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When Physical Layer Simulation Gets Real

Next-Gen Network Modeling

A Technical Paper prepared for SCTE by

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1. Introduction

There are many ways to model advanced broadband networks, and a growing number of ways to simulate their behavior, based on available information about performance characteristics. This paper, co-authored by Comcast's Ramya Narayanaswamy, Karthik Subramanya, Richard Prodan and Larry Wolcott, will explore the intersection of theoretical modeling, practical proactive network maintenance (PNM) and modern data science – a potent combination that is well suited for the sophisticated modeling and simulation needs of cable 10G networks.

Using traditional RF reflection measurements like scattering parameters / S-parameter matrices, fed by real-world field PNM (Proactive Network Management) data into a graph topology, the authors will show how advanced 10G networks, and particular bidirectional signal flows, can be simulated for both existing and proposed networks.

Having a graph topology of the network, along with cable and component specifications, provides a means to apply transmission line theory as never before. Most cable operators have this RF and reflection data readily available for their systems. This simulation of cable systems, as-built, can predict end-to-end performance which can be compared against actual measured performance using PNM tools. This technique can provide a method for the evaluation of full-duplex (FDX), extended spectrum (ES) and traditional RF transmission performance of existing plant, as well as experimental designs. Beyond that, it can solve for fault detection, and many other previously unsolvable RF mysteries within our cable universe.

2. Background of Network Modeling

Scattering parameters or S-parameters are commonly used to model the electrical behavior of radio frequency (RF) communications networks when stimulated by electrical signals. The S-parameters are used to form a scattering matrix or S-matrix, that collectively describe the behavior of a network or circuit at different frequencies. These parameters are convenient for high-frequency RF design because they are easily measured by standard lab equipment such as a vector network analyzer (VNA). Coaxial S-parameters are further described in section 7.2.

There are alternatives to S-parameter modeling, which can be used to understand the transfer function of communications networks. Analysis of the frequency response including signal levels transmitted from the node and received by the cable modem or vice versa is described. This approach allows the use of log magnitude values versus frequency from component spec sheets to model the transfer function of each section of transmission line (i.e., cable) terminated at either end with an impedance (i.e., a tap, node, or cable modem) characterized by its magnitude return loss versus frequency. Note that this formulation provides the complex frequency response with only the scalar amplitude versus frequency of the cable transmission line and the magnitude return loss and insertion loss versus frequency of the tap terminating impedances. This provides a convenient alternative to S-parameter and T-parameter measurements, which are difficult to measure on existing networks.

3. Network Modeling and Simulation

3.1. Physical Layer (PHY) Model

Traditional hybrid fiber/coax (HFC) cable distribution networks have been built as tree and branch networks consisting of a fiber node connecting multiple cascaded amplifier coax cable sections. Each section connects to a series of multiport taps transmitting signal to and receiving signals from drop cables

to customer premise equipment. An example of one coax branch of a conventional Node + N HFC architecture shown in Figure 1. The node span contains multiple amplifier spans, each with multiple taps between amplifiers.

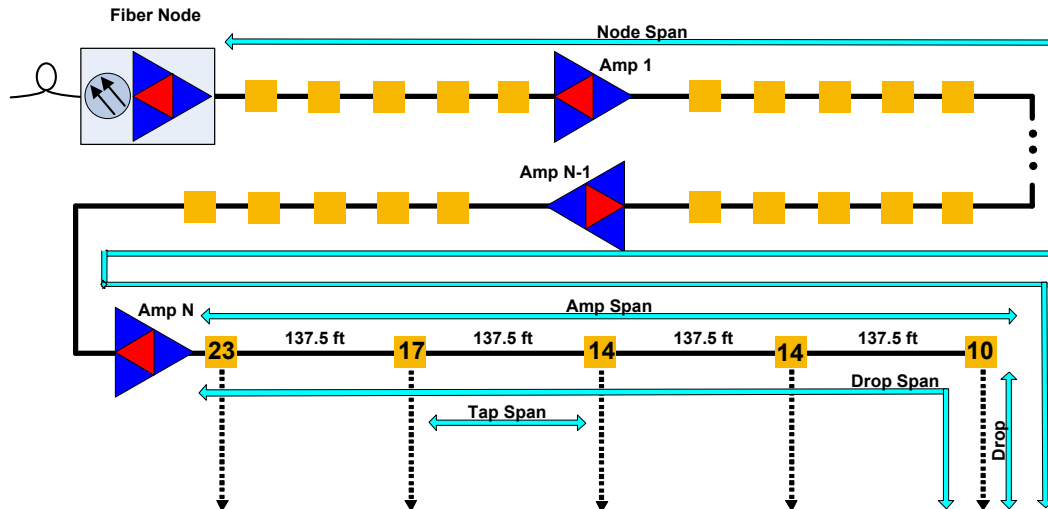


Figure 1 - Conventional node plus N network architecture

This conventional architecture provides two-way signal transmission on separate spectral bands using frequency division duplex operation. Each amplifier section uses diplex filtering to separate upstream transmissions toward the node in the narrower lower frequency band (typically 85 MHz or less) from downstream transmissions from the node in the much wider upper frequency band (up to 1.2 GHz). Such diplex filtering prevents two-way transmission within the same bandwidth. Each multiport tap contains a directional coupler that diverts a portion of the downstream signal to the drops connected to the tap ports and injects the upstream signals present on the tap ports toward the node. The directivity of the directional coupler prevents upstream signals from propagating in the downstream direction or from diverting to other drops upstream from that tap port.

The design of a hybrid fiber coax (HFC) network requires specifying the individual components and their values to obtain consistent signal levels across all the subscriber drops. These include:

- hardline cable (trunk and express feeder) and drop cable type and parameters
- tap parameters (insertion loss, tap loss, return loss, isolation)
- downstream and upstream node and tap equalization type, frequency range, and tilt value
- tap spacing
- drop cable length
- amplifier output (level and tilt)

Analysis of the frequency response including signal levels transmitted from the node and received by the cable modem or vice versa is described. This approach allows the use of log magnitude values versus frequency from component spec sheets to model the transfer function of each section of transmission line (i.e., cable) terminated at either end with an impedance (i.e., a tap, node, or cable modem) characterized by its magnitude return loss and insertion loss versus frequency.

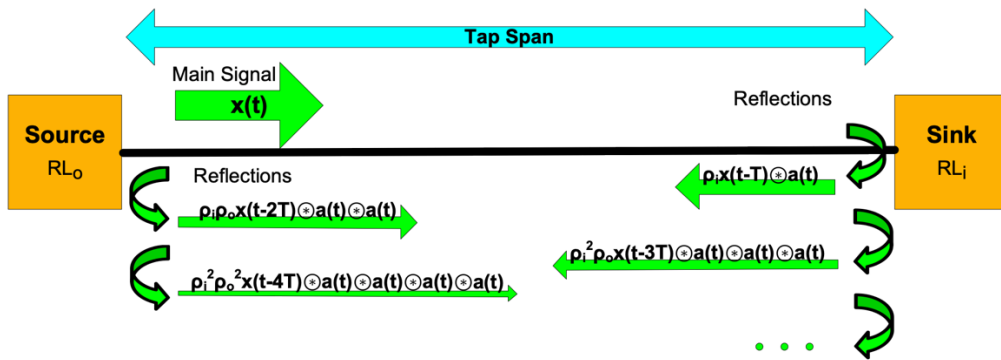


Figure 2 – Signal reflections in a cable between adjacent taps

Consider a signal transmitted downstream from a tap output to the adjacent tap input as shown in Figure 2. with amplitude response $A(f)$ and linear phase response which has the corresponding impulse response of the cable denoted by $a(t)$. The transmitter at the signal source has (nearly) matched impedance to the drop cable but with a return loss RL_o (dB). In general, the return loss is also a function of frequency $RL(f)$. The signal traverses the cable to the tap with propagation delay T which has a (nearly) matched impedance to the cable with return loss RL_i (dB). A portion of the signal equal to the reflection coefficient $\rho_i = 10^{-RL_i/20}$ is reflected back to the source, which in turn a portion of the reflected signal equal to the reflection coefficient $\rho_o = 10^{-RL_o/20}$ is re-reflected back toward the tap, and so on ad infinitum. This can be represented as a sum of the incident signal $x(t)$ and the infinite series of reflections each delayed by the round trip (i.e., twice) the propagation delay T of the cable.

The same analysis applies to a drop cable section between a tap port and the cable modem F-connector port where the attenuation model and propagation delay are specified for the drop cable instead of the hard-line cable and the input and output return losses are specified for the tap port and the cable modem F-connector port respectively. The network simulation will require each of these transmission lines to be cascaded from the node to each cable modem attached to the network.

3.2. Network Topology as a Graph

Cable operators typically have some electronic form of system design, usually rendered by computer aided design (CAD) systems. These electronic representations of the network are used throughout the lifecycle of the network, from design to construction, upgrades, maintenance and so forth.

One of the major goals of the graph system is to map the physical and logical topology of the access network into a graph database that will allow querying paths between the physical and logical objects of the access network. There is no single dataset or database that currently has the required data compiled with common keys, instead the data is distributed across different units of the org among databases currently in use for services with differing goals. The lack of common keys and differing identifiers and definitions for the same objects in different databases present a challenge. The initial goal is to achieve mapping 95% of an access network node in the graph. To overcome this challenge, we develop and implement algorithms that check for availability of an access network object in multiple databases and when found confirm matches to the same vertex, when not-found provides an edge to the nearest vertex with a score that provides the strength of the attribution.

The goal is to map the following path: Device → Household (HH) → Tap → Amp → Buss Leg → Node using the following 3 data sources:

1. *CAD* provides detailed data on the physical design topology of serviceable addresses. Attributes reported include model, type, and location coordinates for CMTSs, Nodes, Taps, Amps and Serviceable Addresses: This is used for HH → Tap → Amp → Buss Leg → connections
2. The node combining plan (*NCP*) provides billing information presenting currently active services serving subscribers with activated devices for each location. Providing for Device → HH connections
3. *CM Topology* provides CMTS registry, and the devices connected to the CMTS at the given moment. Used to validate online inventory from Device → CMTS

While CMTS naming is conventional and CM Topology can be matched to NCP rather easily, the main difficulty is identifying the activated addresses in physical design topology in CAD and matching them with the addresses from the billing data. The matching algorithm consists of the following steps:

Step 1: Filter the CAD dataset for HH's using a node ID and account numbers with corresponding Addresses and Devices in the network data aggregator

Step 2: Use an enhanced fuzzy matching algorithm on the address fields of CAD data and network data aggregator to get high confidence matches for same addresses.

Step 3: Use lon/lat coordinates as reported in CAD data and network data aggregator to match records based on geographical proximity. We match coordinates that are closest to each other using a threshold for maximum distance.

Step 4: When no match can be made between CAD and network data aggregator records, get the records for all the taps with their coordinates attached to the node from CAD. Match all the network data aggregator records to the nearest tap using lat/lon coordinates. All the addresses from network data aggregator gets a match at the end of this step, all the unmatched addresses from CAD are tagged as inactive/passed addresses.

Step 4: Merge all the matched addresses to CM Topology data using media access control (MAC) addresses. Anything unmatched on left and right side are tagged accordingly and will be passed off to related teams for further review.

3.2.1. Graph Topology Data Structure

The graph data used in the simulation consists of a set of vertices/nodes that represent the hierarchical elements of the network, and a set of edges that represents the relationships in this hierarchy. A high-level example is illustrated in Figure 3. Figure 5 shows the details of the graph structure which contains a cable modem (Device), equipment location (HH), drop cable (Drop) and tap (Tap) which are connected. Figure 4 shows a more complex example of the CMTS topology mapping to the RF active ports.

