# Full Duplex DOCSIS over Active (N+X) Cable Networks

A Technical Paper prepared for SCTE•ISBE by

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# Introduction

To enable support for symmetric multi-gigabit services over traditional coaxial cable plants, the new "Full Duplex (FDX) DOCSIS" specification was recently introduced, now morphed into the newly established DOCSIS 4.0 specification effort [1]. This technology increases the upstream (US) capacity by allowing US signals to overlap with downstream (DS) signals in frequency and time, i.e. real Full Duplex operation. At its conception, Full Duplex DOCSIS was expected to only be deployed on fully passive cable plants (N+0), due to the challenges related to making a real full duplex amplifier that is able to simultaneously amplify signals that overlap in frequency in both the US and DS directions. Due to the significant deployment cost related to increasing the fiber penetration depth to reach the last amplifier location (N+0), several concepts have been proposed to deploy FDX DOCSIS technology over an active cable plant, i.e., N+X with X > 0.

In this paper, we will address the trade-offs of several options that allow deploying Full Duplex DOCSIS equipment on an active cable network containing amplifiers. We first briefly introduce traditional amplification methods and Full Duplex DOCSIS, and subsequently discuss three types of amplifiers in greater detail: diplexer-based amplifiers, ALI-cancelling or zero-guard-band amplifiers, and active FDX amplifiers. While the latter aims to support "real" FDX operation across the amplifier, the first two only support FDD operation across the amplifier.

## 1. Traditional Amplifiers

Traditional cable amplifiers simultaneously amplify both US and DS signals while using diplexer filters to separate the US from the DS, leveraging the fact that the US and DS signals occupy non-overlapping frequency bands. An illustration of such an amplifier is shown in **Figure 1(a)**.
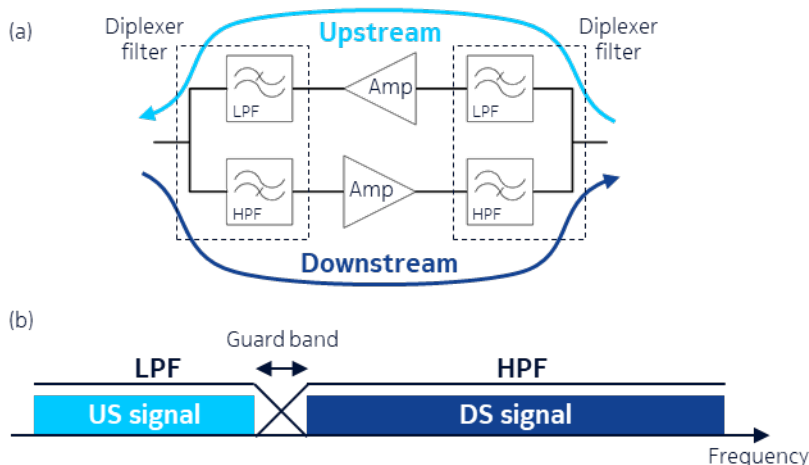


**Figure 1: (a) Illustration of a traditional diplexer-based amplifier. Amp = Amplifier, LPF = Low Pass Filter, HPF = High Pass Filter, (b) Signal spectrum allocation and filter characteristics.**

In legacy Frequency-Division-Duplex (FDD) DOCSIS the spectra of the US and DS signals are non-overlapping, as shown in **Figure 1(b)**. Due to the finite steepness or skew of practical analog filters, a guard band is required between the US and DS bands that doesn't contain any signal. The low and high pass filters attenuate signals coming from the opposite direction and isolate the US from the DS path, ensuring stability and a non-oscillatory condition for the amplifier [2].

Today's Hybrid Fiber Coax (HFC) plants contain cascades of multiple amplifiers combined with passive directional couplers and splitters/combiners, allowing to drastically extend the reach of the copper part of the HFC plant and limit the fiber penetration depth into the access network.

## 2. Full Duplex DOCSIS

The US and DS signal bands as agreed for Full Duplex DOCSIS are (see **Figure 2**):
- 5-85 MHz:           US-only band
- 102-108MHz:         Set-Top Box (STB) DS out-of-band (OOB)
- 108-684 MHz:        FDX band (i.e., both US and DS)
- 684-1218 MHz:       DS-only band

The FDX fiber node, which we will shortly refer to as the FDX node, operates in Full Duplex from 108 to 684 MHz. Outside the FDX band, traditional Half Duplex operation is maintained. Below the FDX band up to 85MHz only US is allowed, and from 102 to 108MHz only DS is allowed. Above the FDX band only DS is allowed.
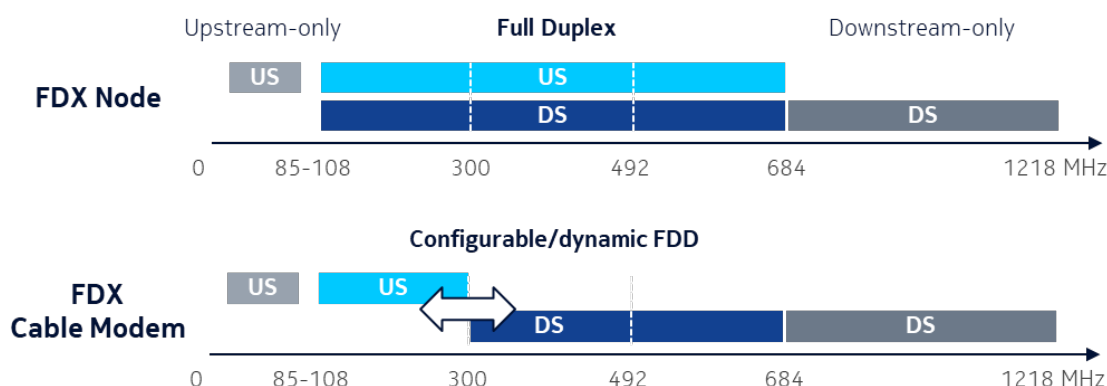


**Figure 2: FDX DOCSIS band plan from the FDX node and FDX cable modem perspective.**

The FDX node leverages in-band echo-cancelation (EC) at the node side to enable simultaneous DS transmission and US reception on the same band. On the user side, however, the FDX cable modem (CM) operates in configurable or dynamic FDD as shown in **Figure 2** in order to manage the interference arising from full duplex operation.
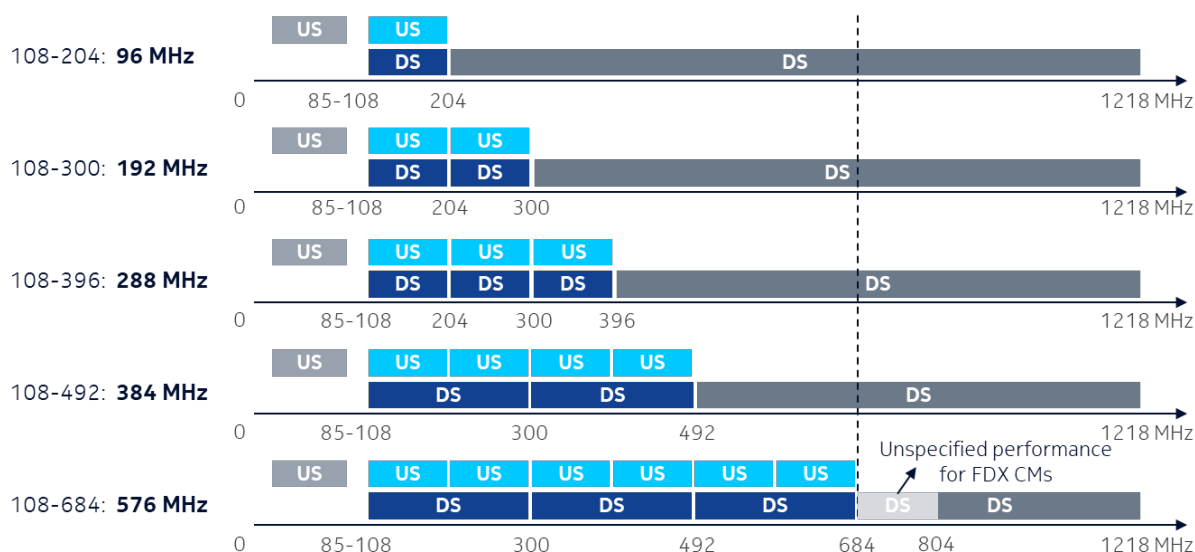
**Figure 3: Five FDX band sets supported by FDX DOCSIS, each with a different total FDX spectrum shown on the left. DS OOB from 102 to 108 MHz is not shown.**

The FDX band is subdivided in either one, two or three sub-bands, depending on how much active FDX spectrum has been configured (see **Figure 3**). Each CM operates in a specific FDD setting, referred to as a Resource Block Assignment (RBA), that defines the communication direction for each sub-band. For example, one CM might operate with RBA 100 (US-DS-DS) while another CM uses an opposite RBA of 011 (DS-US-US). All possible RBA assignments must be supported by FDX CMs (e.g., 8 for 3 sub-bands). Note that there is no guard band between the sub-bands within the FDX band. The FDX CM uses Echo Cancellation to remove the echo arising from out-of-band noise of its own US transmitter, also called Adjacent Leakage Interference (self-ALI), allowing for such a zero-guard-band.

