SUB-BAND DIVISION MULTIPLEXING (SDM) INCREASES BANDWIDTH EFFICIENCY AND PROVIDES HIGHER TOLERANCE OF COMPOSITE DISTORTIONS

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ABSTRACT

Bandwidth needs of customers on cable plants have increased dramatically over recent years and will continue to rise in the near future. Increasing the order of QAM modulation has been the most popular way to satisfying the bandwidth needs until recently when nonlinear distortions and limited dynamic range in HFC systems have proved to be an obstacle for reliable 256 QAM service. Hence, it is of both practically necessary and theoretically interesting to investigate approaches other than QAM to increase bandwidth efficiency and to provide higher tolerance of composite distortions all at the same time. Sub-band Division Multiplexing (SDM) is one of these new approaches. In this paper, we give an introduction of the Sub-band Division Multiplexing (SDM) technique based on filter bank scheme and wavelet mathematics. SDM represents a philosophical change comparing to standard QAM in terms of baseband signal formulation and alphabets selection. To show this change, the fundamentals of SDM will be overviewed. Unique characteristics resulting from the SDM fundamentals will also be presented. It will be shown that these characteristics implying multiple advantages of SDM over equivalent QAM on cable applications, especially the tolerance of composite distortions. Simulation results and measurements on the Broadband Physics prototype system from the lab and field trials also will be presented in the paper to verify the theoretical results.

I. INTRODUCTION

SDM stands for Sub-band Division Multiplexing. It is a technique of dividing RF spectrum into multiple and equal-sized sub-bands using filter bank structures. When the basic sub-band filter is designed to have steep roll-off property, the spectrum of each sub-band has little overlap into the neighboring sub-bands. Combined with SDM's inherent frequency and time orthogonal basis, each sub-band is highly independent of other sub-bands, in both frequency and time. For this independency property, SDM can be used as a digital signaling scheme to transmit data stream in parallel over the multiple sub-bands that it creates without having large inter sub-band interference. Used as such, SDM is a multi-band carrier-less modulation. Each sub-band operates in the same manner, but with a different offset frequency.

SDM can be applied to digital communication over many different types of media, such as cable [1], power line and wireless. The first proposal of applying SDM in high-speed data communication was made almost ten years ago by Miller [2], the founder of Broadband Physics Inc. Later, similar ideas of using filter bank techniques for data communication were proposed in other literatures, such as [3]. However, since the 1990s, Broadband Physics Inc. has been the active leader in developing the SDM technology to be implemented in a variety of applications over different channels. Currently, Broadband Physics Inc. is focusing on developing SDM modems for cable downstream applications. It will be argued in the sequel that SDM has numerous benefits in the cable downstream applications. The most prominent two are increased bandwidth efficiency and higher tolerance of composite distortions. In its implemented form, SDM also demonstrates many advantages over popular digital modulation schemes other than QAM, such as Orthogonal Frequency Division Multiplexing (OFDM). These advantages include but not limited to achievable higher bandwidth efficiency with lower system complexity, less sensitive to phase noise, highly resilient to multi-path impairment. All these advantages make the SDM particularly well suited for wireless applications as well. However, this paper discusses all the said characteristics of the SDM but focuses on its high bandwidth efficiency and tolerance of composite distortions for cable applications. The rest of the paper is organized as the following:

Section II gives an overview of the fundamentals of the SDM and discusses some unique characteristics resulting from those fundamentals. This section further discusses the advantages of the SDM closely connected with its characteristics.

Section III discusses the bandwidth efficiency of SDM and through an example, shows that SDM has higher actual bandwidth efficiency than the theoretically equivalent QAM.

Section IV discusses the resiliency of SDM to composite distortions in depth.

Section V presents the results from a Broadband Physics Inc. prototype system test in a simulated cable plant.

Section VI concludes the paper.

II. SDM AND ITS CHARACTERISTICS

Based on a well-designed band-pass filter with high stop-band attenuation, a filter bank can be constructed by frequency shifting this filter prototype and combining them as a polyphase filter. Assuming that the transfer function of a prototype band pass filter is

$$F_0(z) = \sum_{n=-L}^{L} h_0(n) z^{-n} \tag{1}$$

where $z = e^{sT}$ and T is the sampling time, if we over-sample it at M times the original sampling rate, then the over-sampled version of $F_0(z)$ is

$$F(z) = \sum_{k=-ML}^{ML+M-1} h(k) z^{-k} \tag{2}$$

where $z = e^{s\frac{T}{M}}$.

