VALUE OF TRANSPARENCY IN THE EVOLUTION OF HFC NETWORKS:

HOW OPTICAL TRANSPARENCY ENHANCES CAPACITY, PROMOTES INNOVATION AND PRESERVES INFRASTRUCTURE IN HFC NETWORKS

A Technical Paper prepared for the Society of Cable Telecommunications Engineers
By

Venk Mutalik
VP of Technology and Strategy
ARRIS Access and Transport
15 Sterling Drive, Wallingford, CT
+1 (203) 494-6556
venk.mutalik@arrisi.com

Marcel Schemmann
Principal Engineer
ARRIS Access and Transport
15 Sterling Drive, Wallingford, CT
+1 (203) 494-6556
marcel.schemmann@arrisi.com

Brent Arnold
VP of Engineering
ARRIS Access and Transport
15 Sterling Drive, Wallingford, CT
+1 (203) 494-6556
brent.arnold@arrisi.com

Overview

HFC infrastructure is a significant, scalable asset for Multiple System Operators. The transparency and flexibility of the HFC plant has enabled Cable MSOs to quickly assimilate new technologies and respond to competitive pressure through a series of cost effective extensions and augmentations to their existing networks. Various examples of this include link extensions via the introduction of fiber optics, the activation of bi-directional plant capability to enable interactive services, the introduction of DOCSIS® (and subsequent revisions) to provide high-speed data and telephony services. This has vaulted the Cable MSOs into position as the premier providers of triple play services, usurping the established telephony carriers who, in most cases, had a legacy network that was opaque, restrictive and unable to migrate effectively and efficiently to meet growing capacity and service demands. Transparency of the HFC infrastructure provided the opportunity for Cable MSOs to quickly adapt their networks to meet the growing needs of consumers with minimal, incremental investments. Preserving the transparency of these networks is a critical success factor for Cable MSOs moving forward.

Operators are soon likely to encounter a changing landscape relative to the multi signal environment in the HFC plant. Despite on-going analog reclamation efforts, many operators intend to preserve analog signals in their plants for some time, potentially viewing this as a competitive advantage. In addition to legacy analog content, future lineups could include several telemetry channels, many DOCSIS® 3.0 (D3.0) HSD and J.83 MPEG Video SC-QAM256 channels along with a growing number of DOCSIS® 3.1 (D3.1) OFDM1024, OFDM4096 and even OFDM16384 channels [1,2]. Advanced error correction techniques such as LDPC will allow these D3.1 channels to approach the Shannon limit for data throughput for a given SNR and bandwidth. Further improvements in silicon processing will enable power envelope management of signals resulting in additional RF and DC power efficiency of next generation telecommunication equipment. This benefit in critical infrastructure becomes more important as the forward bandwidth of cable plant is expanded and more RF power is needed to overcome the increasing cable loss budget. As MSOs continue to face competitive pressure and increased consumer demands for capacity, the ability of the existing infrastructure to support new modulation formats and future innovations with a continued minimal incremental investment demands a high level of transparency in the HFC plant.

Analog Optical distribution links have been utilized in HFC Networks for over two decades, with several million of these deployed in MSO networks worldwide. These simple, elegant, high capacity links are completely transparent to any signal format they have been required to transport. In recent years multi-wavelength systems have been deployed to segment nodes and overcome fiber exhaustion, driven primarily by increased bandwidth consumption. Multi-wavelengths networks in the O band and C-band are currently deployed and have been extensively discussed in previous papers

[3,4]. This paper discusses recent advances in optical technologies that enable and promote high capacity transparent fiber links.

Additionally, recent advances in high-speed optics have introduced intensity detection schemes to the 100 Gbps realm in the form of the CFP standards. One 100 Gbps data stream could be comprised of four, 25 Gbps data streams multiplexed onto a single fiber. With advancements in high speed silicon technology that now matches the capability of 25 Gbps optics, it is now possible to implement a solution similar to the transparent baseband digital return technology (introduced over a decade ago) and enable similar transparency in the downstream up to 1.2 GHz. This paper will provide some of the basic design criteria for such operations in the optical and electrical domains and end with a brief, anticipatory glimpse into the future.

Recently, there has been move in the industry to investigate distributed architectures. In some case there is a desire to move physical layer functions such as QAM modulation to the fiber node. In other cases, there is a desire to move critical elements of the CMTS to the node. To counter the transparency arguments, proponents of these solutions point to the fact that D3.1 protocol is so close to the Shannon Limit (more about Shannon later) that additional innovations are unlikely to add capacity. Technology will continue to evolve, and it has been the inherent nature of networks to change and adapt to this evolution. Therefore, before undertaking a radical change of the HFC architecture itself, it may be wise to understand the capability and capacity of deployed networks, especially in the D3.1 environment. Furthermore, it is well known that transition periods in these networks can feature lengthy durations and it is likely that multiple signal environments will coexist for an extended period of time. We have seen many technology and protocol changes over the course of the last two decades. It is critical that any optical network that is envisioned for the future must be transparent in this context. Conversely, an opaque network is one that is dependent upon a protocol rather than the physical layer transmission and therefore runs the risk of stranding significant capital when the protocol changes.

Optical transparency has been a critical parameter for MSOs over the past two decades, enabling them to efficiently and quickly meet the growing needs of their customers and evolving their business model from that of simple video distribution to premier providers of triple play services. This paper is a critical step towards understanding future options available for the continued deployment of transparent networks and provides a roadmap for the future.

Recent Advances in Analog optics

For many years, analog optical systems used O Band transmitters typically located in the 1310nm zero dispersion wavelength region. As Multi-wavelength systems were developed in the early 2000s, O Band transmitters shifted away from the 1310 nm zero dispersion band to eliminate the 4WM effects that are particularly excessive when one of the signal wavelengths are coincident with the zero dispersion of the fiber on hand.

Today it is typical to find 4 or 8 wavelengths used in the 1291nm band or the 1331nm band with a stable and robust 4WM free wavelength plan. While the O band is an attractive solution (primarily due to the lower cost of direct modulated technology (DML)), link budget can be limited due to a lack of optical amplification. As a result, this technology is limited to a maximum distance of around 40 km, with typical links around 25 km.

In recent years, Electro Absorption Modulator Lasers (EMLs) in the C Band (1525-1565nm) have evolved to provide analog grade performance at attractive price points. EML's feature near zero chirp and therefore enable the use of commercial DWDM optical passives to span long links in the C band as they are generally impervious to fiber dispersion. These devices have revolutionized the medium and long distance links. Thus it is common to see up to 16 wavelengths in the downstream and 16 in the upstream with reaches approaching 80 km. As noted in previous papers, the ability to design MWL systems rests upon a complete understanding of optical effects and non-linearities in the transmitters and over the optical fiber. Stable and robust wavelength plans that simultaneously minimize 4WM, SRS and XPM must be considered and have been presented before. A helpful taxonomy of the optical effects is presented below [3].

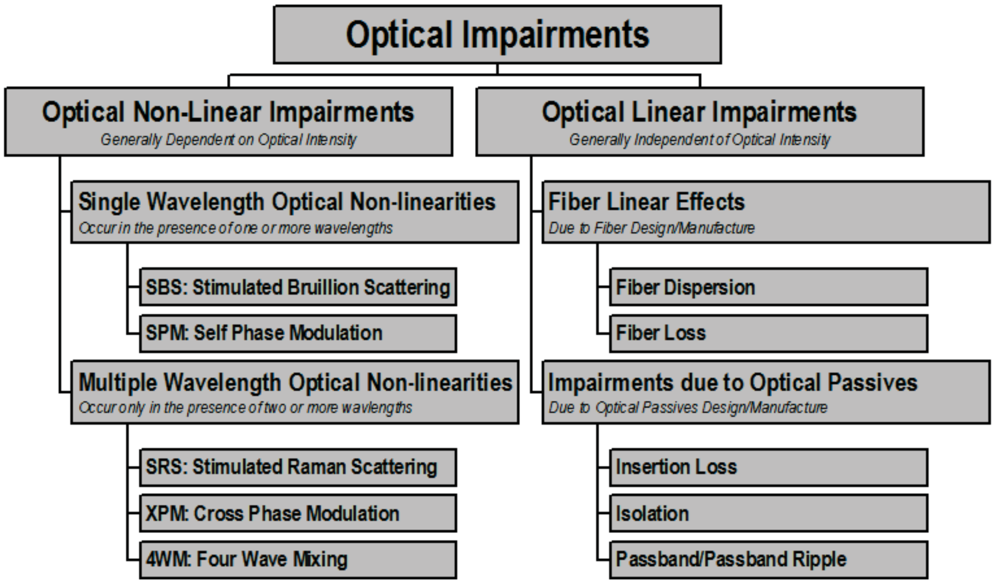


Figure 1 Taxonomy of Optical Impairments

A recent feature available in analog optical receivers is the optical AGC, which enables Plug and Play capabilities for analog links. These receivers have a simple circuit that changes the gain in the receiver in response to the optical input power, thus providing a constant RF power regardless of optical input to the receiver. This feature reduces truck rolls to adjust node output levels which, if left unchecked could compromise plant performance. Additionally, new optical receiver technology features significantly reduced input noise. This lower noise profile enables lower optical input power which provides the system benefit of extending link budgets.

There has been continued focus on critical infrastructure relative to headend/hub optics and optical nodes, with a substantial improvement in the air, power, space and ergonomics. Higher density transmitters with integrated broadcast ports reduce the need for additional RF splitting/combining/amplification networks at the headend that surprisingly, can consume a significant amount of space and power. Often, optical and RF connectors can also limit density improvement. However, smaller form factor LC APC optical connectors have helped increase the density of optical splitters/combiners and multiplexers and similarly innovative RF connector technologies have helped improve density of the opto-electronics devices overall.

Optical Non-Linearities and the RF spectrum

A look at the optical non-linearities as illustrated in the figure below suggests two distinct features. The Stimulated Raman Scattering effect depends upon the optical wavelength separation, and worsens with wider optical spacing and is primarily seen at the lower RF spectrum, generally below 200MHz. The Cross Phase Modulation (XPM) nonlinearity on the other hand worsens with increasing wavelength spacing and is primarily seen at the higher RF spectrum. Both these non-linearities are worse when the wavelengths are all co-polarized and minimal when orthogonally polarized [4].

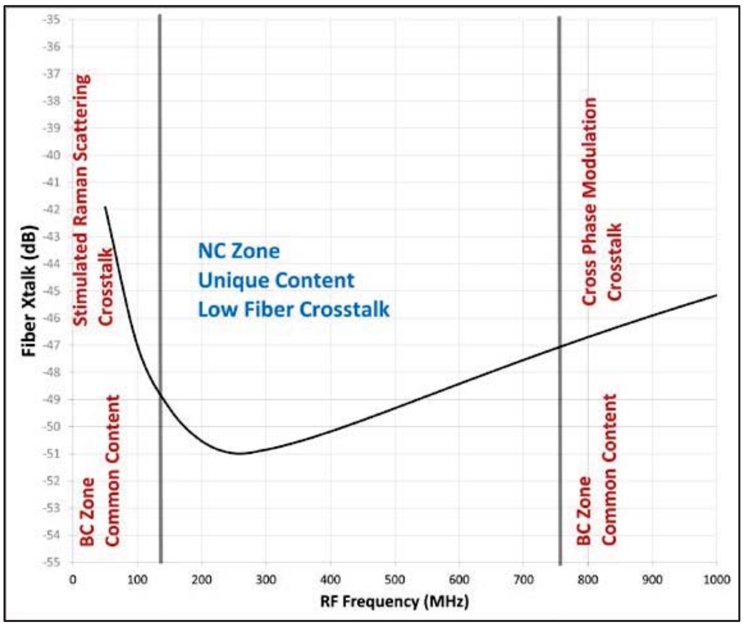


Figure 2 Illustrating worst-case Optical Cross-talk across the RF Frequency domain for a typical C-band wavelength plan. SRS effects shown are for wavelengths that are the farthest and the XPM effects shown are for wavelengths that are closest

In a multi-wavelength environment there is still a fair amount of common content or broadcast content that is common across multiple wavelengths on a single fiber strand. There is also a fair amount of narrowcast content or unique content to each wavelength on the single fiber strand. It is important to remember that the common content is generally impervious to optical crosstalk discussed earlier, while the unique content is generally affected by the optical crosstalk effects. Substantial benefit can be achieved if the RF spectrum is properly harvested.

Presented below is a typical RF spectrum. Traditionally, in North America, the spectrum from 50 to 550 MHz is occupied by Analog VSB channels, with 6 MHz spacing between each channel. For testing purposes, we generally assume these channels to be NTSC CW carriers with the understanding that the modulated carriers used in live systems might be around 3 dB lower in average power as the CW carriers. For all subsequent discussion in the paper, we assume CW carriers as a representative of the analog channels, unless otherwise noted. The spectrum from 550 MHz to 1GHz (sometimes only to 750 MHz or 870 MHz) is usually occupied with 256QAM channels, each of which is 6 dB below the equivalent CW carrier (therefore these may only be around 3 dB (actually between 3 to 8 dB) lower than the average power of an equivalent AM VSB channel). These 256QAM channels today follow the ITU J.83 protocol and could support streaming MPEG Video (for broadcast or narrowcast purposes) or the DOCSIS 3.0 protocols for HSD (generally for narrowcast purposes). The RF levels are presented for the Analog and QAM sections in dBmV and the total composite RF level is indicated as well. It is evident for the transmitter that the maximum amount of RF power is concentrated in the lower part of the RF spectrum.

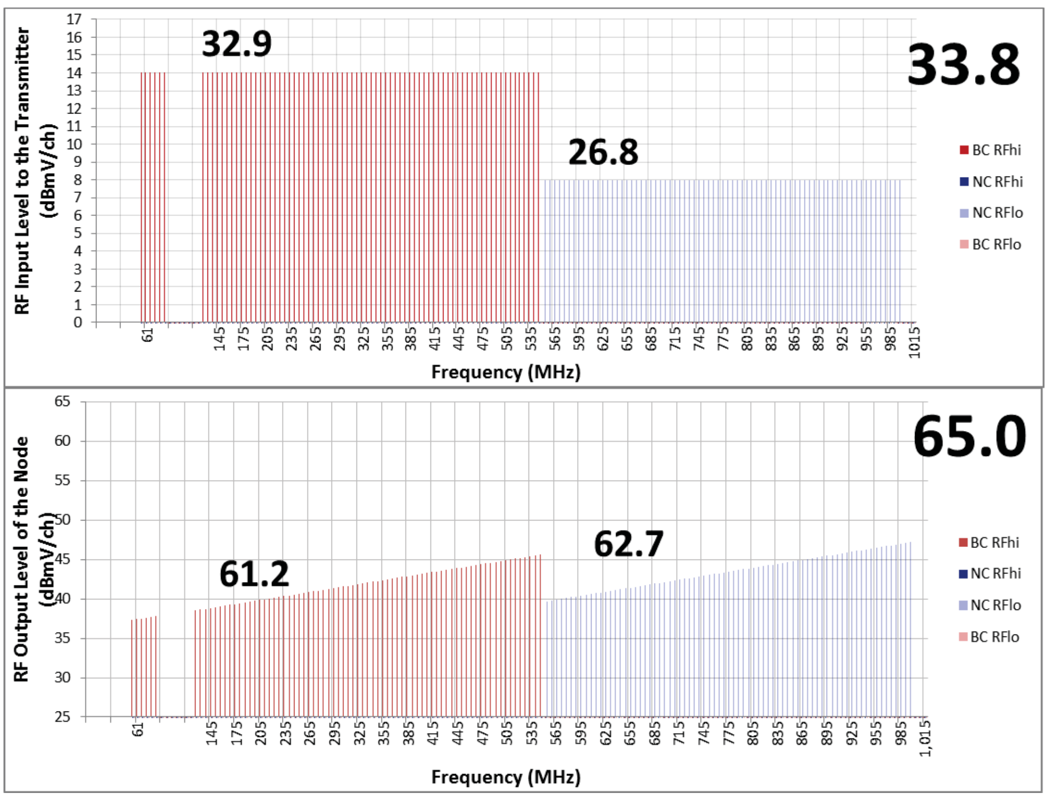


Figure 3 Illustrating RF spectrum and Levels at the headend and the node

However, for the optical node the situation is different. Presented just above is the typical RF output at the Node output. To compensate for higher cable loss at higher frequencies and to equalize carrier-to distortion and carrier-to-noise ratios across the band, it is typical to have tilted RF output levels for the node and the amplifier devices. As can be seen here, the total RF power for the node is dominated by the RF levels at the higher part of the frequency spectrum, even though the RF input to the node has far less power than the analog channels at the lower part of the spectrum. Therefore even a modest increase in the RF levels at the higher spectrum at the transmitter translates to potentially high RF levels in the node due to the tilt. If the node was set just below compression to begin with, any increase in the RF level, especially at the higher frequencies at the transmitter could quickly push the node into compression. One way to alleviate this is to have a truck roll to the affected node, but this is expensive and time consuming.

