Transmit Diversity Code Filter Sets (TDCFS), a MISO Antenna Frequency Pre-Distortion Scheme for ATSC 3.0

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Abstract—**Transmit Diversity Code Filter Sets (TDCFS) are a method of pre-distorting the common waveforms from multiple transmitters in the same frequency channel, as in a Single Frequency Network, in order to minimize the possibility of cross-interference among the transmitted signals over the entire reception area. This processing is achieved using all-pass linear filters, allowing the resulting combination of pre-distortion and multipath to be properly compensated as part of the equalization process in the receiver. The filter design utilizes an iterative computational approach, which minimizes cross-correlation peak side lobe under the constraints of number of transmitters and delay spread, allowing customization for specific network configurations. This paper provides an overview of the TDCFS Multiple-Input Single Output (MISO) antenna scheme adopted in ATSC 3.0, together with experimental analysis of capacity and specific worst-case conditions that illustrate the benefits of using the TDCFS approach.**

Index Terms— **ATSC 3.0, Cyclic delay diversity, DTT, MISO, Single Frequency Network**

I. INTRODUCTION

A TSC 3.0 [1] aims to reliably deliver live broadcast television and accompanying services in the entire coverage area for a range of device types. When television is broadcast from a single tower, as it is commonly done with the first-generation digital terrestrial television standard ATSC, there are many difficult reception environments in any metropolitan area. Whether it is indoor or mobile reception, one common characteristic is often lost Line-of-Sight (LoS), and therefore several transmitters forming a dense Single Frequency Networks (SFN) are desirable. However, in a dense SFN the amplitude of the signals can be similar in magnitude, resulting in severe multipath or possibly weak signal conditions if these signals create destructive interference [2], [3]. ATSC 3.0 has adopted a Multiple-Input Single Output (MISO) antenna scheme to improve the overall performance in SFN, known as Transmit

Diversity Code Filter Sets (TDCFS), in order to minimize the possibility of cross-interference among the transmitted signals over the entire reception area [4], [5]. TDCFS is similar to cyclic delay diversity [6] schemes in that it introduces a frequency pre-distortion of the common waveforms from the different transmitters of an SFN in such a way that special signal processing at the receivers is not necessary, since the frequency pre-distortion is seen by the receivers as part of the channel. TDCFS is also similar to the MISO scheme adopted in DVB-NGH (Digital Video Broadcasting – Next Generation Handheld) [7], known as eSFN (enhanced SFN) [8], [9], but it has better correlation properties. Compared to the MISO scheme adopted in DVB-T2 (Digital Video Broadcasting - Terrestrial 2nd Generation) [10] which is based on Alamouti coding [11], there is no need to double the pilot overhead, and it can be systematically extended to more than two transmitters.

This paper provides an overview of the TDCFS MISO antenna scheme adopted in ATSC 3.0, together with experimental analysis of capacity and specific worst-case conditions that illustrate the benefits of the TDCFS approach.

The paper is structured as follows. Section II introduces the concept of a Maximum Diversity Network to exploit the diversity in SFN. Section III describes the TDCFS technique, and Section IV presents some illustrative performance simulation results. Section V discusses the benefits that the Maximum Diversity Networks offer for the migration to the new ATSC 3.0 standard in the United States. Finally, the paper is concluded with Section VI.

II. MAXIMUM DIVERSITY NETWORK

The proposed Maximum Diversity Network exploits three main sources of diversity: multipath diversity, receiver diversity (small-scale diversity), and transmitter diversity (large-scale macro-diversity).

For multipath diversity gain, the equalizer should not simply cancel multipath signals, but should optimally utilize the additional signal strength. To accomplish this, the equalizer may even use multiple stages, as in a turbo or iterative equalizer.

Receiver diversity requires more than one front-end and an optimal combiner. It benefits reception in indoor environments and so-called urban canyons the most. For handheld devices, however, very small conformal antennas limit the diversity gain, particularly any spatial diversity gain.
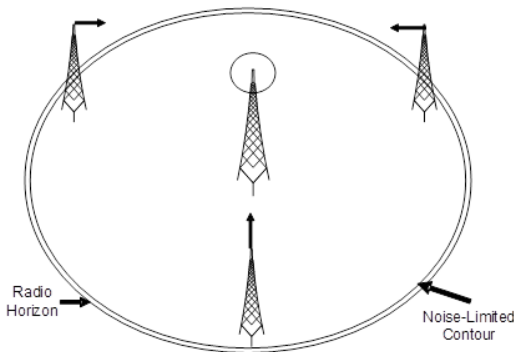
Fig. 1. Idealized single frequency network with significant overlap and large scale transmitter diversity.
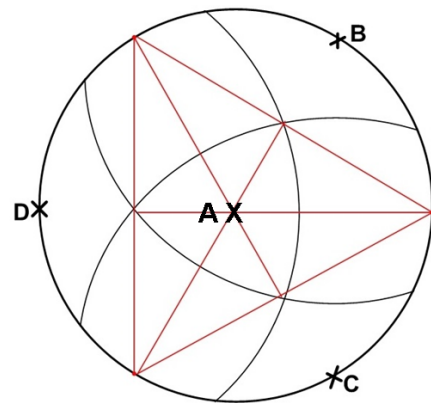

Fig. 2. Overlapping coverage areas for idealized SFN layout, including critical contours in red where signal strength and distance from two of the transmitters may be equal but out-of-phase.

