375 deg rms. Measurements of phase noise showed that for high RF carrier frequencies, typically associated with higher total phase noise, wideband carrier tracking still left about .33 deg rms of untracked error, enough to cause a BER floor to emerge at very high SNR.

The use of degrees rms is sometime easier communicated as a signal-to-phase noise term, and there are some simple rules of thumb to follow and make this simple, starting with 1 deg rms is equivalent to 35 dBc signal-to-phase noise. Doubling or halving entail 6 dB relationships. Thus, we have the following conversions:

4 deg rms = 23 dBc $SNR\varphi$
2 deg rms = 29 dBc $SNR\varphi$
1 deg rms = 35 dBc $SNR\varphi$
.5 deg rms = 41 dBc $SNR\varphi$
.25 deg rms = 47 dBc $SNR\varphi$

The values .33 deg rms and .375 deg rms represent 44.6 dBc and 43.5 dBc, respectively. This is instructive to compare to the SNR for the AWGN case, as it illustrates the more threatening nature of the phase noise impairment on M-QAM modulations of high M.

Recent BER measurements show that error flooring does indeed occur as measured by pre-FEC errors, suggesting that there have not been significant tuner noise improvements or carrier tracking system changes enough to mitigate this effect. However, although phase noise is a slow random process that challenges burst correcting FEC to handle, the combination of the interleaver, the Reed-Solomon encoding, and the relatively low floor, does indeed result in zero post-FEC errors. Note however, that the phase noise alone is requiring the FEC to work to clean up the output data, consuming some FEC "budget" in the process.

We will not carry forth any further analysis on phase noise, except to note that wideband, low-noise frequency synthesis is an art that has been developed in many other applications, at the expense of some cost, of course. It can be costly in particular because wideband and low noise are competing elements in frequency synthesis – high-Q oscillators characterized by low noise do not tune very far. As a result the designs often involve multiple oscillators, frequency multipliers, switches, and more complex direct digital and PLL-synthesis techniques compared to what is done today. The bottom line is if there was a will for improved phase noise, there is a way.

## HFC ARCHITECTURE VARIABLES

### Analog Reclamation

Analog reclamation is a "two birds with one stone" architecture variable, offering benefits to both SNR, through potential power loading adjustment, and distortion. In the case of distortion, the benefit under constant total power loading is in the quantity of analog beat components. Constant power loading allows an operator to take advantage of the opportunity to increase SNR in the digital band with newly available RF power load headroom resulting from extracting analog carriers and replacing them with lower power digital ones. Even with this constant operating point, the effect of analog reclamation is to reduce the total number of analog beats accumulating that can fall beneath a digital channel and create interference. Third-order distortions are the ones that accumulate and cascade most aggressively in the digital band, while worst case second-order distortions populate the low end of the band. Thus, we will focus on the impacts of third-order analog beat distortion, or CTB.

#### *SNR*

The use of analog reclamation to free up bandwidth for digital channels has obvious and well-understood implications for adding service value to the channel line-up. The exposed bandwidth allows for increasing HD content, more digital channels, more niche channels, and more bandwidth for data services. Also, because digital channels typically run at lower power by 6-10 dB relative to analog, replacement of analog channels with digital results in an increase in headroom that can be exploited for SNR purposes while maintaining the same total RF load on the optics or the RF actives. Table 1 shows what this headroom means in terms of the RF power load when compared to a reference power load, under different RF tilts, for the case of occupied forward path bandwidth to 870 MHz. Table 2 shows the same, but for 1 GHz of loaded bandwidth.

It would be ideal if there were 6 dB available to increase the digital levels. However, it is not critical that there is not, as the existing modulations do run with significant link margin under typical cable plant conditions. For example, 45 dB of analog CNR at end of line at a relative back-off of 6 dB it delivers a 39 dB digital SNR. That's 11 dB of margin for 64-QAM and 5 dB of margin to 256-QAM, not including coding gain. When considering coding gain, even in the 256-QAM case there is substantial SNR margin to work with. As previously described, we have the origins of the CATV network as an analog video network to thank for this good fortune, as well as the fact that its origins were as an RF-only plant. Prior to fiber optic carriage, very long amplifier cascades were required, necessitating decent noise properties from the broadband amplifiers.

**Table 1 - Power Loading Effects of Analog Reclamation - 870 MHz**

| | Channel Uptilt @ 870 MHz | | | | | |
|---|---|---|---|---|---|---|
| | Flat | | 12 dB | | 14 dB | |
| | Delta Ref | QAM Increase | Delta Ref | QAM Increase | Delta Ref | QAM Increase |
| **79 Analog** | **Ref Load** | **---** | **Ref Load** | **---** | **Ref Load** | **---** |
| **59 Analog** | -0.7 | 2.5 | -1.0 | 1.5 | -0.9 | 1.5 |
| **39 Analog** | -1.6 | 3.5 | -1.7 | 2.5 | -1.6 | 2.0 |
| **30 Analog** | -2.1 | 4.0 | -2.0 | 2.5 | -1.9 | 2.5 |
| **All Digital** | -4.5 | 4.5 | -2.8 | 3.0 | -2.5 | 2.5 |

**Table 2 - Power Loading Effects of Analog Reclamation - 1000 MHz**

| | Channel Uptilt @ 870 MHz | | | | | |
|---|---|---|---|---|---|---|
| | Flat | | 12 dB | | 14 dB | |
| | Delta Ref | QAM Increase | Delta Ref | QAM Increase | Delta Ref | QAM Increase |
| **79 Analog** | **Ref Load** | **---** | **Ref Load** | **---** | **Ref Load** | **---** |
| **59 Analog** | -0.7 | 2.0 | -0.7 | 1.0 | -0.6 | 1.0 |
| **39 Analog** | -1.5 | 3.0 | -1.2 | 1.5 | -1.1 | 1.5 |
| **30 Analog** | -2.0 | 3.5 | -1.4 | 1.5 | -1.2 | 1.5 |
| **All Digital** | -4.1 | 4.0 | -1.9 | 2.0 | -1.5 | 1.7 |

*Note: For comparison of Tables 1 and 2, the delta of [1 GHz Ref Load – 870 MHz Ref Load] is as follows:*

*Flat: 0.25 dB*
*12 dB Uptilt: 1.27 dB*
*14 dB Uptilt: 1.56 dB*

In the table, the left hand column for each case – flat load, 12 dB tilt, 14 dB tilt – represents the decrease in total RF load compared to the 79-analog channel reference. In every case, the digital carriers run at -6 dB relative level. The right column for each case represents how much *more* power could be allocated to *each* digital carrier in order to maintain very close (within 0.2 dB) to the same total RF power load. This is the added headroom available for SNR that was previously mentioned. These seemingly small available dB become important as we consider increased modulation order, as the move from 256-QAM to 1024-QAM comes with a 6 dB SNR penalty. We will also see how fractions of dBs matter in cases of SNR thresholds vs HFC conditions evaluated later.

In the tables above, the flat case represents the effect on the optical loading of the analog reclamation process. It is apparent than in a mixed multiplex from a single transmitter, the relative digital level allowable will be driven by limitations of the RF plant, where less QAM power can be allocated because of the applied tilt. However, there is nonetheless still headroom that can be exploited by increasing the total power of the analog plus digital multiplex, gaining SNR for all channels. Alternatively, Table 1 and Table 2 show that 1-3 dB of increased QAM power can be applied only to the digital part of the band of the tilted RF outputs, providing some mitigation against the 6 dB increased SNR requirement and subsequent lost margin. It will become clear later as we discuss the effect of cascade depth that these seemingly incremental dBs can have an impact on the ability of the architecture to ensure the desired minimum 1024-QAM SNR objective is met.

Consider simply increasing QAM levels to -3 dB on the 79-channel cascade. This nets a 1.5-2.0 dB total power load increase, which is enough to noticeably impact distortions if nothing else is changed (no cascade shortening). Now consider Table 3. Here, it is postulated that 1024-QAM is run at -3 dBc, with another set of digital channels (half of them) remaining at -6 dBc.

We saw in Tables 1 and 2 that the analog reclamation effort can yield possible QAM level increases for the same power load, but with this typically being less than 3 dB. Nonetheless, to win back margin lost to SNR, in Table 3 we assume that 1024-QAM is run at the -3 dBc level, and that the digital load that is split into -6 dBc and -3 dBc segments. This approach could limit the 1.5-2.0 dB additional power loading effects of having all digital channels increase to -3 dB, but also provide that extra SNR boost to a set of 1024-QAM channels. In one case, the lower half of the digital channels are set at -3 dBc, while in the right-hand columns, the upper half of the digital channels are set at -3 dBc. It is clear from this balance of QAM power that even in the worst case there is a very minor impact on total RF load of < 2 dB using split digital band loading, and essentially no net power increase to the RF load when the lower half of the digital band is used for 1024-QAM. However, it is this part of the band that sees the highest level of CTB beat accumulation, so there is an inherent trade-off between the two.

Note that SNR discussion above refers to the effect of a power increase on a fixed thermal noise floor. However, there is a distortion component referred to as composite intermodulation noise (CIN) that appears like a thermal noise floor, but is in fact a result of distortion products with a digital carrier component. As digital carriers increase, there is more digital contribution to create CIN. In system analysis, the CIN parameter combines with the thermal noise floor to create the parameter known as Composite Carrier-to-Noise, or CCN. CIN looks like thermal noise, and has effects like thermal noise, and is mathematically treated like thermal noise in the calculation of SNR. However, it aggregates as a distortion would aggregate through a cascade, dominated by third order effects. It can be easily isolated in system cascade tests, and the CIN and AWGN components of CCN identified. However, the CIN effect has not been included into the model at this point.