RF Levels in Transition

More and more MSOs are harvesting their analog channels and replacing them today with 256QAM channels. Tomorrow they may replace these with D3.1 channels. As they replace their analog channels, they are faced with the choices below:

1. Use the transmitter AGC to maintain the total RF power constant at the transmitter.
2. Keep the RF level per channel for the 256QAM channels just as today and add new 256QAM channels for the replaced analog channels.

These scenarios are demonstrated below, at the transmitter input and at the node output.

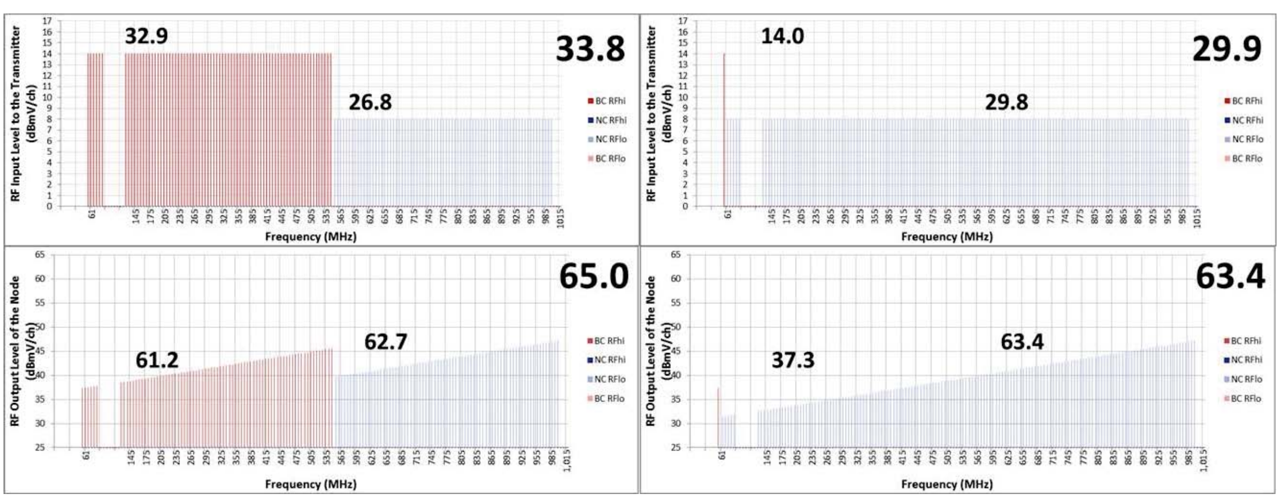


Figure 4a Illustrating RF levels in the MGC mode

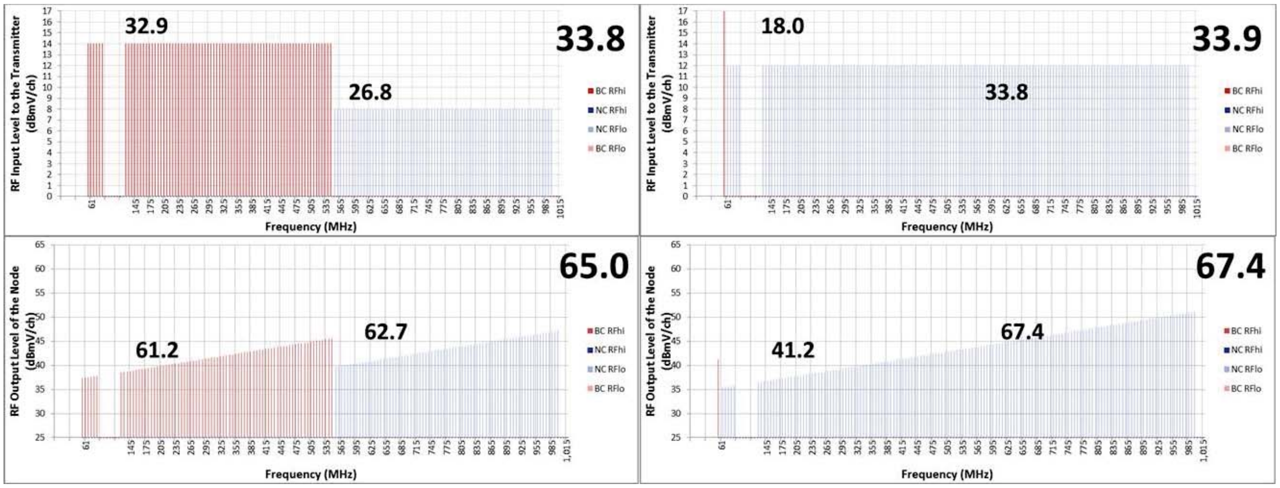


Figure 4b Illustrating RF levels in the AGC mode

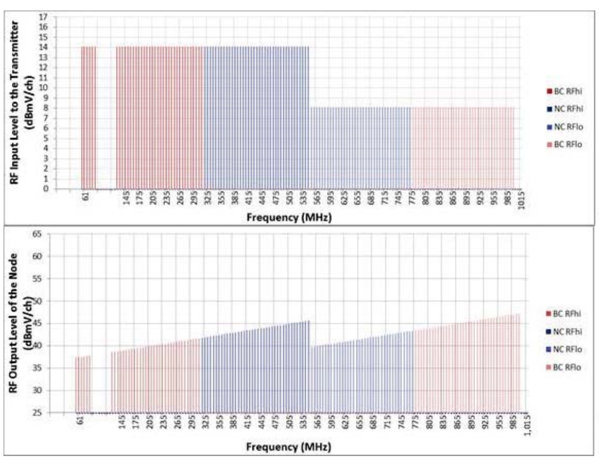
As can be seen from the first figure above, when the optical transmitter is in the MGC mode in the headend, analog harvesting to replace analog channels with 256QAM channels at lower operating levels would always result in lower composite RF power at the transmitter. In this example it is seen that migrating from a 77 Analog +75 QAM channel case to an essentially all 256QAM case results in a net reduction in the composite RF power at the transmitter by about 4 dB. The key benefit here is that the node also experiences a reduction in the RF levels and by definition will not compress if it was not compressing at the initial set up. In this example the node has experienced around 1.5 dB of net RF level reduction, bringing with it the concomitant improvement in CIN [5].

In the second figure above, it is seen that the transmitter is maintained in an AGC mode. In this scenario, the total composite RF power at the transmitters remains the same. In the present example, the RF levels of all QAM channels undergo an increase of around 4 dB while the total composite RF power remains the same. While this may seem like a good tradeoff as it might result in a higher MER or SNR at the transmitter optical link, a look at the node reveals a different story. At the Node it is observed that due to the tilt, the total composite RF power at the node has increased by around 2.5 dB. An increase in the total composite RF power at the node could have a cascading effect across the amplifier chain and could result in substantial compression thereby increasing CIN and reducing effective MER or SNR, thus defeating the purpose of the AGC mode to begin with. The only way to accrue the potential increase in the MER on the optical link, however slight it may be, is to send a truck roll to the node and rebalance the node output. In many cases the additional expense and delay of the truck roll is not commensurate with the potential improvement in the overall MER or SNR. On the contrary, not rebalancing the node is to leave oneself in the danger of compression, which is not a good tradeoff on the whole.

In this paper, it is assumed that the transmitter is always operated in the MGC mode and has been initially set up to a 77 Analog +75 QAM load for a 1 GHz operation. It will also additionally be assumed that the initial set up was with a CW load for the analog. Other analysis using different channel loads and assuming video channels instead of the CW could be done, but would result in numerous options and is therefore not attempted in this paper.

Figures of Merit for the New Network

Before we proceed further with the analysis, it is helpful to understand key figures of merit, such as MER and BER. For many multi-wavelength systems today, it is very common to provide just one figure of merit, that being MER. It has been discussed in previous papers [5] that “Not all MER is created Equal” and the performance of a system can be dramatically different for the same MER but under different conditions such as clipping, phase noise, fiber crosstalk or optical Four Wave Mixing. In a later section, we discuss impairments that might affect a successful 4KQAM transmission while studying the constellation diagrams. It is therefore important to be aware of these impairments and therefore operate the transmitters and the system elements in their linear region and not in their compression or clipping region. Even a superior MER acquired in a clipping region is not as beneficial as a lower MER seen in the linear region from a BER perspective. For all the simulation in this paper, it is assumed that none of the devices are in clipping or compression.



RF Hi	HIGH SNR OPTICAL XTALK Present MID PERFORMANCE	HIGH SNR No OPTICAL XTALK HIGH PERFORMANCE
RF Lo	LOW SNR OPTICAL XTALK Present LOW PERFORMANCE	LOW SNR No OPTICAL XTALK MID PERFORMANCE
	NC QAMs	BC QAMs

Figure 5 Quad Figures of Merit

Consider the figure above. Until recently, analog channels have been a dominant RF load contributor to the transmitters. The 256QAM channels are typically 6 dB below their analog counterparts. However as we have made a case before, as analog channels are harvested and replaced, perhaps by the higher order QAM modulation formats in the DOCSIS 3.1, we recommend that they be placed in the space currently occupied by the analog channels at similar average RF levels as the analogs are today. Doing so would enable trouble free operation of optical transmitters and of the RF chain currently in existence as both transmitters and nodes operate below clipping or compression.

In a future scenario where all of the analog channels are perhaps replaced by D3.1 channels, some of these D3.1 channels could be in the BC mode or be in the narrowcast mode. Similarly some of the QAM channels at the lower RF levels might be in the BC mode while others are in the NC mode. Please note here that the lower RF levels may be operating with D3.0 or D3.1 channels.

With these assumptions, we now have 4 distinct zones

1. A zone with Higher RF levels but with common or broadcast content – BC RF hi
2. A zone with Higher RF levels but with unique or narrowcast content – NC RF hi
3. A zone with Lower RF levels but with unique or narrowcast content – NC RF lo
4. A zone with Lower RF levels but with common or broadcast content – BC RF lo

Based on previous discussion of optical non-linearities such as SRS and XPM, we note that it is advantageous to place the common content or BC content at the spectrum extremities and have the unique content in the middle of the band, perhaps extending from around 150 MHz to 800 MHz.

Based on our previous discussion of RF level at the node, it is advantageous to have RF levels at the higher frequencies lower so as to not compromise the performance of the nodes and RF chains that follow. Therefore if the original systems were all designed for 550 analogs, it is best to limit the RF hi levels to 550 MHz and save additional truck rolls to the field.

With this understanding, it is now possible to analyze the performance metrics for the 4 zones

1. The BC RF hi zone has high SNR and has no effects of Optical Crosstalk. Therefore this zone has the best SNR or MER.
2. The NC RF lo zone has low SNR and is afflicted with the effects of optical crosstalk. Therefore this zone has the lowest SNR or MER.
3. The NC RF hi zone has high SNR but has the effects of Optical crosstalk and therefore this zone has a mid-level SNR. The final MER or SNR depends upon the optical design and the presence of optical crosstalk.
4. The BC RF lo zone has low SNR, but has no effects of optical crosstalk. Therefore it performance is mid-level and its SNR or MER depends upon the optical component design.

Today, if a Figure of Merit were given for system performance, it is generally (2), i.e., where the MER number reported is exclusively the worst case MER for unique content QAM channels. If one neglects the performance of other three zones, this could lead to a very pessimistic estimate of the capacity of these links.

A good understanding of these 4 zones thus enables one to understand how a system could be loaded to achieve the maximum realizable capacity with minimal field intervention. It may be noted here that there are additional ways of manipulating the optical spectrum. These options being far too numerous, are not attempted in this paper.

Analog Harvesting and the Digital Dividend

We now begin to analyze the various options in light of the above discussion. Specifically, we realize that:

1. The transmitters may be in the AGC or MGC mode. For this paper we have assumed that transmitters are always in the MGC mode and have been originally set up with 77 Analog ad 75 QAMs for purposes of this paper
2. Both D3.1 or D3.0 or other signals such as EPoC may be used on the transmitter. In this paper we have made appropriate notations where applicable.

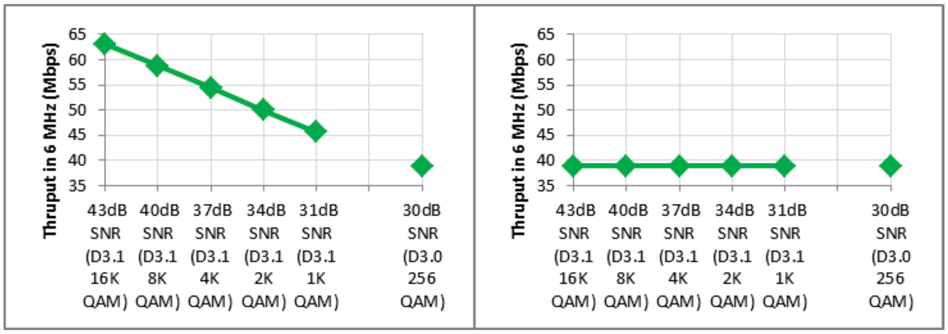


Figure 6 DOCSIS 3.1 and DOCSIS 3.0 Throughput

Specifically, the figure above gives the anticipated throughput for various SNR conditions based on OFDM/LDPC protocols in an equivalent 6MHz bandwidth. It is seen that for the D3.1 case, higher SNR of the demodulated signal actually results in higher throughput, from 1KQAM all the way to 16KQAM. That however is not the case for D3.0, where any SNR above that needed for decoding 256QAM is essentially superfluous and does not result in additional throughput.

3. The D3.1 assumes the use of multi modulation profiles (MMP). A complete description of the MMP is not this paper's focus, but we realize that the role of performance margin is potentially less critical in the case of D3.1. If a design is attempted for 16KQAM and for some reason cannot achieve this performance, the system automatically reduces to a lower, more robust modulation format, such as 8K QAM in this case. This is in stark contrast to D3.0, where losing the SNR for 256QAM results in a wholesale loss of sync and zero data throughput. In the simulations, we calculate the SNR for the various optical, RF cascade and

CPE environments, we then compare the SNR to the throughput capacity graph above to make an estimate of the throughput of the entire wavelength.