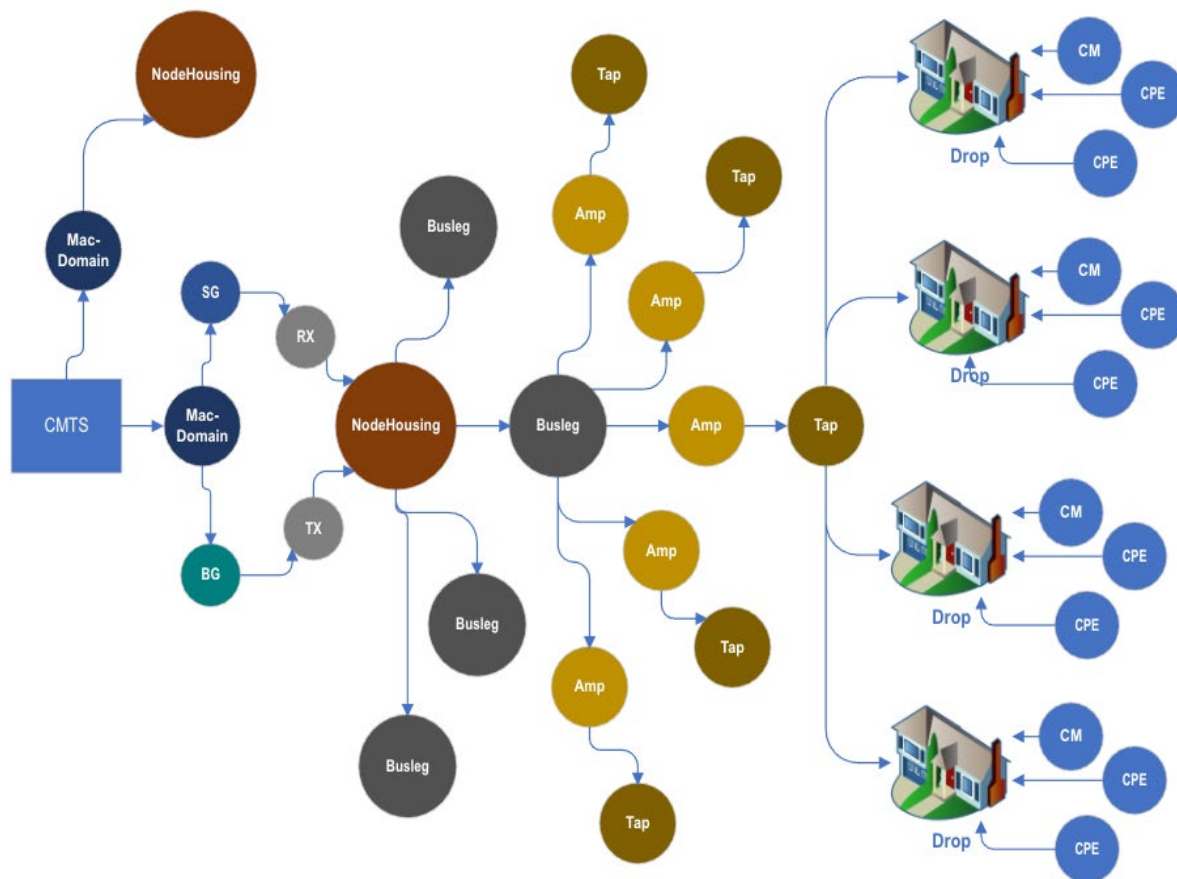


Figure 3 – High-level graph model example



Figure 4 – Example graph terminating at Rfactiveports/busleg

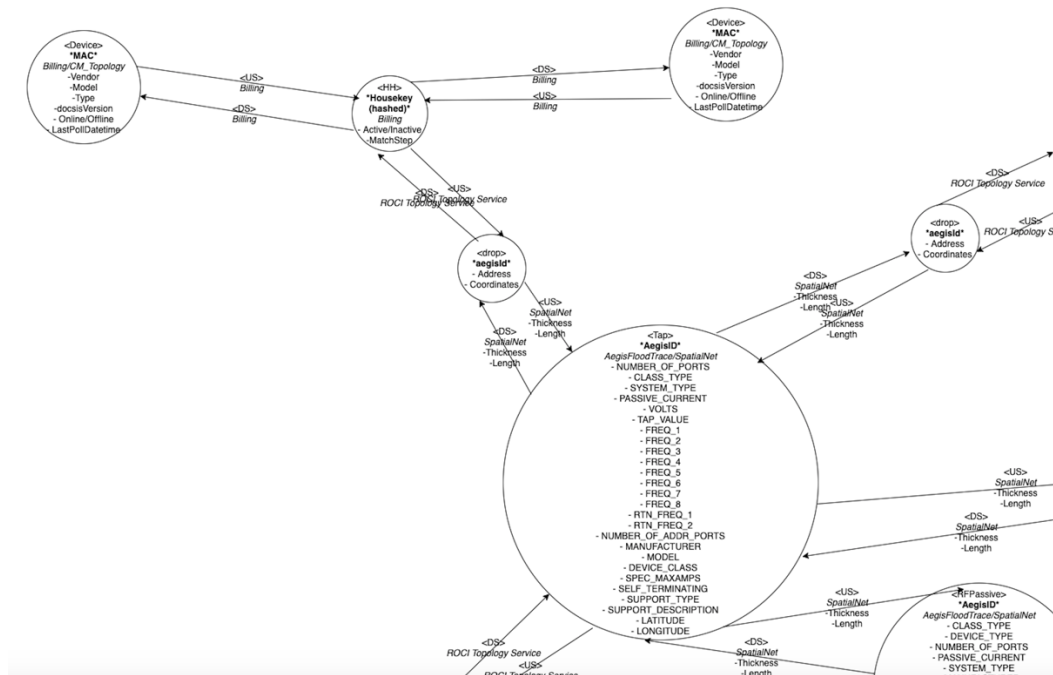


Figure 5 – Detailed graph model (tap, drop, modem)

1. Vertices

- The graph vertices are elements from the Division to a site, a CMTS in the site down to the individual devices in a subscriber's home.
- Division, Region, Site, CMTS, Headend Tx/Rx. US/DS Port, MAC Domain, Service/Bonding Group, Clamshell (node housing), Bussleg, Active (Amp), Passive, Tap, Design Address/Drop, Billing Address/HH, and Device.
- Labeled as:
'division', 'region', 'cmts', 'headendTx', 'headendRx', 'usport', 'dsport', 'macDomain', 'serviceGroup', 'bondingGroup', 'clamshell' is the node housing, 'bussleg' is the fiber node in the node housing, 'amp', 'passive', 'tap', 'drop', 'billingaddress', 'device'

2. Edges

- Edges exist between vertices in the graph that denote the relationship between the vertices.
- The edge type depends on the data source that creates the edge, the context of the edge in terms of direction, or physical nature of the relationship between vertices.
 - US - an upstream edge. This exists from device upstream through the flood trace all the way to MAC Domain and directly from MAC Domain to the CMTS.
 - DS - a downstream edge. This exists from the CMTS directly to Mac Domain down through the flood trace to device.
 - NCP_US - this is an edge created based on a confidence metric that we can map NCP reports of CMTS US ports to RX to the correct MAC Domain and bonding groups.
 - NCP_DS - this is an edge created based on a confidence metric that we can map NCP reports of CMTS DS ports to TX to the correct MAC Domain and Service Groups.

- v. `resides_in` - Used for busslegs and optical rx/tx that physically exist inside a node housing.

3.3. Software Simulation

After reviewing the physical layer model and graph topology, we now have sufficient context to begin the simulation. Each of the previously discussed models and network elements are implemented as software components. The software objects are used to instantiate the network topology and gather the required parameters to build transmission line segments. These transmission line segments can then be cascaded to simulate a complete network that is redrived from the network design and specifications.

3.3.1. Database of Component Specifications

The graph network topology provides information that identifies the equipment types, which are used to assemble the required specifications. The product specifications are usually provided in the form of human-readable documentation such as portable document format (PDF). These specifications need to be converted to a database format that can be referenced by the software. To support this simulation exercise, the scope of equipment was limited and generalized to reduce the amount of document translation required. The list network elements and their required specifications are as follows:

- **Cables** - Diameter, length, structural return loss, nominal velocity of propagation (NVP) and attenuation vs frequency
- **Taps** – Return loss vs. frequency, attenuation vs. frequency, tap loss, port-to-port isolation, port-to-tap isolation, tap-to-tap isolation
- **Conditioners and pads** – Attenuation vs. frequency
- **Amplifiers** – Return loss vs frequency, gain vs. frequency
- **Splitters (and directional couplers)** – Through loss vs. frequency, tap loss vs. frequency, return loss vs. frequency

3.3.2. Network Object Model

The software object model looks a lot like the network physical model in terms of equipment and their connections. All the typical network elements have corresponding classes which are extended from a parent class of `NetworkElement`. The abstract class of `NetworkElement` contains all the functionality that is common to all network elements, such as their graph relationships (parent-child or node-edge), attenuation vs. frequency and return loss vs. frequency.

The abstract class of `NetworkElement` is extended by all the subclasses that provide their extended implementations. Our model contains `Cable`, `Node`, `Splitter` (inclusive of `Directional Coupler`), `Tap`, `Amplifier` and `CableModem`, which each have unique characteristics beyond the abstract `NetworkElement`. The `NetworkElement` is an abstract class which is extended by several inheriting subclasses including cables, taps and so forth. An example of this class diagram is shown in Figure 8.

3.3.3. Transmission Line Object Model

Figure 6 shows a common schematic symbol for a coaxial transmission line. The source impedance is Z_S , characteristic impedance of the transmission line is Z_0 and sink impedance is Z_L .

Having the instantiated network elements, equipment specifications and connection topology (Figure 7) allows for transmission lines to be modeled and simulated. These are the minimum parameters required to perform a simulation using the proposed model: cable length, cable velocity of propagation, cable attenuation vs. frequency, source output return loss vs. frequency and sink input return loss vs. frequency.

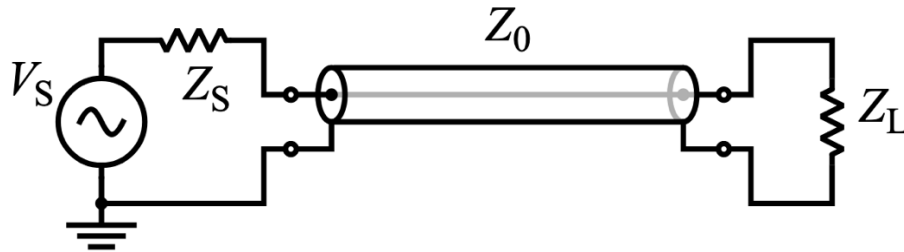


Figure 6 – Schematic symbol for a coaxial transmission line

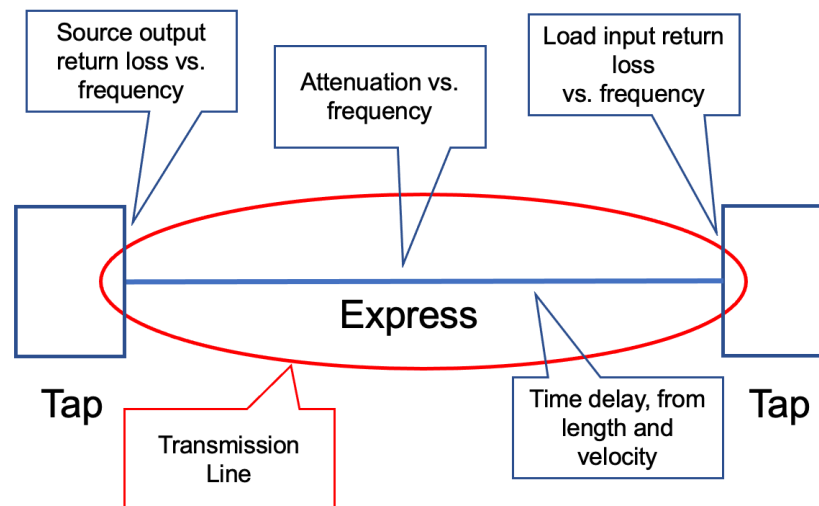


Figure 7 – Parameters required for S21 equivalent of a transmission line

To support the assembly and cascading of the transmission line parameters, a transmission software package is created to provide the implementation. Our TransmissionLine objects are constructed by traversing the fully assembled network graph, extracting the connected network elements (example, Figure 7) and copying the required parameters (Figure 7) to the TransmissionLine object.