FDX DOCSIS introduces new MAC functionality at the Cable Modem Termination System (CMTS) that deals with the interference between US and DS that arises in the network due to the FDX operation. There are two crucial components to that interference management:

1. Grouping CMs in Interference Groups (IGs). An IG is a group of CMs that highly interfere with each other. The CMTS uses a standardized "sounding" protocol to measure CM-to-CM interference on which interference grouping decisions can be based.

2. Grouping IGs into Transmission Groups (TGs). TGs are solely a logical grouping of IGs to simplify traffic scheduling. All CMs in the same TG group must use the same FDD allocation or RBA (i.e., their US and DS bands are identical).

Because CMs in the same TG use identical RBAs, interference between CMs in the same TG is avoided since there is no spectral overlap between the US transmission on the one hand, and the DS reception on the other hand. An example illustrating TG grouping and RBA assignment is illustrated in **Figure 4.**
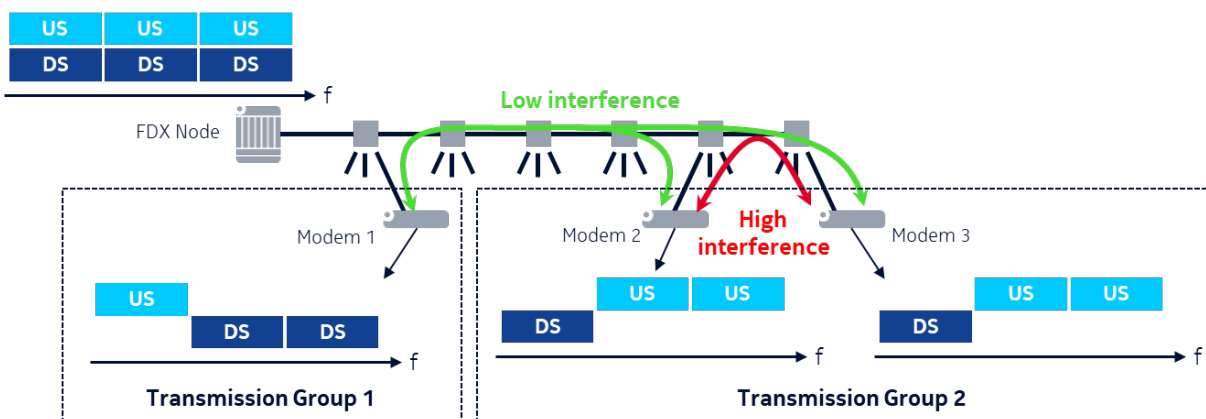
**Figure 4: Modems that don't interfere with each other (1 and 2, 1 and 3) can be assigned to different TGs and use different RBAs. Highly interfering modems (e.g., 2 and 3) are assigned to the same TG and therefore use the same RBA.**

## 3. Sounding and Echo Cancellation Training

To enable FDX functionality, two components are critical: sounding and echo cancellation training. **Sounding is used to measure the interference from one CM to another CM**, which is done by temporarily interrupting traffic, and ordering a test CM (or multiple test CMs) to send out an US signal, while other measurer CMs listen and report back to the CMTS what they have received. This way, the CMTS can create a map of the inter-CM interference and assign TGs and RBAs accordingly.

The interference addressed by sounding is the so-called Co-Channel-Interference (CCI), which represents interference between spectrally overlapping signals. Two other types of interference also exist within Full Duplex networks: Adjacent Channel Interference (ACI), and Adjacent Leakage Interference (ALI). ACI refers to the impact of the power present in an adjacent channel on the receiver chain of the desired channel. ALI refers to the impact of OOB noise from US transmitters leaking into receivers of adjacent channels. FDX CMs cancel their own ALI (self-ALI) using Echo Cancellation. However, ALI arising from different FDX CMs cannot be cancelled, and can therefore have an impact on receivers of other FDX CMs [3].

**Echo Cancellation training is required both at the FDX node as well as at the CMs to continuously track the echo channel**, which varies over time (e.g., due to temperature or wind). For both sides, there are two modes of EC training - one that interrupts data traffic and one that doesn't. For FDX CMs, these are referred to as "foreground" and "background" training. Foreground training requires assigning US grants to the CM to be used for sending proprietary EC training signals with regular power, possibly complemented by and synchronized with dummy or zero-bit-loaded (ZBL) traffic in the DS direction [1]. Background training, on the other hand, doesn't consume US bandwidth and transmits low-power signals in DS sub-bands. It is nevertheless also scheduled by the CMTS, which needs to limit the amount of CMs simultaneously engaging in background EC training. Although the use of either method is governed by requests from the CM and is vendor proprietary, it should be noted that background training by itself is insufficient for cancelling self-ALI and self-ACI. Some foreground training will always be required [1].

# Marrying FDX and Amplification

There are different scenarios that can be considered for using FDX equipment in an active cable network containing amplifiers, and the crucial difference is the amplifier type. We will consider three amplifier variants in the following sections:

- **Diplexer-based amplifiers,** which don't use Echo Cancellation and which only support FDD operation. They leverage "business-as-usual" diplexer filters.
- **ALI-cancelling amplifiers,** which do use Echo Cancellation and which only support FDD operation. Echo Cancellation is leveraged to reduce the size of the required guard band between the US and DS signal bands.
- **Active FDX amplifiers,** which do use Echo Cancellation and which can also support FDX (i.e., overlapping US and DS signal bands). Since FDD is a subset of FDX, these amps can also support FDD (i.e., non-overlapping US and DS signal bands).

In the following sections, we will examine how these three amplifier types could be used in combination with FDX nodes and FDX CMs to obtain higher US capacities over active cable networks, and will shed some light on related challenges and requirements.

Some other amplifier types have been suggested in the industry as well. For example, leveraging a switching amplifier that dynamically switches the amplification direction of the entire FDX band (either amplifying the whole FDX band in the US direction or amplifying it in the DS direction, corresponding to RBAs 111 and 000). This amplifier variant basically performs Time-Division-Duplexing (TDD) inside the FDX band. However, this proposal would effectively insert a TDD-amplifier into a system that was designed to support FDX in combination with FDD, requiring a major upheaval of the FDX specification in order to be practically feasible. Thus, we omit further discussion of this type of amplifier.