4. It is possible that varying amounts of common content (BC) or unique content (NC) may be allocated over the RF hi and the RF lo regions. For purposes of this paper, we have always assumed an equal spread of BC and NC signals.
5. It is also possible that the BC and NC signals may be allocated to multiple areas of the spectrum. In this paper we have assumed that the BC is at the spectrum extremities and the NC is ensconced in the middle.
6. Various RF chains are possible. In this paper we have only analyzed N+3 and N+6 cases.
7. Various CPE inputs levels are possible, considering that the common expectation is from -15 to +15 dmV/ch. In this paper we have analyzed a CPE input of -10 dBmV for a CPE NF of 10 dB. In addition, we realize that a move to D3.1 would also entail more of a gateway type of an architecture, where the CPE is fed directly from the drop to the house. In such a case the CPE does not encounter the RF splitting at the input panel and the RF level into the CPE is much higher. Therefore we have also analyzed a case for the CPE input of a nominal 0 dBmV/ch. with a 10 dB NF (noise figure). It is to be noted that the RF levels to the CPE generally vary, especially when it is considered that the RF level difference itself causes a 6 dB change (as assumed in this paper)
8. Finally, it is highly likely that there will be the simultaneous transmission of multiple formats in the Cable plant. We have assumed a single scenario for this paper where the RF lo signals are exclusively D3.0 256QAM and the RF hi are exclusively D3.1 according to their SNR profiles. This scenario is the one that enables one to see the power of transparency and the additional capacity possible during the transition timeframe

We will now begin with a C-Band 16 wavelength analog system. As described in previous papers, it is critical to have a stable and robust wavelength plan with respect to 4WM and one that simultaneously minimizes SRS and XPM. Using such a stable and robust plan, one may use 16 DWDM externally modulated transmitters (using EML devices) and pass through a multiplexer port, and traverse 40 km of fiber followed by a de-multiplexer. Individual wavelengths of this demux output arrive at the node receiver. One may use the optical AGC node receiver to maximize or stabilize further the RF levels as described in one of the sections above. The Node output, which is tilted by approximately 16 dB, then passes thru the RF chain. While N+0 is one option, most MSOs have typically between 3 to 6 RF elements before the home. At the home, the RF drop is either let thru the RF splitter network before it goes thru to the CPE. Alternatively it may be directly fed to a home gateway device.

C-Band 16 WL 40 km D3.1 Design:

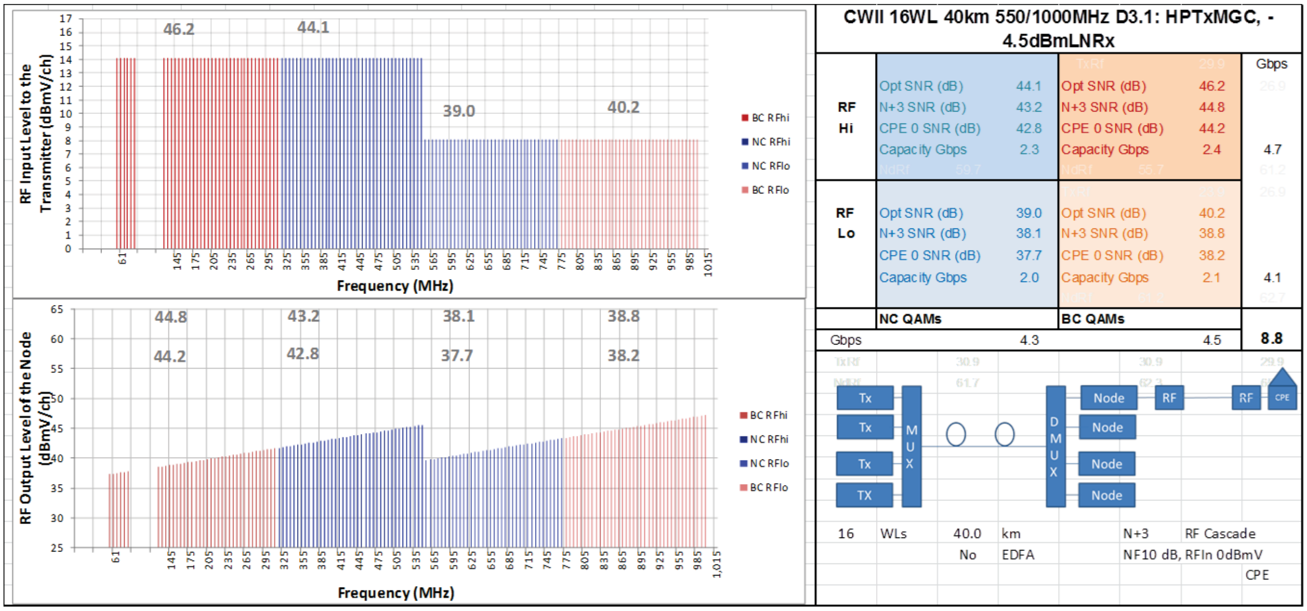


Figure 7 C Band 16 WL 40 km D3.1 Design: At Max RF hi Frequency of 550MHz

The figure above represents the case of a 77 Analog + 75 QAM link being harvested and replaced by a D3.1 QAM Load with all channels below 550 MHz having similar power levels as the CW channels they have replaced and the remaining RF load above 550 MHz at the same RF level as the 256QAM channels they have replaced. As mentioned previously, we have assumed half of the load to be common content (BC) and is spread out to the extremities of the spectrum and indicated in Red. The NC or unique content is in the middle and is shown in Blue. A block diagram of how the system is put together is shown towards the bottom right. Critical performance parameters are shown towards the top right in a Quad Chart. A look at the estimated SNR or MER numbers when converted to capacity based on the D3.1 throughput values indicates that this link has a total capacity of 8.8 Gbps. As a quick reminder, the total capacity of a 1 GHz system, all carrying 16KQAM is around 9.6 Gbps. A system carrying D3.0 would have a capacity of 5.9 Gbps.

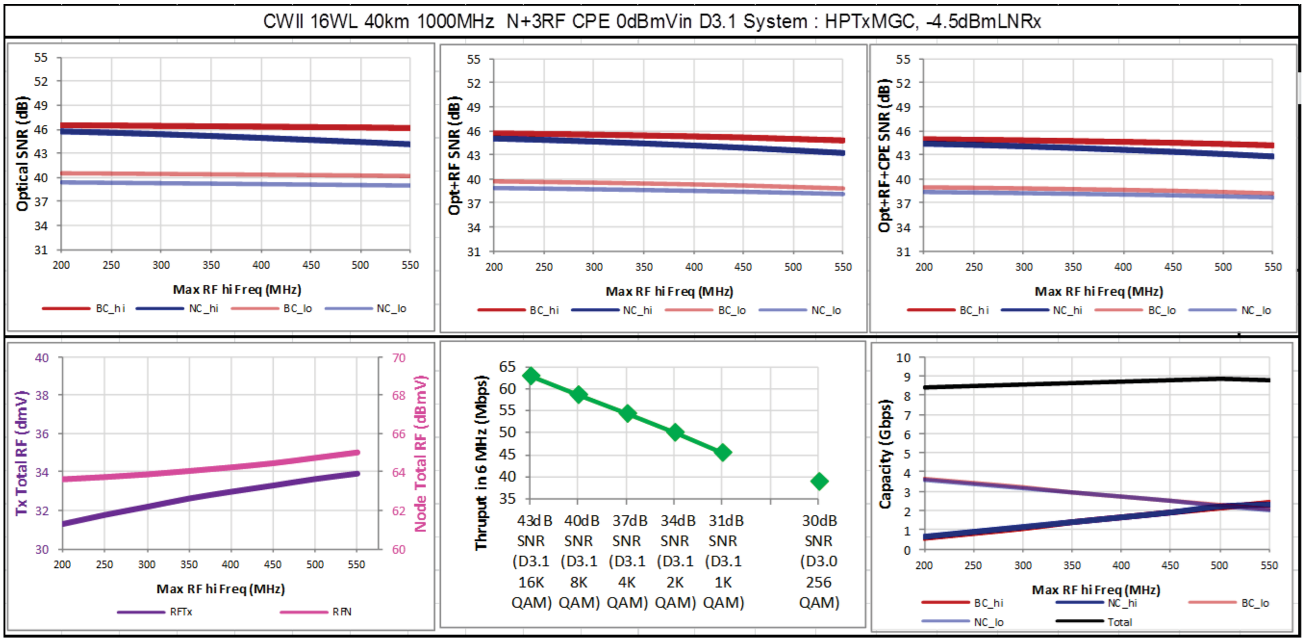


Figure 8 C Band 16 WL 40 km D3.1 Design Detailed Analysis

The set of figures above depict what happens when this system is loaded with different maximum RF Hi frequencies. The SNR or MER variation is shown for the optical link, thru the RF chain and at the CPE with a 0 dBmV/ch. input. One can see on the bottom left that the RF level at the transmitter changes as the transmitter is in the MGC range and the RF level at the node changes too, but is never above the nominal value of node compression. The bottom middle indicates graphically the D3.1 throughput ranges and the bottom right describes that this link always has a capacity exceeding 8.5 Gbps.

It is critical to know that this significant amount of capacity is approached by a well-designed conventional link using standard optical components and over existing RF cascades. We now proceed to analyze various additional configurations that give a more rounded view of the capability of existing links.

C-Band 8 WL 80 km D3.1 Design:

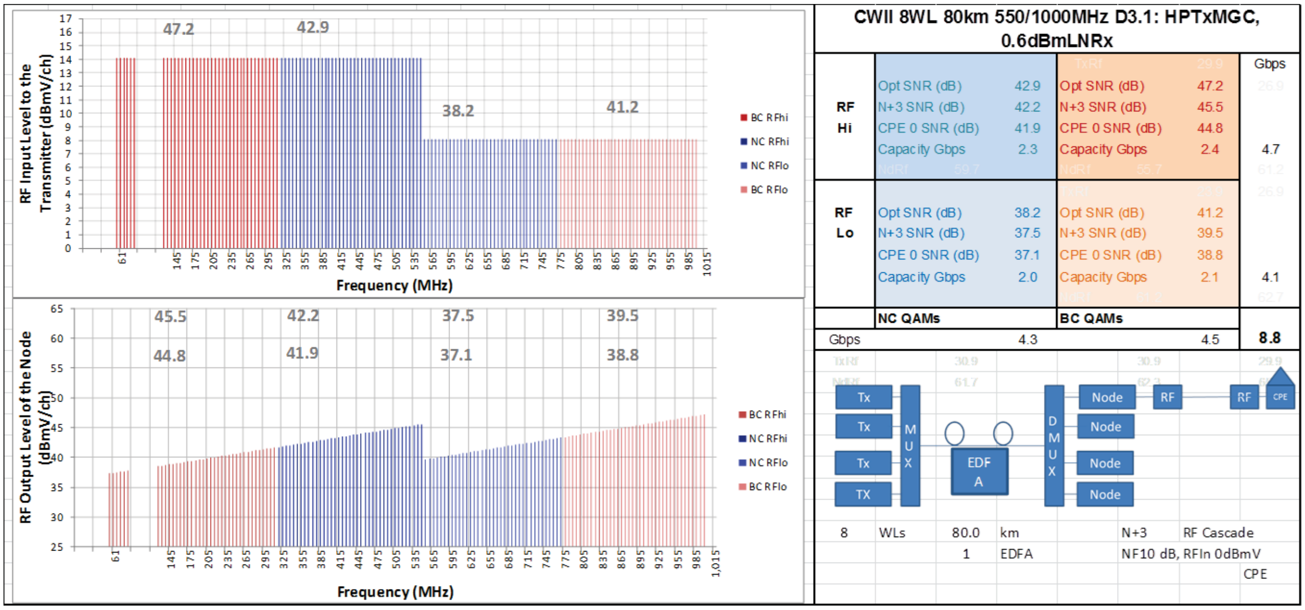


Figure 11 C-Band 8 WL 80 km D3.1 Design: At Max RF hi Frequency of 550MHz

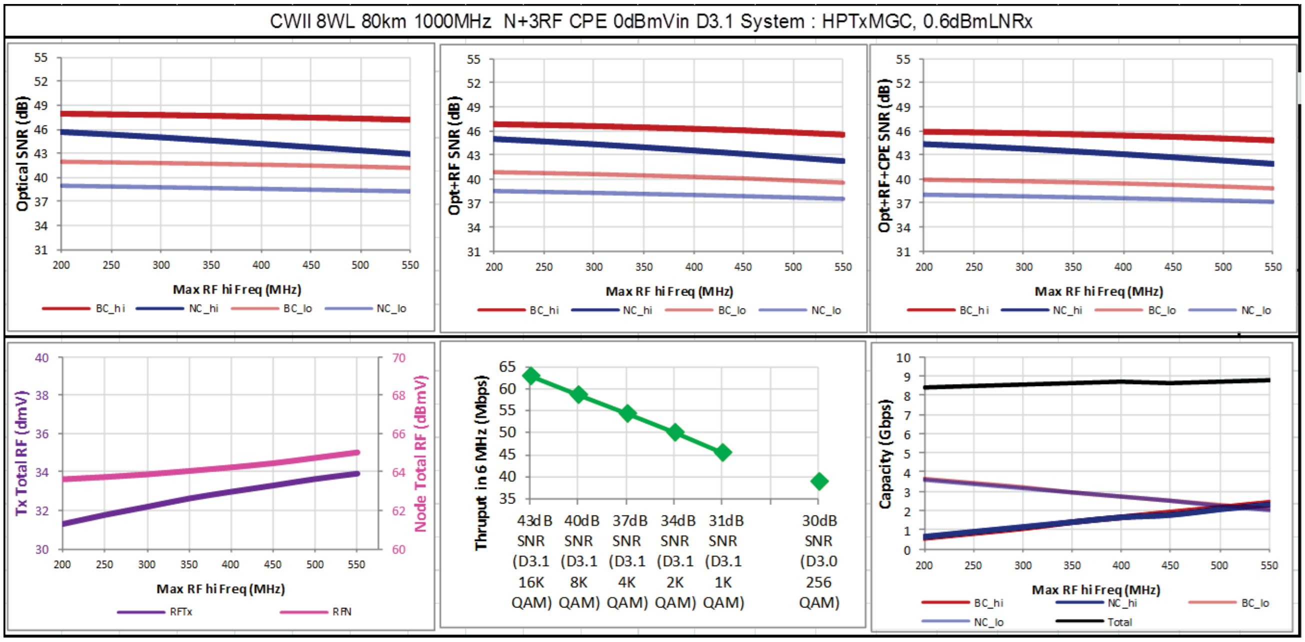


Figure 12 C-Band 8 WL 80 km D3.1 Design: Detailed Analysis

Since the C Band wavelength plan is stable and robust, it may be augmented further with an intermediate EDFA. Placing an EDFA 40 km away enables one to drive an additional reach, thus this figure shows the performance for such a link with a cumulative 80 km reach. It is seen that there is no diminishing of capacity, but the link expense has gone up due to the added EDFA. It is seen here that the total capacity is similar to that shown earlier. In other words, stable and robust wavelength plan along with the EDFA placement have succeeded in a substantial reach extension.

C-Band 16 WL 40 km D3.1/D3.0 Design:

We realize however that the power and benefit of analog optics is its ability to transport any of the signals imposed on it. To this end, we simulate here the system with D3.1 over the RF Hi section and D3.0 over the RF Lo section of the spectrum. This is to simulate the situation where video QAMs following the conventional ITUJ.83 standard continue to be used in the plant for quite some time into the future.