For $k = nM, n = -L,...,0,...,L,$ we have

$$h(k) = h(nM) = h_0(n). \tag{3}$$

Separating $F(z)$ into its polyphase components, we obtain

$$\begin{aligned} F(z) = \\ &\sum_{k=-L}^{L} (h(kM) z^{-kM} + h(kM+1) z^{-(kM+1)} \\ &+ ... \\ &+ h(kM+M-1) z^{-(kM+M-1)}) \\ &= F_0(z) + z^{-1} F_1(z) + ... \\ &+ z^{-(M-1)} F_{M-1}(z), \end{aligned} \tag{4}$$

where $F_0(z)$ is the same as defined in (1) because of (3), and its polyphase versions are

$$F_1(z)=\sum_{k=-L}^{L}h(kM+1)z^{-kM},$$
$$F_2(z)=\sum_{k=-L}^{L}h(kM+2)z^{-kM},\ldots, \tag{5}$$
$$F_{M-1}(z)=\sum_{k=-L}^{L}h(kM+M-1)z^{-kM},$$

for $z=e^{s\frac{T}{M}}$.

Here we can construct a filter bank using the prototype $F_0(z)$ and its frequency-shifted versions of $F_1(z),F_2(z),\ldots,F_{M-1}(z)$ by choosing the prototype so it's frequency shifted versions form an M-band quadrature-mirror filter bank, the impulse responses for each sub-band filter are wavelets orthogonal in both time and frequency. The polyphase construction makes it computationally efficient. With this filter bank, we can transmit data in parallel through all or part of the M branches of the filter bank with each branch being delayed by one sample clock $\frac{T}{M}$ from the previous branch. Due to the clock delay between each branch, we can combine the outputs of each branch to form a single transmitting signal. To express this combined transmitting signal in equation, we assume that branches

$$F_j(z),\ldots,F_{j+l-1}(z),\ 0\le j,..,j+l-1\le M-1,$$

are used to transmit data. The data streams with rate $\frac{1}{T}$ for these branches are defined as

$$\begin{aligned}&A_j : a_{j0}, a_{j1},\ldots\\&A_{j+1} : a_{(j+1)0},\ldots\\&\ldots\\&A_{j+l-1} : a_{(j+l-1)0},\ldots\end{aligned} \tag{6}$$

These data streams can come from different alphabets or constellations. For instance, A_j comes from a four state Amplitude Modulation (4-AM) alphabet, taking values from {-3, -1, 1, 3}, while A_{j+1} comes from 8-AM, taking values from {-7, -5, -3, -1, 1, 3, 5, 7}, etc. Delaying A_{j+d} with respect to A_{j+d-1} by $\frac{T}{M}$ for $0\le d\le l$, and combining with all zero data streams for the unused branches, we have the following combined data stream with rate $\frac{M}{T}$

$$\begin{aligned}&A:0,\ldots,0,a_{j0},a_{(j+1)0},\ldots,a_{(j+l-1)0},0,\ldots,0,\\&0,\ldots0,a_{j1},\ldots,a_{(j+l-1)1},0,\ldots,0,\\&0,\ldots,0,\ldots\end{aligned} \tag{7}$$

Transmitting this combined data stream through the complete filter bank $F(z)$ is equivalent to transmitting $A_j,\ldots,A_{j+l-1}$ (expanded to rate $\frac{M}{T}$ through expanders) separately through the corresponding branches, delaying by one clock from each other and combining them at the output as shown by the equation below

$$F(z)[A]=\sum_{k=j}^{j+l-1}z^{-j}F_k(z)[A_{ke}]$$
(8)

where the square bracket denotes the filtering operation and $A_{ke}, j\le k\le j+l-1$ are the expanded $A_j,\ldots,A_{j+l-1}$.

We can also illustrate this process in spectrum plots.

Figure 1-1 shows an example spectrum of the output of a single branch.