## 4. Upstream Amplification Requirements

Before diving into the details of the different amplifier types discussed in this paper, we will first shortly address the tilt requirements for the US gain, which are common to all variants. **Figure 5** shows the loss of 100 ft of a common P3 0.500 inch trunk cable (solid aluminum outer conductor, foamed dielectric with $\varepsilon_r \sim 1.32$, copper-coated aluminum center conductor) [2]. If one considers 1000 ft of such trunk cable in between two amplifiers, the cable loss between the amplifiers is equal to 10 times the losses shown in **Figure 5**. For example, at 5 MHz there would be 1.6 dB cable loss, while at 684 MHz there would be 20.2 dB of cable loss, resulting in a loss difference between those frequencies of 18.6 dB. For comparison, current low-split (5-42) or mid-split (5-85) cable plants would experience a cable loss difference between the lowest and highest US frequency of 3.2 and 6.1 dB, respectively. Although the absolute numbers vary with the length of cable between the amplifiers, the relative increase in US tilt (in dB) will be the same and amount to a factor of 3 vs. mid-split (18.6 vs. 6.1 dB) and a factor of 6 vs. low-split (18.6 vs. 3.2 dB).

| f<br>[MHz] | Loss<br>[dB/100ft] |
|---|---|
| 5 | 0.16 |
| 42 | 0.48 |
| 65 | 0.59 |
| 85 | 0.68 |
| 108 | 0.77 |
| 204 | 1.07 |
| 300 | 1.31 |
| 396 | 1.51 |
| 492 | 1.70 |
| 588 | 1.86 |
| 684 | 2.02 |
| 1218 | 2.76 |
| 1794 | 3.41 |
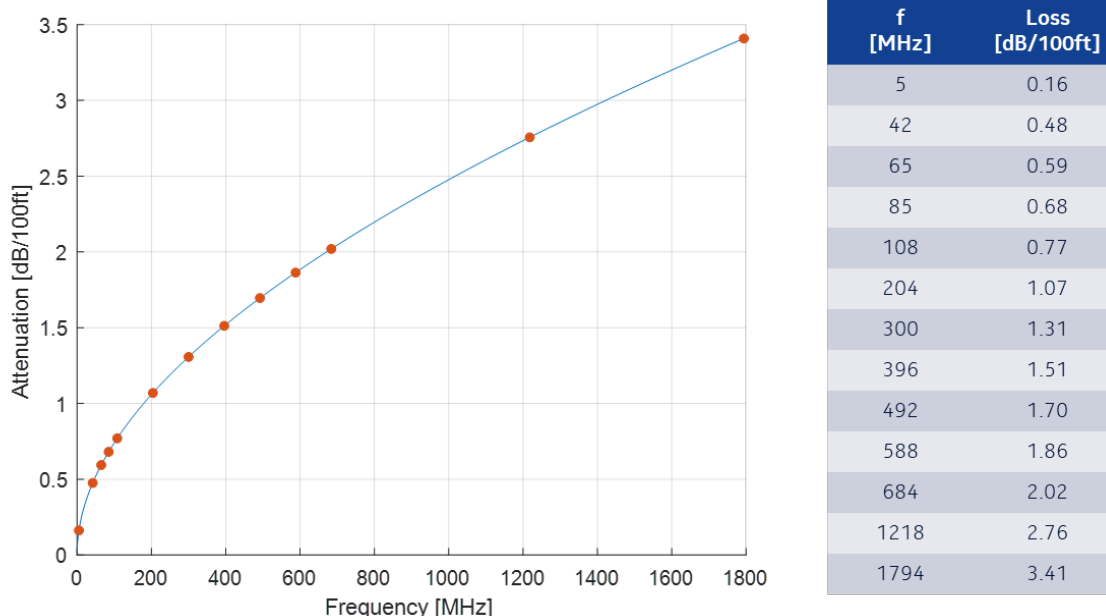


**Figure 5: Attenuation in dB/100ft  vs. frequency of P3 0.500 inch trunk cable [2]. The dots highlight the attenuation at the frequencies shown in the table on the right.**

## 5.  Diplexer-based Amplifiers

Diplexer-based amplifiers are the most straightforward type of amplifier, and can be considered "business as usual" (see **Figure 6**). They leverage diplexer filters at input and output ports to isolate the non-overlapping US and DS signals. The use of analog filters entails the requirement of a guard band between US and DS. A reasonable approximation - assuming a similar amount of isolation or stopband attenuation as commonly used today, is that traditional diplexers using lumped-element inductors and capacitors have a minimum guard band size equal to 20% of the maximum US frequency. In **Table 1** we show the size of these guard bands and the related loss in US and DS MAC-rates, assuming an efficiency of 75% and 10 (12) bit constellations for US (DS).
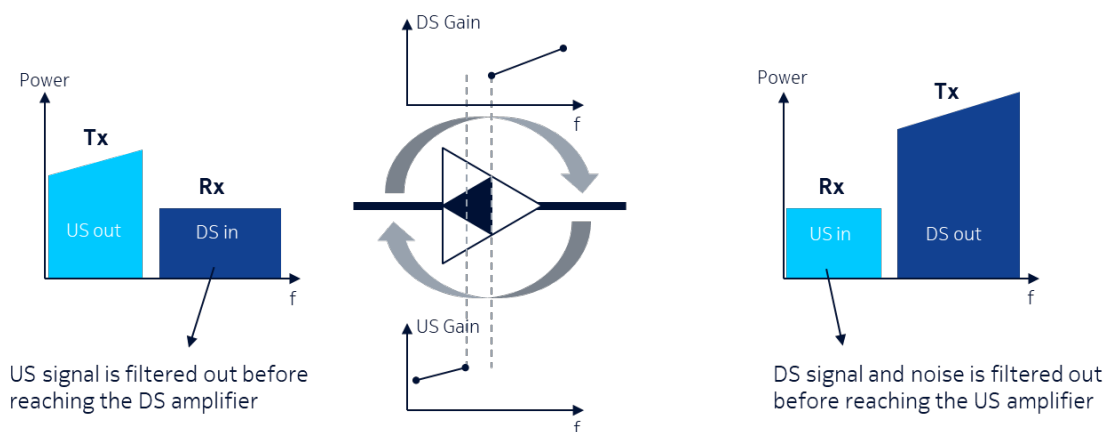


**Figure 6: Illustration of a diplexer-based amplifier, filtering out signal and noise from the opposite direction using a diplexer filter.**

**Table 1: Minimum guard band size for different splits using traditional diplexer technology. Capacity losses are calculated assuming an efficiency of 75% and 1024(4096)-QAM constellations for US(DS).**

| Max. US frequency [MHz] | Guard band size [MHz] | US capacity loss [Mbps] | DS capacity loss [Mbps] |
|---|---|---|---|
| 204 | 41 | 310 | 370 |
| 300 | 60 | 450 | 540 |
| 396 | 79 | 590 | 710 |
| 492 | 98 | 740 | 890 |
| 588 | 118 | 880 | 1060 |
| 684 | 137 | 1030 | 1230 |

From an operational perspective, one could deploy diplexer-based amplifiers in combination with FDX nodes and CMs. The amplifiers themselves can be made to support multiple splits. This can be done by manual replacement of the diplexer modules, or pre-equipping them with multiple diplexer filters and remote configurability to switch the diplexer filter. The latter will naturally result in a higher CAPEX and OPEX cost for the amplifiers (OPEX increased by power consumption of the remote configurability). The US/DS split of the cable plant can in principle be upgraded without a need to replace the nodes and CMs.

For diplexer-based amplifiers, there is no such thing as full duplex interference, and everything is "business as usual" due to the isolation provided by the filters and the existence of a sizeable guard band between US and DS spectral bands.

The main disadvantage of diplexer-based amplifiers is that there is a loss of capacity vs. echo-cancelling amplifiers that either allow for zero guard band FDD operation, or possibly active Full Duplex operation.

# 6. Echo-cancelling Amplifiers

Both the ALI-cancelling amplifier and the Active FDX amplifier leverage Echo Cancellation and hence share some common concepts that we will first address before comparing and contrasting these two types of amplifiers.

## 6.1. Commonalities

### 6.1.1. Echo Cancellation Training and Tracking

It is important to continuously track echo channel changes to ensure adequate EC performance. Even very small changes in echo channel can lead to big changes in SNR, especially when the echo signal is much larger than the desired input signal.

We denote the desired receive signal at the receiver as $u$, the echo signal at the receiver as $x$, and the receiver noise without echo channel drift as $n$. The achieved SNR after EC if the echo channel $H$ has drifted by an amount $\alpha$ to $(1 + \alpha)H$ (i.e. the EC cancels the echo channel $H$ instead of the drifted echo channel $(1 + \alpha)H$) can then be written as (see **Appendix A: SNR calculation with a varying echo channel**)

$$SNR = 10 \log_{10} \left( \frac{|u|^2}{|\alpha x|^2 + |n|^2} \right)$$

This is the resulting SNR if the EC would not account for the drift $\alpha$ in the echo channel. If the initial SNR before the channel drift, $|u|^2/|n|^2$, was 40 dB, and the echo signal level is equal to the receive signal level ($|x|^2/|u|^2 = 0$ dB), then an echo channel amplitude change of 0.5 dB ($\alpha \cong 0.06$) would reduce the SNR from 40 to 24.4 dB. This means that an echo channel amplitude change of 0.5 dB would lead to a 16 dB hit in SNR, if not corrected for (i.e., tracked). In FDX DOCSIS, the echo signal level is typically seen to be higher than the receive signal level ($|x|^2/|u|^2 = +10$ to $+20$ dB) [1]. For an echo level that is 10 dB larger than the receive signal level, the SNR is reduced from 40 dB to a mere 14 dB, i.e., an impact of 26 dB in case of a 0.5 dB drift in echo channel. For larger SNRs, the impact of an untracked drift in echo channel increases even further.