To the extent that D3.0 256QAM channels continue to be used in the plant it is to our strategic advantage to allocate more of the power to the RF Hi section that carries the D3.1 and maximize its SNR performance and therefore increase the throughput capacity of the system. Indeed such a plan has been attempted and has been simulated below.

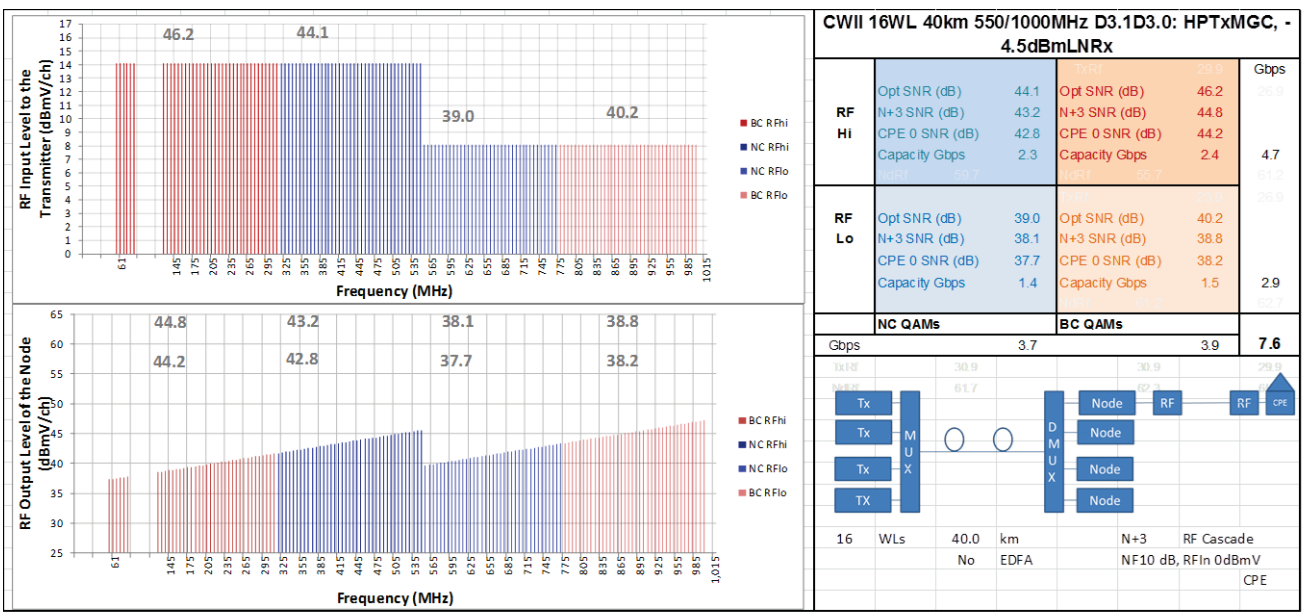


Figure 17 C-Band 16 WL 40 km D3.1/D3.0 Design: At Max RF hi Frequency of 550MHz

It is observed here that both the D3.1 channels as well as the D3.0 channels can have a mixture of broadcast (common content) as well as narrowcast (unique content) and achieve a decent throughput.

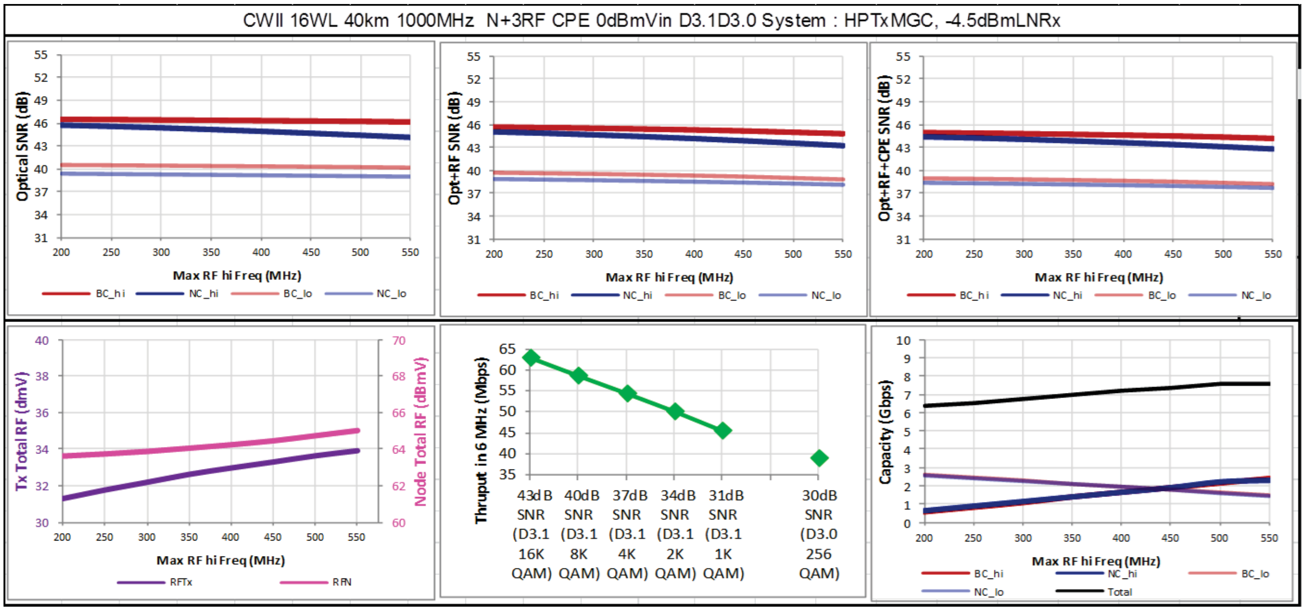


Figure 18 C-Band 16 WL 40 km D3.1/D3.0 Design: Detailed Analysis

The above picture has the same SNR or the MER calculated earlier, however this demonstrates a situation where the capacity is proportional to the RF hi frequency split. The larger the number of D3.1 channels, the higher the capacity.

O-Band 4 WL 25 km D3.1 Design:

By far the largest deployed optics base is the O Band. From the mid-1990s to the present day, major Cable deployments have been characterized by modest optical links featuring single wavelength1310 nm optical transmitters. By the early 2000s, CWDM and closely spaced O Band MWL systems have been deployed the world over.

For transmission systems to work in the O Band, it is extremely important to eliminate 4WM egress. This can be easily achieved by vacating the zero dispersion region of the optical fiber which typically extends from 1300nm to 1320 nm. Therefore stable O Band systems are either in the 1291nm band or in the 1331 nm band. O Band transmitters benefit by minimal dispersion and are therefore easier to design and deploy. Since these are typically Direct Modulated Lasers (DMLs) it is critical to have optical passives with small instantaneous slope, else the composite second order performance of the system is affected severely. Such optical passives have been available for the better part of a decade and are stable and cost effective.

Since no convenient and cost effective amplification exists for the O band, the reach of the system is limited primarily by the launch power and also by the optical passives. Typical maximum reach of a MWL O band system is generally around 25 km.

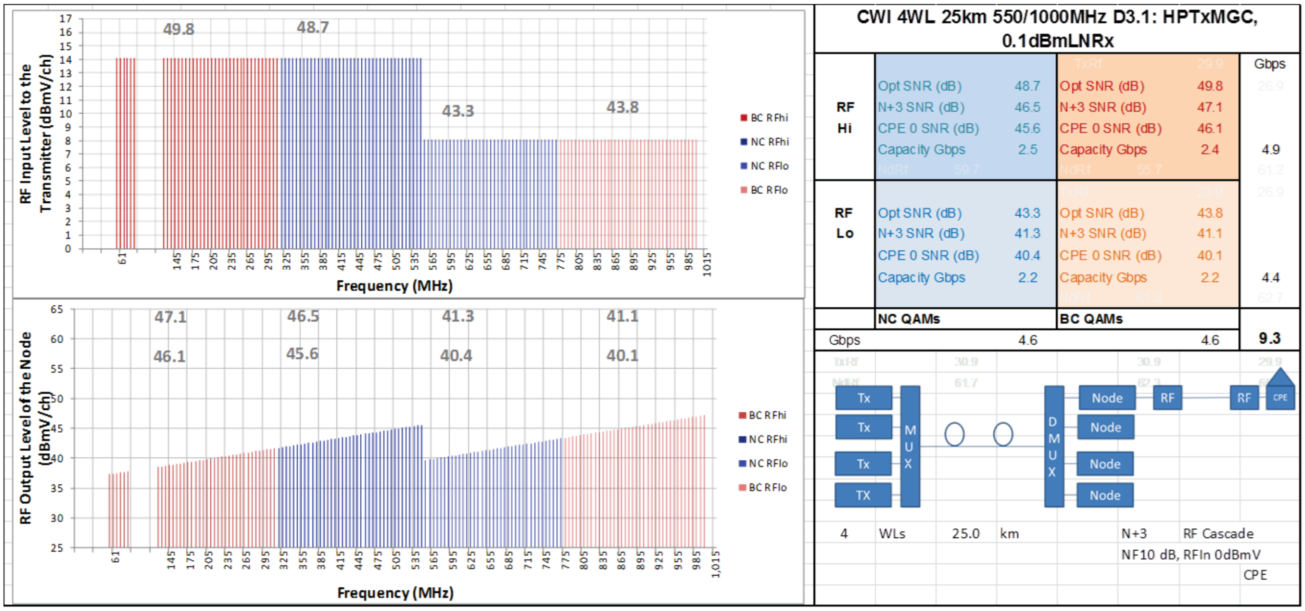


Figure 19 O-Band 4 WL 25 km D3.1 Design: At Max RF hi Frequency of 550MHz

The figure above shows that the O Band system is indeed very high performing, easily approaching the maximum capacity of a 1 GHz link.

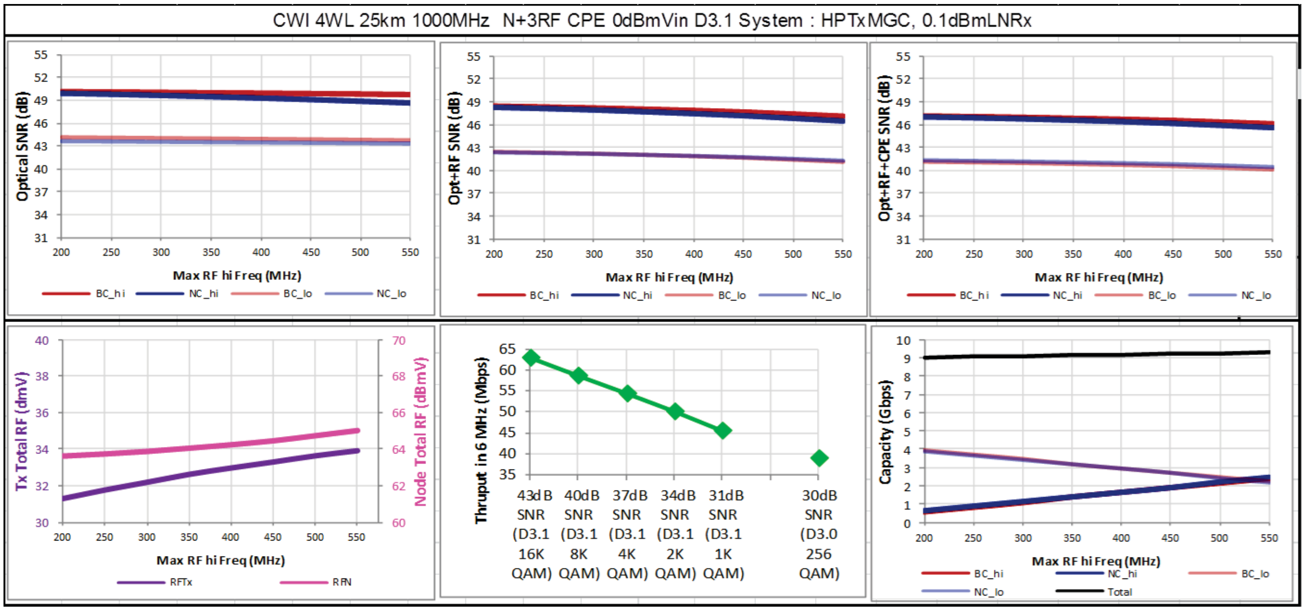


Figure 20 C-Band 4 WL 25 km D3.1 Design: Detailed Analysis

The high performance of the O band link when combined with its extraordinary cost effectiveness and its large installed base ensure that it would be transparent to the D3.1 deployments. Indeed it is possible that these characteristics of the O Band would enable one to increase the reach further or add additional wavelengths or both. State of the art O Band solutions now support up to 8 wavelengths over around 20 km of reach.

Analog Link Discussion

We have seen thru many different examples that a well-designed analog link supports higher order QAM modulations and approaches the maximum possible link capacity. Such a result is not that surprising since we are all aware that analog links are generally unforgiving of impairments. To summarize, for down-stream capacity we have analyzed

C-Band 16WL 40km D3.1 Design	8.8Gbps/WL	140.8Gbps/Fiber
C-Band 8WL 80km D3.1 Design	8.8Gbps/WL	70.4Gbps/Fiber
C-Band 16WL 40km D3.1/D3.0 Design	7.6Gbps/WL	121.6Gbps/Fiber
C-Band 4WL 25km D3.1 Design	9.3Gbps/WL	37.2Gbps/Fiber

The table above gives one an indication of the power and the limitation of these links. Since many of today's links are limited by the CPE input levels, improvements in CPE levels via the gateway architecture would then place the limitation on the D3.0 capacity. With the advent of the D3.1 and the modulation profiling that limitation is eliminated to a large extent thus opening up the space for additional capacity. We have further seen that when the link lengths extend out there is a proportional reduction in the wavelength counts and possible increases in link cost due to additional EDFAs. But the main message her is that the well-designed analog links support higher order QAM modulation formats without the need for a fundamental change in HFC architecture.

Introducing Baseband Digital Forward

At long link budgets or designs requiring high wavelength counts, analog transmission potentially has limitations as discussed in the previous section. Digital optics on the other hand is suitable for transmission over long reach and can reach Tbps (terabits-per-second) transmission rates in DWDM systems. As Moore's law progresses, the cost of silicon for AD and DA conversion and subsequent digital signal processing of several 100 MHz to a GHz worth of RF spectrum has reduced considerably and now is cost competitive against that of analog transmission. Here we investigate the requirements on AD/DA converters and the digital line rates to meet the performance equivalent of all digital QAM256 systems, mixed analog and digital systems and future operation with QAM4096 for D3.1.