**Table 3 - Split Loading for 1024-QAM - 1000 MHz**

| | 14 dB Channel Uptilt @ 870 MHz | | | |
|---|---|---|---|---|
| | Lower Digital @ -3 dBc | | Upper Digital @ -3 dBc | |
| | Delta Ref | QAM Increase | Delta Ref | QAM Increase |
| 79 Analog | 0.7 | --- | 1.7 | --- |
| 59 Analog | 0.1 | --- | 1.4 | --- |
| 39 Analog | -0.4 | 0.5 | 1.0 | --- |
| 30 Analog | -0.6 | 0.5 | 1.2 | --- |
| All Digital | -1.0 | 1.0 | 1.1 | --- |

In our examples, performance analysis for nominal and increased output levels lead to the conclusion that CIN contribution is always smaller than the thermal noise contribution, and in some cases negligibly so. Reduction occurs as expected as the cascade shortens. During analog reclamation, adding digital carriers adds more potential CIN contributors as described above. However, the analog carriers drive the highest CIN3, generated as (2A+D), and the removal of the highest level analog carriers more than offsets this increase. Nonetheless, to consider the impact of a CIN effect, we evaluate the SNR threshold developed for the cascade over a range of link noise performance. From these curves, the impact of allowing for increased CIN degradation of SNR can be observed by considering the plotted SNR and adding 2-3 dB maximum degradation.

*Distortion*

In addition to its positive effects on digital SNR, analog reclamation offers benefits in the distortion domain as well. Some modeled relationships are shown and discussed below.

Consider a load of 79 analog, with a digital load to 870 MHz as a reference for this example. Observing an 870 MHz upper band will allow us to demonstrate the primary impact of constant loading when bandwidth extension is also considered.

Now observe Table 4, and the flatly loaded (optic) cases first. As can be seen, as a 79 channel analog load is reduced, the relative effect on worst case CTB is for it to drop. It also moves lower in frequency. However, the digital channel band extends lower as well, and thus this movement of worst case CTB does not provide much assistance to the QAM channel. The actual CTB peak extends just into the analog band, and there is about a 1 dB difference in the worst case part of the digital band – near the analog crossover. This difference increases slightly as the total number of analog channels is reduced.

The CTB beat count impact on the optical link is significant when the number of analog drops to its minimum. However, typically the optical link of an HFC cascade will dominate the SNR aspect, but not be the primary driver of distortion. That would be the RF cascade, including the RF drive from the node. The effect on third-order digital distortion, CIN3, which adds to the optical link's noise characteristics, is less significant. CIN3 is the dominant digital distortion that accumulates in mixed cable multiplexes.

Finally, note that the CIN3 improvement of analog reclamation can be nearly washed out with the addition of the digital band to 1 GHz on the optical link (flat loading). Obviously, this is just more digital spectrum to add to the nonlinear mix, at the highest amplitude levels, so this is not unexpected.

Now consider the tilted (in this case 12 dB to 870 MHz) examples. The impact of removing analog is magnified, because the analog channels being removed are the highest level channels, and thus the strongest contributors to CTB. This is quantified as up to 15 dB improved worst case CTB for minimal analog channel count, a significant gain. We can observe in Table 5 that nearly this full gain is achieved at the node output – the optical link and one RF amplification *in* the node (N+0 case).

**Table 4 - 3rd Order Distortion vs Analog Channel Count**

| | CTB | | CIN3 | |
|---|---|---|---|---|
| **Analog Channels** | **Flat (Optics)** | **Tilted (RF)** | **Flat (Optics)** | **Tilted (RF)** |
| **79** | 0 dB (Ref) | 0 dB (Ref) | 0 dB (Ref) | 0 dB (Ref) |
| **59** | -2 dB | -5 dB | -1 dB | -2 dB |
| **30** | -8 dB | -15 dB | -2 dB | -9 dB |
| **Extend Digital to 1 GHz** | | | Add 1.5 dB | Add 3.5 dB |

On the CIN3 side, we see more of the effect of removing the high analog carriers for the tilted multiplex, which drive the highest digital distortion levels as (2A + D). While the reclamation adds digital channels, it adds them at lower power than the existing channels on the uptilt, and, combined with the reduction of highest analog levels, net gains in CIN3 are achieved.

The extension of the band to 1 GHz has a magnified effect in a tilt versus flat case, obviously a result of the digital channels now being relatively higher when installed at the high end of the band.

Cascade Shortening

It is of curse not a secret that each additional amplifier placed beyond the node creates degradation. There are very well-understood rules in the RF world for cascaded degradations of equivalently performing amplifiers:

SNR $\rightarrow$ 10 Log N
CSO $\rightarrow$ 15 Log N
CTB $\rightarrow$ 20 Log N
CIN $\rightarrow$ 20 Log N

There are caveats to these rules that have to do with the mix of optical and RF distortions, noise and digital distortion, and the active technology used. It is with these rules and caveats, adjusted by offsets associated with the variables introduced in the paper – primarily channel loads and cascade depths, that we can estimate distortion effects and the likelihood it will impact 1024-QAM. While we will use particular numerical examples that represent typical characteristics, it is certainly the case that these dBs can also be subject to variation across product types and models. The intent was to deliver some practical conclusions rather than accumulate worst case assumptions that deliver a skewed result.

Table 5 shows modeled performance for a particular 1310 nm link, followed in the N+6 case by two amplifier types in a 2+4 configuration. The simulation uses mathematical models derived through hardware verification of the particular laser and receiver family, and of the individual amplifier characteristics. Of course, parameters of different lasers, receiver, amplifiers, etc can vary across product families, vendors, implementation, etc., so a "typical" arrangement using nominal levels and link lengths were chosen for a reference point. The data underscores the impact on noise and distortion of decreasing analog channel loads, and shorter RF cascades.

Again, CCN represents Composite Carrier-to-Noise – a combination of the CNR or SNR (optical link dominated) and digital distortions.

**Table 5 - Noise and Distortion @ 550 MHz vs Analog Channel Count**

| | CCN | | CTB | | CSO | |
|---|---|---|---|---|---|---|
| **Analog Channels** | **N+6** | **N+0** | **N+6** | **N+0** | **N+6** | **N+0** |
| **79** | 48 | 51 | 58 | 70 | 56 | 64 |
| **59** | 48 | 52 | 60 | 70 | 59 | 65 |
| **30** | 48 | 52 | 68 | 74 | 67 | 70 |

This table shows important trends in two important directions that we will further quantify to develop threshold rules for 1024-QAM. Along the rows, the advantages of going from a 6-deep RF cascade to an N+0 architecture, with the node as the last active, are on display for each parameter. This is also sometimes called Fiber-to-the-Last-Active (FTLA). Gains in both noise and distortion are clear, both of which more ably support the ability to handle higher order modulation such as 1024-QAM. Moving down columns, the benefits of doing analog reclamation becomes clear.

From the perspective of noise, the improvements available are of most significance in the shortening of the cascade, and the impact of the lack of accumulation of amplifier noise. There is 3-4 dB additional optical SNR available relative to a typical line-up and cascade depth of today. When coupled with possible loading adjustments with the larger digital tier and added distortion headroom available (from distortion improvements), there are ways to claw back close to the 6 dB of SNR – the amount of increased sensitivity of 1024-QAM compared to 256-QAM.

For both cascade shortening and minimizing the analog loading, Table 5 shows large available distortion gains. These analog beat distortions look like narrowband interferers, but with random amplitude and phase properties and a measurable bandwidth which, as previously described, makes them more difficult to cancel compared to static CW interference. Cascade shortening reduces the 20LogN accumulation of CTB distortion through the RF amplifiers, and these are the HFC elements that tend to drive link distortion. Analog reclamation reduces the amount of beats entered into the mixing process through which the CTBs and CSOs accumulate. Both are aided by changes in the directions shown in Table 5. We will discuss what these results mean more quantifiably in a later section.

## TEST RESULTS

Based on field testing previously described, and lab performance characterization testing, it has been clear for some time that the output right at the end of the optical link (an N+0 link) was well-suited to 1024-QAM. The SNR and distortions at this point are as good as they are going to get n the cable, and the cascaded RF link sitting between the node output and the STB can only serve to degrade this. It is this fact that suggests that the N+0 architecture is an ideal HFC evolution supporting 1024-QAM, but also that any cascade shortening works in favor of the higher order scheme. Testing was performed to verify this prediction.

### Optical Link Testing

Consider Table 6, which shows the results of typical 1310 nm optics through 20 km of fiber, with 79-channel analog loading and digital loading to 1 GHz. The QAM level is ranged over -4 dBc to -8 dBc, and the 1024-QAM channel inserted in several locations in the loaded digital band, where it's MER and pre-RS FEC BER are measured.

It is apparent, in particular given the -4 dBc data and analog CNRs in the low 50 dB range, that there is a measurement floor associated with the link, reflected in both BER and MER.

**Table 6 – 1024-QAM Performance on Fully Loaded Optical Link**

| | | 1024-QAM Carrier Frequency | | |
|---|---|---|---|---|
| | | 603 MHz | 747 MHz | 855 MHz |
| QAM @ -4 dB to Analog | MER | 39.6 | 39.2 | 38.9 |
| | BER | 6.1E-08 | 1.12E-07 | 3.76E-07 |
| QAM @ -6 dB to Analog | MER | 39.0 | 38.9 | 38.6 |
| | BER | 1.5E-07 | 2.6E-07 | 2.5E-07 |
| QAM @ -8 dB to Analog | MER | 38.3 | 38.2 | 37.7 |
| | BER | 4.30E-07 | 2.02E-06 | 3.48E-06 |

The digital SNR in the -4 dBc cases should all be in the high 40 dB's, and, even with reasonable implementation losses, should be error free or nearly so. Since they are not, there is evidence of an impairment affecting the result, with possibilities in this setup being phase noise, beat distortions of 79 channels, I/Q imbalance (modem implementation loss), STB-limiting noise figure, or a combination thereof. However, in all cases in Table 4, the post-FEC BER was zero – it was error-free at the output.

Not shown was another set of data run with 1550 nm optics. The essential result of that link was that all pre-FEC BERs were in the 1e-6 order of magnitude. Based on the above data, and similar noise and analog distortion numbers, it is likely that this result stems from the effect of optical dispersion on the 1024-QAM channel when operating at 1550 nm. In this case also, however, operation post-FEC was error free.