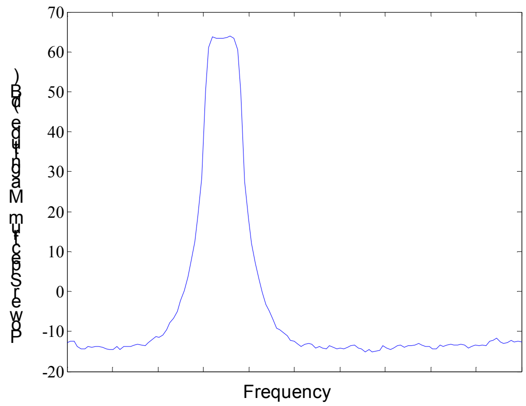

Figure 1-1: Single sub-band spectrum

Figure 1-2 shows an example spectrum of the combined outputs from all active branches.

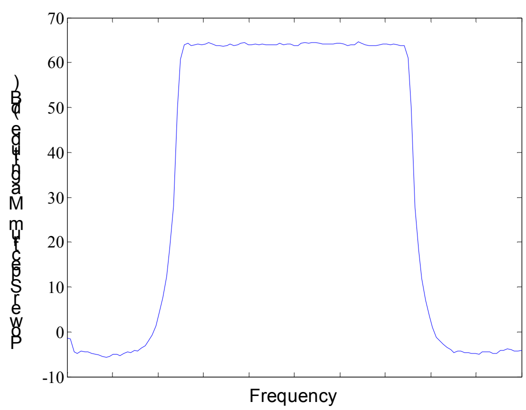

Figure 1-2: Combined multi sub-band spectrum

In view of the spectrum plots in Figure 1-1 and 1-2, we see why the name "Sub-band Division Multiplexing", each branch creates a sub-band of the entire spectrum; the overall spectrum is the combination of all the sub-bands. The performance shown in Figure 1-2 has been achieved in digital hardware by Broadband Physics, Inc.

With such transmitter structure, we can easily see that the receiver structure is essentially the same as the transmitter with the same filter bank structure and correspondingly matched filters filtering the received signal into separate sub-bands. The orthogonality of the filters can be compromised if, for example, there is a group delay variation across the subbands. The resulting self-interferences are relatively easy to remove if the prototype filter has very steep roll-offs Another useful view of the above transmitter and receiver structure is the wavelet transform. The transmitting filter bank is a form of inverse wavelet transform.. Each data symbol is amplitude modulating a wavelet, the data stream to be transmitted is a signal vector comprised of modulated wavelets or equivalently, a vector expressed by a set of base functions in the wavelet transform domain. Hence, the receiving filter bank just needs to be a wavelet transform [2], [4].

Comparing to other typical digital modulation schemes, such as QAM, SDM has several unique characteristics. All the characteristics of the SDM are the results of the basic construction of the SDM as shown above. Furthermore, these characteristics are the reasons behind the advantages of SDM comparing to other typical digital modulation schemes. Here we discuss some of them. For more detailed comparisons between SDM and other modulation schemes, e.g. QAM and OFDM, please see [5].

First, depending on the design of the prototype filter, 50dB plus stop-band attenuation is achievable by the SDM spectrum. This characteristic of the SDM makes an extra pulse-shaping filter unnecessary at the transmitter. With typical additional 10% bandwidth required for a pulse-shaping filter, (e.g. raised cosine filter) for QAM, SDM with the same theoretical bandwidth efficiency has more effective bandwidth efficiency.

Second, each data symbol is confined to its respective sub-band by the filter bank as seen in Figure 1; consequently, the abrupt transitions of the data symbols do not cause ringing in the channel unlike the multi-carriers in OFDM. Since no ringing effect exists, cyclic prefix is not needed in SDM. In OFDM the length of the cyclic prefix can sometimes exceed 30% of the symbol duration. Thus SDM can achieve significantly higher effective bandwidth efficiency comparing to OFDM. Like OFDM, the bandwidth of each SDM sub-band and the number of sub-bands are design parameters and can be optimized accordingly for the channel. Furthermore, by selectively choosing active and inactive sub-bands, we can tailor the entire transmitting spectrum to the actual operating channel. The benefits of such flexibility are numerous, including easy fitting under emission masks and easy mitigating narrow band interference [4]. Because of the very high stopband attenuation of SDM subbands there is an advantage compared to OFDM, which typically has only 13 dB attenuation between adjacent bands.

Third, SDM can transmit a real signal, i.e. signal with only I component but no Q component without being limited to lower bandwidth efficiency. This can be achieved by choosing the data stream for each sub-band from some amplitude modulation (AM) constellations (same constellation is not required for different sub-band). Since the transmitted signal is a real signal, the decision region of the receiving slicer is single dimension, it can stand up to higher phase noise than signals with both I and Q components carrying the information.
Fourth, SDM is orthogonal in time allowing the overlap in the transmission of one symbol with previously transmitted symbols [5]. The amount of symbol to symbol overlap is a design choice.