The attainable SNR of an AD converter for a given ENOB (Effective Number of Bits that combines the effects of quantization noise and AD converter distortions) that is driven full scale with a sine wave is given as (at the Nyquist sample rate of 2x the maximum frequency to convert):

$$SNR_{dB} = 1.76 + 6.02 \cdot ENOB$$

However, in multi-channel RF systems the input signal is not a sine wave, instead it consists of a large number of both amplitude and phase modulated carriers. This composite signal features amplitude where the probability distribution of amplitude samples closely resembles a Gaussian distribution width w:

$$P(x,w) := \exp\left[-\left(\frac{x}{w}\right)^2\right]$$ Gaussian signal x probability density function (not normalized)

$$Pn(x,w) := \exp\left[-\left(\frac{x}{w}\right)^2\right] \cdot \frac{1}{w \cdot \sqrt{\pi}}$$ The normalized distribution (integral=1)

The rms (or effective) value of the signal x is:

$$\mu_{rms}(w) := \sqrt{\int_{-\infty}^{\infty} x^2 \cdot Pn(x,w)\, dx}$$ That is rewritten as: $$\mu_{rms}(w) := \frac{w}{\sqrt{2}}$$

Such a probability distribution is shown below (log scale) wherein the ADC converter input window is also shown for a channel plan featuring 155, 6 MHz wide QAM4096 channels:

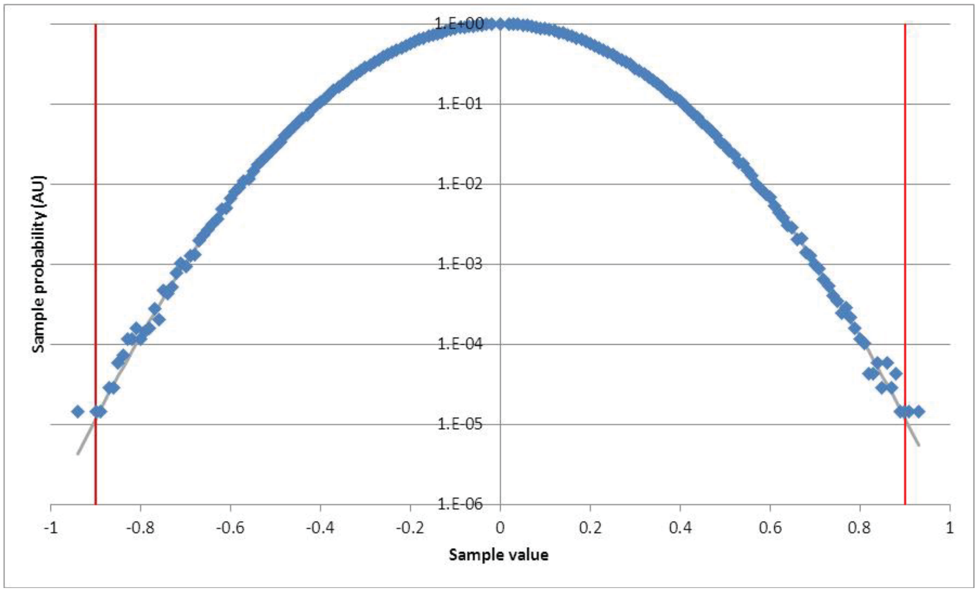


Figure 21 Input sample probability of 155 QAM4096 channel configuration sampled at 3.2 Gbs for 3.2 Msamples (blue markers) and Gaussian fit (grey) with practical ADC sampling window (red lines) for 1E-6 cumulative sample clip probability

Note that the ADC converter input window is smaller than the actual signal range. This leads to clipping of the input signal. For analog optical transmitters there is an equivalent process when lasers are clipped or external modulators are driven into saturation. Clipping causes signal distortions with an impulse noise character, a comparison of AWGN (average white Gaussian noise) and a clipping noise signal as a function of time is shown below for the same signal level and 40 dB SNR:

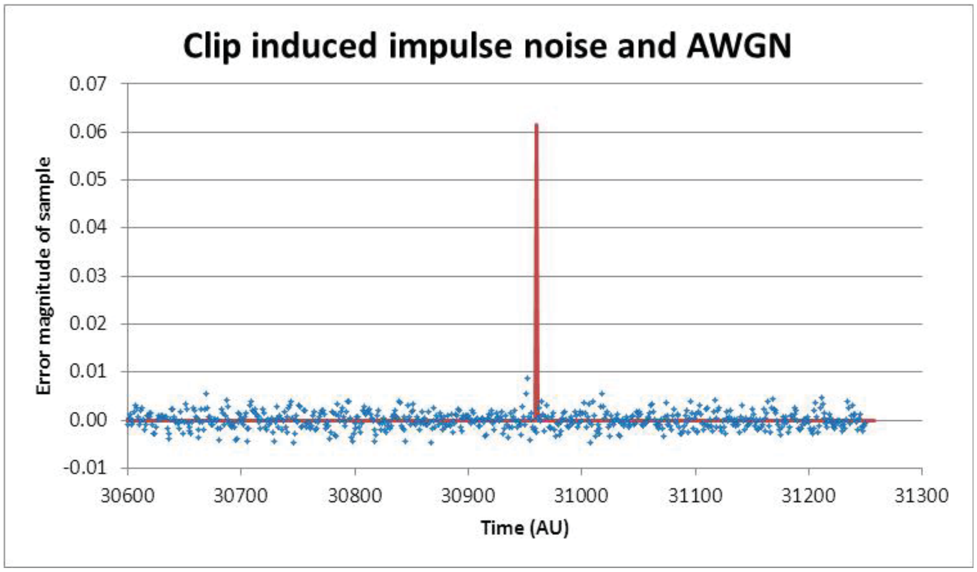


Figure 22 Clip Induced Impulse Noise and AWGN

Clipping noise is characterized by very high and short impulses that occur relatively rarely in time. The time averaged noise power of this impulse noise is much less than the AWGN due to optical link SNR or quantization noise in the case of a digitized forward or return link; here the time averaged SNR due to clipping induced noise is around 76 dB. However, when the clip events occur they generally lead to bit errors as the average SNR due to clipping induced noise is only 16 dB. AWGN based error correction schemes are not robust in the presence of impulse noise. This also applies to D3.0 that is more sensitive to clipping noise than AM-VSB. The impulse noise character of the clipping noise implies that a wide spectrum is generated affecting many QAM channels simultaneously thus spreading the energy, however the impulse energy is very high such that errors still occur. In particular AWGN based noise correction that combines information of many (OFDM) QAM channels (such as LDPC designed for AWGN) may struggle with impulse noise as multiple channels will be affected simultaneously by clip events (or for that matter amplifier compression in RF chains).

For this reason the clip probability must be limited; the figure below shows the cumulative probability that a clip event occurs within a symbol time as function of the output peak/rms amplitude ratio where the peak amplitude represents the highest amplitude that the optical transmitter can handle (or encode in the case of a DA converter). Input signal peaks exceeding this ratio are not represented in the output signal that is limited to an output peak/rms ratio and cause clip induced impulse noise.

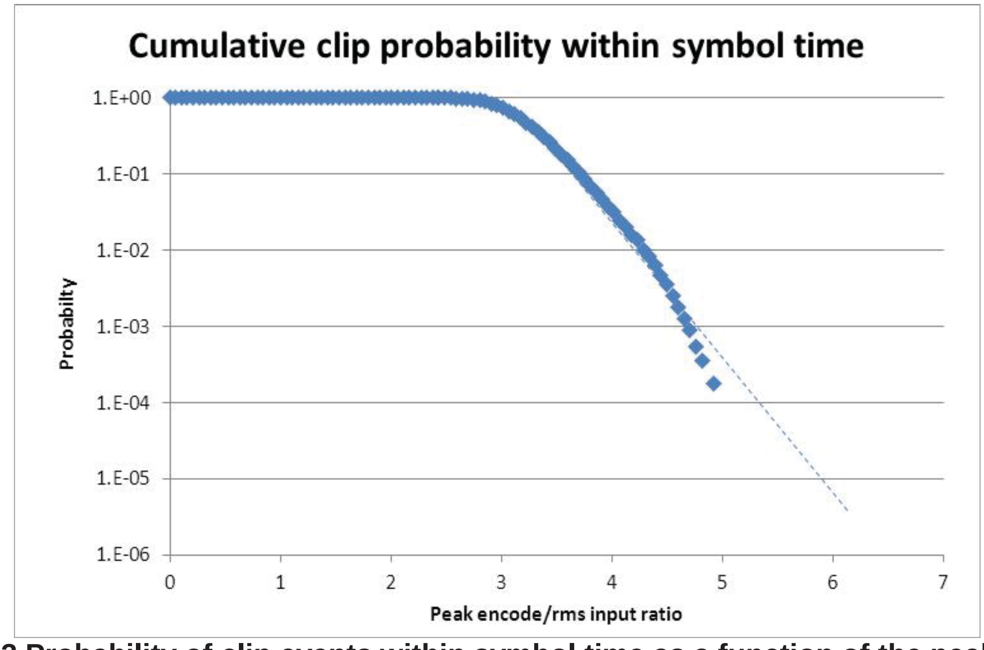


Figure 23 Probability of clip events within symbol time as a function of the peak encode capability over rms input signal ratio

The figure explains that in order to keep the BER due to clipping events low, typically a peak to rms ratio of 5 to 7 is used in optical transmission systems, be it analog or AD converted systems.   This is much less than the peak/rms ratio of a sine wave (3dB) and thus the SNR that can be expected from an AD converter loaded with a mixed channel RF load is:

$$SNR_{dB} = 1.76 + 6.02 \cdot ENOB + 3 - 20 \cdot \log\left(\frac{peak}{rms}\right) + 10 \cdot \log\left(\frac{f_s}{f_{Nyquist}}\right)$$

Note that the last term is added to represent the over-sampling ratio when the ADC and DAC are driven with a sampling rate f_s exceeding the Nyquist sampling rate $f_{Nyquist}$.

An over-sampling ratio typically between 10% and 50% may be assumed; here we will assume 25% for the figure below. The figure shows AD converter SNR as a function of ENOB at 25% over-sampling rate that can be expected from a given AD for encoded peak/rms ratios of in the range of 5 to 7 covering the range that can reasonably be accepted to maintain low BER due to impulse noise.

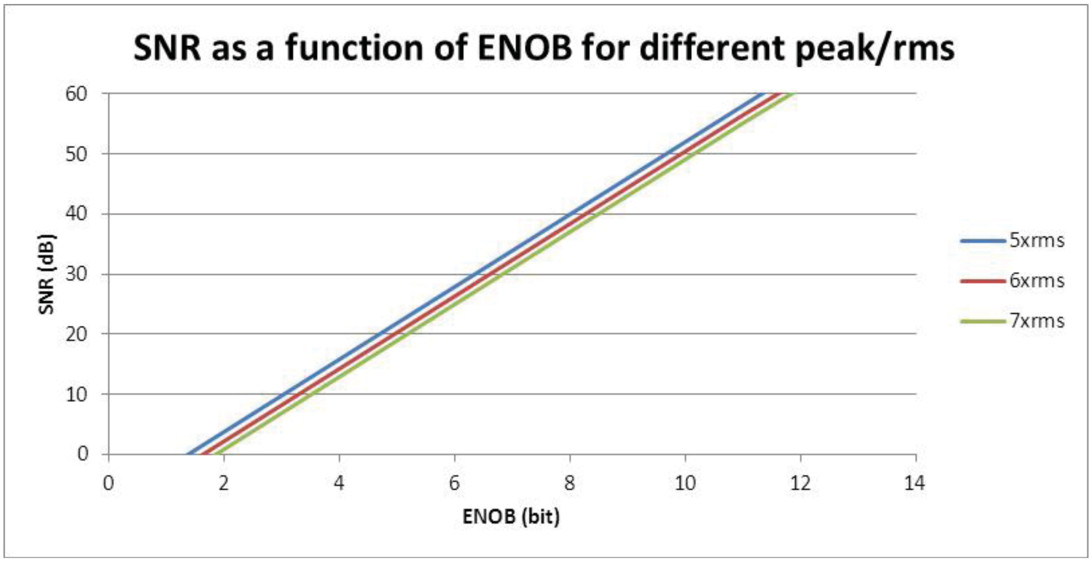


Figure24 SNR available from an ADC given an effective number of bits for an input peak range equal to 5 to 7x the rms value of the input signal.

Note we do not discuss the DAC performance here as current DA converters tend to significantly outperform ADC converters.

For a practical system a maximum RF frequency of 1 to 1.2 GHz is needed. With an ADC oversampling ratio of 25% and 12 bit ADC output width this would result in a very high serial link line rate of 36 Gbps. However the number of bits carried on the serial link can be reduced by scaling the ADC output down and truncating fractional bits. This leads to additional quantization noise that must be added to the quantization noise due to the ADC ENOB. In order to reach 38 dB SNR for D3.0 256QAM, 41 dB SNR for D3.1 4KQAM or 50 dB SNR for AM-VSB in a system with 1.2 GHz f_{max}, the minimum required line rate can be calculated as a function of the available ENOB from the ADC. These required line rates are shown in the figure below:

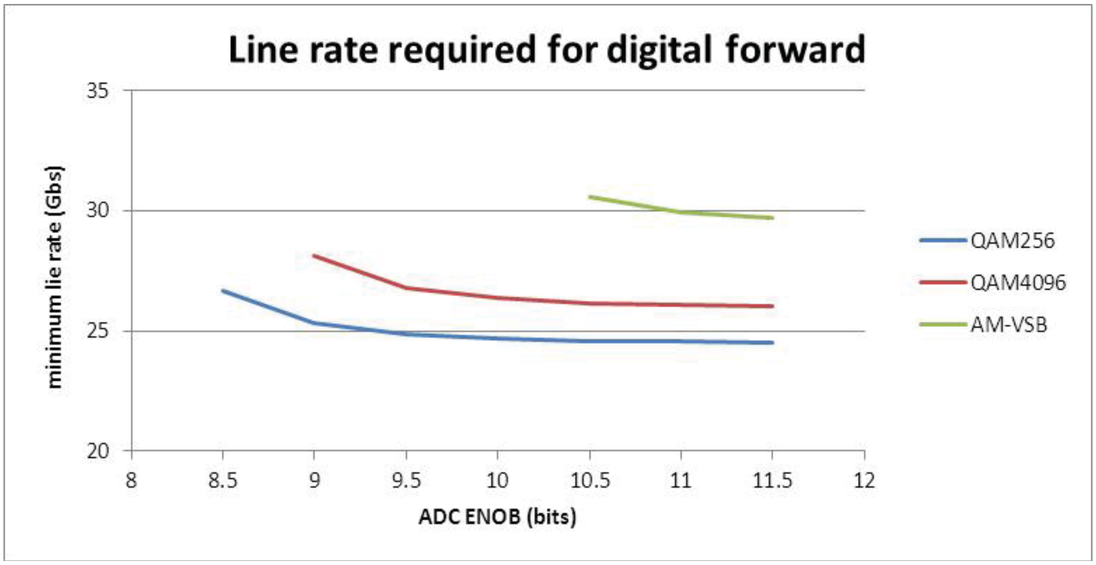


Figure 25 Line rate requirement for uncompressed digital forward as a function of ADC ENOB for 1.2 GHz f_{max}, 25% oversampling and ADC peak input range at 7xrms input voltage

It can be therefore be concluded that 25 Gbps optics are required to support a digital forward system to 1.2 GHz; up to 30 Gbps for the more challenging modulation formats but this number is up to 20% less for a 1 GHz system and can also be reduced when more clipping is allowed at the ADC input.

Note the previous discussion is on SNR and impulse noise, however for an MER measurement, amplitude and phase noise is important (see constellation plots with phase noise presented later). The highest frequency channels are most susceptible to phase noise. In order to keep the impact of phase noise on MER significantly lower than that of SNR the phase angle clock jitter of the regenerated DA clock relative to the ADC clock should be kept below:

$$\mathrm{Rad}_{rms} < 10^{-0.05 \cdot \mathrm{SNR}_{dB}}$$

(rms jitter in radians)

This means

$$dt_{rms} < \frac{Rad_{rms}}{2 \cdot \pi \cdot f_{max}} \qquad \text{(rms jitter in seconds)}$$

Note that this is a simplified approximation. In reality, the spectrum of the phase noise is important and its effect on MER can often be mitigated for long symbol times. However it provides the relevant order of magnitude that is on the 1 ps rms scale or better.