This 1310 nm optical link performance is encouraging in that it was completely error free post-FEC, but with minor flags on taking pre-errors even at very high SNR (BER flooring). It was also completely expected that performance would be good, as the node output – essentially the N+0 case, eliminates any cascade effects. With noise performance at the node output in the low 50 dB range, and distortion outputs in the high 60's, conditions for 1024-QAM appear quite good. Of course, quantifying what distortion levels are acceptable is part of the objective of this paper. Nonetheless, given these excellent RF characteristics at the node output, coupled with previously described characteristics of the tuner performance, it points to phase noise flooring and modem implementation loss (transmit fidelity) as the most likely sources of the pre-FEC BER floor.

While this N+0 scenario presents a clear case as a sound environment for 1024-QAM, unfortunately, as the RF cascade is lengthened, noise degrades at 10LogN per amplifier to combine with this optical SNR. Furthermore, third-order analog distortion cascades as 20LogN to combine with the node output. It is these cascaded impacts that eat into margin for successful 1024-QAM.

RF Testing

To investigate thresholds of distortion interference as a function of SNR, a laboratory test bed was put together, as shown in Figure 6. While not a full cascade, power adjustment variables in the setup allowed it to accommodate a range of distortion amplitudes at different SNRs.

The RF testing performed include varying SNR under conditions of CW and *live video* CTB interference (NCTA practices define CTB with CW carriers), across multiple frequencies in the digital band (three chosen), and observing the pre-FEC and post-FEC performance. Table 7 shows the results of this testing. Note that all dB values are relative to a digital QAM signal, not an analog reference. Also, note that the "Pre-FEC Error threshold" used was 1E-7. Since the receiver bottoms out around 1E-8 (slightly below), this was selected as a point at which non-floor related errors are being taken. The "Post-FEC Error Threshold" is simply the point at which that metric, representing data after it has been through its full clean-up, is still non-zero. It may be the point at which an operator decides that CTB beyond this is, or some acceptable dB higher, is consuming enough of the FEC budget, which then becomes unavailable for other impairments. The post-FEC 1E-6 threshold is logged as a reference point. Post-FEC curves tend to be very steep, and once this level of error accumulation begins, the system is not very far from operating with unacceptable error counts or not at all. It also represents a point of visual threshold-of-visibility (TOV) in some circles. Finally, the

**Table 7 - Error Threshold of Interference on Most Sensitive Channel**

| | CW Interference | | CTB Interference | | | |
|---|---|---|---|---|---|---|
| SNR | Pre-FEC Error Threshold | Post-FEC Error Threshold | Pre-FEC Error Threshold | Post-FEC Error Threshold | Post FEC > 1E-6 | Post FEC Broken |
| **50 dB** | 34 dB | 33 dB | 55 dB | 55 dB | 55 dB | 45 dB |
| **45 dB** | 35 dB | 33 dB | 55 dB | 60 dB | 55 dB | 46 dB |
| **40 dB** | 36 dB | 34 dB | 60 dB | 60 dB | 55 dB | 49 dB |
| **37 dB** | 38 dB | 37 dB | 60 dB | 60 dB | 55 dB | 50 dB |

point at which the system breaks under the weight of excessive CTB is identified.

Note CTB referred to in Table 7 is as it is measured in this setup, *not* per referenced definition and typical practices. That is, rather than CW beats under an analog virtual carrier, it is live video beats underneath a QAM channel power. Beat distortion from live video beats are typically assumed 8-12 dB below "CW" CTB due to the downward analog modulation, with most references identifying 12 dB of "help" by turning on live video. And, QAM power is, of course, derated by some amount to an analog reference – typically 6-10 dB, but as we have been arguing here, this could perhaps be only 3-4 dB for 1024-QAM.

Also note that in this table, the performance noted is the *worst* performing of the three frequencies across the set of cascade measurements taken. These are important nuances, because the average measurement across the frequency set was typically 3-5 dB better than the worst measurement. And, certain cascade combinations actually showed slightly *better* CTB at N+1, for example, compared to N+0. This phenomenon has been observed before in sample test of amplifiers employing E-GaAs technology, which has different distortion-generating mechanisms than, for example, Silicon. The effect has been attributed to this reality of the device physics, and random phasing relationships. While this has encouraging benefits, it is not considered a system "rule" that these amplifiers are immune to degradation in a general sense, and system design goes according to the safe assumption that distortion is normally assumed to degrade.

Table 7 indeed proves a postulate made here and originated in the 2002 paper – that mitigation of CTB is more difficult that CW interference. Again, CTB's random properties challenge the equalizer. In addition, CTB is an averaging measurement of a noise-like waveform. Thus, it has a high peak-to-average ratio, with 12-14 dB often used for practical noise. Theoretical analysis would predict higher, while measured values tend to this range or lower. Meanwhile, the peak-to-average of a CW signal is 3 dB, and, it is of constant envelope amplitude.

Recall, the prediction that a threat could exist for 1024-QAM was based on the CW interference case for 256-QAM, coupled with these factors as follows:

*CTB min (FCC)* = 53 dBc
*CTB pk* ~ 40 dB
*CTB pk, relative to QAM* ~ 34 dBc

In 2002, 256-QAM showed BER sensitivity, pre and post-FEC, for S/I ratios in this mid-30's range. It is encouraging to note that in testing, now seven years hence, that the receiver evolution has led to narrowband ingress cancellation of 1024-QAM that is comparable to what once was recorded to be the limits for 256-QAM. In [5], an ingress cancellation example is shown that displays roughly 30 dB of interference mitigation of a static interferer in a noise-free environment (>50 dB SNR).

From this simple example, it is clear by comparing this 34 dBc value to the left-hand side of Table 7, that the CTB impairment's peak is achieving levels in the range that CW carrier amplitudes were seen to be of concern, but now for the more sensitive 1024-QAM case. It is this relationship that led to the conclusion that plants that were ok (but not great) could be a concern for 256-QAM at the time, and thus even average systems could create issues for 1024-QAM. Now, however, the 1024-QAM CW thresholds are much improved, comparable to prior 256-QAM thresholds. This is much better situation than had the S/I sensitivity been as previous for 256-QAM, knowing the added sensitivity created by 1024-QAM.

Table 7 reveals some possible threshold pairing of noise and distortion for a quality 1024-QAM link. Again, note that these are the most sensitive readings recorded, with averages across the board running 3-5 dB less sensitive. Nonetheless, let's point out the key result: the 1024-QAM receiver saw error degradation under some CTBs that are not inordinately high levels of distortion.

## HFC PERFORMANCE THRESHOLDS

### SNR

Taking a look at SNR, we can draw some conclusions about the depth of the RF cascade as a function of the expectation from the STB noise figure contribution, and the optical CCN contribution, as a function of a pre-defined overall SNR link objective. Consider Figure 7 and 8. These figures illustrate the nature of small SNR margin increments in their effect on the tolerable cascade depth over a range of optical SNRs, for two chosen link requirements: 38 dB and 40 dB. For simplicity, we have assumed equivalent RF amplifier noise and similar gain, and based these estimates on the results shared in Table 5. The effect of typical gain and noise differences do not substantially changing the qualitative conclusion to be drawn. The net noise figure impact of the gain variation of, for example, splitting the gains between trunk amplification and line extenders, is a dB or so in this N+6 case. This is small, but, as we will point out below, adjusting the RF noise contribution by a dB or two could make a difference in some regions of the curves. Thus, it is desirable *not* to be operating in such regions.

Each curve in Figure 6 and 7 represents a different value of SNR as set by the STB alone, associated with the noise figure and digital level (derated from analog) at its input. Note from the figures that there is a wide range of SNR combinations that essentially offer no practical limit to RF cascade depth as it relates to noise degradation. Clearly, tolerating a 38 dB link requirement, which would be relying very strongly on FEC for just thermal noise aspects, provides a comfortable range of noise performance of the link contributors to work with. However, the range still includes conditions that could lead to a sharp reduction in cascade acceptable. From a sensitivity analysis standpoint, such conditions hinge on small dBs and even fractions thereof. This makes it more valuable to be able to earn back SNR in the analog reclamation process by taking advantage of the load being lightened by the expanded digital, or by taking advantage of headroom in distortion offered by the dwindling beat counts. To ensure operation above the desired threshold, it is not practical to rely on the outdoor plant to maintain key noise and distortion parameters to within fractions of dBs over time and temperature, so operating in these sensitive regions should be avoided.

Again, the 40 dB SNR value represents approximately the 1E-8 point for uncorrected 1024 QAM – a noble objective, setting up the link budget to be acceptable without FEC under the otherwise ideal conditions, and reserving the FEC to deal exclusively with the set of channel issue. However, this may not be practical without lower noise contributors on the CPE side (STB).

CTB Distortion

We will make the similar simplifying assumption on the RF cascade as in the noise case – similar CTB performance and gains across the actives at nominal RF output levels. In this case, splitting the cascade as above among some typical performing amplifier types results in about a 2 dB increase (lower distortion – thus a conservative evaluation) in the third order compression point – the metric that sets CTB. Again, the point here is to recognize that around the data presented below, consider that the exact CTB may not be easily counted on to the dB level, and therefore understanding how performance behaves over a range around the anticipated CTB is important to ensure a robust actual system. The CPE end of the contribution is extracted from the analysis, because the threshold results will be based on measured data that includes that receiver contribution when BER is measured, but the recorded CTB value is that delivered to the receiver input. This leaves us with two contributors – the optical link and the RF cascade – and the variables that modify the contributions to make the assessment.

For the optical link, the beat map effect of flat loading and a fixed transmitter RF load can be relatively easily quantified mathematically and applied as such during analog reclamation. While the link distortion is dominated by the RF cascade, the optics will be included because, as the cascade shortens, it becomes more relevant, in particular with the magnified gains of the tilted channel load on CTB, compared to the flat loading gains. Thus, we will consider the optical link to be the node receiver output, where the multiplex is flat, and bundle the RF stages of the node into the RF cascade for the analysis. Mathematically, then, our N+x case will be modeled as N+x+1. We will assume actives of the same characteristics per the original assumption across the link, including the node RF, established at some nominal reference output level. Again, this is a simplifying assumption, and has single digit IP3 effects for typical ranges of parameters. However, it allows creation of a model that provides a feel for some basic relationships. Of course, the model can be extended to consider any number of individual contributors. Note that there is not a simple formulaic relationship for the RF cascade, as the tilt variation and channel maps both contributing to CTB variation.