3.3.4. Simulation of a Transmission Line

Transmission line transfer function is described earlier by Dr. Prodan in Figure 2. Recall that there is an infinite series of reflections interacting with incident signals. The series of reflections might imply that a recursive software routine could be useful. Fortunately, because of attenuation, the loss associated to the infinite recursions does eventually become insignificant (or having a practical limit). Because of this, there is a math shorthand which can be used, known as the Euler summation. This is a handy way to accelerate the summation of an infinite series.

for a length of trunk cable with cable propagation delay T , cable amplitude response $A(f)$, and tap input/output port return loss $RL = -10 \log(\rho)$ where ρ = reflection coefficient, that the transfer function $H(f)$ for the tap span cable transmission line is given by:

$$H(f) = \frac{A(f) e^{-j2\pi fT}}{1 - A^2(f) 10^{-\frac{(RL_i + RL_o)}{20}} e^{-j4\pi fT}}$$

This formulation provides the complex frequency response with only the scalar amplitude versus frequency of the transmission line and the magnitude return loss versus frequency of the terminating impedances. This avoids the need to measure complex valued S-parameters versus frequency for each component. It also avoids the conversion to t-parameters for transmission frequency responses that would be cascaded to compute the end-to-end transfer function between any two points in the network.

3.3.5. Running the Simulation

3.3.5.1. Instantiating the Network Graph

To be useful in the network simulation, the network objects need to be instantiated and assembled dynamically, driven from the graph model discussed in section 3.2 and section 3.3.

Fortunately, coaxial cable networks adhere to a tree-and-branch structure, so graph models are easily traversed without concern of recursion or circular referencing (circular dependencies). This cardinality is strictly enforced by the `NetworkElement` object having only a singular, mandatory parent with zero-to-many child relationships. Each graph vertex can be used to instantiate the object from its respective class using a factory software pattern. The unified modeling language (UML) diagram in Figure 8 helps illustrate the factory pattern.

Converting to-and-from graph and tree-branch networks may seem like an anti-pattern. However, a good way to think about the network topology is having a physical vs. logical duality. The graph model better serves the network's logical model with properties such as multiple combining of ports, service groups and MAC domains. At the same time, HFC's physical network elements are designed and built in a tree-branch structure with elements such as nodes, cables, splitters, and taps.

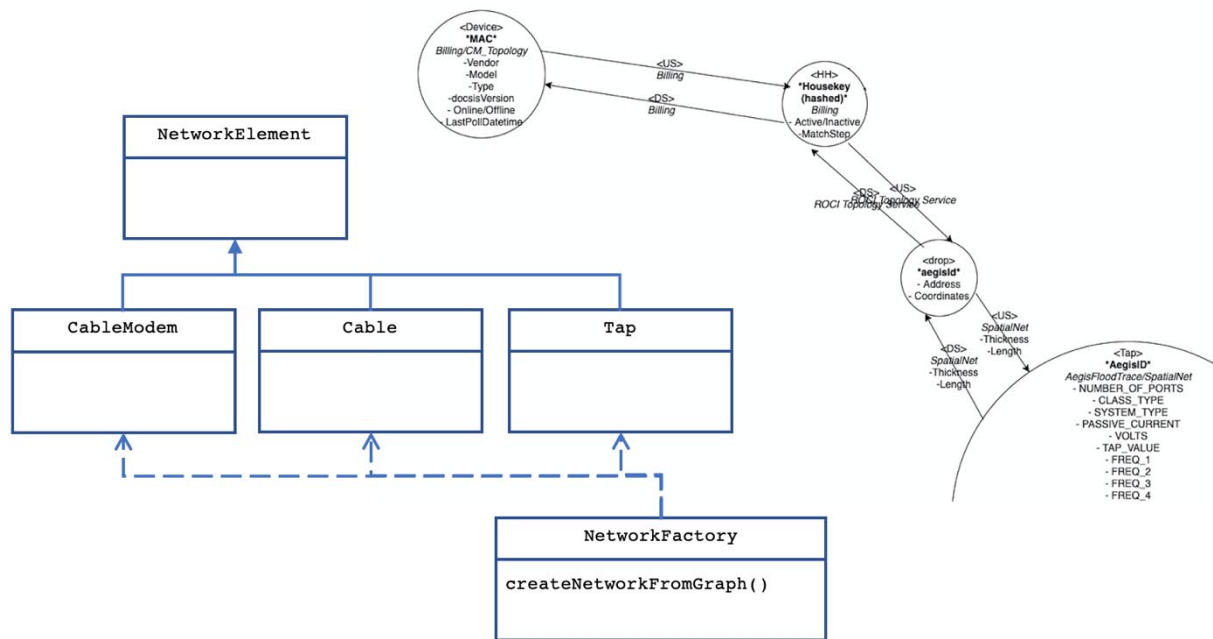


Figure 8 – UML diagram of network factory from graph

3.3.5.2. Operational Power Profiles and Launching

When simulating the transmission of RF signals on a cable network, the power levels and tilt are required. There are several different variables that go into operational launch power configurations. Especially when considering existing cable networks, different types of equipment, lengths, and types of cables, taps and architecture all influence the launch power selection. At Comcast, nearly 50 different variations of operational power launch profiles exist. As a practical matter in the simulator, the top 6 profiles will be generalized, allowing for additional power profiles to be added as needed.

The operational launch power is expressed as power levels at given frequencies, resulting in a positive tilt. This tilt is designed to accommodate the anticipated attenuation of the passive components. When designing the systems, the operational tilt is optimized to deliver a flat RF spectrum at the end of the line(s). Additional tilt can be removed by conditioning and equalizing the signal, which are also accommodated in the software model, described later.

Figure 9 and Figure 10 show two examples of operational power profiles. The tilt is measured as the difference between the 6 MHz channel power measured at 57 MHz and 873 MHz respectively. Notice that the frequency spectrum is fully occupied up to 1 GHz. In both examples, the full-spectrum tilt is higher than the specified profile tilt value. When comparing the two figures, note the power per-division on the y-axis is 2 and 5, Figure 9 and Figure 10 respectively.

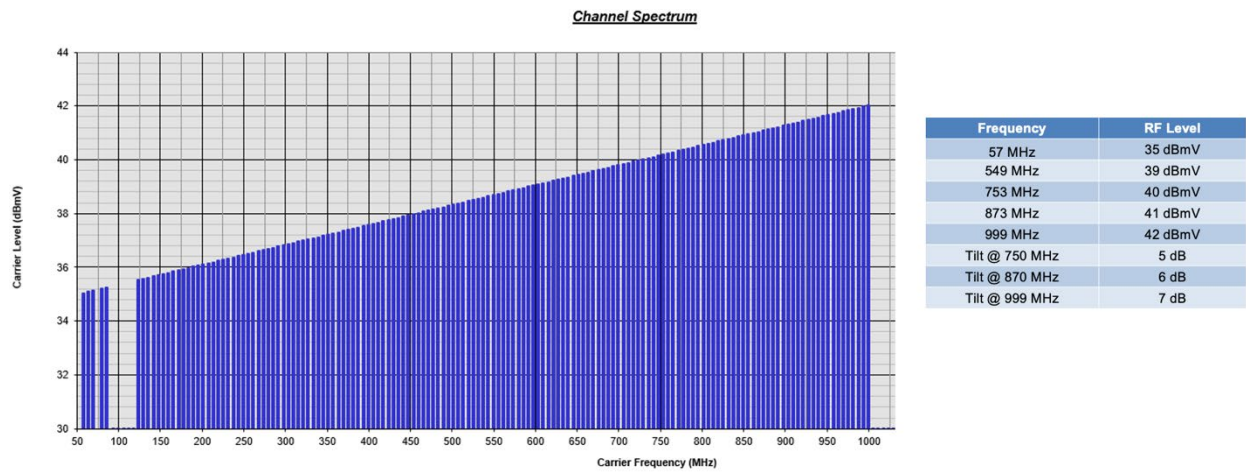


Figure 9 – Operational power profile with 6 dB of tilt

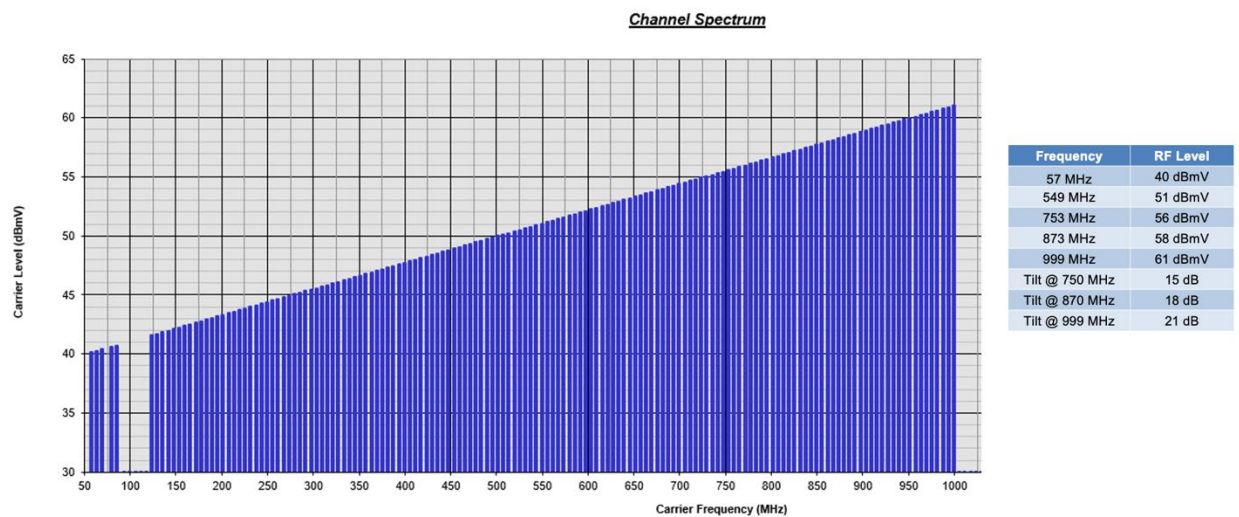


Figure 10 – Operational power profile with 18 dB of tilt

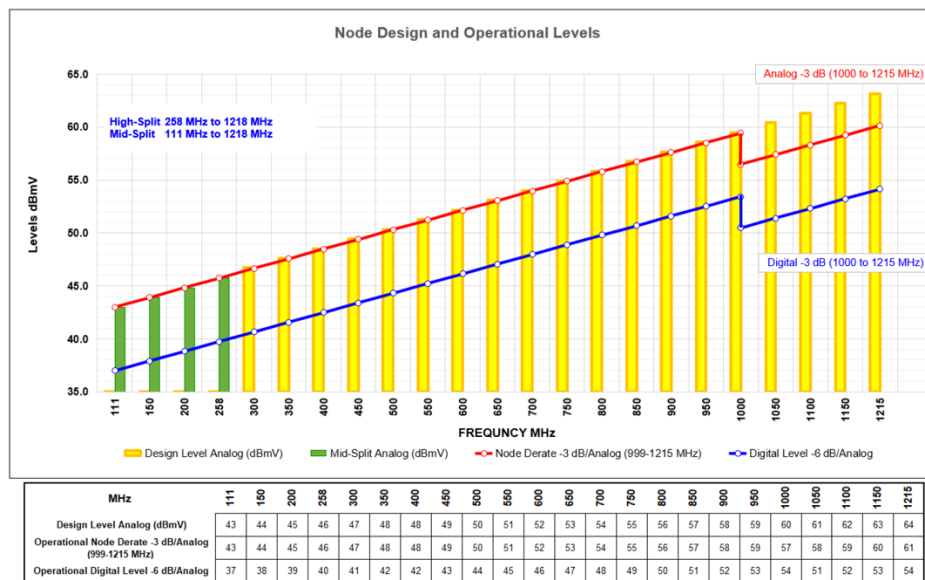


Figure 11 – Operational power profile with 1 GHz derating of -3 dB

Operational power profiles become further complicated when considering designs using frequency spectrum above 1 GHz. In contemporary system designs, RF output ports can be set with a 19 dB linear tilt from 258 MHz to 1,215 MHz (measured from CTA carrier center frequency to CTA carrier center frequency).