Recent advances in 25 Gbps Optics

In the previous section it was concluded that uncompressed digital forward would require 25 Gbps optics to provide a transparent optical link with binary optics. Several optical techniques have become available in the last two years as a part of the 100Gbps modules. While some of these are coherent systems, many are direct detection systems and they typically provide offering in the O band the C band wavelength ranges. And while many 100 Gbps modules are built with multiple 10Gbps components, more and more 100Gbps systems are now built with 25Gbps components. It may be noted here that the CFP, CFP2 and the CFP4 are Multi-source Agreement defined modules for 100Gbps transport.

1. The 100 Gbps O Band LR Option: There is a recent standard that enables 4, 25 Gbps array of transmitters and receivers in the O band set at 1295, 1300, 1305 and 1310 nm. These have a purported link budget of 20 to 40 km and are available in the CFP and the CFP2 form factor (100GBASE-LR4). Unfortunately the wavelengths selected for this standard are in the zero dispersion region and as such are prone to excessive 4WM egress. Unless the wavelength region changes it would be hard to launch enough light for reasonable link distances
2. The 100Gbps ODB Option: ODB stands for Optical Duo Binary. Here a standard 10 Gbps externally modulated device is over driven with overlapped signals at 25 Gbps. At the receiver end, the signal is received and reconstructed. While the transmitter is a 10 Gbps device the receiver continues to be a 25 Gbps device and that adds to the cost of the system. ODB systems are however mature and are available from multiple vendors in the CFP form factor. The ODB generally widens the optical spectrum and would therefore limit transmission distance if appropriate counter measures are not taken. The general limit to transmission is around 40 km
3. The 100 Gbps MLSE: MLSE stands for Maximum Likelihood Sequence Estimation. In this system, commercial 10 Gbps EMLs are overdriven with 25Gbps line rates. Unlike the ODB, where a MZM is overdriven in a particular way, here regular 10Gbps devices are overdriven. When this overdriven signal travels over the fiber, the effects of dispersion act upon the signal. Due to the high chirp induced, the signal and the sequence together affect the end results at the receiver. At the receiver a commercial 10 Gbps APD (avalanche photodiode) receives the data which is completely garbled. The data is immediately digitized

and by means of DSP (digital signal processing) algorithms the original data is regenerated using the MLSE (maximum likelihood sequence estimation) algorithms. The data so generated is error prone so therefore one would have to invest in an FEC over head to get error free data subsequently. Currently the MLSE transmission devices are cost effective and they operate in the C Band. However, the receivers need additional overhead of an expensive DSP chip which may necessitate a Quad receiver design to save on the DSP cost.

4. The 100 Gbps CWDM QSFP (this is a Quad SFP form factor) is a relatively new entrant to the game. This is similar to the QSFP in vogue for 40 Gbps short distance optics. The idea is very similar to (1) above, however, since the form factor is small and the wavelengths are standard CWDM wavelengths, it helps with the optical 4WM and makes it a viable candidate.

DSP and Data Compression Limits

Perfect reconstruction filter banks have been developed [7] that permit decomposition of an input signal into sub-bands. This permits processing of information in the RF spectrum on a per channel basis. Unless specific information on signal properties of information in a channel is available the theoretical limit for the minimum number of bits required to represent that channel with a given SNR is equal to the theoretical information content of that channel for that SNR; the Shannon limit. In a transparent communication system this condition generally applies; the communication system provides a signal to noise ratio to a channel without further dependence on the types of modulation formats applied. Thus the minimum number of bits per second per Hz to encode a signal spectrum with a given SNR for a transparently encoded channel is

(Shannon limit):

$$\text{bps\_per\_Hz}_{\text{min}} = \frac{\log\left(1 + 10^{0.1 \cdot \text{SNR}_{\text{dB}}}\right)}{\log(2)}$$

The figure below compares the minimum data rate needed for transparent compression of spectrum with a given SNR with the data rate available from D3.0 QAM256 and D3.1 OFDM with LDPC.

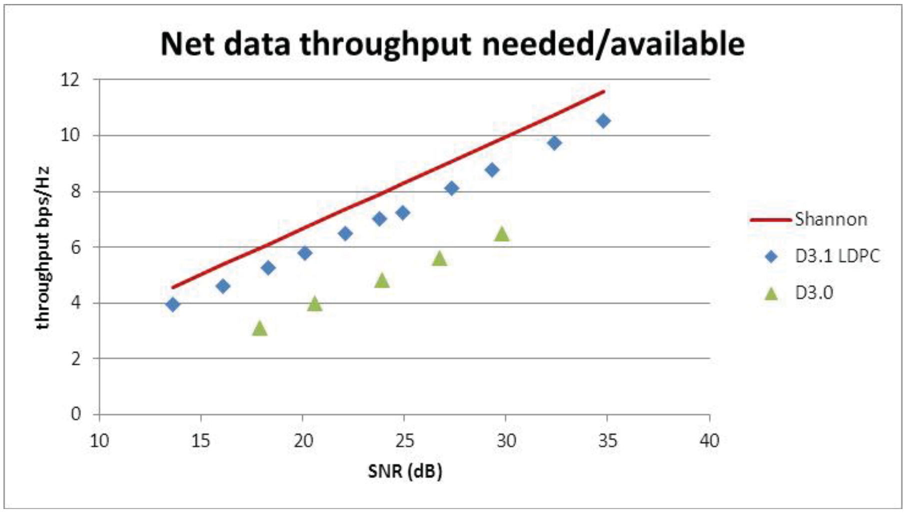


Figure 26 Minimum required data rate after compression per Shannon limit (red line) compared to available data rate with D3.1 (blue markers) and D3.0 (green markers)

The figure shows how D3.1 is close to the Shannon limit; within 2-3 dB. Note that the Shannon limit as shown here is the minimum required line rate of a compressed digital forward system. This line rate will be used as the reference to evaluate how efficient a compressed digital forward system could be. The actual implementation of the compression algorithm will determine how closely this line rate can be approached, but modern compression algorithms are very efficient.

The next figure shows the attainable efficiency (net data rate available/minimum compressed data rate required) of a compressed digital forward system. The unit-less efficiency is the ratio between throughput in bps/Hz available with a given modulation format and the minimum line rate needed to encode the modulation format in bps/Hz.

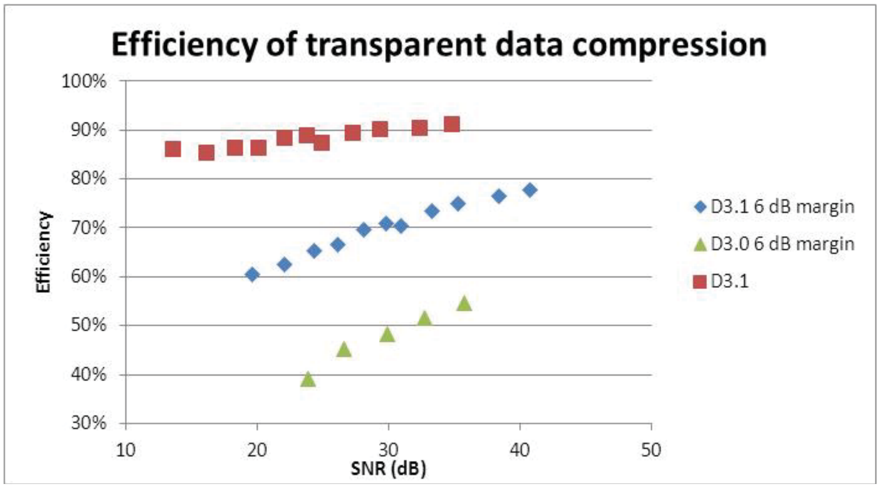


Figure 27 Efficiency of operation of system with D3.1 (red markers) and D3.1 with margin (blue markers) and D3.0 with margin (green markers).

The attainable efficiency for D3.1 is high, around 90%. However in a practical system some margin will be required for operation and other system impairments. With 6 dB margin the compressed signal will need to deliver a higher SNR than strictly needed for operation and thus the minimum compressed data rate required increases (by 1 bps/Hz for a 6 dB improvement). As a result the efficiency is lower in a practical system with margin but still approaches 80% for complex modulation formats. D3.0 is not as efficient a modulation format at providing a net data rate to the end user (about 6.5 bps/Hz for QAM256) and can attain an efficiency just over 50% for compressed data. However this efficiency is still sufficient to support most existing 750 or 860 MHz systems currently deployed today with 10 Gbs optics as will be shown later.

A key takeaway from this figure is that as modulation formats become more complex (and this generally holds whether it is D3.1 or another modulation format) the compression efficiency of a transparent compression method increases to 80% or more. The efficiency can approach 90% when OFDM systems that use the MMP (multi modulation profiles) protocol of D3.1 automatically select the highest modulation format that the SNR still supports, thus reducing the margin required. Hence a transparent compression method will be able to support any new high performance modulation format very efficiently. These conclusions apply equally to the return band where current uncompressed digital return transmitters have rather limited efficiency, typically well under 25%. Transparent digital return systems can achieve good efficiency with compression and it gets better as the return band spectrum gets used more efficiently.

The minimum serial link rate can now be calculated, the figure below shows the minimum serial link rate for a compressed digital forward link with 38 dB SNR for D3.0 QAM 256, 41 dB SNR for D3.1 QAM4096 or 47 dB SNR for QAM16k in a system with 250 MHz to 1.2 GHz forward bandwidth. Note that AM-VSB is not shown because the relatively limited bandwidth (4.2 MHz) of unique content in AM-VSB signals permits more efficient compression. Basically, AM-VSB is not as challenging for a compressed digital forward system as QAM16k.

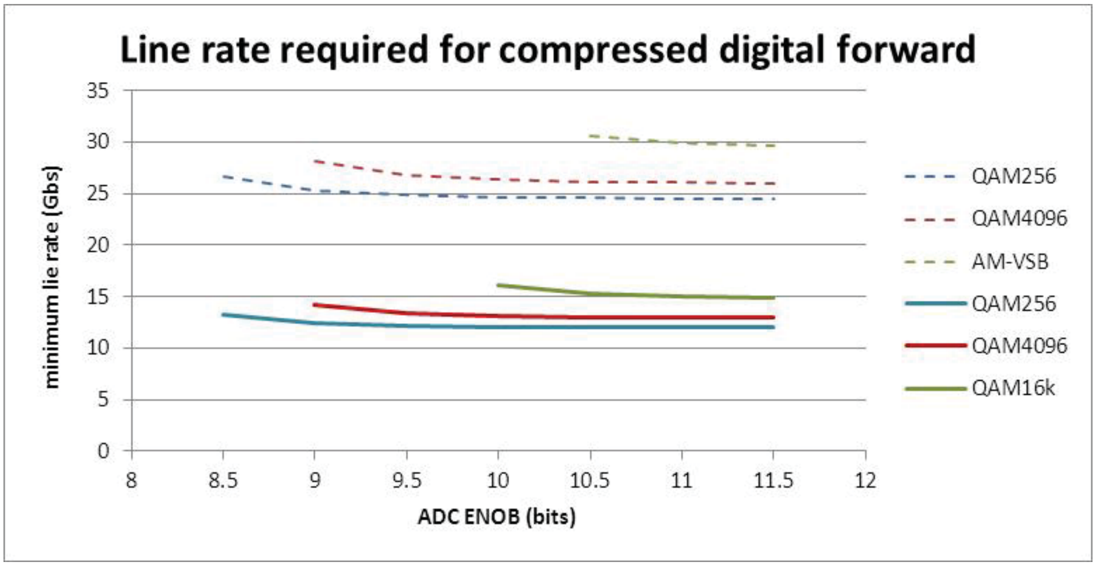


Figure28 Compressed digital forward minimum required line rate as a function of ADC ENOB, dashed lines are repeated from non-compressed digital forward plot, solid lines are for QAM256, QAM4096 and QAM16k performance levels in 250-1.2 GHz RF bandwidth, all with 6 dB of MER margin

As shown in the figure compression provides almost a factor 2 reduction in the required line rate and brings the optical line rate of a digital forward system within the range of 10 Gbps optics (11.3 Gbps OTN, or 16 Gbps fiber channel). The table below lists the minimum required line rate for compressed digital forward for operation with 6 dB MER margin

System f_max (MHz)	750	870	1000	1200	1200	1700	200
System f_min (MHz)	54	54	54	85	250	250	5
system BW (MHz)	696	816	946	1115	950	1450	195
Line rate legacy (Gbs)	8.9	10.4	12.1	14.2	12.1	18.5	2.5
Line rate D3.1 (Gbs)	9.7	11.3	13.1	15.5	13.2	20.1	2.7
Line rate QAM 16k (Gbs)	11.0	12.9	15.0	17.7	15.1	23.0	

The table shows that currently deployed 750 MHz and 860 MHz system loads can be supported with 10 Gbps optics in an efficiently compressed digital forward system. 16 Gbps fiber-channel can support all envisioned applications with complex modulation formats to 1 GHz. Finally 25 Gbps optics are sufficient to support any future application or bandwidth envisioned. Also shown in the last column is the line rate needed per compressed 200 MHz return port (where the “D3.1” row would support as much as QAM4096 in the return). Current 10 Gbps optics (operating at 10-11 Gbps) can support up to 4 fold segmentation in the return on a single optical link.

Architecture Options with Compressed Digital Forward

A compressed digital forward system will generally use compression on each separate RF channel to optimize performance of each individual RF channel. This implies that individual RF channels are available as encoded data before transmission and therefore these channels can be directed to one or more destinations as needed. Broadcast and narrowcast traffic for instance can be directed to one or multiple destinations on a channel by channel basis. The figure below gives an example of full node segmentation using a single 25 Gbps optical link:

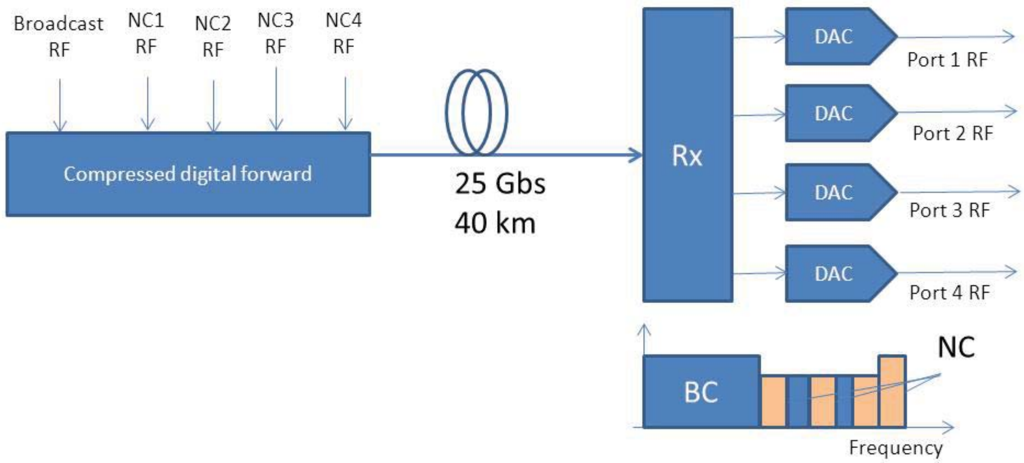


Figure 29 Segmented node with 25 Gbps optics with 48 narrowcast channels per port and broadcast spectrum. Broadcast output spectrum shown in blue, narrowcast in red.