Any echo-cancelling amplifier needs to account for this and ensure proper training and tracking under all circumstances. The larger the echo signal w.r.t. the receive signal (larger Echo Cancellation depths), the more stringent the requirements on EC training.

### 6.1.2.  Echo Cancellation Depth

The "Echo Cancellation depth" is a crucial factor in assessing the EC feasibility. We define the EC depth as the difference between the level of the uncancelled echo and the desired input-referred noise level after EC. The larger it is, the more stringent the requirements on the EC.

A large EC depth requires large dynamic ranges for hardware components resulting in, e.g., high resolution requirements for components such as ADCs and DACs, and high resolution EC filter coefficient representations.

An EC depth comparable to or larger than typical useful SNR values would mean the echo level is roughly equal to or larger than the desired input signal, putting stringent requirements on the stability of the EC training as described in **Section 6.1.1** (cf. examples of $|x|^2/|u|^2 = 0/+10/+20$ dB).

### 6.1.3.  Distribution System Design with Echo-Cancelling Amplifiers

Distribution system design for active cable plants using diplexer-based amplifiers is a well-established discipline for cable operators. Thermal noise and nonlinear signal distortion contributions of the amplifiers (e.g., CTB, CSO, CIN [2]) are considered when designing an amplifier cascade. When using EC amplifiers, however, a third noise contribution needs to be considered due to the amplifiers' finite EC depth that will limit the achievable signal fidelity (SNR) at the output of the amplifier.

Similar to FDX nodes, the performance of EC amplifiers will depend on the level of the echo, i.e., the return loss of the cable plant attached to the amplifier. The larger the echo-to-useful-signal ratio, the lower the signal fidelity at the output can be expected to be. How the echo level impacts achievable SNR and how strong that coupling is, depends on the actual EC implementation.

For example, consider an EC amplifier that fully digitizes the incoming useful signal and digitally regenerates an echo-cancelled signal at the other end. The ADC used inside the EC amplifier will have a certain Effective Number Of Bits (ENOB) and sample frequency, which can be translated into an ADC noise floor. Even when the EC itself introduces no extra noise (best case assumption), the SNR of the signal at the input of the ADC represents an upper bound on the achievable SNR at the output.

Similar to thermal noise and distortion, these EC-related noise contributions add up in an amplifier cascade. For K identical amplifiers with an individual output SNR of X dB, the end-to-end SNR is $X - 10*\log_{10}(K)$ dB in case of uncorrelated noise contributions. This SNR limitation needs to be considered when redesigning the distribution system using echo-cancelling amplifiers in order to make useful predictions about end-to-end performance. **Table 2** shows the number of amplifiers that can be cascaded as a function of the single amplifier output SNR and the end-to-end SNR target. For example, in order to maintain an end-to-end SNR above 41 dB (i.e., the minimum input CNR for supporting 4k-QAM in DS in DOCSIS 3.1 [1]), a single amplifier SNR of at least ~47 dB is required to achieve an amplifier cascade depth of three.

**Table 2: Number of amplifiers that can be cascaded as a function of the single amplifier output SNR and the end-to-end SNR target**

| Single amplifier SNR | Max. # amplifiers yielding an end-to-end SNR above … | | | |
|---|---|---|---|---|
| | 47 dB | 44 dB | 41 dB | 38 dB |
| 41 | 0 | 0 | 1 | 1 |
| 44 | 0 | 1 | 1 | 3 |
| 47 | 1 | 1 | 3 | 7 |
| 50 | 1 | 3 | 7 | 15 |
| 53 | 3 | 7 | 15 | 31 |
| 56 | 7 | 15 | 31 | 63 |

### 6.1.4. Modulation-agnostic Echo Cancellation

Since amplifiers will initially be deployed in networks that will also contain a lot of non-OFDM signals like DVB and DOCSIS 3.0 signals, the amplifier will need to be modulation-agnostic. If the amplifier uses Echo Cancellation, this means the **EC needs to be modulation-agnostic**, i.e., it needs to be able to amplify and cancel echo of SC-QAM, DVB and OFDM signals. This imposes a time-domain EC implementation, which is more complex to implement - for the same EC performance - than a frequency-domain approach that is able to exploit signal modulation properties.

### 6.1.5. Breaking the Analog Chain

One can define two types of echo-cancelling amplifiers, based on whether they "break" the analog path from input to output:
1. **Non-receive-chain-breaking** have a direct analog path from amplifier input to amplifier output. The echo cancellation happens at least partially in the analog domain, but can be complemented by digital echo cancellation. Any ADCs and DACs required for the latter will be connected in parallel to the direct analog path, and do not "break" the signal path.
2. **Receive-chain-breaking amplifiers** do not have a direct analog path from amplifier input to amplifier output. The input signal is fully digitized using an ADC and regenerated after echo-cancellation using a DAC. The echo cancellation happens at least partially in the digital domain, but can be complemented by analog echo cancellation.

One difference between the two types of echo-cancelling amplifiers is the time delay the signal experiences between the input and output ports. In non-receive-chain-breaking amplifiers this delay is similar to the delay in diplexer-based amplifiers, while receive-chain-breaking amplifiers will introduce a fixed additional delay due to the conversion stages between the analog and digital domain and the processing in digital domain (this could range from tens over multiple hundreds of nanoseconds, to even microseconds depending on the processing in between the ADC and DAC). Insofar as this delay is equal in US and DS,

it will manifest itself as an additional propagation delay between the FDX node and the CM. A cascade of several such amplifiers will proportionally increase this additional (but artificial) propagation delay, and cable networks containing receive-chain-breaking amplifiers could therefore experience propagation delays significantly larger than the "physical" propagation delay corresponding to the cable length.

For non-receive-chain-breaking amplifiers, since there is a direct analog path from input to output, they also have a "closed loop" analog path going through the forward and the reverse amplifier. This requires special consideration given its criticality for the amplifier stability.

## 6.2. ALI-cancelling Amplifiers

ALI-cancelling amplifiers use Echo Cancellation to remove the echo of the OOB noise arising from the transmitter in the opposite direction (see **Figure 7**). The US and DS signals are still not allowed to occupy overlapping frequency bands and hence both the FDX node and the CMs operate in FDD mode, but the advantage compared to diplexer-based amplifiers is that ALI-cancelling amplifiers can allow for a zero guard band. The absence of a guard band would allow to regain the system capacity lost due to the diplexer guard band shown in **Table 1**.
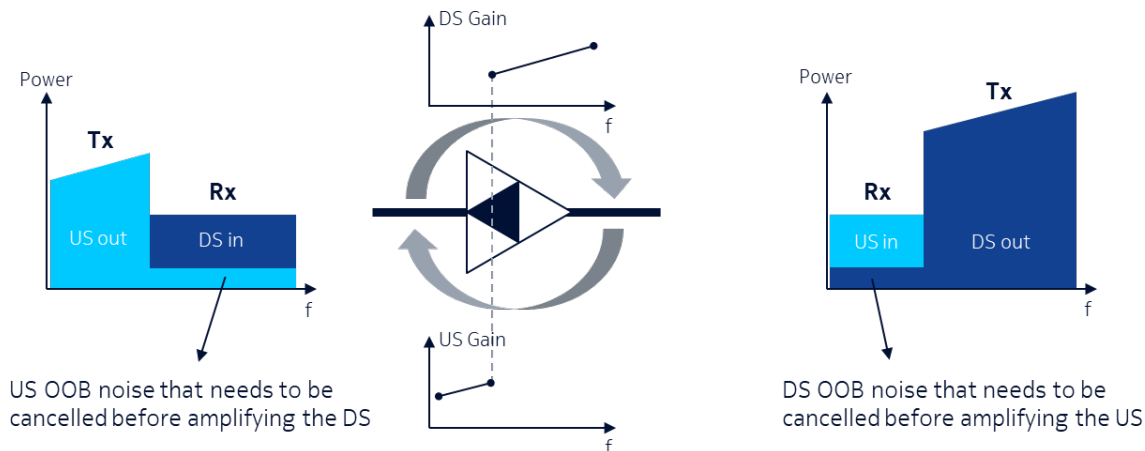