In this example (Figure 11), the node operating levels include a 3dB derate for all signals above 1000 MHz and for each RF output port set to:

- 43 dBmV analog-equivalent NTSC level at 111 MHz. Single-channel, 6-MHz 256-QAM level shall be 6 dB lower, or 37 dBmV at 111 MHz for initial mid-split deployments.
- 46 dBmV analog-equivalent NTSC level at 258 MHz. Single-channel, 6-MHz 256-QAM level shall be 6 dB lower, or 40 dBmV at 258 MHz.
- 60 dBmV analog-equivalent NTSC level at 1000 MHz. Single-channel, 6-MHz 256-QAM level shall be 6 dB lower, or 54 dBmV at 1000 MHz.
- 61 dBmV analog-equivalent NTSC level at 1,215 MHz. Single channel, 6-MHz 256-QAM level shall be 6 dB lower, or 55 dBmV at 1,215 MHz.

3.3.5.3. Channel Plans and Lineups

For modeling purposes, it's typical to use a standard channel plan and fully populate the RF spectrum model with 6 MHz power equivalent SC-QAM channels. In Figure 9, Figure 10, and Figure 11, it can be seen that the operational power profiles are often expressed as such. However, the simulation will be required to support all common forms of cable signal transmissions such as SC-QAM, OFDMA, SCTE-51 out-of-band (OOB), analog channels and continuous wave (CW) pilots. This makes it possible to simulate and compare channel lineups and power loaded scenarios that are found in deployed networks. Figure 12 shows 2 examples of different channel lineups. The top graph shows a simulated RF spectrum, fully loaded with SQ-QAM from 57 MHz to 1 GHz. The bottom graph shows the same frequency spectrum, mixed with SC-QAM and OFDM which are sparsely loaded. In this example, there are 5 channel blocks of SC-QAM and 1 block of OFDM.

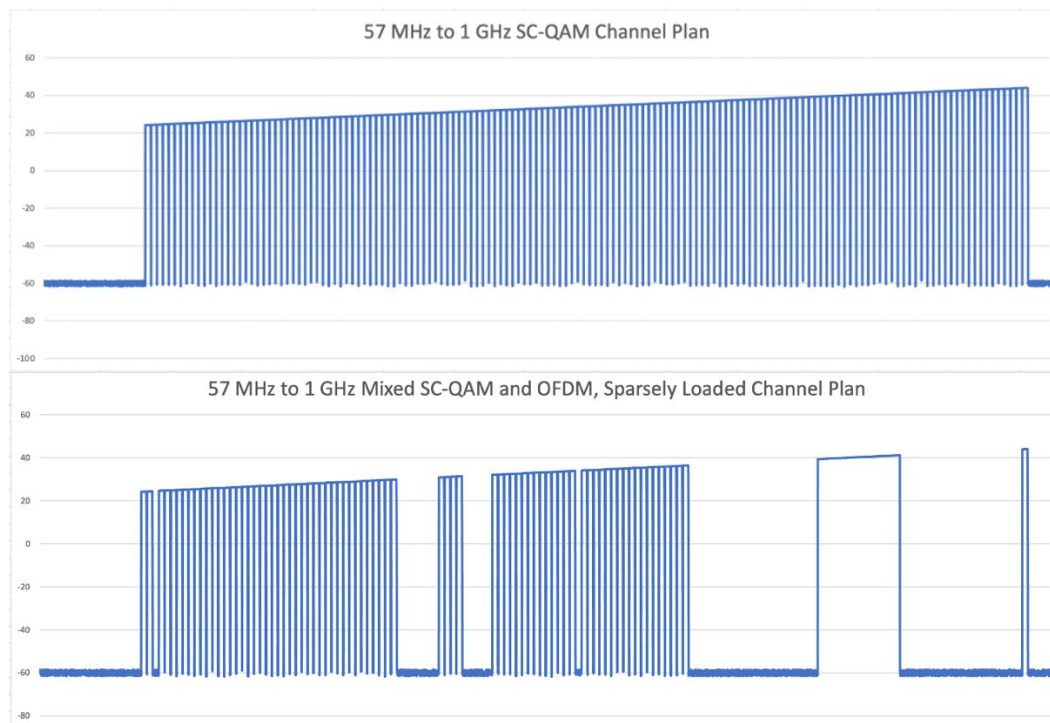


Figure 12 –Examples of channel plans and lineups

3.3.5.4. Echo Transfer Function

Consider a signal transmitted downstream from a tap output to the adjacent tap input as shown in Figure 2 with amplitude response $A(f)$ and linear phase response which has the corresponding impulse response of the cable denoted by $a(t)$. The transmitter at the signal source has (nearly) matched impedance to the drop cable but with a return loss RL_o (dB). In general, the return loss is also a function of frequency $RL(f)$. The signal traverses the cable to the tap with propagation delay T which has a (nearly) matched impedance to the cable with return loss RL_i (dB). A portion of the signal equal to the reflection coefficient $10^{-RL_i/20}$ is reflected back to the source, which in turn a portion of the reflected signal equal to the reflection coefficient $10^{-RL_o/20}$ is re-reflected back toward the tap, and so on ad infinitum. This reflected signal can be represented as a sum of the infinite series of reflections each delayed by the propagation delay T plus multiples of the round-trip time (i.e. twice the propagation delay or $2T$) of the cable.

3.4. Simulation vs. Measured Performance

Having the measured values that can be compared with simulated values enables several use-cases. Implementors can evaluate designs vs. specifications, locate significant impedance mismatches and identify unanticipated path loss or gain. In the simplest form, this can be used to improve the classification if impedance mismatches such as faulty or damaged cable and equipment. In more complex scenarios, new designs can be simulated and validated, comparing with pre-existing network designs.

3.4.1. PNM Spectrum measurements

As of November of 2012, the DOCSIS 3.0 specification was expanded to include spectrum analyzer-like functionality in cable modems. This feature is known as full band capture (FBC) and is supported by most DOCSIS 3.0 and all 3.1 cable modems. Since this time, most cable operator-deployed modems now have the FBC spectrum analysis capability. This offers a convenient tool for measuring RF spectrum performance at the location of the cable modems and CMTS.

3.4.2. Configuring, enabling, and measuring the RF spectrum

Information about controlling PNM spectrum measurements is provided in [CM-OSSIV3.1], Section D.2.4, "CM Spectrum Analysis Objects." The spectrum analyzer control parameters, defined by Table 1, are used to configure and enable the FBC. Most of the implementations have default values, so simply enabling the spectrum analyzer should result in usable information for testing the basic functions of TFTP and SNMP. The measurements are configured and enabled by using SNMP with the following MIB objects.

Table 1 - Spectrum analyzer control parameters

Attribute Name	Type	Access	Type Constraints	Units	Default
Enable	Boolean	R/W			False
InactivityTimeout	UnsignedInt	R/W	0..86400	seconds	300
FirstSegmentCenterFrequency	UnsignedInt	R/W		Hz	93000000
LastSegmentCenterFrequency	UnsignedInt	R/W		Hz	993000000
SegmentFrequencySpan	UnsignedInt	R/W	1000000..900000000	Hz	7500000
NumBinsPerSegment	UnsignedShort	R/W	2..2048	bins-per-segment	256
EquivalentNoiseBandwidth	UnsignedShort	R/W	50..500	hundredths of bin spacing	150
WindowFunction	Enum	R/W	other(0), hann(1), blackmanHarris(2), rectangular(3), hamming(4), flatTop(5), gaussian(6), chebyshev(7)		
NumberOfAverages	UnsignedShort	R/W	1..1000		1
FileEnable	Boolean	R/W			False
MeasStatus	MeasStatusType	R/O			
FileName	AdminString	R/W	SIZE(1..255)		

The spectrum measurement output is obtained either through SNMP or TFTP file transfer. Table 2 describes the structure of this file, including byte offsets for each of the discrete values reported by the test. See [CM-OSSIV3.1], Section D.2.4, "CM Spectrum Analysis Objects," for the MIB details for implementation.

Table 2 - Spectrum analysis response format

Element	Size
Size File type (value = 504E4D09)	4 bytes
Major Version (value = 1)	1 byte
Minor Version	1 byte
Capture Time	4 bytes
Channel ID	1 byte
CM MAC Address	6 bytes
FirstSegmentCenterFrequency	4 bytes
LastSegmentCenterFrequency	4 bytes
SegmentFrequencySpan	4 bytes
NumBinsPerSegment	2 bytes
EquivalentNoiseBandWidth	2 bytes
WindowFunction	2 bytes
Length (in bytes) of SpectrumAnalysis Data	4 bytes
SpectrumAnalysisData	BinAmplitudeFileData

Described in Table 2, the PNM spectrum measurements are expressed as log magnitude power, per bin. Bins represent a range of frequency spectrum, given a center frequency and frequency span. The frequency span is sometimes referred to as the resolution bandwidth (RBW) and is used to specify the width of the measured spectrum.

For further details about the PNM spectrum capture capabilities, refer to [PNM Best Practices Primer: HFC Networks (DOCSIS® 3.1)], Section 5.2.8, “How to Implement”.

3.4.3. Simulated launch of the RF spectrum

The input of the simulator begins with a channel lineup, which has an operational power profile. Then, as the mathematical models are applied and cascaded, a resulting amplitude vs. spectrum is calculated. The amplitude vs. spectrum can be calculated anywhere within the network graph, at the input or output of the transmission line segments. Figure 13 shows the output of the simulated launch of a channel lineup, using the 18 dB tilt operational power profile shown in Figure 10. The simulated channel lineup includes both SC-QAM and OFDAM channels, fully occupying a 57 MHz to 873 MHz channel plan.

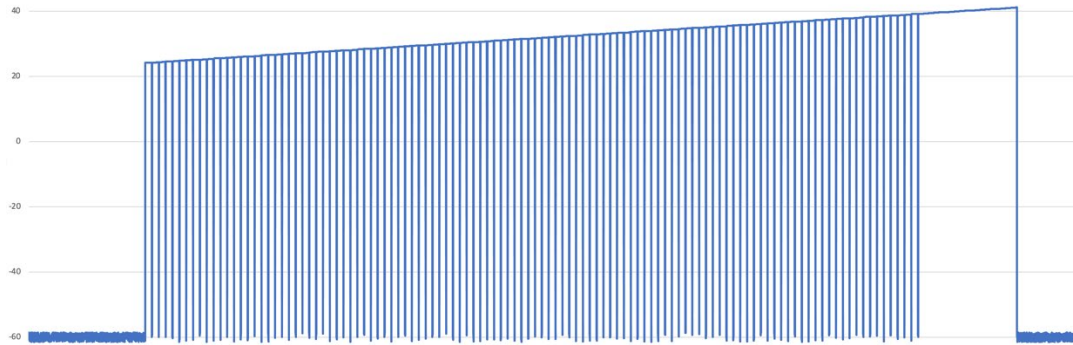


Figure 13 – Shows the simulated launch of an RF channel lineup with 18 dB of tilt

To simplify the comparison of the measured vs. simulated RF spectrum, the log magnitude power values should be scaled to the same RBW. This is a straightforward exercise using the following conversion equation. To change from a power spectral density (PSD) measured in RBW1 (i.e., dBmV/RBW1 MHz) to RBW2: $PSD2 \text{ (dBmV/RBW2 MHz)} = PSD1 + 10 \cdot \text{LOG}(\text{RBW2/RBW1})$

The process for generating simulated bins is as follows (including code examples). Note that this example does not contain any accommodating for spectrum shaping, such as a square-root raised cosine (RRC) filter. This may be a feature available in a future version.