Similarly the same 25 Gbps optical link could serve multiple optical nodes through passive optical splitting. The narrowcast and broadcast channels can be selected on individual node ports; there is no need for defining specific narrowcast and broadcast bands. Also each individual channel can be put out at a level and SNR performance as needed in the application without affecting the compression efficiency of other channels. While the communication system remains fully transparent for RF modulated signals it offers the benefits of software reconfigurable networks.

QAM4096 constellation diagrams under different operating conditions

In order to investigate if analog or digital forward transmitters can support higher order modulation formats 156 6 MHz wide QAM4096 channels on the NTSC channel plan were simulated and subjected to various impairments including average white Gaussian noise (AWGN), phase noise, second order and third order distortion and clipping to represent optical transmitter clipping and amplifier compression. 20 channels were demodulated, constellation diagrams were created and BER and MER results were determined for the various impairments. The figure below represents a QAM4096 constellation with 57 dB MER and no bit errors after accumulation of 100 k symbols.

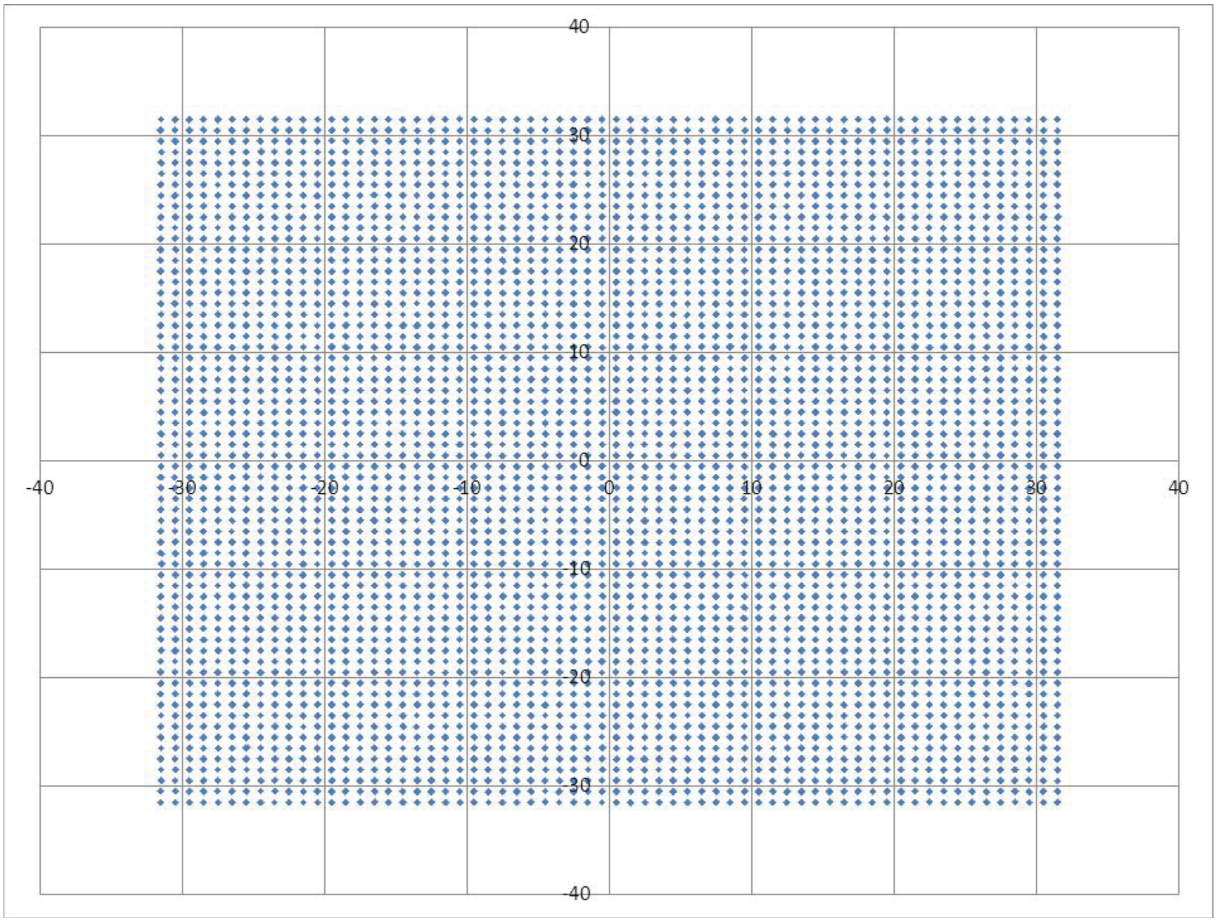


Figure 30 QAM4096 constellation with good SNR, no errors
57 dB MER is more than enough for error free QAM4096 operation.

The first impairment is due to AWGN; an impairment present in any communication system limited by SNR, the next figure shows the constellation degradation due to AWGN.

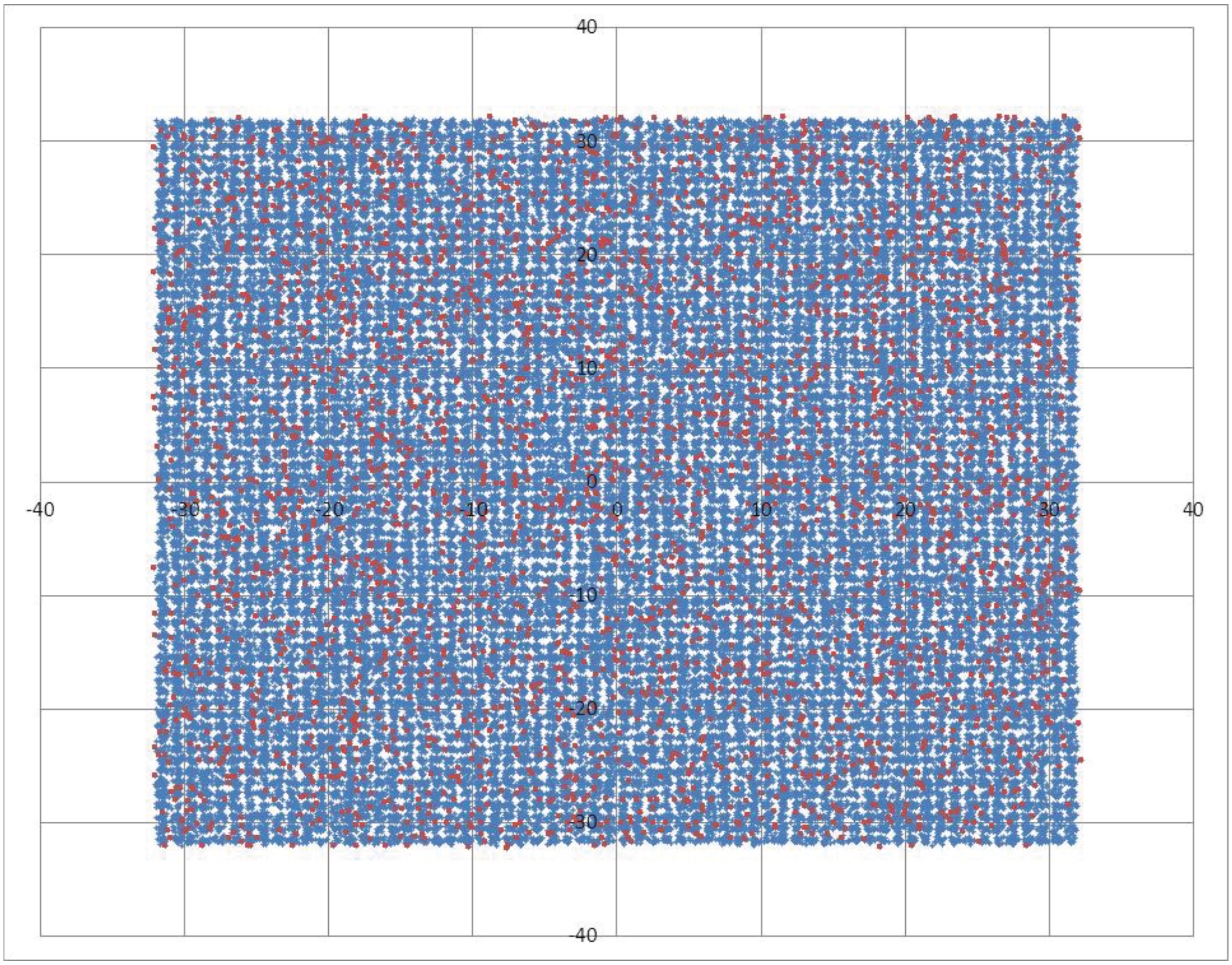


Figure 31 QAM4096 constellation with AWGN for MER=38.6 dB, BER=3.6E-3. Symbols with errors are marked red

At an MER of 38.6 dB the bit error rate is close to 0.36%, this is a high error rate but LDPC (Low Density Parity Check) error correction as planned for the next generation DOCSIS (D3.1) is designed to handle even higher AWGN induced error rates with an MER of 35 dB [6]. At this MER the theoretically limited raw BER is high, around 2% but with around 12% overhead LDPC is able to handle that raw error rate. All constellation diagrams were simulated at 3 dB margin above this minimum MER. While 3 dB margin appears low next generation DOCSIS has provisions to automatically select the highest modulation profile that the system supports failing which the next lower modulation format is selected at a loss of merely ~10% in throughput (MMP, Multiple Modulation Profile). Hence MER "failures" do not result in hard failures where channels are lost, instead there is a soft transition in the channel performance as MER gets worse. As a consequence the system can generally be operated at the limit of what the MER permits

instead of reserving a large margin (such as commonly used with QAM256 on D3.0). Most current analog and digital CATV links support MER of 38 dB and higher.
Unlike analog communication links without frequency conversion, systems with frequency conversion (including ADC and DAC systems) are degraded by the phase noise of local oscillators or of ADC and DAC clocks. This phase noise generally leads to both amplitude and phase noise after a DAC output has been passed through a reconstruction filter. For illustration the phase noise component alone is shown below for the same MER:

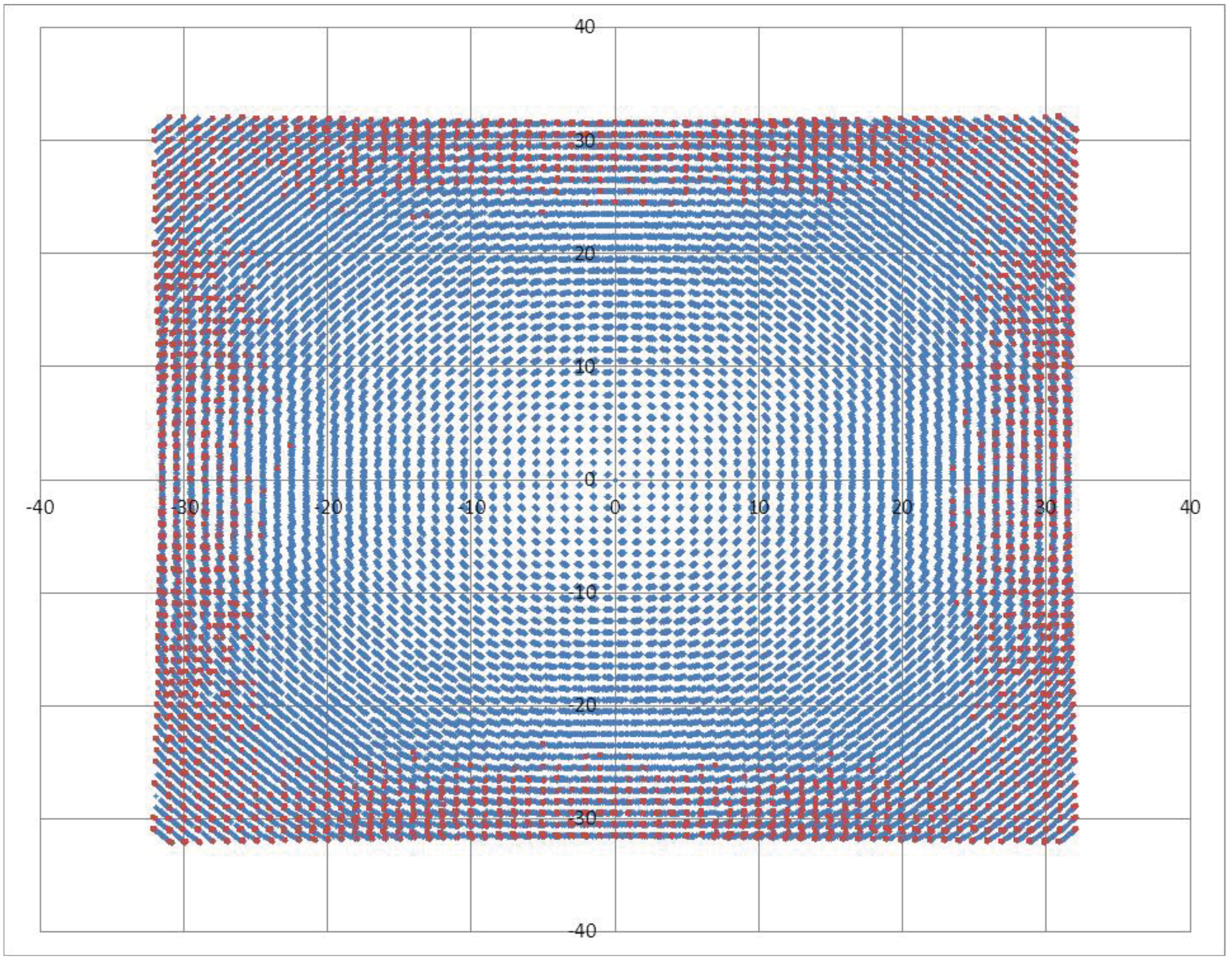


Figure 32 Constellation diagram with phase noise, MER=38.7 dB, BER=4.2E-3, error symbols marked red

The constellation looks rather different with phase noise, the errors occur for the larger vector amplitudes. Even though the error rate is similar as for AWGN the error characteristics are very different and performance of an AWGN error correction scheme may be affected.