**Figure 7: Illustration of an ALI-cancelling amplifier, applying echo cancellation to get rid of the echo of the out-of-band (OOB) noise from the opposite direction.**

Echo Cancellation requirements for an ALI-cancelling amplifier are illustrated in **Figure 8**. The hybrid circuit (diamond), typically a power splitter or directional coupler, serves to interconnect the transmit amplifier, the receive circuitry and the RF output port interfacing the cable plant. It can be characterized by three loss numbers it introduces between its ports: **(1)** the loss between the amplifier and the cable, **(2)** the loss between the cable and the receiver, and **(3)** the isolation it provides between the transmitter and the receiver port. Note that **(3)** is the total echo from transmitter to receiver, including external reflections occurring after the hybrid (which add coherently to obtain peak echo level values, i.e., as $20 \times \log_{10}$), like the connector and cable plant return loss. Knowing the input and output power and noise levels at the amplifier and a target SNR to be achieved after EC, allows to derive the required EC depth, as shown in **Figure 8**.

For example, assuming:
- Line-to-Rx loss **(1)** = 7.5 dB
- Tx-to-Rx isolation **(2)** = 17 dB to 26 dB (depending on frequency)

- o This has been calculated using the reference cable plant return loss used in the CableLabs Full Duplex DOCSIS specification effort [1], as shown in **Figure 9**, combined with an assumed 40 dB of hybrid isolation and a connector return loss of 25 dB between the hybrid and the cable plant.
- Tx-to-line loss **(3)** = 1.5 dB
- US OOB Noise level at RF port = -5 dBmV/6MHz
  - o With an US output MER of 50 dB this would correspond to a +45 dBmV/6MHz US output power level
- DS Input Signal level at RF port = 10 dBmV/6MHz

yields an echo-to-signal ratio before EC of -23 to -32 dB. Assuming a target SNR of over 50 dB (to allow a decent cascade of amplifiers), this translates to a modest maximum EC depth of roughly 18 to 27 dB.

With respect to EC training, the negative echo-to-signal ratio before EC (corresponding to the $|x|^2/|u|^2$ ratio in **Section 6.1.1**) results in a reduced sensitivity to echo channel changes compared to positive echo-to-signal ratios. A -32 to -23 dB echo-to-signal ratio would result in a 0.1 to 0.7 dB impact on an SNR of 40 dB for a 0.5 dB change in echo channel, significantly relaxing EC training requirements and increasing stability compared to positive echo-to-signal ratios.

Finally, because there is no overlap in US and DS spectrum, the ALI-cancelling amplifier doesn't need to schedule silences in US and DS to allow for effective EC training and therefore doesn't need to participate in any CM EC training protocols.
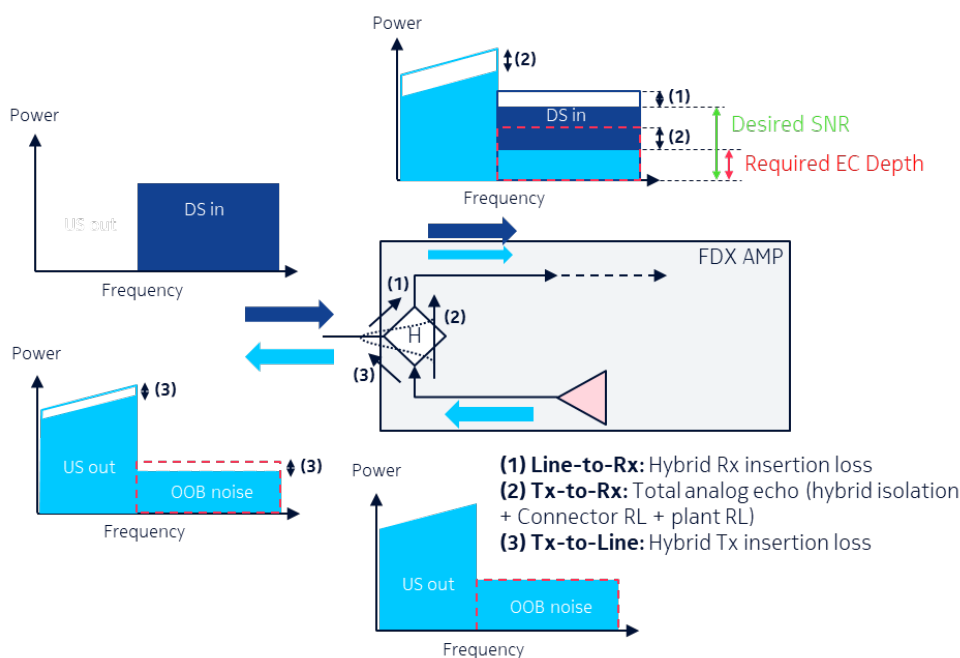


**Figure 8: Illustration of the required Echo Cancellation Depth for an ALI-cancelling amplifier in the forward direction (reverse direction is similar)**
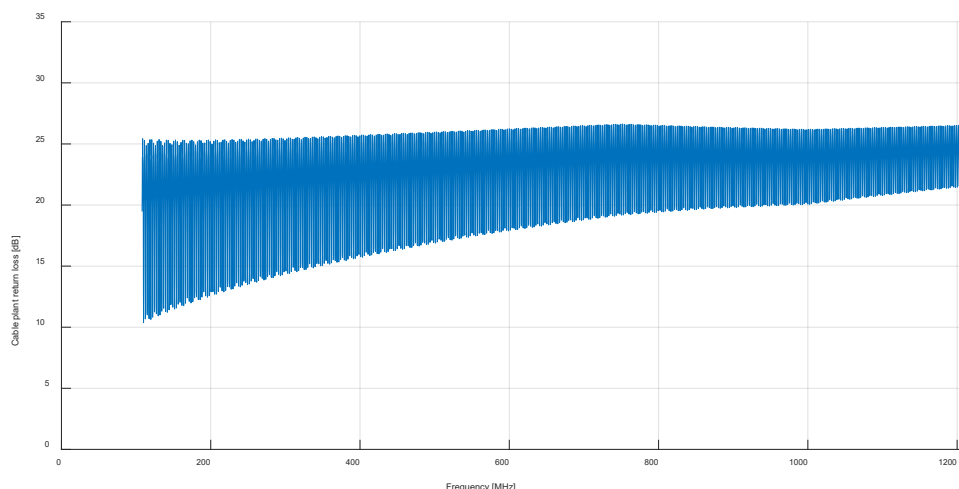
**Figure 9: CableLabs reference model for the cable plant return loss [1]**

### 6.2.1. *FDX interference with ALI-cancelling amplifiers*

Because the FDD-operation ensures spectral separation between US and DS signals, deployments leveraging ALI-cancelling or zero-guard-band amplifiers have no CCI in the network. Therefore, **for ALI-cancelling amplifiers there is no need to measure CCI (using sounding) on the cable plant**.

A cable plant containing ALI-cancelling amplifiers will, however, still experience ALI arising from US OOB noise of CMs that are transmitting coupling into the DS CM receiver of neighboring CMs. CMs are able to cancel ALI from their own transmitter, but not that of other CMs. This ALI impact will be situated at the lowest DS frequencies, right above the highest US frequencies where the US OOB noise is expected to be largest.

Previous studies [3] have shown that ALI interference from CMs on other taps ("trans-tap ALI") is negligible, while the most severe impact of ALI interference comes from CMs attached to the same tap and is mainly determined by the tap-to-tap port isolation of the tap. The worst-case isolation is typically the isolation between two adjacent tap ports that emanate from the same final power splitter inside the tap (the isolation between other tap ports should experience an isolation that is larger by at least twice a power splitter insertion loss, i.e., ~7 dB). The total attenuation experienced by the OOB noise from one CM to another CM is equal to the sum of the tap-to-tap isolation and twice the drop cable attenuation (first drop from CM to tap and second drop from tap to CM). It has been shown that ALI interference between CMs attached to the same tap can yield an impact of a couple of dB [3], which can potentially reduce the spectral efficiency in the lowest DS band by ~1 bit/s/Hz. However, it is important to note that the exact impact is highly dependent on the tap-to-tap isolation values and the CM US signal fidelity.