1. Construct an array of bins using the frequency span divided by the target RBW. Our examples will use 6 MHz to 1026 MHz (1020 MHz total width) at 117187.5 kHz which results in 8704 total bins.

```
Bin[] bins = new Bin[(int)((endFreq - startFreq) / resolutionBandwidth)];
```

2. Initialize the bins with a simulated random noise floor.

```
for (int i = 0; i < bins.length; i++){
    bins[i].setLogPower(randomNoise);
    ...
}
```

3. Enumerate the channels from the simulated launch plan and render the bins from each, interpolated at the center frequency from the operational power profile. Note that different channel types may have different rendering rules, such as SC-QAM and OFDM. For example, SC-QAM will have a simple roll-off factor and OFDM will have raised pilots.

```
bins[i].setLogPower((double) interpolatedBinPower);
```

3.4.3.1. Attenuation vs. Frequency

The linear attenuation versus square root frequency characteristic of coaxial cable is predictable. This useful property simplifies modeling the behavior of signal attenuation of the interconnecting hardline cable between taps or between a tap port and the terminating device on the drop cable.

Tap parameters including insertion loss, tap loss, return loss, and tap-to-output isolation are also specified as attenuation versus frequency. Figure 14 shows the proposed attenuation model (left) compared to the simulated output of the same model (right). Both charts show the attenuation vs. frequency on 100 meters

of series 11 coaxial cable, and they agree. Further, the chart in Figure 15 shows a channel lineup using the 18 dB launch profile (green trace), compared with simulated attenuation from 100 meters of P3 .875 (amber trace), P3 .500 (blue trace) and RG6 coaxial cables (grey trace). The attenuation parameters were extracted from the component parameter database discussed in section 3.3.1.

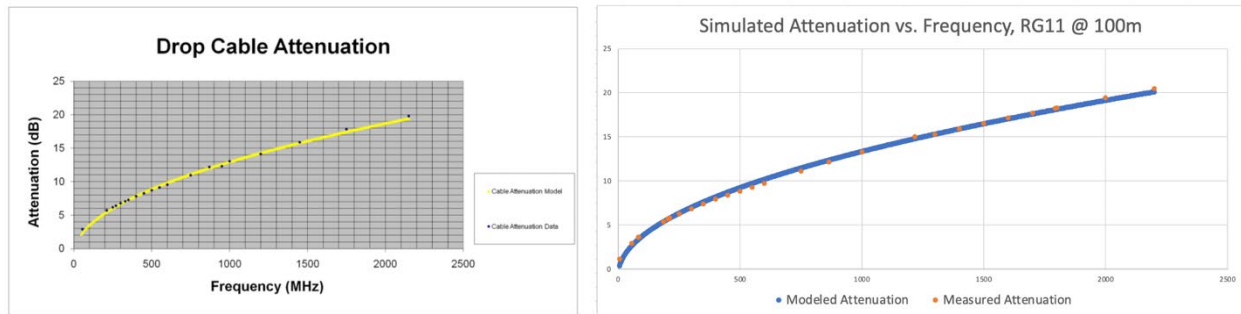


Figure 14 – RG11 attenuation model compared to RG11 simulated attenuation

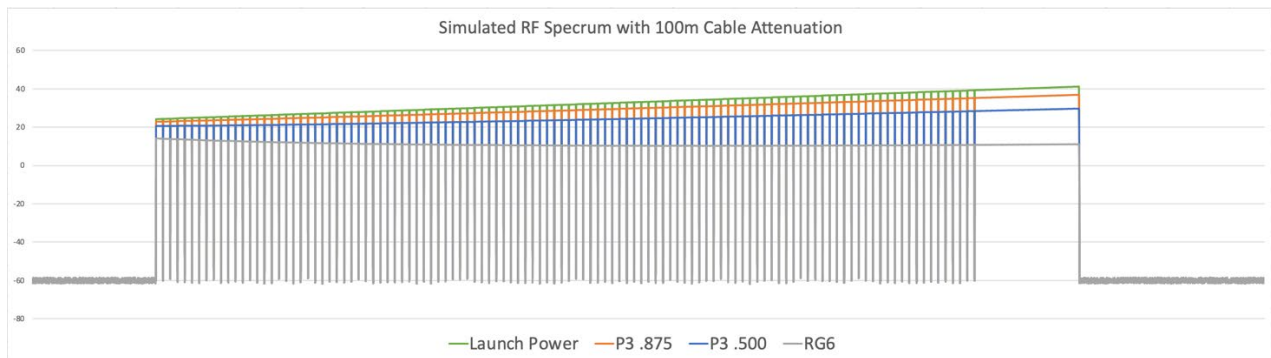


Figure 15 – Simulated launch RF compared to attenuation on P3 .875, P3 .500 and RG6 cables

3.4.3.2. Tap Equalization and Signal Conditioning

An important consideration for supporting the wider bandwidth on the passive fiber deep plant is the use of tap plug-in equalizers. Such equalizers provide either upward or downward attenuation (tilt) with increasing frequency. The amount for each tap is determined to provide approximately the same output levels versus frequency at all tap ports. An example of tap equalizer magnitude frequency (amplitude) responses is shown in Figure 16 and the simulator response in Figure 17.

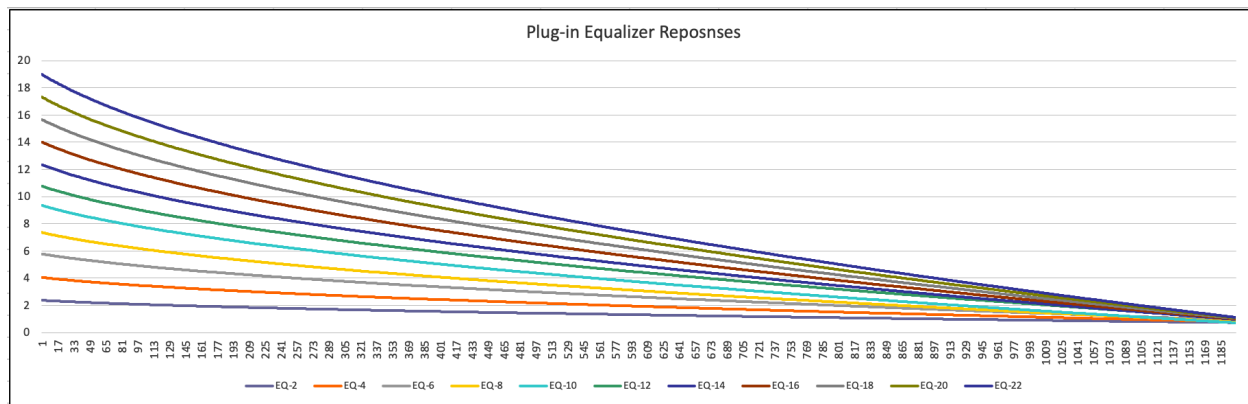


Figure 16 – Tap equalizer response vs. frequency

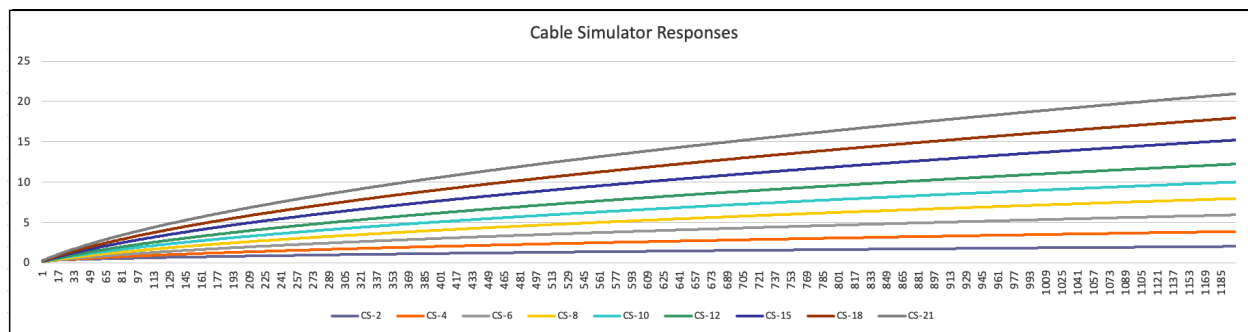


Figure 17 – Cable simulator response vs. frequency

3.4.3.3. Return Loss vs. Frequency

As seen in Dr. Prodan's model, tap parameters including insertion loss, tap loss, return loss, and tap-to-output isolation are also specified as attenuation versus frequency. Figure 18 shows an example of such tap specifications.

2-Way Maximum Insertion Loss (dB)											
Number of Ports (#-Way), Tap Value (dB)											
Frequency ≤(MHz)	2,4	2,7	2,10	2,12	2,14	2,17	2,20	2,23	2,26	2,29	
5		3.9	2	1.8	1.3	1.2	1	0.7	0.7	0.7	
10		3.6	1.8	1.6	1.2	1.1	0.9	0.6	0.6	0.6	
50		3.5	1.8	1.6	1.2	1.1	0.9	0.7	0.7	0.7	
100		3.6	2	1.8	1.3	1.2	1	0.8	0.8	0.8	
450		4.2	2.4	2.1	1.6	1.5	1.3	1.1	1.1	1.1	
550		4.2	2.4	2.3	1.8	1.5	1.4	1.2	1.2	1.2	
750		4.2	2.4	2.4	1.8	1.6	1.5	1.3	1.3	1.3	
870		4.3	2.7	2.7	2	1.6	1.6	1.4	1.4	1.4	
1000		4.3	3	3	2.3	1.6	1.6	1.6	1.6	1.6	
1218		5.2	3.6	3.6	2.7	2.2	2	2	2	2	

Minimum Return Loss (dB)		
For ALL Tap Values		
Frequency ≤(MHz)	Minimum Return Loss In-Out (dB)	Minimum Return Loss Tap (dB)
5	16	16
10	16	17
750	16	17
1000	16	17
1218	16	16

2-Way Nominal Tap Value (dB)										Tap Value Tolerance ±(dB)			
Number of Ports (#-Way), Tap Value (dB)										Number of Ports (#-Way)			
Frequency ≤(MHz)	2,4	2,7	2,10	2,12	2,14	2,17	2,20	2,23	2,26	2,29	2	4	8
5	4	7	10	12	14	17	20	23	26	29	2	2	3
1000	4	7	10	12	14	17	20	23	26	29	2	2	3
1218	4	7	10	12	14	17	20	23	26	29	2.5	2.5	3.5

2-Way Tap-to-Output Isolation (dB)											
Number of Ports (#-Way), Tap Value (dB)											
Frequency ≤(MHz)	2,4	2,7	2,10	2,12	2,14	2,17	2,20	2,23	2,26	2,29	
5		20	20	22	22	26	29	32	35	38	
10		20	20	22	22	26	29	32	35	38	
85		25	25	25	26	30	33	36	38	40	
300		21	22	23	26	30	33	36	38	40	
750		22	22	23	26	30	31	34	36	39	
900		20	20	22	23	28	30	33	35	37	
1218		20	20	20	22	25	29	31	33	35	

Tap-to-Tap Isolation (dB)	
For ALL Tap Values	
Frequency ≤(MHz)	Isolation (dB)
5	20
10	25
85	27
300	27
750	23
1218	20

Figure 18 – Tap specifications for insertion loss, tap value loss, return loss, and tap-to-output isolation versus frequency for each tap value

4. Implementation and Results

4.1. Examples

When the simulation is implemented as described in the previous sections, a number of use cases can be satisfied. At the time of writing, the simulation software and analysis were just being completed so we will review 3 preliminary results. Further development will continue, including material described in Section 5. Future

4.1.1. Plant fault isolation

PNM continues to be one of the most important tools for operators to proactively monitor and maintain their networks. By enhancing PNM with an RF transmission line simulation, its classification and accuracy can sometimes be improved.