Figure 9 shows a sample of an analog beat map for 79 analog channels on a 12 dB tilt to 870 MHz. Tools such as this are used to calculate the impact of varying channel line-ups on *relative* distortion level (tone another and versus frequency) with results as depicted in the Table 4 examples. Note that the peak distortion is third-order (CTB), and that it peaks just below the start of the digital band. As previously described, the worst digital band CTB is about 1 dB lower than this peak. Figure 10 shows the case of flat loading, such as would be carried on the optical link. In this case, the maximum CTB beat location shifts lower, and the difference relative to the worst digital band is closer to 2 dB. These results are used in lieu of an available formulaic representation to evaluate the reclamation process on the titled multiplex distortions.

We now evaluate the ability of the HFC cascade to support a CTB threshold derived from the data provided in Table 7. We will evaluate over a range of given RF amplifier CTBs specified at some typical RF output levels to which these CTB values apply, and evaluate for the varying analog channel counts used above. Two performance thresholds are used, each of which is derived by using the measured case of -60 dBc interference as the objective. The -60 dBc measurement threshold was selected because

- We care most about post-FEC performance
- Post-FEC above 1E-6 is too high, beyond a video TOV, and crashing occurs shortly thereafter on a Post-FEC curve
- It correlates with the range of the SNRs we would expect to see delivered to the home

From the -60 dBc measured value of live video underneath a QAM channel, we make the following adjustments to reference that value to a standard CTB measurement. A standard measurement would be referenced to an analog carrier level, and it would be using CW carriers as load signals. The net results of the former is a 6 dB offset (typical), and we will also evaluate a 3 dB offset to consider the case where more power is given to the QAM load to support the higher SNR requirements of 1024-QAM. When considering live video carriers are the source of the distortion beats and not CW carriers, we have another offset to account for (in the opposite direction – live video is more benign) in the range of 8-12 dB. We will evaluate the endpoints of these cases starting with a -60 dBc measurement threshold, of which we will no longer refer to as "CTB" given the assumptions that the term entails. The endpoints are:

CTBth, min = 60 + 6 – 8 = 58 dBc
CTBth, max = 60 + 3 – 12 = 51 dBc

Figures 11-14 display the results for these cases for the depth of RF cascade, across a range of RF amplifier CTBs, or net effective CTB performance, as a function of the number analog channels from a 79-channel system and going through two stages of analog reclamation.

In Figure 11, it is clear that relying on a 12 dB CTB drop with live video and boosting the QAM power by 3 dB, there is virtually no practical cascade limitations for the cases of analog reclamation employed. In the 79-channel case, only poor or malfunctioning RF amplifiers appear to be a threat to the 1024-QAM system. However, note that there is no loading impact assumed of the added QAM level. We know from prior discussion that this amount of increase in relative level, when applied to the entire digital load, can nudge the total RF level up measurably, increasing distortion. For the cases employing analog reclamation, it has been shown in prior sections that increased QAM level can be obtained with no effect on total RF load, but there is not enough headroom for a full 3 dB increase. There are also techniques such as the split loading of 256-QAM and 1024-QAM that could limit the loading impact for the existing 79-channel case.

For Figure 11, then, the assumption of maintaining the total RF load means either the nominal power load is referenced to the multiplex as described here (it includes margin headroom for this small increase). Or, only a subset of the channels employs 1024-QAM (such as the HD tier for more per-QAM efficiency, or the HSD tier). Or, analog levels are compensated downward. In any case, we will remove the constraint regarding loading in Figure 14 and recalculate this case.

Now consider Figure 12, where the CTBth,min = 58 is used as the guideline. It is clear the power of complete analog reclamation to support virtually any practical RF cascade depth in this case also. However, it also becomes clear how for 79-channel systems and 59-channel systems, some limitations are beginning to appear. This is consistent with the expectation that the march to shorter and short cascades supports the move to 1024-QAM, albeit not as powerfully as what analog reclamation creates in margin. Figure 13 removes the 30-channel analog case, where analog reclamation is complete or nearly so, thus adding granularity to the figure such that reasonable cascade depths can be better quantified as the requirement for them becomes shorter and shorter. Note that by the nature of the requirement chosen compared to the RF CTB range, the curves are unable to cross the x-axis, which represent N+0 (i.e. the combined CTB of the optics and the CTB range of RF used is better than the worst case requirement of 58 dBc).

Finally, re-consider the CTBth,max of 51 dBc under the conditions that the increase in QAM power to a -3 dB derate applies to the whole load, and that the whole load is 1 GHz of bandwidth. The total RF load then is increased about 2 dB, increasing distortions, and in particular third order distortions by 4 dB. This case is shown in Figure 14. Note the shift of a tolerable N+5 cascade to now what might be an N+3 to maintain performance against this threshold if not paying attention to the total power loading. Note also that this assumes a starting derate of QAM at -6 dBc and increasing to -3 dBc. If the starting derate is lower, than the impact on total power is smaller.

With the curves above, operators are able to evaluate the likelihood of 1024-QAM can run successfully on the plant around some basic assumptions. Furthermore, a model has been put in place that allows evaluation on a case-by-case basis for given combinations of equipment, channel plans, and alignment.

## SUMMARY

The use of 256-QAM is commonplace for DTV and HSD, and operators have learned what it takes to implement it successfully on mixed multiplexes. Its performance relative to 64-QAM is well-understood, and the additional 33% bandwidth efficiency can be achieved now with minimal cost and pain. There is 25% more bandwidth efficiency available if 1024-QAM is deployed. Modem ICs exist, and this paper sets out to get a feel for the pain part of the equation, understanding that this scheme is yet another level more sensitive to impairments than 256-QAM, and rivaling in many ways the sensitivity of analog video itself.

The good news is that interference analysis conducted in 2002 based on findings extrapolated from 256-QAM performance of that era now seems to have been addressed in the receivers available today. Narrowband cancellation performance seems robust, and a measurable improvement over 2002. The performance against narrowband interference today for 1024-QAM is roughly what it once was for 256-QAM – the level of interference that used to disturb 256-QAM performs to roughly the same error rate now, but on the more advanced modulation. This is very important for analog beat distortion mitigation, which appears as noise modulated narrowband interference. The random element of it makes the receiver have to work harder, and as a result it is less capable of attenuating the interference. The improved performance makes what once looked to be a potentially troublesome CTB problem now a manageable one with some constraints on the HFC link, with performance characteristics plotted against CTB and channel line-up variations.

In terms of noise performance, not much appears to have changed, and the system display a pre-FEC floor likely associated with modem and RF imperfections, such as phase noise. The system relies on FEC to obtain error-free performance under typical plant conditions because of the 40 dB pre-FEC SNR threshold for 1E-8 performance. For 64-QAM and 256-QAM, significant link margin existed under normal conditions to set reasonable thresholds to qualify link acceptability. Because of the increased SNR requirement of 1024-QAM, it is up to the operator to make a different type of determination with respect to link margin and acceptable threshold under normal plant conditions. Acceptable margin will have to be looked at differently for 1024-QAM. The figures in the paper are meant to help operators see the trade space they are working with for SNR in that respect.

Given the results shown here, an awareness of the key drivers to link performance, and a modeling approach that can be used to assess HFC readiness to accept 1024-QAM signals, operators can start netting themselves that extra 25% of bandwidth efficiency that is currently going unused in already occupied spectrum. That is, all the tools and knowledge are in place to start getting new Megabits from the same Megahertz.

## ACKNOWLEDGEMENTS

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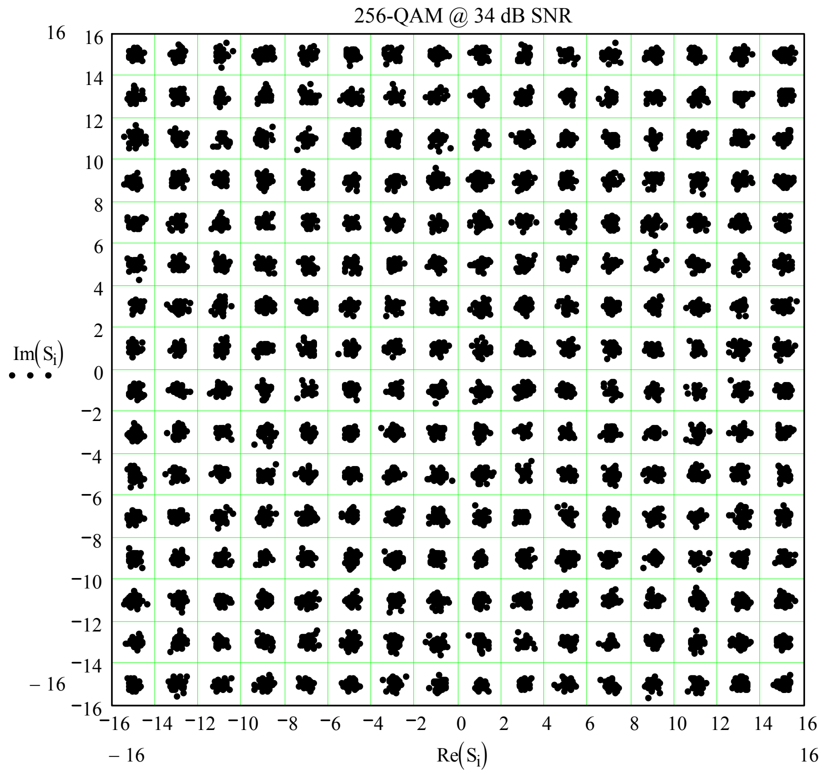