Fifth, recalling that the SDM transmitter and receiver pair is essentially a wavelet transform pair. Any equalization to be done in the receiver while following the receiving filter bank is not in time domain. Thus the equalizer structure can be made significantly less complicated than the time domain adaptive equalizer typically seen in QAM systems.

The above five properties and advantages of SDM are not the only benefits of using SDM, rather they are the five most essential and also most intuitive to describe at this time.

A note of notation is helpful here and for the rest of the paper, depending on the number of bits a symbol in each subband represents, the corresponding SDM scheme is called L-SDM for L bits per symbol in a single subband, or equivalently, for an alphabet of size $S = 2^L$ for the subband.

In the next two sections we will discuss advantages of SDM in cable applications, especially its capability of providing actual higher bandwidth efficiency and mitigating composite distortions.

III. SDM PROVIDES INCREASED ACTUAL BANDWIDTH EFFICENCY

Since one of the current major driving forces of digital cable technology development is the need of bandwidth, any modulation scheme that can provide higher than current standard bandwidth efficiency (measured in bits/second per Hz or bps/Hz) will increase the raw digital capacity of the spectrum without increasing the actual bandwidth. SDM happens to be such a modulation scheme.
To study the bandwidth efficiency of the SDM, we can start with an example for cable applications. Each current cable

channel is 6 MHz wide. To formulate a baseband transmitting signal on one channel using SDM, we can first choose the digital sampling frequency to be $2f$ ($f > 6$ MHz to satisfy the Nyquist theorem). Then, we divide the spectrum of 0-f Hz into M subbands, each subband has bandwidth of f/M and the symbol rate for each subband is $2f/M$. For convenience, f can be chosen such that f/M is an integer. We can turn on any block of continuous $6*10^6*M/f$ subband to create a 6 MHz cable channel. Two of the active subbands at both edge of the channel can be turned off to avoid interfere with the neighboring channels.

Again, for the convenience of presentation, we assume that each active subband has the same alphabet that has S states. So the bandwidth efficiency of each subband is $log2(S)*2f/M/(f/M) = 2*log2(S)$ bps/Hz, therefore, the bandwidth efficiency of the combined channel is also $2*log2(S)$ bps/Hz if we just ignore the two inactive guard subbands at the channel edges for now.

With the general derivations above, we can look at some actual numbers for the example. Assume that the chosen digital sampling frequency is 51.2 MHz, the entire useable bandwidth for SDM is $51.2/2 = 25.6$ MHz. By using SDM technique, we can divide 25.6 MHz into 256 subbands; each subband is 100 kHz wide with a symbol rate of 200 kHz. To create a 6 MHz wide cable channel, we can choose any 60 continuous subbands out of the entire 256. Please note that the entire 256 subbands are for signal formulation purpose only, no actual energy being put on them except the 60 chosen subbands, the real occupied spectrum is still only 6 MHz wide, the other unoccupied spectrum is free for other uses. Suppose we choose 60 subbands from 9 MHz to 15 MHz, two edge subbands being turned off as guard bands, so the actual number of subbands for data transmission is $60 - 2 = 58$. For the convenience of presentation here, assume that each subband uses the same 16-AM alphabet with 16 states on a single real axis. Each subband has bandwidth efficiency of $log2(16)*2*100,000/100,000 = log2(16)*2 = 8$ bps/Hz, the same as 256 QAM. The actual overall combined channel data rate (accounting the two inactive guard subbands) is $58*8*100,000 = 46.4$ Mbps, the actual overall channel bandwidth efficiency is $58*8*100,000/(60*100,000) = 58/60*8 = 7.7$ bps/Hz, which is less than 4% lower than the theoretical bandwidth efficiency of 256 QAM. However, considering the excess bandwidth of QAM from the pulse shaping filters such as root raised cosine filter, the actual 256 QAM bandwidth efficiency can be 10% less than the theoretical value of 8 bps/Hz. The other possibility is that each subband uses the same 32-AM alphabet with 32 states on a single real axis. With this set up, the actual channel data rate is $58*log2(32)*2*100,000 = 58$ Mbps, the actual bandwidth efficiency is $58*2*log2(32)/60 = 9.7$ bps/Hz, which is about 3% less than the theoretical bandwidth efficiency of 1024 QAM, and could be more than the actual bandwidth efficiency of 1024 QAM if the 10% additional bandwidth required for pulse-shaping filter is considered.