Analog communication systems suffer from impairments due to distortions, second and third order distortion components were investigated. An example constellation diagram is shown below:

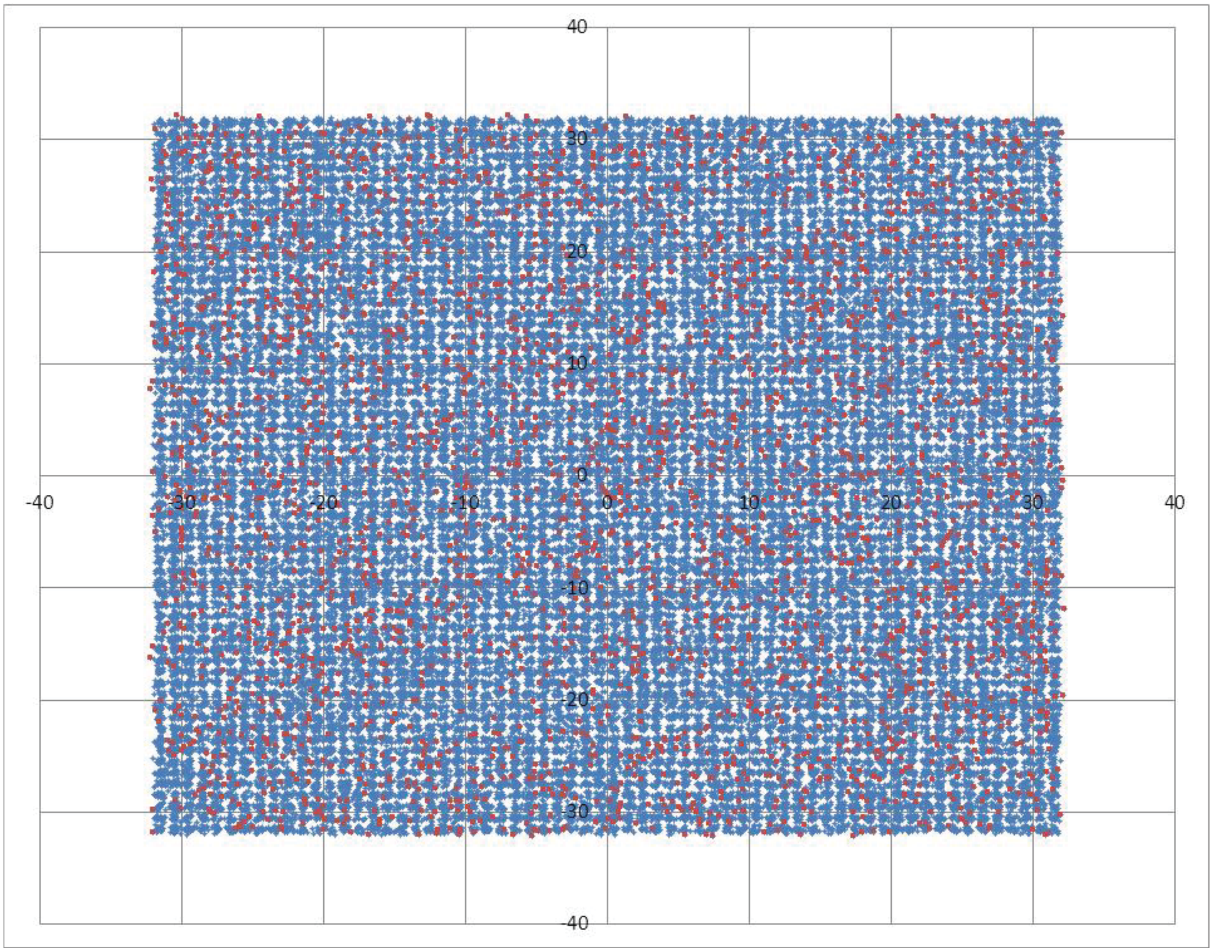


Figure 33 Constellation with second order distortion, MER=38.8 dB, BER=3.3E-3

The figure shows an apparently random error distribution that is amenable to an AWGN error correction scheme, albeit that there will be some correlation between errors of different channels that may be sufficiently random. For an MER of 38.8 dB the BER is virtually the same as for AWGN. For third order distortion set for an MER of 38.6 dB a BER of 1.6E-3 was found. The equivalent second and third order distortion performance numbers of an analog link would have been ~41 and ~47 dB for CSO and CTB respectively for an NTSC77 load with the same overall modulation index as was used for this simulation of the 156 QAM4096 channels. Adding 6 dB for a possible future transition from QAM4096 to QAM 16k and another 6 to 8 dB for additional margin, the preferred CSO and CTB specs would be on the order of 55 and 60 dBc respectively, well within range of most current analog transmission equipment.

All practical communication equipment with actives has a limited dynamic output range. In analog laser transmitters laser clipping occurs, on ADC/DAC systems the converters have a limited range and in amplifiers the output swing is limited by the voltage supply and bias point of the amplifiers. This leads to clipping (or compression in the case of amplifiers) of the signal in case the input/output voltage swing exceeds the dynamic range. A constellation with clipping induced errors is shown next.

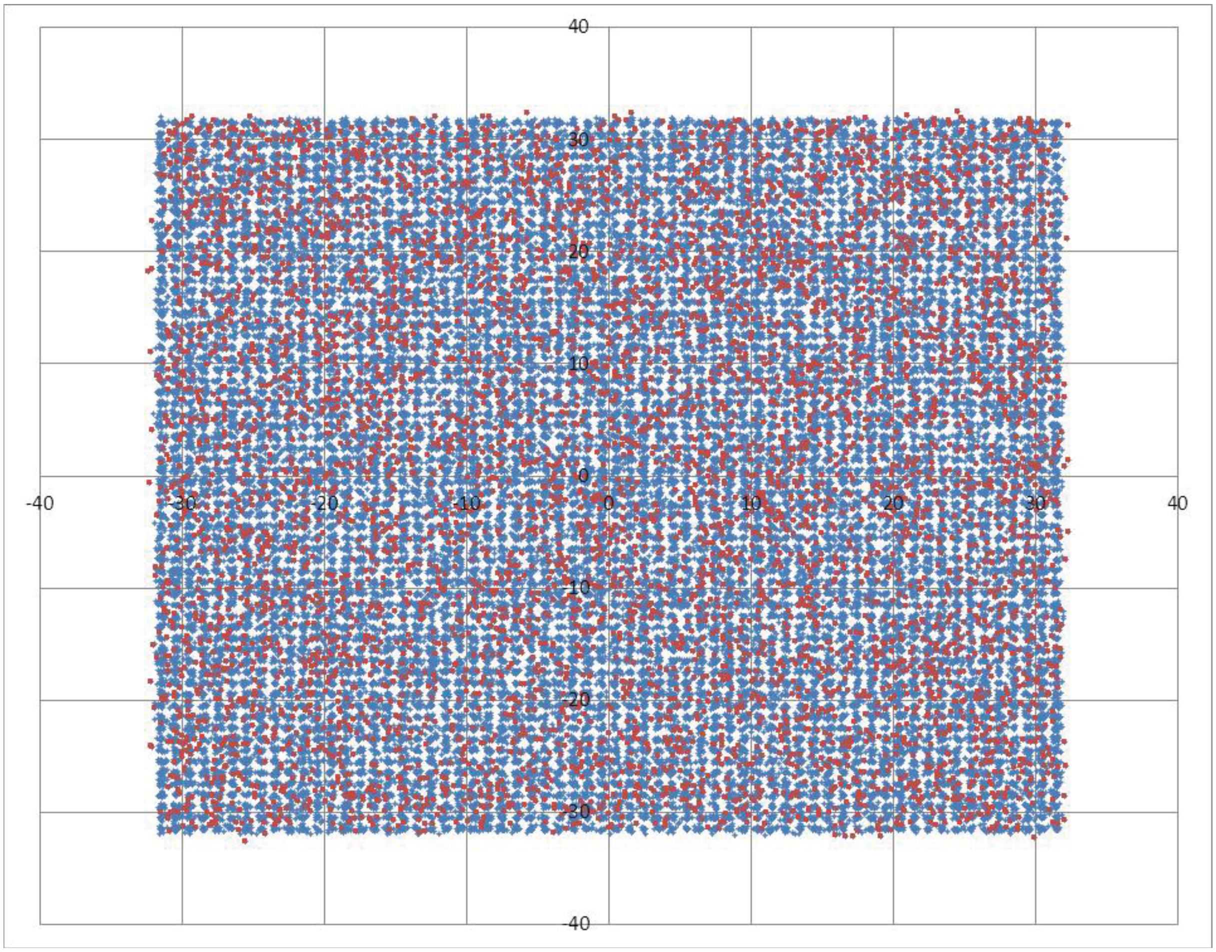


Figure 34 Clipping induced MER degradation to 38.7 dB, BER is 6.7E-3, error symbols marked in red.

Note that the constellation diagram does not look "compressed" as one might expect because this is a multi-channel simulation where the peak amplitudes of individual channels have limited correlation with the composite signal peak amplitude. It is the composite signal amplitude rather than the individual channel amplitude that determines clip events.

Whereas the clipping induced BER at this MER is only about a factor 2 higher than AWGN induced errors it is known that the errors across multiple channels due to clipping are correlated for such high intensity impulses [8,9]. Unless an error correction scheme is specifically designed for this purpose there is no guarantee that it could handle the high error rates generated. The different nature of clipping induced errors from AWGN induced errors is also illustrated with the BER versus MER curve shown below:

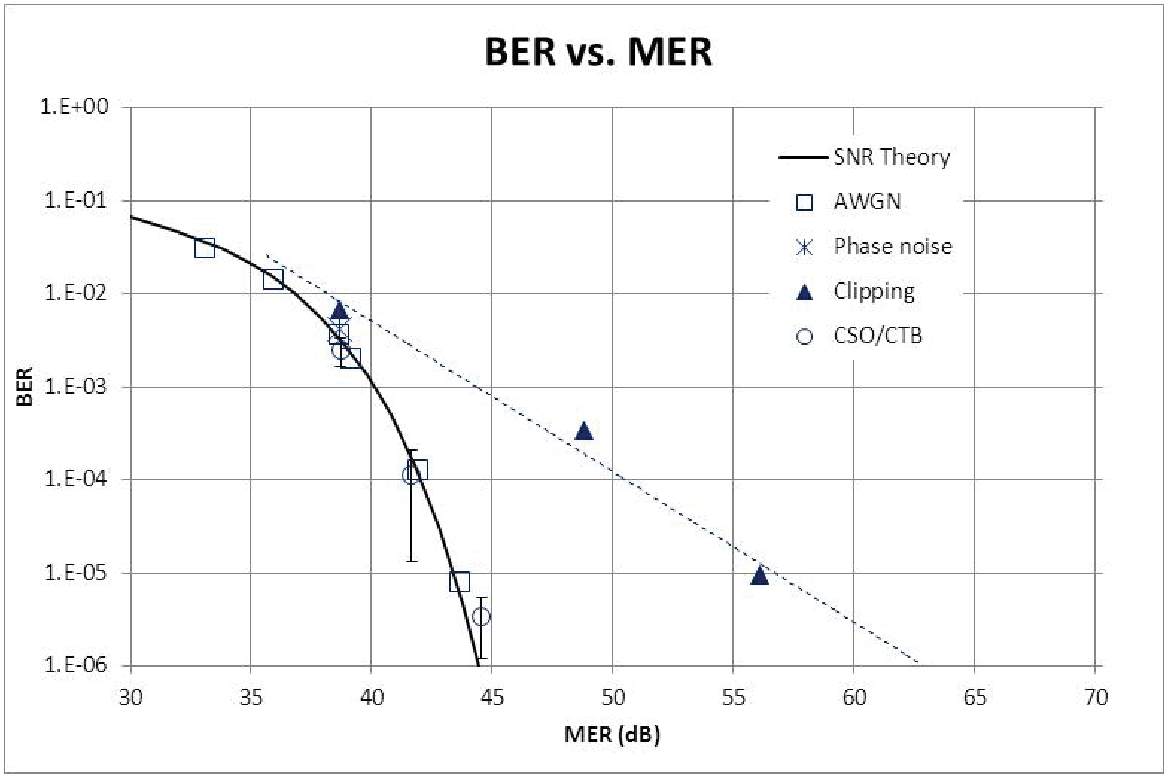


Figure 35 Comparison of BER-MER curves

The simulated AWGN induced BER perfectly matches the theoretical curve for QAM4096. Error rates induced by phase noise and distortions do not deviate significantly from the error rate predicted from the MER by theory for AWGN.

For clipping induced errors however a much higher MER is needed to obtain a negligible error rate. The clipping induced errors have very different properties from AWGN induced errors. Analog links are typically operated at a modulation index where the MER due to clipping is 70 dB or more such that the raw error rate due to clipping remains well under 1E-6. It can safely be assumed that clipping of analog links will not cause error rates that require novel error correction methods specifically designed for clipping noise although in the long term such methods can be useful in a different context discussed in following sections.

Conclusions

We have presented an analysis of transparent optical solutions for HFC networks, demonstrating that current analog transmitter technology supports new digital modulation formats with performance requirements similar to those that have been applied to transmitters supporting AM-VSB signals. For long reach or high wavelength counts, where the design of analog transmission systems is challenging, we have shown that current ADC and DAC technology permits realistic implementations of uncompressed digital forward transmitters using 25 Gbps optics. More importantly, we have shown that transparent compressed digital forward transmitters can support current system channel loading and bandwidth requirements with 10 Gbps optics and can support software configurable system segmentation with 25 Gbps optics without the need to install additional fiber or wavelengths. It was further demonstrated in these compressed digital forward solutions that complex modulation formats are particularly suitable for compression with 80-90% efficiency.

The key message of this paper was twofold. The first message is to impress upon the reader the value of transparent networks. This is demonstrated by the technology and service growth of MSOs over the past twenty years with minimal, incremental investments to their existing infrastructure. Secondly, that currently available, transparent optical technology can continue to provide high value and preservation of investments moving forward as new modulation formats are deployed.

Bibliography

1. "Mission is Possible: An Evolutionary Approach to Gigabit-Class DOCSIS", Authors J. Chapman, M. Emmendorfer, R. Howald, & S. Shulman, (Cisco, ARRIS, Motorola, and Intel) published paper in the Proceedings of the May 2012 NCTA Cable Show

2. Michael J. Emmendorfer and Tom Cloonan, Examining the Future Evolution of the Access Network, NCTA 2013

3. Shamim Akhtar, Doug Weiss, Venk Mutalik, Zoran Maricevic, Marcel Schemmann, David Heisler, Don Wesson, Liyan Zhang (Comcast and C-COR), "Gigabit Ethernet and Analog/QAM Traffic Compatibility in HFC Networks: A case for physical layer convergence" SCTE ET 2006

4. Fernando Villarruel, Venk Mutalik, Kuang-Yi Wu and Marcel Schemmann, (Cisco and ARRIS) "When Wavelengths Collide, Chaos Ensues: Engineering Stable and Robust Full Spectrum Multi-Wavelength HFC Networks" SCTE 2011

5. Venk Mutalik, Marcel Schemmann and Brent Arnold, When All Digital Requires All Analog Performance: Preparing the Access Network for Next Generation QAM, SCTE 2011

6. ETSI TS 102 991 v1.2.1 (2011-06) Digital Video Broadcasting (DVB); Implementation Guidelines for a second generation, digital cable transmission system (DVB-C2), table 20, figure 71

7. Multirate Systems and Filter Banks, P.P. Vaidyanathan, Prentice Hall 1993, Chapter7

8. Ayham Al-Banna, "OFDMA vs. SCDMA: Performance Analysis in the Presence of Impulse Noise", Cable Congress, (March, 2013).

9. Ayham Al-Banna, "Performance Analysis of DOCSIS® 3.1 PHY Technologies in the Presence of Impulse Noise", ANGA Cable Conference, (June, 2013).

Abbreviations and Acronyms

Abbreviation	Description
SC-QAM	Single Channel – Quadrature Amplitude Modulation
D3.0	DOCSIS 3.0
D3.1	DOCSIS 3.1

Acronym	Description
DOCSIS	Data over Cable System Interface Specification
HSD	High Speed Data
OFDM	Orthogonal Frequency Division Multiplexing
LDPC	Low Density Parity Check
EML	Electro Absorption Modulated Laser
AGC	Automatic Gain Control
MGC	Manual Gain Control
4WM	Four Wave Mixing
SRS	Stimulated Raman Scattering
XPM	Cross Phase Modulation
ITU	Internal Telecommunication Union
NTSC	National Television System Committee
MWL	Multi-Wavelength
QAM	Quadrature Amplitude Modulation
CIN	Composite Intermodulation Noise
CWDM	Coarse Wave Division Multiplexing
ADC	Analog to Digital Converter
DAC	Digital to Analog Converter
AWGN	Additive White Gaussian Noise
SNR	Signal to Noise Ratio
ENOB	Effective Number of Bits
CFP	C Form Factor Pluggable