**Figure 1 – 256-QAM @ 34 dB SNR**

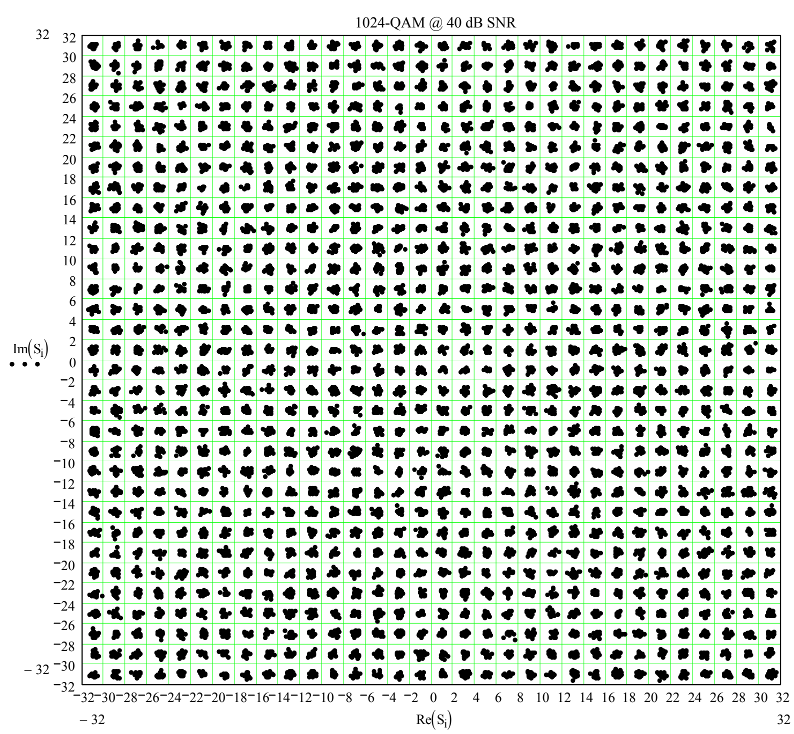



**Figure 2 – 1024-QAM @ 40 dB SNR**

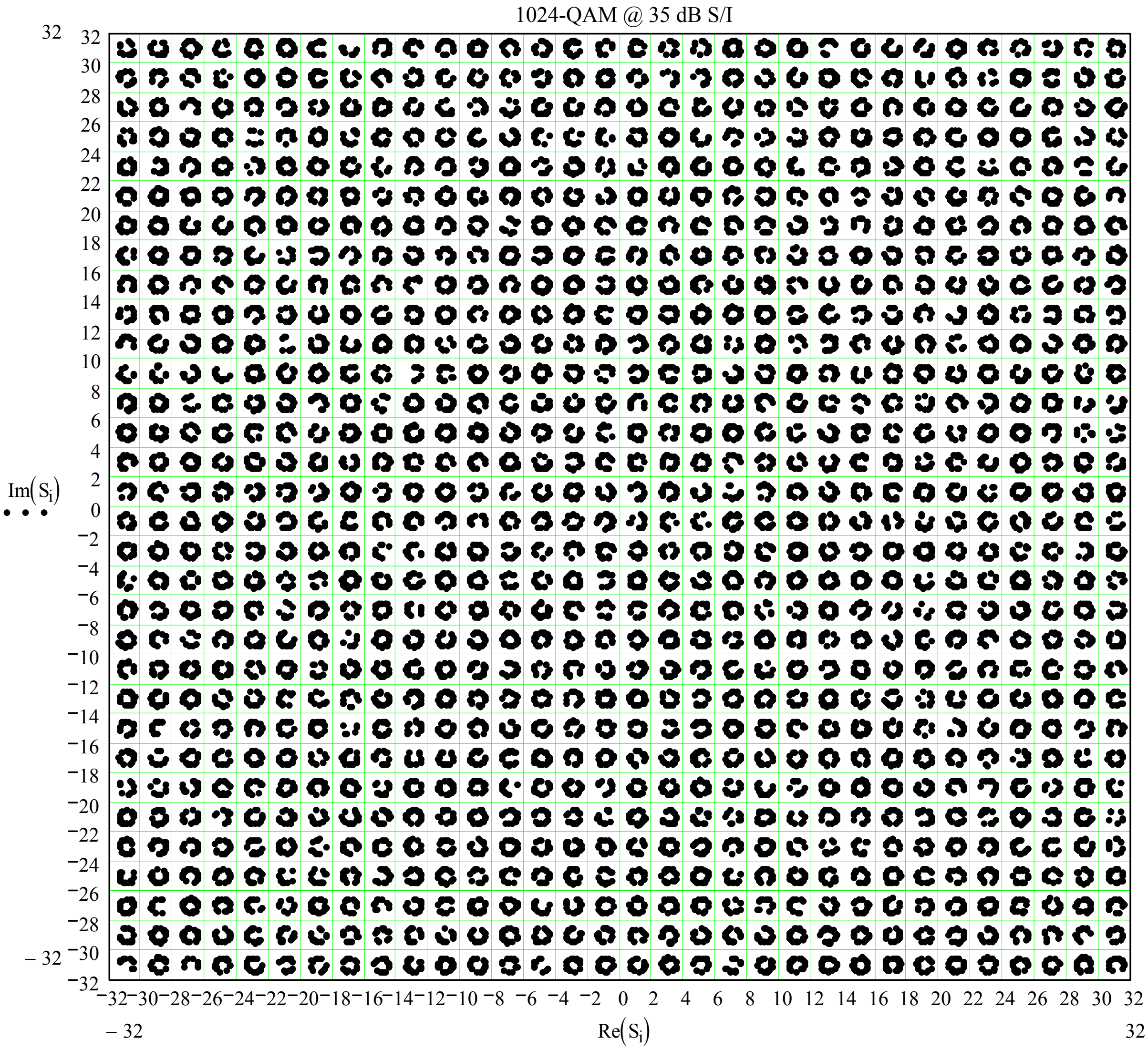
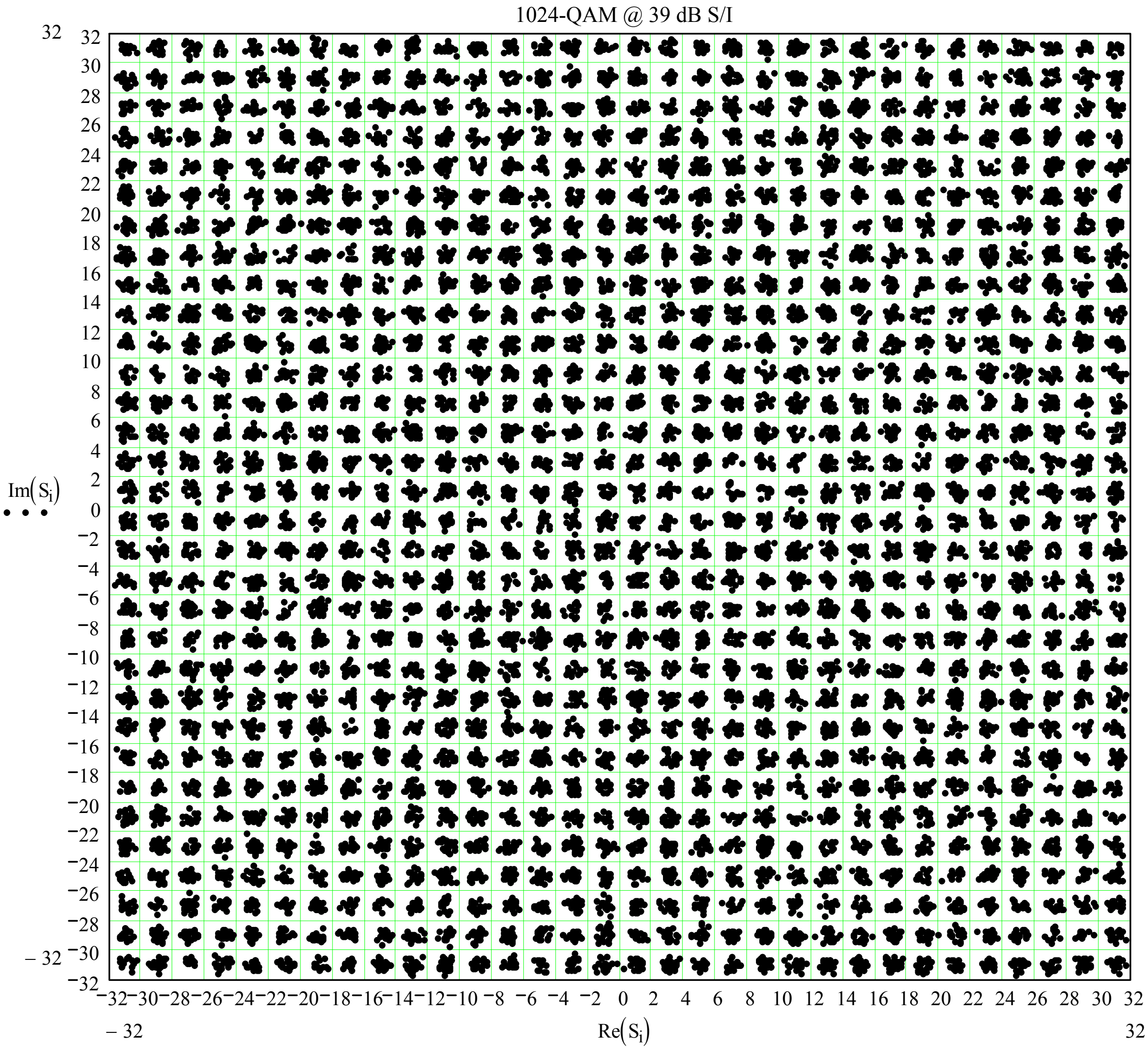