Our observation from the example is that based on the first characteristic of the SDM, no pulse-shaping filter is needed at the transmitter; hence SDM can achieve higher actual bandwidth efficiency than equivalent QAM.

In view of the example derivation above, we have

*SDM actual channel bandwidth efficiency = Number of active subbands * subband bandwidth * bandwidth efficiency of each subband / Channel bandwidth = Number of active subbands * bandwidth efficiency of each subband / (Number of active subbands + Number of guard-bands)* (9)

Using (9) and noting that the number of the guard-bands is always 2 regardless of the channel bandwidth due to the extremely high stop-band attenuation of the SDM subbands, we can see that the actual bandwidth efficiency will get better when the number of active bands increases with any increment of the digital cable channel bandwidth (e.g. 12 MHz or 18 MHz).

Thermal noise performance for SDM can be calculated and simulated by Additive White Gaussian Noise (AWGN) model. Since the theoretical AWGN performance for a SDM subband with *S*-AM alphabet (*S = 2, 4, 8, 16, 32, 64, …*) is equivalent to a *S^2* QAM, so 1SDM, …, 5SDM and 6SDM with corresponding 2-AM, …, 32-AM and 64-AM alphabets have the same AWGN performance as QPSK, 16 QAM, …, 1024 QAM and 4096 QAM, respectively [5].

Bit-true simulation results and lab-measured data on the baseband prototype system in Figure 2 show that both BERs are close to the theoretical values.

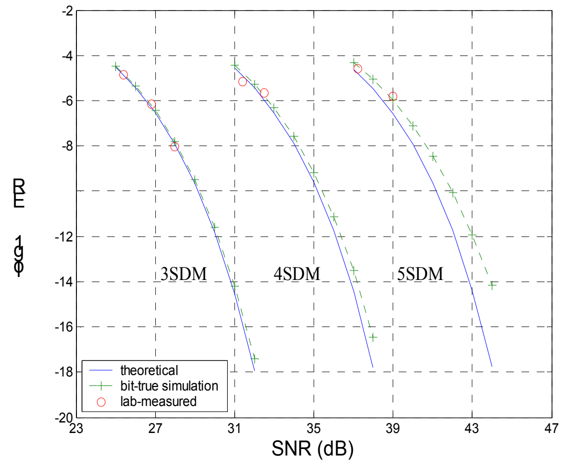

Figure 2: AWGN BER performance of SDM

Noting that the choice of alphabet for each subband is independent of other subband, we can in fact choose the alphabet for each subband differently. The freedom of doing so allows us to set the bandwidth efficiency/modulation density for each subband according to the channel condition, thus finely tune and optimize the trade off between error performance, channel signal to noise ratio and over all bandwidth efficiency. As we will see in the next section, the ability of changing bandwidth efficiency on a fine frequency scale instead of on a whole channel scale will help SDM providing higher tolerance to composite distortions.

SDM channel also has higher capacity/cost ration than combination of logical QAM channel. One of the major reasons is that SDM receiver has less complexity than comparable single QAM receiver for the fifth characteristic of SDM. In the final version of the paper, we will give more detailed comparisons between SDM and logical combination of QAM channels.

IV. SDM MITIGATES COMPOSITE DISTORTIONS

Composite distortions are produced by amplifier nonlinearity caused intermodulation of analog TV carriers. The dominant components of the distortions are Composite Triple Beats (CTB) and Composite Second Order (CSO) [7] [8]. These distortion components typically have average power levels 12~15 dB below the thermal noise level. Even though their low average power levels appear to be harmless, due to their statistical properties, the random peak envelope power can be significantly higher to cause large performance degradation for 256 or higher order QAM [8].

Other than asking operators to control and improve CSO/CTB levels through carefully choosing channel frequency offsets and maintaining head-end transmitter aggregate noise power at low levels, etc., the main methods to mitigate composite distortion at the baseband digital modulation level that have been proposed include increasing interleaver depth and improving adaptive equalizers [8]. Just by looking at the performance data, these two methods appear to be adequate. However, if we look closely, there are problems associate with each of them.

For longer interleaver depth, first we know that the prices of increased interleaving depth include increased latency, which affects the quality of service in another way. Second some longer interleaving depth required to handle CSO/CTB transients are not even supported by lots of set-tops.