## 6.3. Network-level FDX operation with FDD amplifiers

It is possible to achieve Full Duplex operation in a Service Group (SG) consisting of several coax legs connected to the same FDX node, using amplifiers that operate in FDD mode (like the diplexer-based or ALI-cancelling amplifiers). Similar to the interference/transmission grouping in **Figure 4**, if CMs on different coax legs of a multi-port FDX node (that are part of a single SG) use complementary RBAs, FDX operation can be obtained at the node, provided that the interference from one leg to another is sufficiently low that CMs on different coax legs are members of different IGs. With an US Rx PSD at the FDX node

of 7 dBmV/6.4MHz and a minimum DS Tx PSD of 37 dBmV/6MHz, an isolation of barely ~11 dB between node RF ports would be sufficient to guarantee that the interference of one coax leg would be more than 41 dB below the DS Rx PSD at the CMs of another coax leg. Since isolation levels between node RF ports can easily be made larger than 11 dB, interference between different coax legs can typically be considered negligible. Therefore, the use of complementary RBAs on different coax legs is actually a very plausible mode of operation for multi-port FDX nodes in N+0 deployments.

For this scheme to be compatible with amplification, however, amplifiers with complementary FDD capabilities would be required. **Figure 10** illustrates the concept using only two FDX sub-bands. The amplifiers in the top leg support the 01 (or DS-US) RBA, while the amplifiers in the bottom leg support the complementary 10 (or US-DS) RBA. Given the existence of an US-only band below 85 MHz and a DS-only band above 684 MHz, an amplifier supporting the DS-US RBA would contain three US-DS transitions, instead of a single one for the US-DS RBA. For example, if the desired FDD split would be at 396 MHz, that amplifier would require guard bands at 85-102 MHz, 396-475 MHz and 684-821 MHz (assuming a 20% filter skewness as in **Table 1**), totaling a whopping 233 MHz of spectrum lost to guard bands.
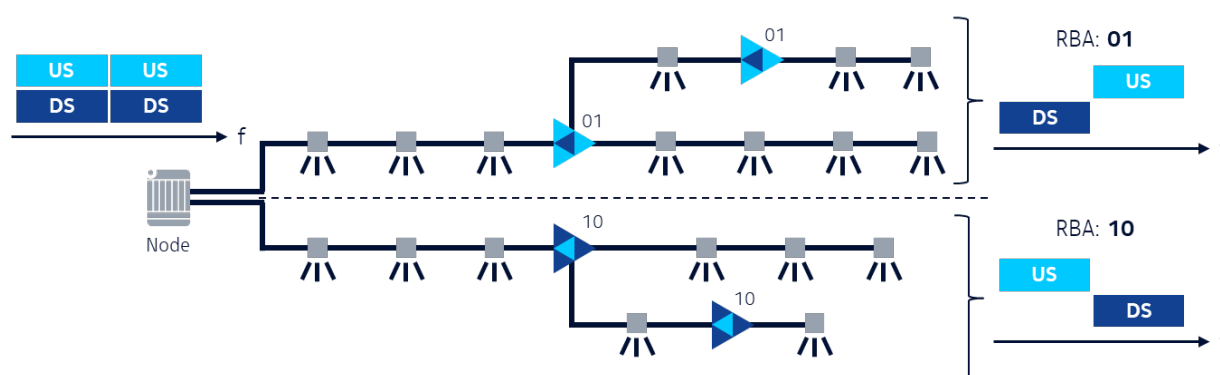


**Figure 10: A N+2 example network using FDD amplifiers, but achieving network-level FDX operation at the FDX node. The amplifiers in the top leg support the 01 or DS-US RBA, while the amplifiers in the bottom leg support the complementary 10 or US-DS RBA.**

The support for the complementary RBAs with a DS band at a lower frequency than an US band is therefore not attractive for approaches leveraging analog filtering. For ALI-cancelling amplifiers, the multiple US-DS transitions pose similar challenges to their implementation, making it very difficult to exploit the spectral separation of US and DS signals using analog filtering to alleviate the EC. Due to the multiple transitions, their implementation challenges would become more comparable to those of active FDX amplifiers (which we will discuss in the next section). This is important because a more challenging EC will reduce the depth of the cascade that's attainable using this type of amplifier (cf. **Table 2**).

From an operational perspective, network-level FDX operation using FDD amplifiers requires having two different types of amplifiers in the field whose different locations need to either be mapped out before deployment (two amplifier hardware types) or somehow set at node bring-up (single amplifier hardware that is remotely configurable).

So although it is possible to achieve FDX on a network level using such FDD amplifiers, implementation (excessive guard bands in case of diplexer-based amplifiers and reduced EC feasibility for ALI-cancelling amplifiers) and operational challenges make a FDD-based operational model on the network level much more realistic and attractive.

## 6.4. Active FDX Amplifiers

An active FDX amplifier, as illustrated in **Figure 11**, needs to apply Echo Cancellation at each receiver to remove the spectrally overlapping echo of the high-power transmit signal and noise that is transmitted in the opposite direction before amplification. The calculations illustrated in **Figure 8** also apply to active FDX amplifiers, except for the fact that the echo signal now is the actual US signal instead of the US OOB noise.
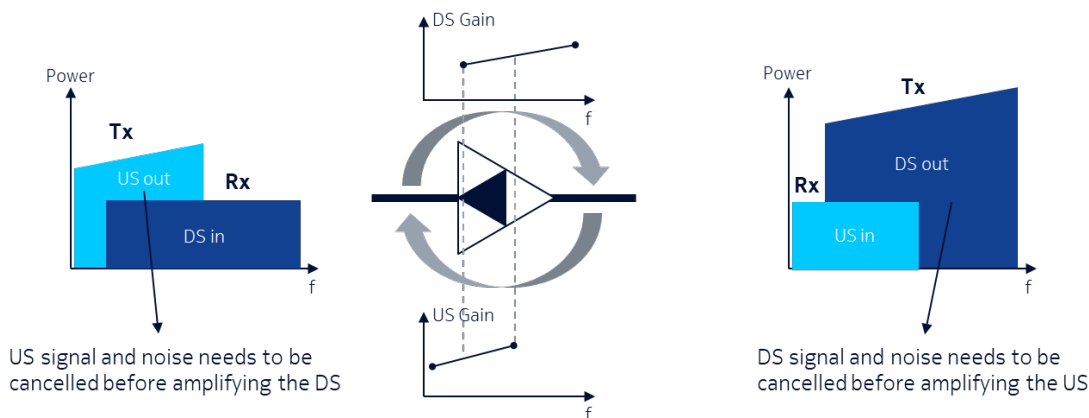


**Figure 11: Illustration of an active FDX amplifier, applying echo cancellation to get rid of the echo of the high-power signal and noise from the opposite direction.**

Adjusting for that difference would add approximately 40-50 dB (the US signal fidelity), to the required EC depth and the echo-to-signal ratio. This means the required EC depth increases from somewhere in the 20s/30s to somewhere in the 60s/70s dB, and the echo-to-signal ratio increases from negative 20s to positive 10s/20s dB, drastically tightening the EC performance requirements vs. ALI-cancelling amplifiers.

With respect to EC training, the positive echo-to-signal ratio in combination with the high SNR desired in echo-cancelling amplifiers puts very stringent requirements on the stability of the echo channel tracking (see **Section 6.1.1**). Since US and DS signals overlap, the residual error the EC algorithms try to observe for training is buried deeply below those signal levels. In combination with the very stringent requirements on EC stability it is very likely for them to require scheduled silences in US and DS for performing EC training (i.e., foreground EC training), which requires synchronization and interaction between the active FDX amplifier and the CMTS, and which implies frequent disruptions to US and DS data traffic on the plant.