Among the most commonly detected PNM signatures using FBC are amplitude ripples, often caused by standing waves. Standing waves are caused by impedance mismatches somewhere between the node and the modem. Having a minimum of 2 impedance mismatches, energy reflects off the second mismatched impedance, back to the first, and then some of that energy continues downstream, but delayed. This causes an impact to the signal that appears as a standing wave in the spectrum amplitude. When observed in a spectrum display, the standing wave will appear as a periodic amplitude change. This is precisely what the physical layer modeling and simulation are doing. However, in an ideally designed and deployed cable plant, there should be no amplitude ripple.

This example (Figure 19) shows log magnitude frequency response obtained using the FBC feature of PNM. In this example, the PNM measurement data has 8704 total bins representing 6 MHz to 1026 MHz frequency spectrum. Each bin spans 117.1875 kHz, or .1171875 MHz of frequency spectrum. In this case, an amplitude ripple is clearly present throughout the occupied frequency spectrum. Then in Figure 20, the

network response is simulated including transfer function of the distribution cables, tap and drop cables. The simulation is configured to match the channel plan, launch-power profile and similar output resolution bandwidth of 117.1875 kHz.

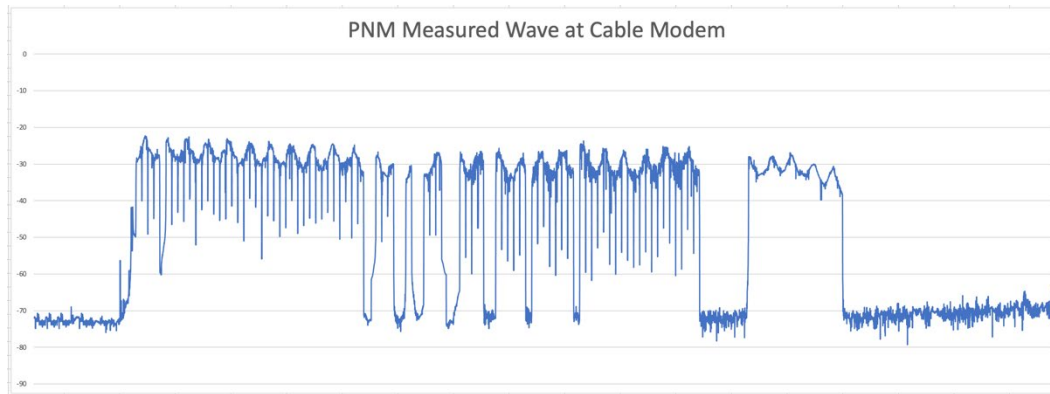


Figure 19 – PNM measurement of amplitude ripple from cable modem using PNM

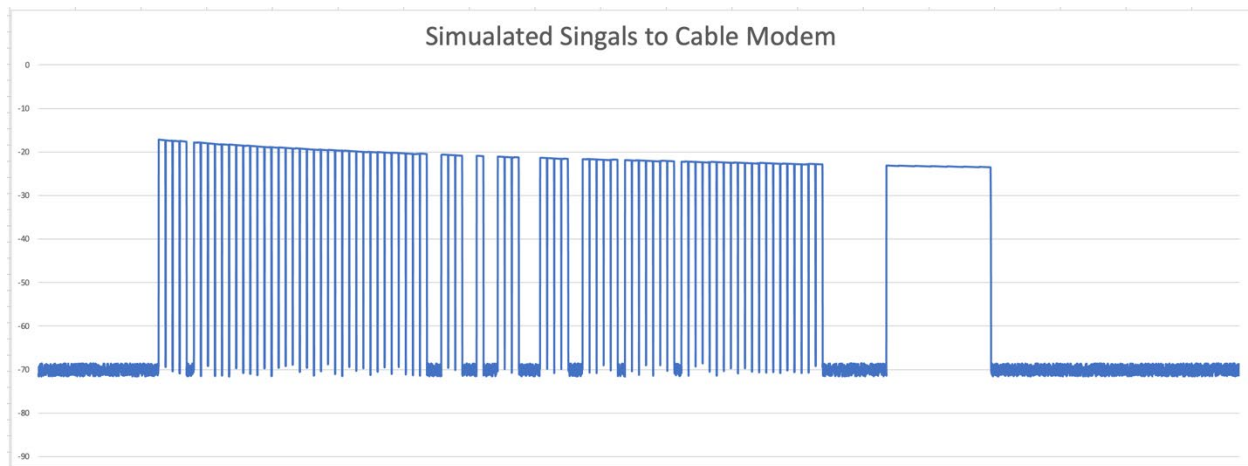


Figure 20 – Simulation of launched signal with cable, tap and drop attenuation

The resulting log magnitude bins from the PNM measurement and simulation can be processed using common digital signal processing (DSP) techniques using fast Fourier transformations (FFT). The following results were achieved by projecting magnitude-only amplitude bins to the Complex number plane. This can be achieved in a number of ways, demonstrated here and discussed further in Appendix 7.1.

This method uses “zero stuffing” of the phase component which achieves a satisfactory result, assuming that the cable plant exhibits minimum-phase characteristics.

First, the bins need to be converted from logarithmic to linear form.

```
double power = Math.pow(10.0, (0.1 * decibels));
```

Next, an array of Complex numbers is instantiated by constructing each Complex value with the magnitude and a zero-value phase component.

```
Complex[] c = new Complex[linearBins.length];

for (int i = 0; i < linearBins.length; i++){
    c[i] = new Complex(linearBins[i], 0);
}
```

The resulting Complex arrays can then be used to perform Complex division, resulting in the frequency-domain Complex quotient seen in the top section of Figure 21.

```
Complex[] quotientArray = new Complex[complexArray.length];

for (int i = 0; i < complexArray.length; i++){
    quotientArray[i] = complexArray[i].divide(denominatorArray[i]);
}
```

Finally, using the inverse fast Fourier transformation (IFFT), a time-domain response of the Complex quotient is produced (bottom of Figure 21). In this case, the frequency and magnitude accuracy of the impairment is increased by referencing the impaired signal with the simulated signal, which are otherwise unknown to the PNM analysis.

```
Complex[] complexTimeDomain = fourier.transform(getPaddedArray(binsComplex),
TransformType.FORWARD);
```

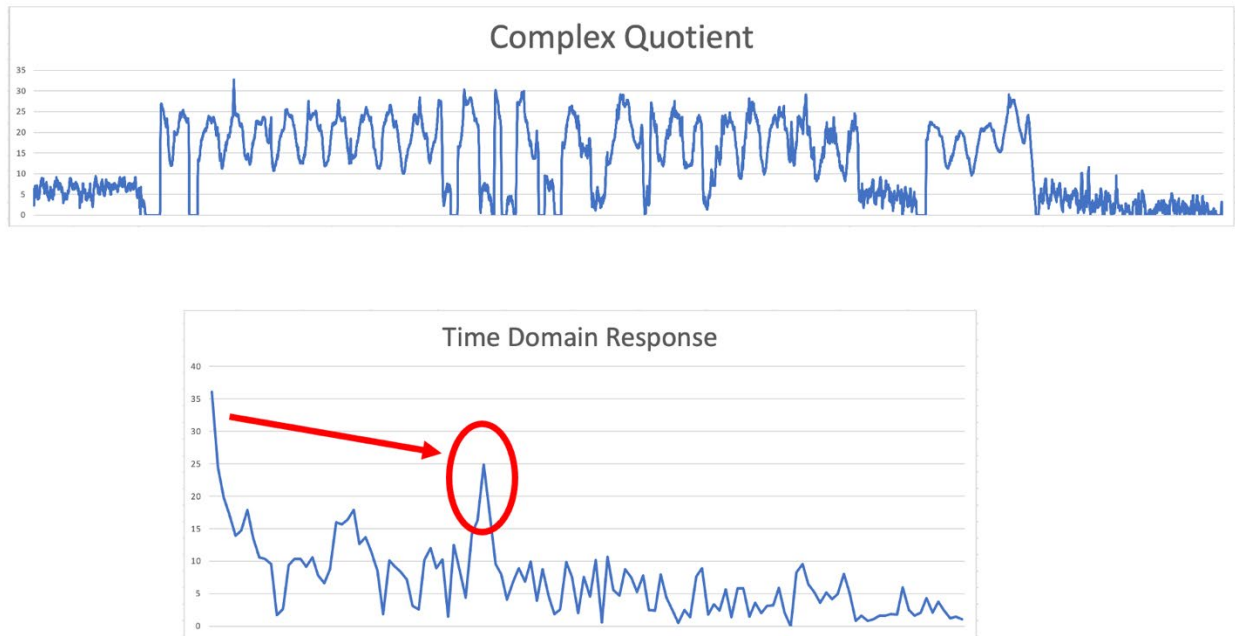


Figure 21 – Frequency(top) and time domain (bottom) analysis of Complex quotient

Having the improved frequency and time accuracy of the fault results in more precise fault lengths which can be calculated in either frequency or time.

This example was calculated, starting with the speed of light in feet per microseconds, rounded for simplicity to 983.571. The number is then divided by 2 for round-trip time and multiplied by the velocity factor of the drop cable, which is 0.85. Finally, the number is divided by the peak-to-peak ripple length of 19.28 MHz. The resulting calculation reveals that the faulty cable length is 21.68 feet, or perhaps the distance of the reflective fault from the cable modem F-connector.

4.1.2. System Design Verification

Another interesting and potentially valuable use case for simulation is to evaluate the quality and completeness of system designs. This simulation can predict end-to-end performance which can be compared against the design specifications. By traversing the network graph and comparing against PNM measurements, deviations and threshold violations can help identify inconsistencies in the designs.

4.1.3. In-home (drop) Amplifier Detection

The third example of the simulation is to help operators identify the presence of in-home (drop) amplifiers. These small amplifiers have been around for decades, used by operators and customers to add gain to the signal received by the modem. There are many different types of amplifiers including those with active and passive returns (upstream), passive, balanced and unbalanced active port configurations. In some cases, depending on equipment and configuration, it will be difficult or impossible to detect the presence of an amplifier. This method does not suggest an all-inclusive way of detecting every amplifier, rather it does propose a method that will work in some common configurations.

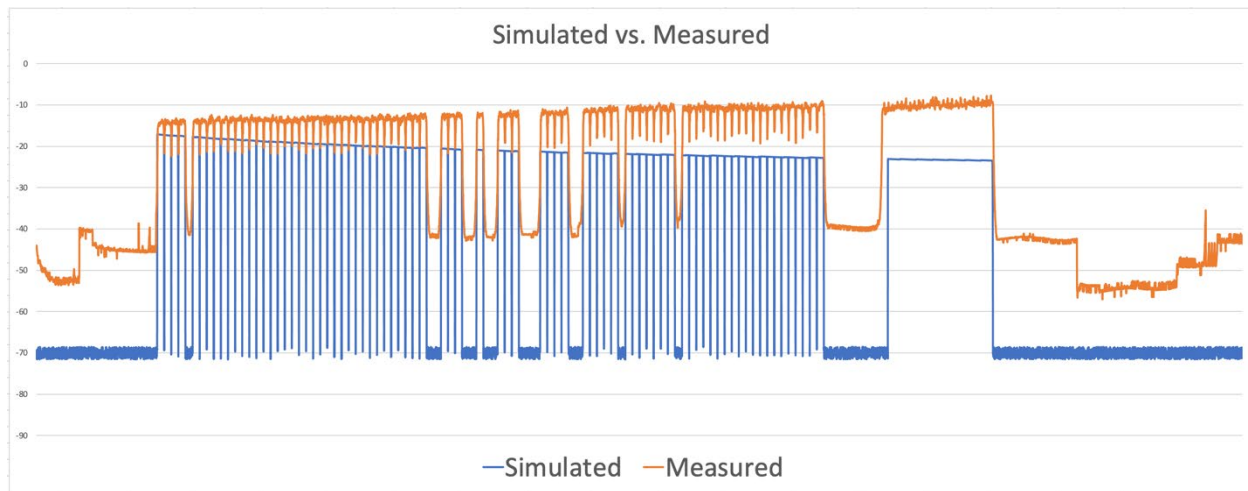


Figure 22 – Drop-amplifier detected: 15 dB gain measured vs. simulated

5. Future

5.1. FDX and ESD PNM extensions

Future versions of FDX DOCSIS technology will be supporting architectures beyond node+0. To support amplifier cascades in FDX, the amplifier technology will be equipped with echo cancellation. This complicates the physical model, which is well suited for this simulation exercise. There are several different amplifier design proposals, each making different claims about bit loading vs. distance, cascade depths and other significant performance implications.