For improved adaptive equalizers, according to [8], it is possible to have adaptive equalizers to converge to a state that forms a sharp notch at the interferer frequency. However, to achieve a sharper notch through adaptive equalization, higher number of equalizer taps is required. Naturally, higher number of equalizer taps requires more demodulator complexity and more system throughput delay. In the case of non-blind equalization, which uses a training sequence to obtain the optimal equalizer taps, higher number of equalizer taps also requires longer training sequence. The longer training sequence again causes more throughput delay and overhead.

Adopting the SDM approach for baseband modulation will achieve actual bandwidth efficiency higher than 256 QAM and obtain inherent capability of mitigating composite distortion effect. In the event of excessive composite distortions, as we will discuss in the following, SDM does not need to fall back to a lower bandwidth efficiency mode completely unlike 256 QAM has to fall back to 64 QAM for the entire channel. Instead, SDM can fall back to a lower modulation density only at the subbands being affect by the composite distortions most severely, thus allowing reliable service without significantly lowering the bandwidth utilization. In addition, SDM is not contradicting with those proposed improvement done by the operators or proposed longer interleaving depth. When these improvements are available, SDM can work with these methods to provide an even higher tolerance to the composite distortions. When the limits of the HFC plants or other restrictions render these methods unusable, SDM alone still can provide a more reliable service.

To fully understand the effect of composite distortions on SDM, we first look at some properties of CSO/CTB.

Since CSO/CTB are produced by intermodulation of analog carriers, they are of narrow frequency nature. The typical power bandwidth of an individual beat is about 10- 20 kHz. Remember that the bandwidth of each subband of a SDM channel is a design choice, we can choose it to be convenient for overall system requirements yet wider than a typical composite beat component. In the example of Section III, we used 100 kHz as the subband width. It is indeed wider than the typical bandwidth of a composite beat component.

Another important property of the CSO/CTB distortions is that the locations of all the beat components can be calculated [1], [9]. Using the information of the beat components locations, we can check the composite distortion locations within a 6 MHz cable channel. Calculation shows that with 100 kHz wide subbands, only 14% of sub-bands will experience direct beats products. Due to the narrow bandwidth nature of the composite distortions and high independency of SDM subbands, only the two immediate neighboring subbands will be affected, thus the maximum number of potential neighboring subbands to be affected will be no more than 24% of the overall subbands, leaving a minimum 62% of sub-bands clear of any CSO or CTB beat impact [5]. So even without protection of the error correction codes and/or interleaver, in the event of excessive composite distortions, we can lower the bandwidth efficiency on those 38% affected subbands without changing the 62% unaffected subbands. As an example, we can lower the bandwidth efficiency of the 38% affected subbands from 256 QAM equivalent 8 bps/Hz to 64 QAM equivalent 6 bps/Hz to maintain the reliability, while leaving the rest 62% subbands still at 8 bps/Hz. The end results is that the overall channel capacity is about 6/8*38% + 8/8*62% = 90% of the normal channel capacity, but still gain better reliability.

To look further at the error correction and interleaving protection, we note that the duration of a random composite distortions pile up burst is often inversely proportional to the distortion power bandwidth [8]. Given a 10 kHz power bandwidth, a distortion burst can be 100 μs long. Since these bursts are highly localized in subbands, they only affect the symbols in one subband. Recall from Section II, the interval between two consecutive symbols in the same subband is 5 μs for 100kHz wide subbands. So a distortion burst of 100 μs only covers about 20 symbols in the same subbands. The fact that a distortion burst only affects a low number of symbols implies that an appropriately chosen Reed-Solomon (RS) type error correction coding can correct most errors caused by composite distortions bursts. In the final version of the paper, we will have more detail about the effect of RS code on SDM and we will also show that a simple standard length interleaving will have the similar results as well.

V. TEST RESULTS OF A SDM PROTOTYPE SYSTEM

A Broadband Physics, Inc. SDM prototype system has been tested over RF channel with some simulated composite distortions. The system modulation density was set at 3SDM, which according to the previous sections provides a theoretical 64 QAM equivalent bandwidth efficiency of 6 bps/Hz. The actual system bandwidth efficiency is 5.8 bps/Hz (see Section III). The 6 MHz channel occupies RF spectrum of 582~588 MHz and is centered at 585 MHz. An approximate “spread” interference tone was generated with an occupied bandwidth of 20 kHz. This interference tone was modulated at different frequencies within the channel. The overall uncoded channel BERs were measured under different levels of interference tone and Additive White Gaussian Noise (AWGN) at constant level of –37dBc. Parts of the results are shown here in the chart below.