### 6.4.1. FDX Interference with Active FDX Amplifiers

For active FDX amplifiers, CCI is the predominant source of interference due to the overlap in US and DS signal spectra. Since active FDX amplifiers amplify both US and DS on the same frequency, there can be CCI impact caused by transmissions from CMs on one side of the amplifier to CMs on the other side of the amplifier. **Figure 12** shows a schematic of an N+1 cable plant that we will use to illustrate the impact of CCI in a cable plant leveraging active FDX amplifiers. Consider a CM attached to a specific tap "A" US from the active FDX amplifier. $X_1$ is the loss from the node to the input of tap A, **IL** is the insertion loss from input to output of tap A, $X_2$ is the loss from the output of tap A to the input of the amplifier, **TL** is the loss between the tap input and the tap port, **ISO** is the isolation between the tap output and the tap port, **D** is the drop cable loss, **P** is the node DS transmit power level, and **R** is the node US receive power level.
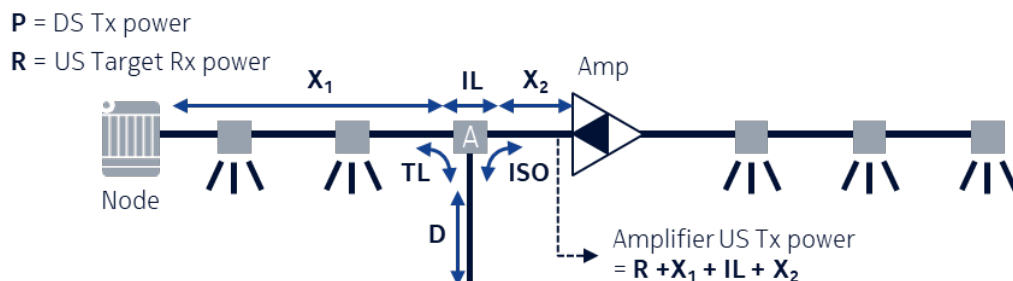
**Figure 12: Example of an N+1 cable plant to illustrate the impact of CCI. $X_1$ = loss from node to tap A input, IL = loss from tap A input to output, $X_2$ = loss from tap A output to amplifier input, TL = loss from tap A input to tap, ISO = loss from tap A output to tap, D = drop cable loss, P = node DS transmit power level, R = node US receive power level. The amplifier US Tx output power level is assumed to be configured to reach the FDX node at the US target Rx power level R.**

If we define the directivity of tap A as the difference between the output-to-tap isolation and the tap loss value (input-to-tap), i.e., $DIR = ISO - TL$, the signal-to-interference ratio **SIR** (i.e., the DS Rx MER of CMs attached to tap A when CMs after the amplifier are transmitting in the US direction) is given by (see **Appendix B: CCI calculation with Active FDX Amplifiers**):

$$SIR = P - R - 2X_1 + DIR - IL \qquad (1)$$

This expression is an upper-bound on the Rx MER achievable under interference since it doesn't consider any other noise contribution than the interference itself. The SIR needs to at least be larger than ~41 dB (CNR requirement for supporting 4k-QAM in DS [1]) to be considered interference-free. The last term, the tap insertion loss **IL**, is often smaller than 2 dB and therefore only has a small impact. Only for taps with small tap loss values and/or high port count will the **IL** be larger than 2 dB. **Equation (1)** shows that the SIR of the CMs attached to a tap decreases with increasing tap **IL**.

Since some of these values are frequency-dependent, we consider two different frequencies in the FDX band in **Figure 13** (the middle of the first and third sub-band for a 576 MHz-wide active FDX band): (a) 204 MHz and (b) 588 MHz. We assume a DS Tx PSD of **P** = 37 dBmV/6MHz at 108MHz to **P** = 58 dBmV/6MHz at 1218 MHz, a target Rx PSD at the node of **R** = 7 dBmV/6.4MHz = 6.7 dBmV/6MHz (the value required for 1k-QAM support in US [1]), and an US power on the reverse output port of the amplifier sufficient to achieve the desired target Rx PSD at the node (i.e., equal to **R + $X_1$ + IL + $X_2$**). The tap **IL** is assumed to be 1 dB.

For each frequency, the SIR is plotted in dB as a function of the tap directivity **DIR** (different lines, varied from 5 to 30 dB) and the loss from the node to tap A (horizontal axis), which can be considered proportional to the amplifier spacing. The 41 dB SIR threshold is shown as a horizontal black line, as reference. According to datasheets of several tap manufacturers the typical directivity **DIR** of taps, as defined above, is of the order of ~5 to ~15 dB (corresponding to the blue, orange and yellow lines) [4-6]. For example, in order to achieve a SIR above 41 dB at 204 MHz, the loss between the node and the tap cannot exceed ~2.5 dB for a tap directivity of 15 dB. Referring to **Figure 5**, this corresponds to a length of P3 0.500 cable of roughly 250 ft. Similarly, at 588 MHz, this loss cannot exceed ~6 dB, which corresponds to a length of P3 0.500 cable of roughly 320 ft.
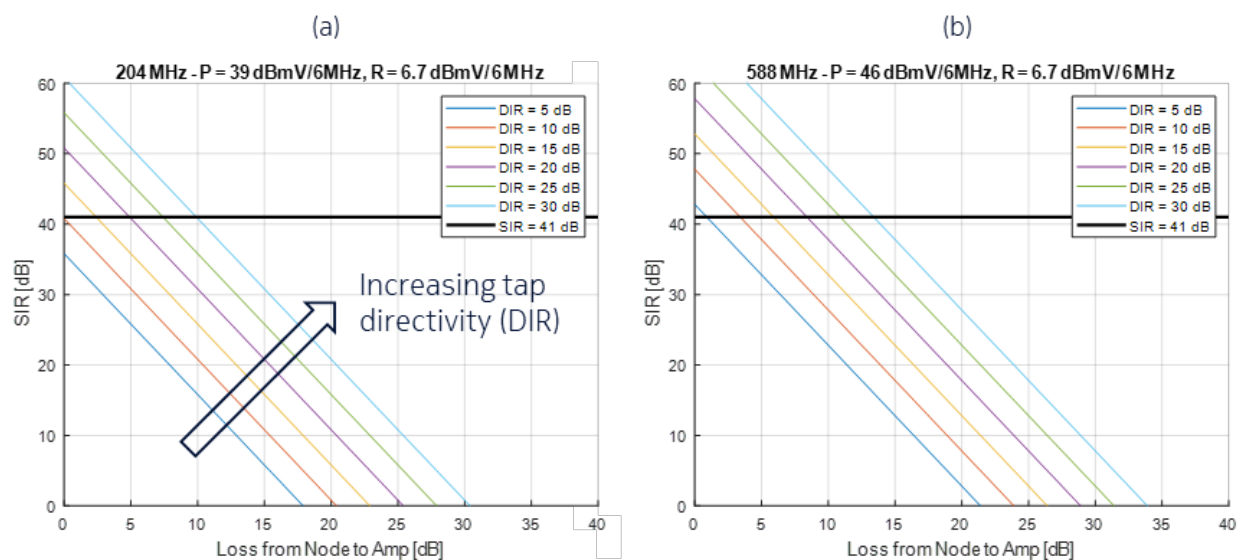
(a)

(b)



**Figure 13: SIR as a function of tap directivity (DIR) and loss from node to tap ($X_1$), at (a) 204 MHz, and (b) 684 MHz.**

Note that these trunk cable lengths are much smaller than typical amplifier spacings in today's networks, and don't yet take into account additional insertion losses of taps, splitters and couplers between the amplifiers. From the above, it is clear that **a significant number of CMs attached to taps before the amplifier will experience very high interference of CMs after the amplifier (SIR much smaller than 41 dB) and will therefore end up in the same IG**. In N+X cable networks using a cascade of FDX amplifiers, this means all CMs in the cascade would be assigned to the same interference group. This would result in half-duplex FDD operation on nearly the entire cascade with the exception of some CMs before the very first amplifier, essentially wasting the FDX capability of the amplifier. So even if active FDX amplifiers would be feasible in terms of EC capabilities, **drop-in replacement of existing amplifiers by active FDX amplifiers is not useful due to the impact of full duplex interference, making their active FDX capabilities redundant**.

To enable deployment of an active FDX amplifier network, **Equation (1)** shows that the SIR can be increased by decreasing the amplifier spacing (decreasing $X_1$) and/or increasing the directivity of the taps (increasing **DIR**). The former decreases the amplifier spacing in the network, and the latter requires a new brand of taps with significantly increased directivity properties. Since directivity is specified with all ports terminated, it would also be crucial to adopt field practices to terminate any unused tap ports and properly deal with corrosion that can degrade isolation [7].