**Figure 3 – 1024-QAM with 35 dB Signal-to-Interference (S/I)**




# Figure 4 – 1024-QAM with 40 dB SNR and 38 dB S/I

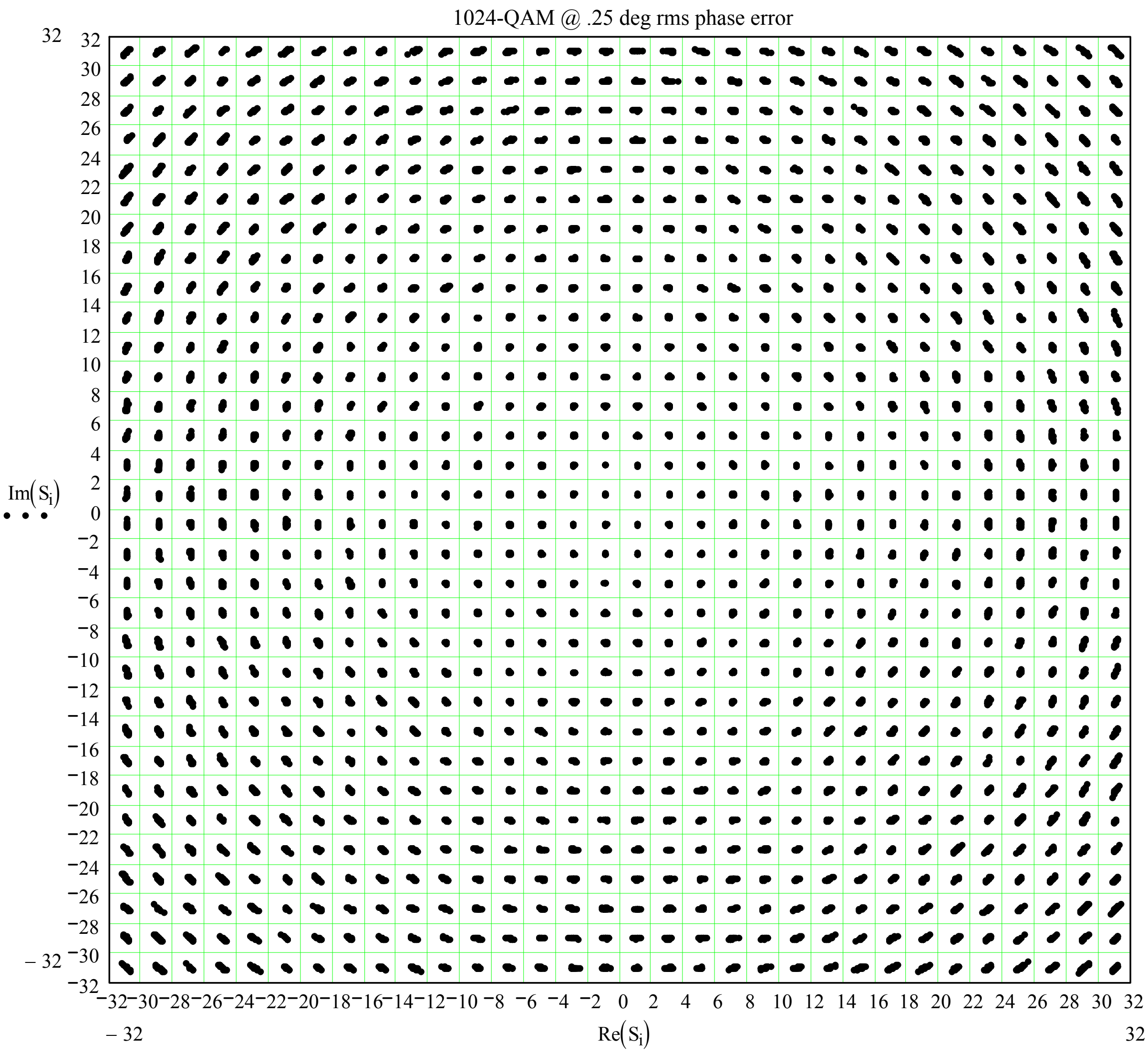



**Figure 5 – 1024-QAM with .25 deg rms Gaussian Phase Noise**

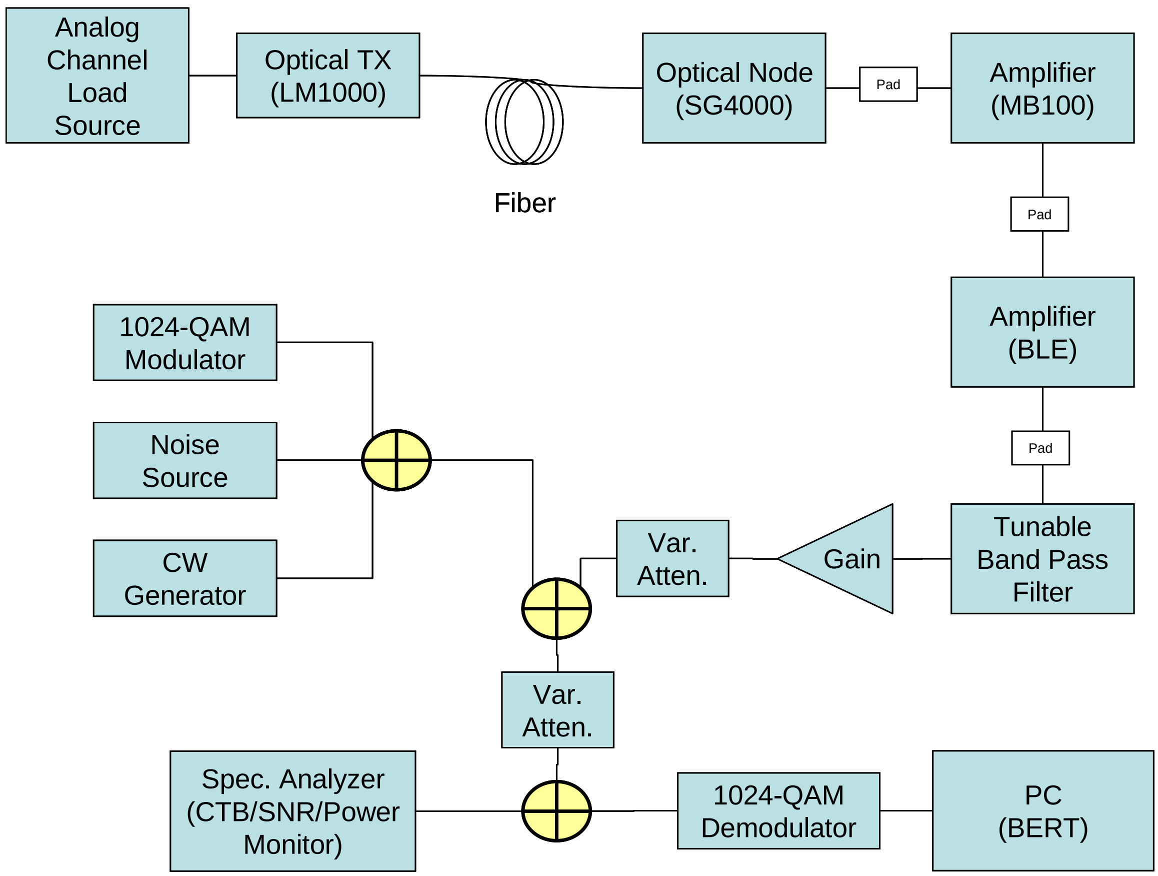



**Figure 6 – Distortion Interference Test Bed**

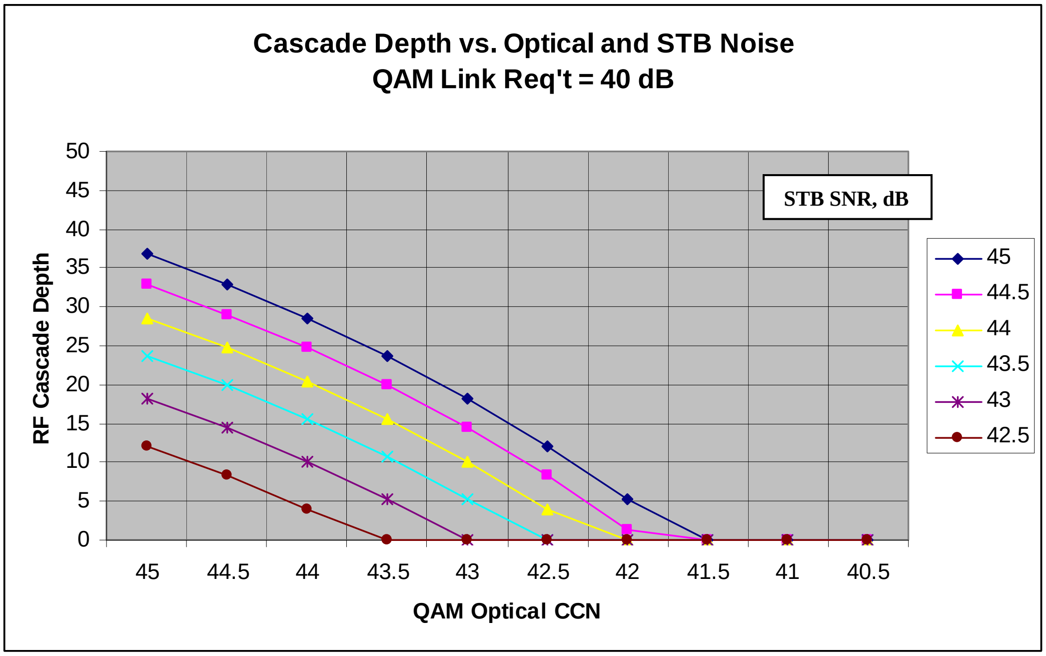



**Figure 7 – Sensitivity of Cascade Depth to STB and Optical Noise Contributions for a 40 dB Link Requirement**

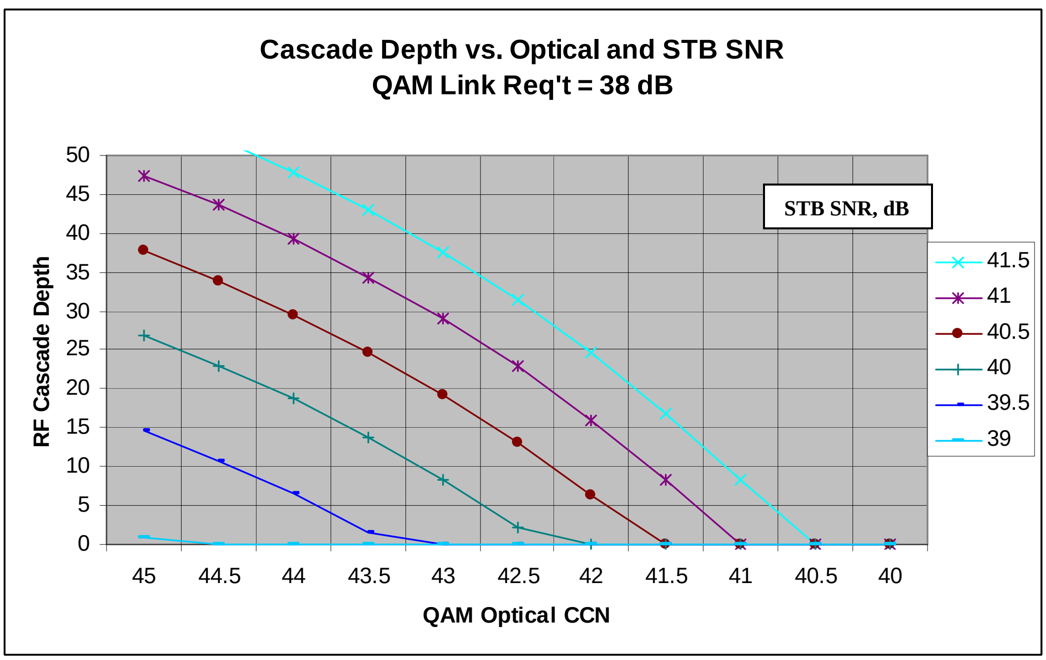