ESD DOCSIS technology is far simpler to model, having very little change in the simulation other than the equipment specifications. Most of the cables already specify upper frequencies in the range of 2 GHz to 3 GHz. However, many of the active and passive components such as nodes, splitters and taps remain to be designed. When complete, adding their specifications to the existing simulator database is trivial.

In Figure 23, the simulated launch and attenuated signals can be compared when expanding from DOCSIS 3.1 to DOCSIS 4.0 frequency spectrum. The top chart is 5 MHz to 1 GHz of fully occupied spectrum using SC-QAM and one 192 MHz wide OFDM channel. The bottom chart shows the same lineup, adding 4 additional 192 MHz wide OFDM channels up to 1.8 GHz. This simulation was run using the 18 dB launch profile (green trace), compared with simulated attenuation from 100 meters of P3 .875 (amber trace), P3 .500 (blue trace) and RG6 coaxial cables (grey trace).

The TCP of the simulated node launch and cables are easily calculated by summing the log magnitude bins at the output of each.

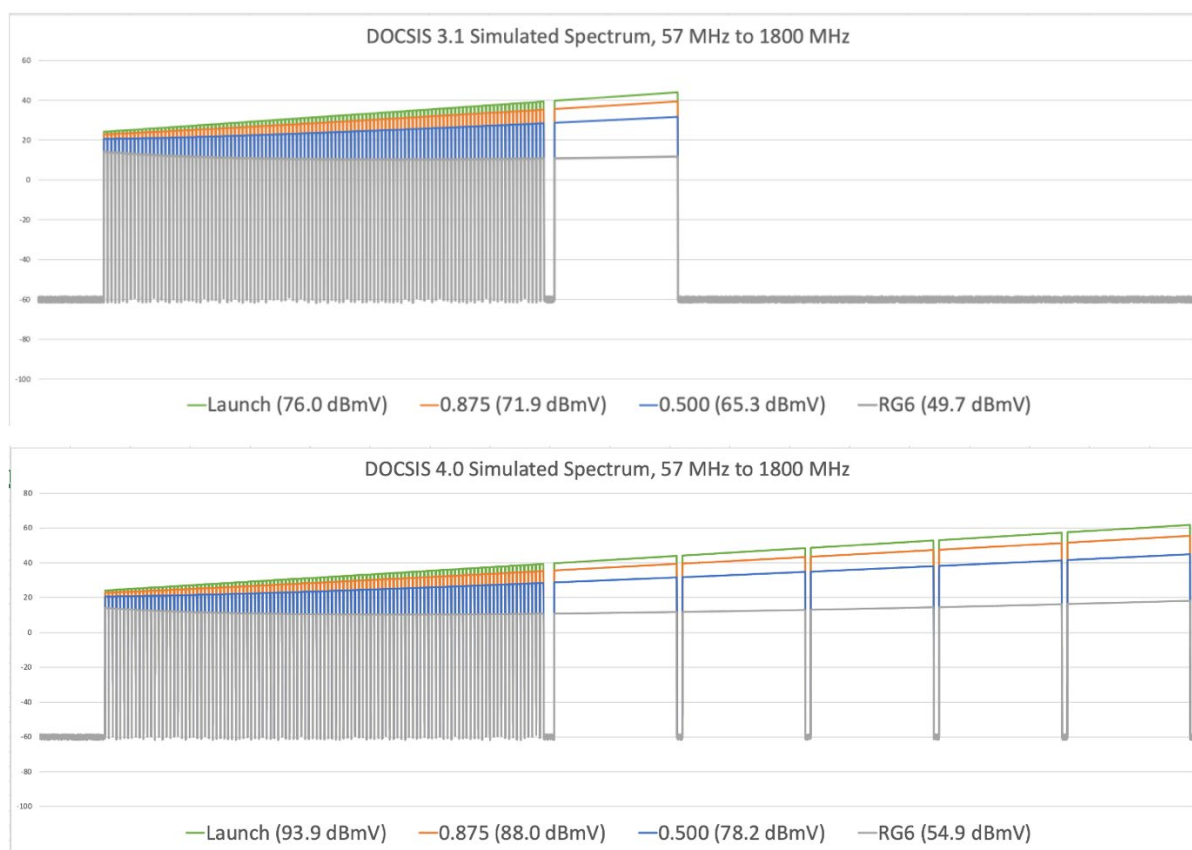


Figure 23 – Simulated spectrum of 1 GHz and 1.8 GHz compared, with TCP

Table 3 – DOCSIS 3.1 and DOCSIS 4.0 simulated output TCP compared

	18 dB Launch	P3 0.875 (100m)	P3 0.500 (100m)	RG6 (100m)
1 GHz	76.0 dBmV	71.9 dBmV	65.3 dBmV	49.7 dBmV
1.8 GHz	93.9 dBmV	88.0 dBmV	78.2 dBmV	54.9 dBmV

6. Conclusion

As our 10G networks continue to evolve and increase in complexity, so does our need to improve the sophistication of our modeling and simulations. This paper by Comcast’s Ramya Narayanaswamy, Karthik Subramanya, Richard Prodan and Larry Wolcott, examine a practical approach to doing that. They prove that software, data science and graph theory can be used to advance the theoretical FDX and ESD physical-layer models previously proposed by Dr. Prodan.

By using a graph topology of the network, paired with PNM software, cable specifications, and data science, simulating cable networks just “got real”. The simulation examples show that evaluating as-built

vs. simulated RF performance will be valuable for a number of use cases. Among them are the evaluation of full-duplex (FDX), extended spectrum (ES) RF performance. Also, traditional RF transmission performance of existing plant, as well as experimental designs. Lastly, it improves PNM fault detection, classification, and many other previously unsolved mysteries of RF.

7. Appendix

7.1. Phase recovery of magnitude-only measurements

Common digital signal processing (DSP) techniques exist to facilitate the analysis of digital signals. The PNM and simulation measurements discussed in this document are represented as magnitude-only power measurements. These measurements do not contain the constituent phase and amplitude information required to instantiate a Complex number, required for many DSP routines. There are several methods for projecting a magnitude value on to a complex plane, one of them is the Hilbert transformation.

Let $G[k]$ be the minimum phase magnitude response. First convert magnitude to nepers using the natural (base- e) logarithm.

$$H[k] = \ln(G[k]) \quad 0 \leq k \leq \frac{N}{2}$$

Next mirror the first half into the latter half (the latter half of the DFT corresponds to negative frequencies or negative times):

$$H[k] = H[N - k] \quad \frac{N}{2} < k \leq N - 1$$

There are 3 steps to compute the Hilbert transform.

$$h[n] = \mathcal{DFT} \left\{ H[k] \right\}$$

$$= \sum_{k=0}^{N-1} H[k] e^{-j2\pi \frac{nk}{N}}$$

Then, multiply every positive time index ($n < N/2$) with $-j = e^{-j\pi/2}$ (or spin those complex values by -90°) and multiply every negative time index ($n > N/2$) with $+j = e^{+j\pi/2}$ (or spin those complex values by $+90^\circ$). $h[0]$ and $h[N/2]$ (if N is even) should be set to 0.

$$h[n] \leftarrow \begin{cases} 0 & n = 0 \\ -j \cdot h[n] & 1 \leq n < \frac{N}{2} \\ 0 & n = \frac{N}{2} \\ j \cdot h[n] & \frac{N}{2} < n \leq N - 1 \end{cases}$$

Finally inverse transform that result and negate.

$$\phi[k] = -IDFT \left\{ h[n] \right\}$$

$$= -\frac{1}{N} \sum_{n=0}^{N-1} h[n] e^{j2\pi \frac{nk}{N}}$$

$\phi[k]$ is the phase, in radians, of the minimum-phase system. The complex transfer function is:

$$G[k] e^{j\phi[k]}$$

7.2. About S parameters (Courtesy of Ron Hranac)

Consider a component or device being evaluated as a “network” with some number of ports N , and a characteristic impedance Z_0 . For instance, a terminator can be considered a one-port network, as shown in Figure 1.

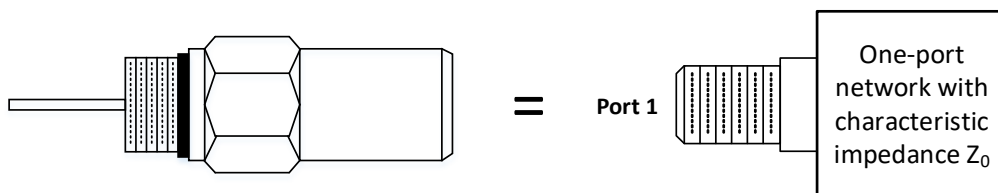


Figure 24 - Cable equipment chassis terminator represented as a one-port network.

Likewise, an in-line attenuator (or amplifier, tap, etc.) can be considered a two-port network, as shown in Figure 2.

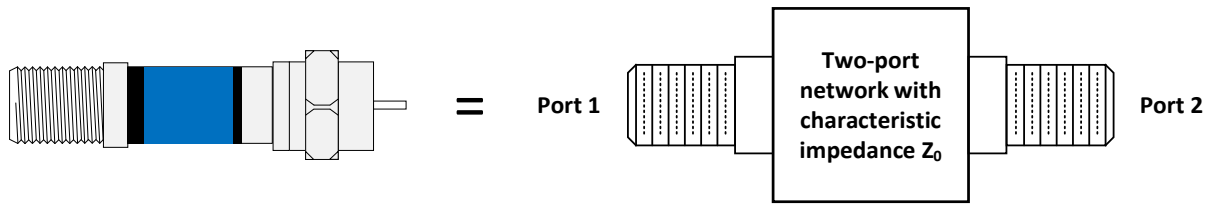


Figure 25 - In-line attenuator represented as a two-port network.

Among the metrics that can be used to characterize N-port networks are scattering parameters, or S-parameters. S-parameters are complex numbers usually expressed in the format S_{mn} , where m is Port 2 (the output port) and n is Port 1 (the input port). The remainder of this discussion focuses on two-port networks.

Assume that a two-port network such as an attenuator is being characterized. A test signal is applied to Port 1, and that signal measured at Port 2. From an S-parameter perspective, this gives us S_{21} , the forward voltage gain or transmission coefficient. See Figure 3.

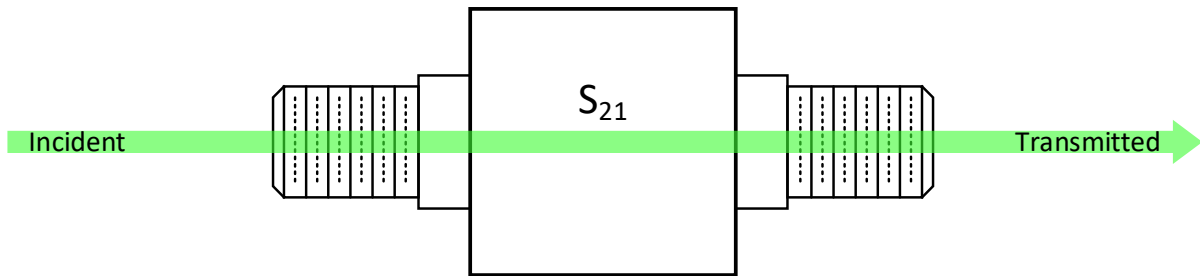


Figure 26 - Testing the two-port network from Port 1 to Port 2 gives the S-parameter S_{21} .