# Conclusion

Full Duplex DOCSIS is a technology that was designed to operate on passive N+0 cable plants, and which combines Full Duplex operation at the node side with dynamic FDD at the modem side to manage the interference arising from Full Duplex operation. We've shown in this paper that the marriage between Full Duplex operation and amplification is not a straightforward one. We've considered three different approaches to using FDX DOCSIS equipment on active cable plants: traditional diplexer-based amplifiers, ALI-cancelling or zero-guard-band amplifiers, and active FDX amplifiers.

Active FDX amplifiers are very challenging to realize due to the stringent requirements on their Echo Cancellation performance because of the spectral overlap between US and DS signals. But even if the desired EC performance can be achieved at an adequate complexity and cost, a "drop-in" replacement of current amplifiers will lead to half-duplex (FDD) operation on nearly all the CMs in the amplifier cascade (as shown in **Section 6.4.1**), due to the full duplex interference occurring across the active FDX amplifier, essentially wasting the FDX capability of the amplifier. The only way to avoid this is to decrease the amplifier spacing (which would further increase the EC-requirements of the individual amplifiers due to their larger number) and to leverage taps that have a higher directivity (difference between the tap loss value and the isolation between the output and the tap) than taps available today. Therefore, the use of active FDX amplifiers in active cable plants is not deemed an attractive solution.

Diplexer-based and ALI-cancelling amplifiers by definition cannot support FDX across the amplifier. A positive consequence of an FDD-based operational model on the network level is the absence of any full duplex interference in the network and no need for FDX processes like sounding and RBAs. Whether or not to adopt an FDD-based operational model is essentially an operator choice that would likely be driven by a desire to ultimately transition from an FDD-based operational model to an active FDX operational model using the same nodes and CMs. If active FDX is not the end game, then purpose-built FDD-based nodes, amplifiers, and CMs may well be the better choice and significant aspects of the functionality of the FDX spec could be ignored for this environment (e.g., RBA switching, sounding).

Diplexer-based amplifiers suffer from a guard band penalty, which makes a part of the spectrum between US and DS unavailable for communication. Therefore, the larger the guard band in diplexer-based amplifiers, the more attractive ALI-cancelling amplifiers become, and vice versa. An ALI-cancelling amplifier that supports a reduced guard-band loss could potentially also delay the need to transition from 1.2 GHz to 1.8 GHz ESD.

In conclusion, the best option for supporting increased US data rates on active cable networks would either leverage diplexer-based amplifiers or ALI-cancelling amplifiers with FDD operation across the network. The EC required for ALI-cancelling amplifiers is technically and operationally much more feasible than the EC required for active FDX amplifiers. This would still leverage the EC knowledge acquired in the FDX DOCSIS research and development effort, but would not require significant parts of the FDX specification that would become useless due to the absence of active FDX operation in the network.

# Appendix A: SNR calculation with a varying echo channel

Let's denote the total Rx signal on a specific subcarrier by

$$y = u + Hx^{tx} + n$$

Where **u** is the desired receive signal coming from the far-end transmitter and attenuated by the channel between that transmitter and the receiver, **x** is the transmit signal from the near-end transmitter echoing back into the Rx, **H** is the echo channel and **n** is the receiver noise. Applying Echo Cancellation to this Rx signal (no matter whether it is analog, digital, or both) can be represented as subtracting a signal $\widehat{H}x$ from the receive signal, yielding an echo-cancelled Rx signal

$$y' = u + (H - \widehat{H})x^{tx} + n$$

If the EC would be perfect, $\widehat{H} = H$, and

$$y' = u + n$$

with a resulting SNR

$$SNR = 10 \log_{10}\left(\frac{|u|^2}{|n|^2}\right)$$

If we now suppose the echo channel has drifted to $(1 + \alpha)H$, without updating the EC coefficients $\widehat{H}$, this would yield a Rx signal

$$y' = u + (1 + \alpha)Hx^{tx} - Hx^{tx} + n = u + \alpha Hx^{tx} + n$$

Where the term $\alpha Hx$ now represents residual (i.e., uncancelled) echo and hence noise. The SNR therefore becomes

$$SNR = 10 \log_{10}\left(\frac{|u|^2}{|\alpha Hx^{tx} + n|^2}\right) = 10 \log_{10}\left(\frac{|u|^2}{|\alpha Hx^{tx}|^2 + |n|^2}\right)$$

The last equality exploited that the Rx noise and the residual echo are independent. If we now define the echo signal at the Rx $x = Hx^{tx}$, we obtain the following formula for the SNR after EC when the EC coefficients have not been updated to the channel drift $\alpha$

$$SNR = 10 \log_{10}\left(\frac{|u|^2}{|\alpha x|^2 + |n|^2}\right)$$

where **u** is the desired receive signal coming from the far-end transmitter attenuated by the channel between that transmitter and the receiver, **x** is the undesired echo coming from the near-end transmitter attenuated by the echo channel, and **n** is the receiver noise.

# Appendix B: CCI calculation with Active FDX Amplifiers

Referring to **Figure 13**, the DS signal level **S** at the CM attached to tap A is given by

$$S = P - X_1 - TL - D$$

The level of the interference **I** arising from the US transmissions of CMs after the tap is given by

$$I = (R + X_1 + IL + X_2) - (X_2 + ISO + D) = R + X_1 + IL - ISO - D$$

Where $R + X_1 + IL + X_2$ is the US transmit power level at the reverse output port of the amplifier, which allows to reach the FDX node at the target Rx PSD level of **R**.

The signal-to-interference ratio then becomes

$$SIR = S - I = P - X_1 - TL - D - (R + X_1 + IL - ISO - D) = P - R - 2X_1 + ISO - TL - IL$$

Defining the directivity of the tap as $DIR = ISO - TL$, this becomes

$$SIR = P - R - 2X_1 + DIR - IL$$

This formula yields the expected signal-to-interference ratio experienced by the CM attached to tap A at a specific frequency (all variables are frequency-dependent).

Note that the expression for the SIR experienced by CMs after the amplifier due to US transmissions of CMs before the amplifier is identical to this formula. This can be calculated as the ratio of the DS signal level over the interference level at the forward input of the amplifier, since both will experience the same gains and losses DS from that point. The DS power at the forward input of the amplifier is

$$S = P - X_1 - IL - X_2,$$

while the interference power at the forward input of the amplifier is

$$I = R + D + TL + X_1 - D - ISO - X_2 = R + X_1 + TL - ISO - X_2$$

Where $R + D + TL + X_1$ is the US transmit power level of the CM, which allows to reach the FDX node at the target Rx PSD level of **R**, yielding a signal to interference ratio

$$SIR = S - I = P - X_1 - IL - X_2 - (R + X_1 + TL - ISO - X_2) = P - R - 2X_1 + ISO - TL - IL$$

which is identical to **Equation (1)**.

# Abbreviations

| ADC | Analog to Digital Converter |
|---|---|
| ALI | Adjacent Leakage Interference |
| Amp | Amplifier |
| CCI | Co-Channel Interference |
| CIN | Composite Intermodulation Noise |
| CM | Cable Modem |
| CNR | Carrier-to-Noise Ratio |
| CSO | Composite Second Order |
| CTB | Composite Triple Beat |
| dB | Decibel |
| DIR | Directivity |
| EC | Echo Cancellation |
| ECD | Echo Cancellation Depth |
| ENOB | Effective Number Of Bits |
| ESD | Extended Spectrum DOCSIS |
| FDD | Frequency Division Duplex |
| FDX | Full Duplex |
| HPF | High Pass Filter |
| IG | Interference Group |
| IL | Insertion Loss |
| ISO | Isolation |
| LPF | Low Pass Filter |
| MER | Modulation Error Ratio |
| MHz | Megahertz |

| | |
|---|---|
| PNM | Proactive Network Maintenance |
| PSD | Power Spectral Density |
| RBA | Resource Block Assignment |
| Rx | Receive |
| SIR | Signal-to-Interference Ratio |
| SNR | Signal-to-Noise Ratio |
| TDD | Time Division Duplex |
| TG | Transmission Group |
| TL | Tap Loss |
| Tx | Transmit |

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