**Figure 8 – Sensitivity of Cascade Depth to STB and Optical Noise Contributions for a 38 dB Link Requirement**

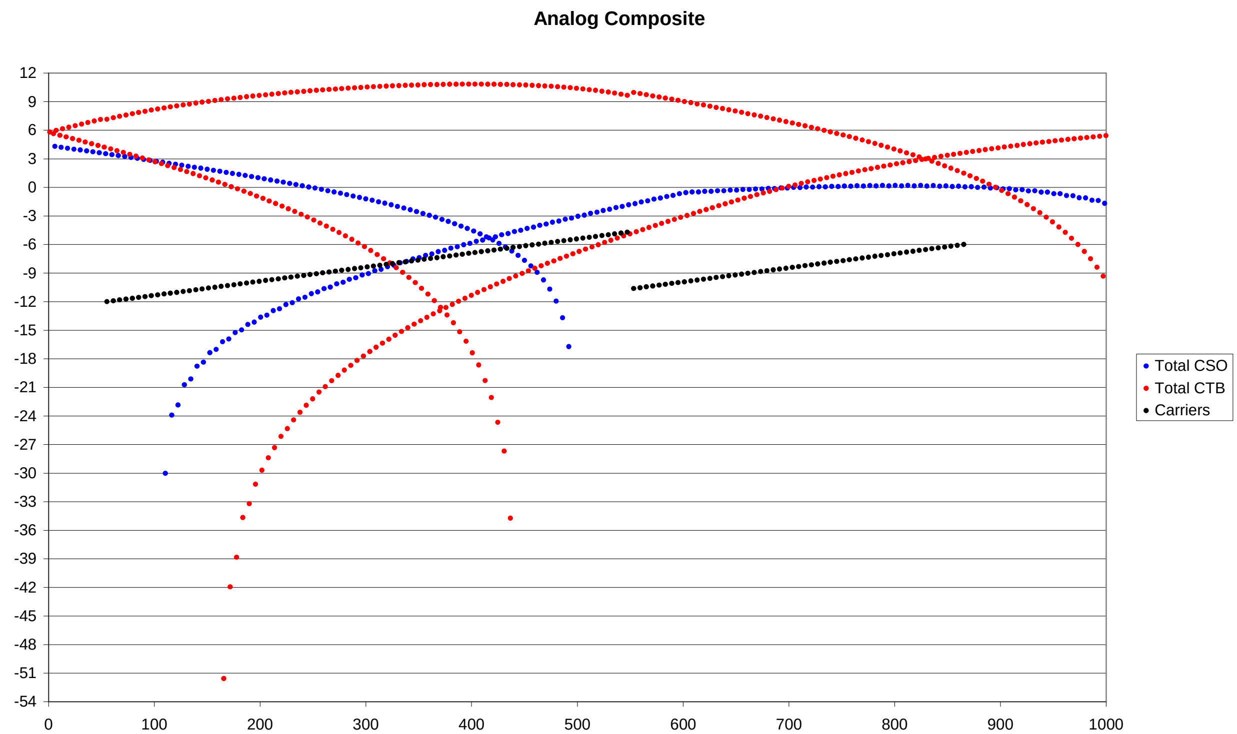



**Figure 9 – Beat Distortion Map of 79 Analog Channels on 12 dB Uptilt**

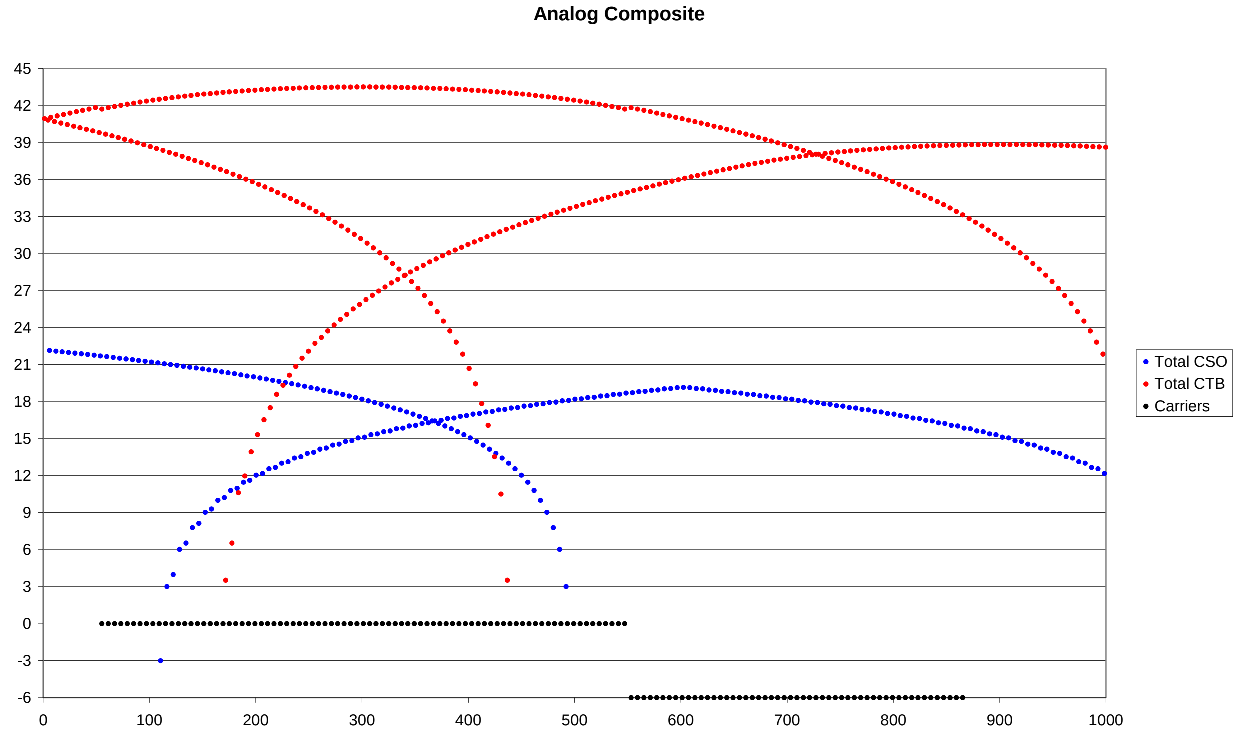



**Figure 10 – Beat Distortion Map of 79 Analog Channels Flatly Loaded**

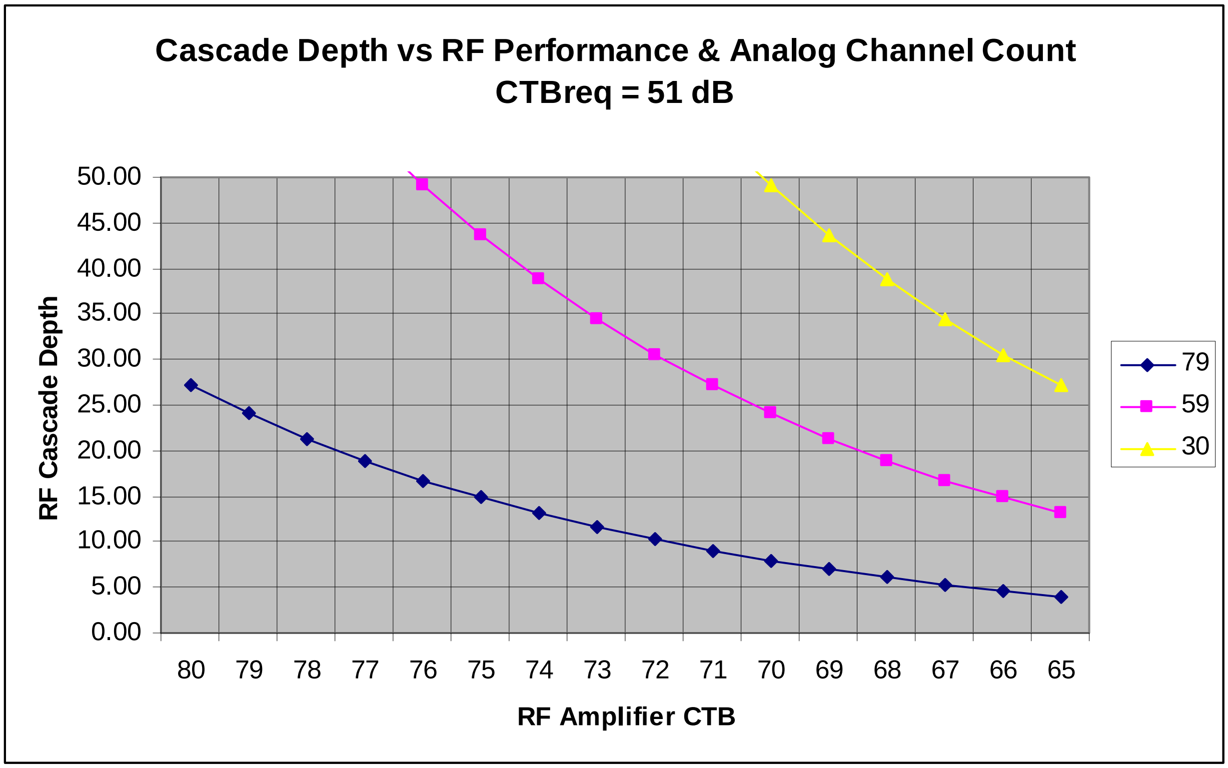



**Figure 11 – Cascade Depth Thresholds vs. CTB Performance and Analog Channel Loading, CTBmax = 51 dB**

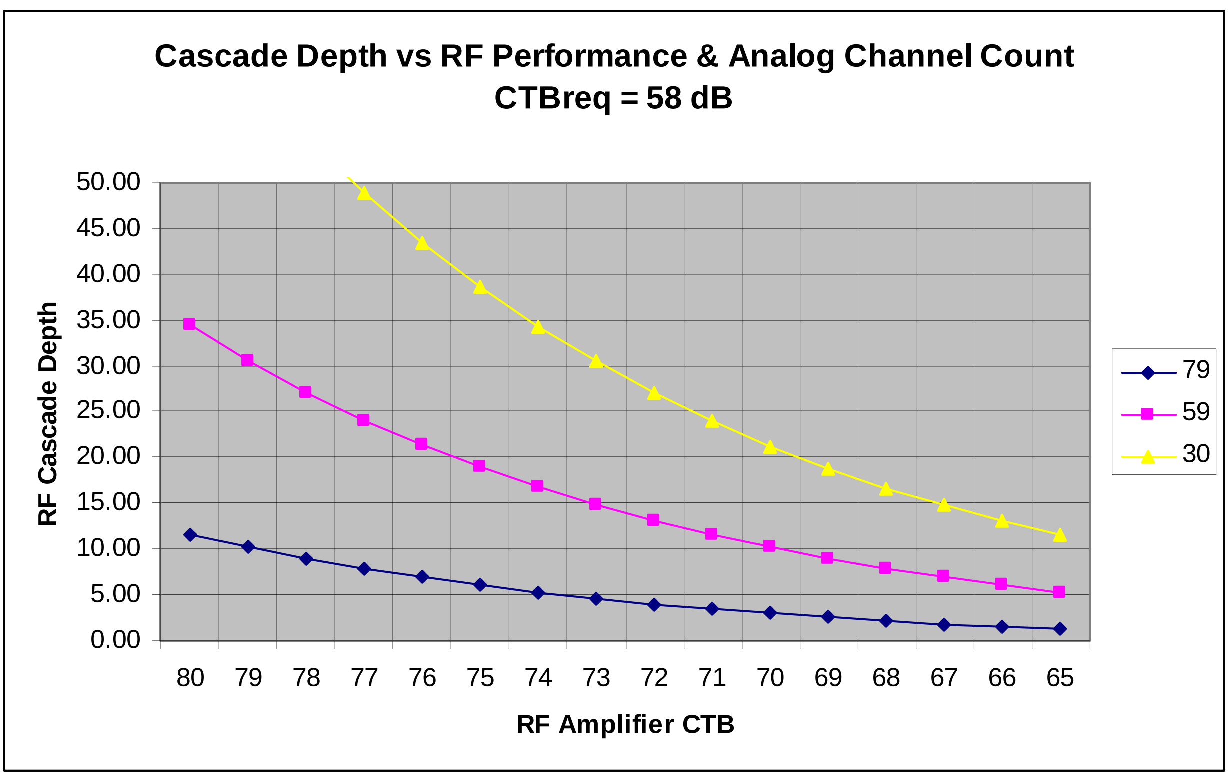