While the incident signal is being applied to Port 1, we can also measure any reflection from the two-port network. The reflection is a combination of reflections from Port 1's connector and all of the components inside of the two-port network. Here, the S-parameter is S_{11} , the voltage reflection coefficient for Port 1. See Figure 4.

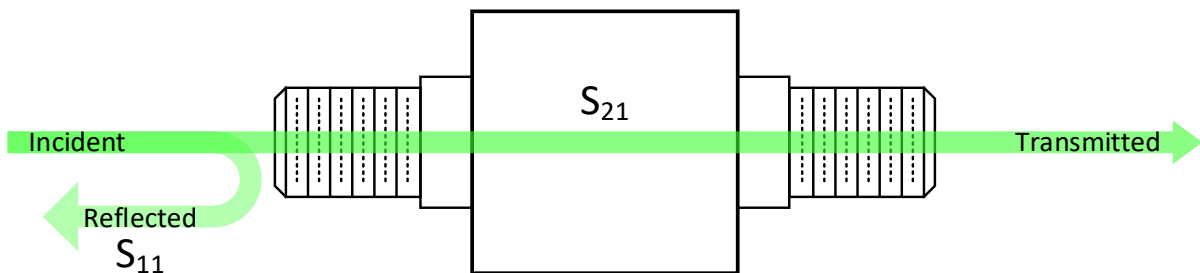


Figure 27 - Measuring the reflection from Port 1 gives the S-parameter S_{11} .

Note: When performing a measurement of S-parameters, the port(s) other than the one(s) being measured should be terminated in the network's characteristic impedance (typically 75 ohms for cable systems). An example is shown in Figure 5 for the case of a reflection measurement on Port 1, with Port 2 terminated.

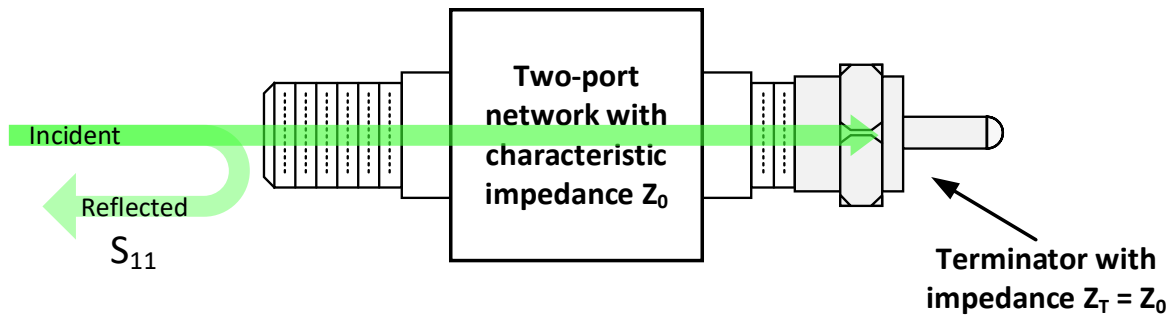


Figure 28 - It is common when measuring S_{11} to terminate Port 2 in an impedance equal to Z_0 .

Next, apply a test signal to Port 2, and measure at Port 1. This gives us the S-parameter S_{12} , the reverse voltage gain or transmission coefficient. See Figure 6.

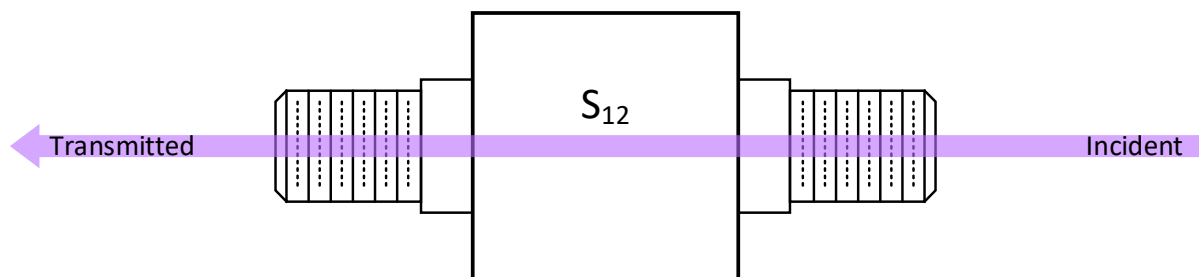


Figure 29 - Testing the network from Port 2 to Port 1 gives the S-parameter S_{12} .

Here, too, a reflection measurement can be made at Port 2, as shown in Figure 7. This gives us the S-parameter S_{22} , or Port 2's voltage reflection coefficient.

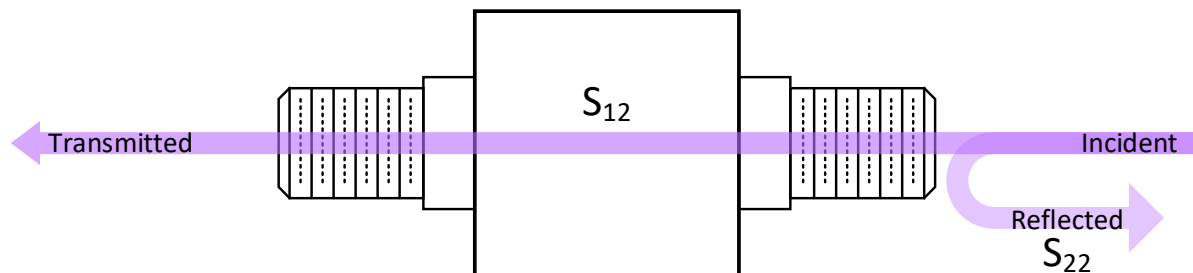


Figure 30 - Measuring the reflection from Port 2 gives the S-parameter S_{22} . In many cases Port 1 would be terminated in an impedance equal to Z_0 while performing this measurement (not shown).

In an operational FDX DOCSIS network, the FDX node's echo cancellation circuitry characterizes the network from the perspective of S_{11} , similar to the example shown in Figure 8.

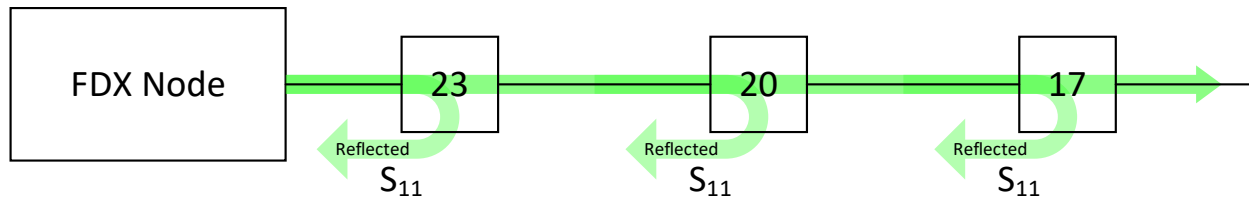


Figure 31 - FDX DOCSIS node S_{11} characterization of the cable network.

The following table summarizes the S-parameters just discussed.

Table 4 – Two-port S-paramters

Two-Port S-Parameters	
S_{11}	Port 1 (input port) voltage reflection coefficient
S_{12}	Reverse voltage gain or reverse transmission coefficient
S_{21}	Forward voltage gain or forward transmission coefficient
S_{22}	Port 2 (output port) voltage reflection coefficient

For S_{11} and S_{22} , the voltage reflection coefficient is the ratio of reflected voltage $E_{\text{reflected}}$ to E_{incident} . For S_{12} and S_{21} , the transmission coefficient is the ratio of transmitted voltage $E_{\text{transmitted}}$ to incident voltage E_{incident} .

Figure 9 shows the relationships of E_{incident} , $E_{\text{reflected}}$, and $E_{\text{transmitted}}$ when measuring from Port 1 to Port 2. Here, $E_{\text{transmitted}}/E_{\text{incident}} = S_{21}$ and $E_{\text{reflected}}/E_{\text{incident}} = S_{11}$.

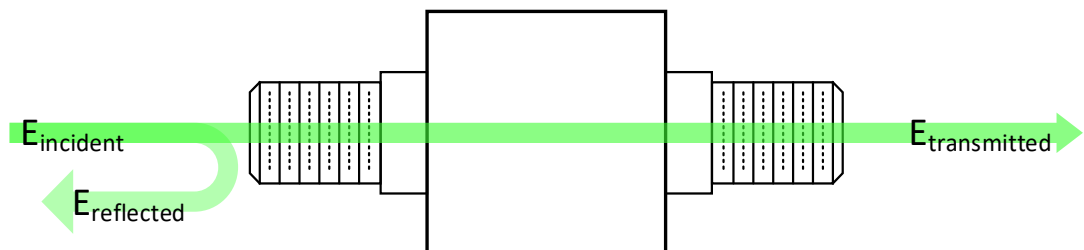


Figure 32 - Relationships of E_{incident} , $E_{\text{transmitted}}$, and $E_{\text{reflected}}$ when measuring from Port 1 (input port) to Port 2 (output port).

Going the other direction, from Port 2 to Port 1, $E_{\text{transmitted}}/E_{\text{reflected}} = S_{12}$ and $E_{\text{reflected}}/E_{\text{incident}} = S_{22}$. See Figure 10.

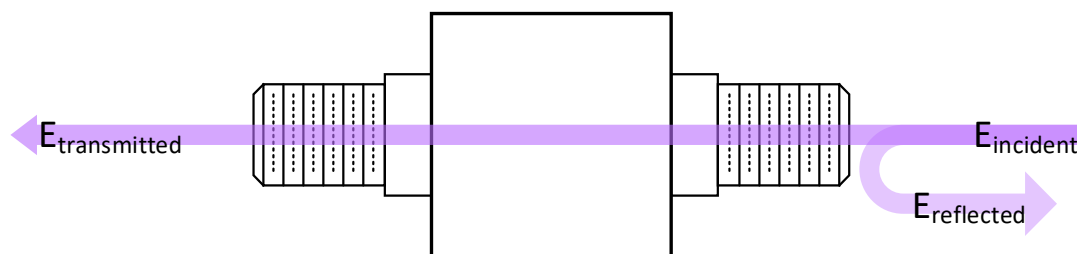


Figure 33 - Relationships of E_{incident} , $E_{\text{transmitted}}$, and $E_{\text{reflected}}$ when measuring from Port 2 (output port) to Port 1 (input port).

Cable operators are usually more familiar with characteristics such as gain, insertion loss, and return loss. Here are some common relationships derived from S-parameters (referenced to a two-port network, and assuming Port 1 is the input port and Port 2 is the output port).

Input return loss in decibels

$$R_{\text{in}} = -20\log_{10}|S_{11}|$$

Output return loss in decibels

$$R_{\text{out}} = -20\log_{10}|S_{22}|$$

Gain in decibels

$$G_{\text{dB}} = 20\log_{10}|S_{21}|$$

Insertion loss in decibels

$$L_{\text{dB}} = -20\log_{10}|S_{21}|$$

Abbreviations

CAD	computer aided design
CTA	Consumer Technology Association
CW	continuous wave
dB	decibel
dBmV	decibel millivolt
DOCSIS	Data-Over-Cable Service Interface Specifications
DSP	digital signal processing
ES	extended spectrum
FDD	frequency division duplex
FDX	full duplex
FFT	fast Fourier transformation
GHz	gigahertz
HFC	hybrid fiber-coax
HH	household
ID	identifier
IFFT	inverse fast Fourier transformation
IG	interference group
ISBE	International Society of Broadband Experts
MAC	media access control

MER	modulation error ratio
MHz	megahertz
NCP	node combining plan
NVP	nominal velocity of propagation
OFDM	orthogonal frequency division multiplex
OOB	out of band
PDF	portable document format
PHY	physical
PNM	proactive network maintenance
PSD	power spectral density
QAM	quadrature amplitude modulation
RBW	resolution bandwidth
RL	return loss
RRC	root raised cosine
SCTE	Society of Cable Telecommunications Engineers
SNR	signal-to-noise ratio
UML	unified markup language
VNA	vector network analyzer
VP	velocity of propagation

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