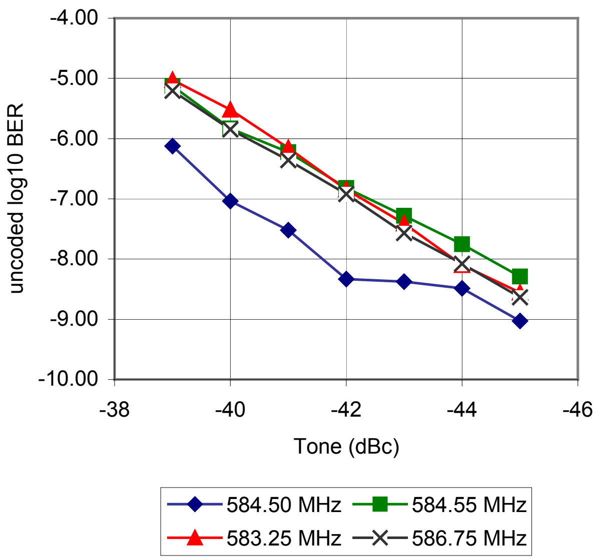

Figure 3: BER Performance of 3SDM over RF Channel with AWGN and Narrowband Interference

The most interesting feature of the chart is that for interferences with the same power level, their effects are also dependent on their locations within the channel. This distinctive feature, which is not available with single band modulations, is a direct result of the multi-band approach and the subband independency of SDM.

Recalling that the 6 MHz RF channel is divided by 60 SDM subbands, each subband occupies 100 kHz bandwidth, we find that three of the four interferences, namely the tones at 583.25, 584.55 and 586.75 MHz, are in the middle of a subband and the other one (at 584.5 MHz) is on the boundary of two neighboring subbands. The chart shows that the interference on the boundary of two neighboring subbands causes less BER degradation than interferences falling in the middle of a subband. Because of the SDM subband independency, the increased bit errors due to the narrow band interferences are from one and two subbands, respectively, for the interferences falling in the middle of a subband and on the boundary of two neighboring subbands. Let us assume that the average number of increased errors caused by an interference with a certain power level falling in the middle of a subband is *Ed*. Another interference with the same power level but falling on the boundary of two neighboring subbands will actually have half the interfering power in each affected subband, or equivalently, 3 dB higher signal to interference ratio. In view of this, the average number of increased errors in both affected subbands will be $Ed1 << Ed/2$, and the overall number of increased errors will be $2*Ed1 << 2*Ed/2 = Ed$. Hence, the overall BER degradations caused by narrow interferences falling on the boundary of two neighboring subbands are less than degradations caused by those falling in the middle of a subband.

The simple analysis above is not valid for the single band modulation or multi-band modulation without subband independency. In those cases, error degradations are not determined by the interference power within an individual subband or a spectrum subsection, so the type of results in Figure 3 can only come from multi-band modulation schemes, with highly independent subbands, such as SDM.

In the near future, we will present more test results to further verify the CSO/CTB mitigating capability of SDM.

VI. CONCLUSION

In this paper, we have introduced the basic concept of Sub-band Division Multiplexing (SDM). Several essential characteristics of the SDM have been presented. As a digital modulation scheme, advantages of the SDM associated with its characteristics were also discussed. These advantages include:

1. Effective bandwidth efficiency is improved, as no pulse-shaping filter is needed.
2. Transmitting can be tailored easily to fit under transmission masks since sub-bands are highly independent, nearly orthogonal, and can be turned active or inactive easily.
3. Transmitted signal can be made real to have higher phase noise resiliency.
4. Less receiver complexity can be achieved, as no time domain equalizer is needed.

In the latter part of the paper, we focused on the application of SDM in cable channels. We discussed that due to its characteristics, SDM can provide higher actual bandwidth efficiency than QAM with the same theoretical bandwidth efficiency. We further discussed that also due to various advantages of SDM, especially its subband independency, SDM can effectively mitigating composite distortions caused by CSO/CTB, thus relax the constraints on the RF channels and lessen the burdens on operators.

In Section V, we also presented the test results for a prototype SDM modem to verify the theoretical results.

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