**Figure 12 – Cascade Depth Thresholds vs. CTB Performance and Analog Channel Loading, CTBmax = 58 dB**

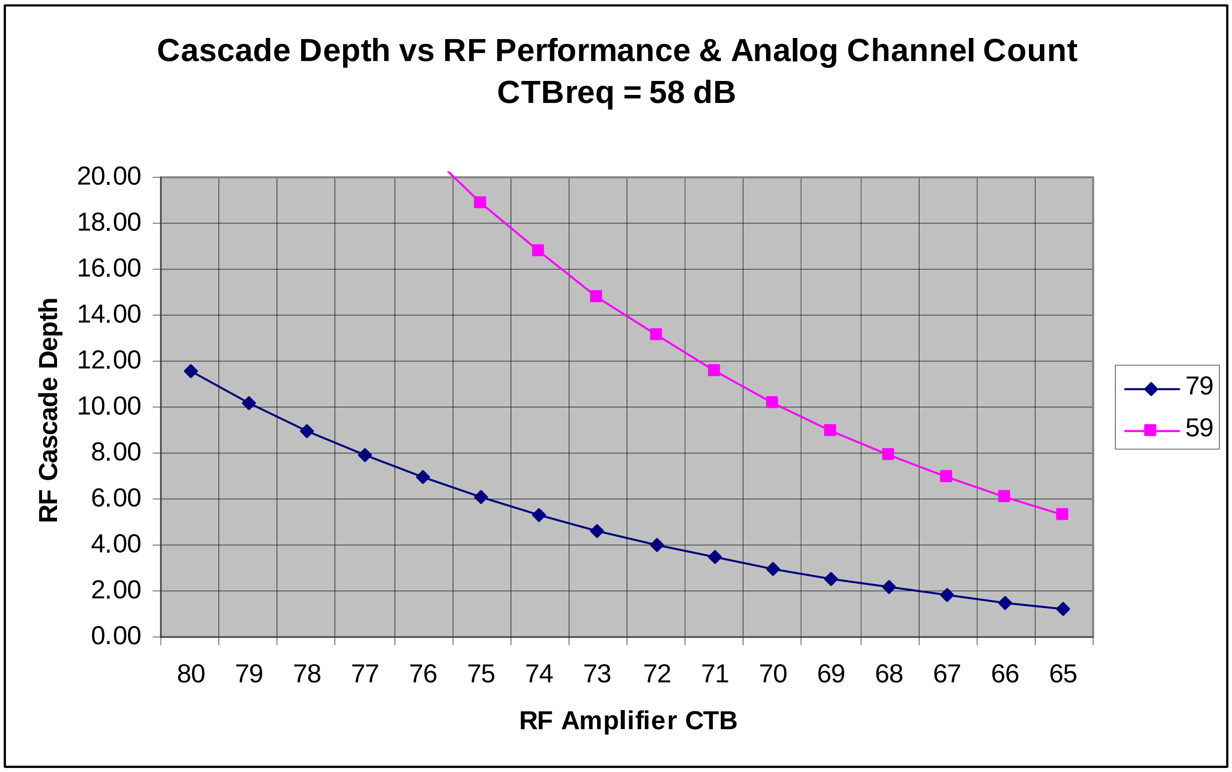



**Figure 13 – Cascade Depth Thresholds vs. CTB Performance and Analog Channel Loading, CTBmax = 58 dB, Expanded for 59 & 79-channels only**

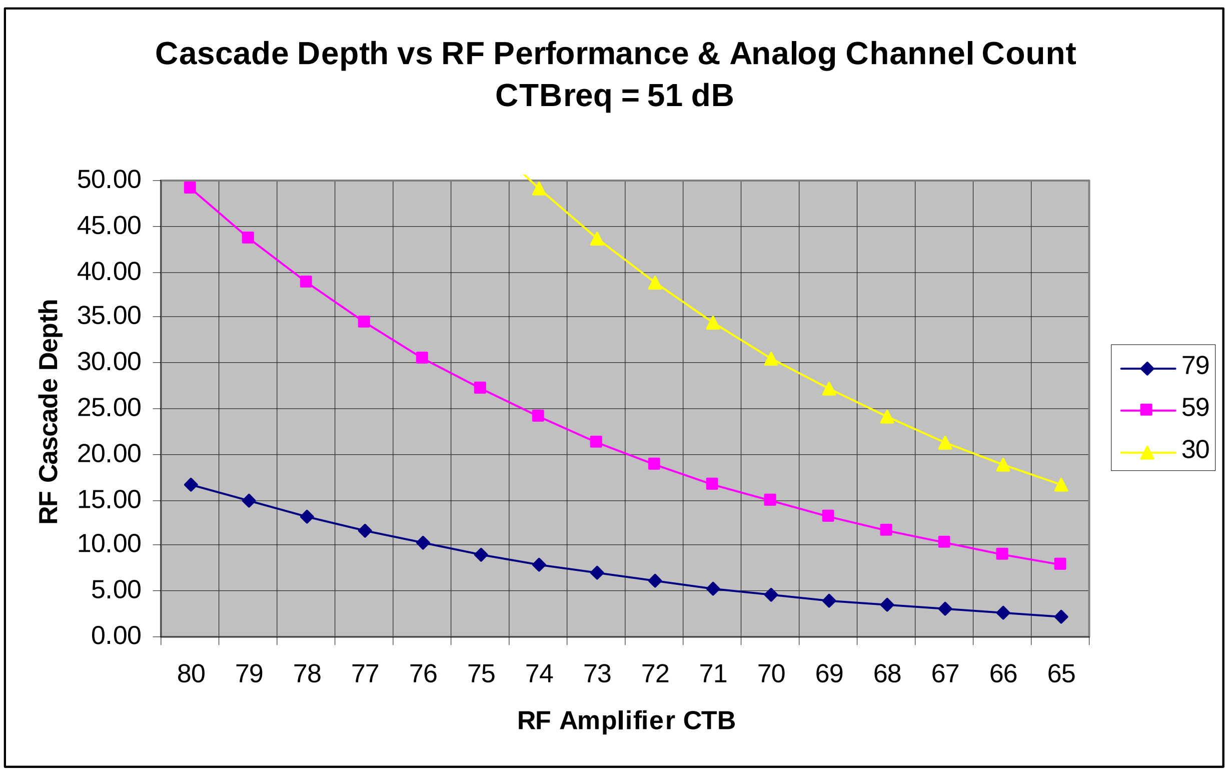



**Figure 14 – Cascade Depth Thresholds vs. CTB Performance and Analog Channel Loading, CTBmax = 51 dB, QAM load @ -3 dBc**