For large scale transmitter diversity, wide space transmitters form a large cell the size of the noise-limited contour as shown in Fig. 1, with a standard main tower transmitter with high power and height at the center. Three or more additional transmitters within the cell provide maximum overlapping coverage. All locations within the cell are covered by at least two transmitters and most are covered by three, as indicated by the contour lines in the idealized network of Fig. 2. The overlapping coverage attempts to produce uniform field strength over the entire metro area which would create a large improvement in white space protection ratios. Also, cellphone base station protection ratios would be improved. The most difficult environments would experience the largest diversity gains.

The secondary transmitters should be configured so that they have a well-bounded area of coverage. They would have a high gain of -5 to 0 dB of the main transmitter, as well as a height in the range of 50-100% of the main transmitter height. By using a 140-degree directional pattern and down tilt, the coverage boundary can be sufficiently controlled. Polarization should be 50%-50% horizontal-vertical since this provides balanced coverage for handheld receivers which are expected to be used in all possible orientations. Furthermore, cooperation among broadcasters to share both primary and secondary transmitter locations will benefit from performance, logistical, and economical standpoints. Shared sites, towers, and antennas will be cost effective and provide the best adjacent channel ratios.

In many locations the signals from two transmitters would be close in both amplitude and delay, as is a common issue in an SFN. In the most severe case, if the transmitted signals are the same but opposite in polarity the two signals could significantly cancel. In the basic SFN, these locations are shown in Fig. 2 by the six red lines, one for each pair selected from the four transmitters. While more complicated networks with different terrain will not be as well-defined as the example, such zones still exist and need to be considered.

In order to prevent the possibility of cancellation, the transmitter signals should be encoded using a transmitter diversity code. One problem of using the Alamouti code is that the original implementation is limited to only two transmitters, and proposals to support more than two transmitters have

compromised performance [9]. Moreover, receivers need to estimate the channel from the two transmitters, which implies an increase of the pilot overhead. TDCFS can be optimized for two transmitters or more, and with this uniquely designed code, the decoder complexity does not increase as the number of transmitters is increased.

III. TRANSMIT DIVERSITY CODE FILTER SETS

The Transmit Diversity Code Filter Set is a way of coding the waveforms from multiple transmitters that are transmitting the same information in the same frequency channel. This coding is done using linear filters so that the decoding in the receiver can be implemented as part of the traditional equalizer process. Furthermore, the coding filters are designed in such a way as to provide robust performance at the receiver over a wide range of expected multipath behavior in the transmission environment.

A. Signal Model

A transmit diversity code filter set is a set of unique transmit filters used to filter a common modulated signal. These filtered signals are transmitted from separate transmitters in an SFN. The design of the transmit diversity code filter set is based on creating all-pass filters using a minimized metric, called the Peak Sidelobe (PSL), over all filter pairs within the constraints of the number of transmitters M and the length of the filters L. Use of such filter code sets create improved overall signal condition at the receiver taking into account the likelihoods of multipath transmission conditions.

Fig. 3 shows a multi-transmitter OFDM system with FFT block size of N_B using transmitter diversity across M transmitters and applying the designed filter set. The common modulated signal $S[k]$ is individually filtered through M strategically optimized filter vectors $\{C_1[k], C_2[k], \ldots, C_M[k]\}$ to created uniquely pre-distorted signals $\{F_1[k], F_2[k], \ldots, F_M[k]\}$. The code filters are implemented as a multiplication of fixed coefficients across all carrier frequencies, indexed within each block by k. Each pre-distorted signal block is transformed through IFFT (Inverse Fast Fourier Transform) and cyclically extended with a guard
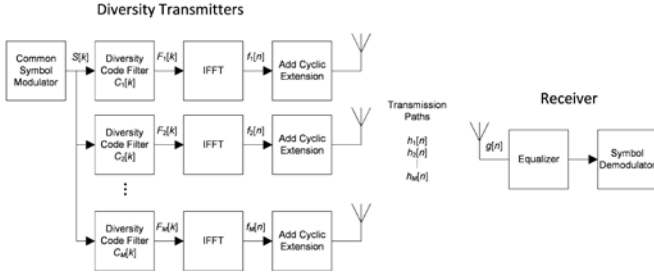
Fig. 3. The TDCFS system that will be deployed for ATSC 3.0 is an OFDM system with code filters applied in the frequency domain and a cyclic extension applied in the time domain.

interval in an OFDM manner before being transmitted through their respective transmit antenna. These transmitted signals will then be modified through different channel conditions represented in the time domain by the channel impulse responses $\{h_1[n], h_2[n], \ldots, h_M[n]\}$. The superposition of these signals combined with AWGN $w[n]$ are received by an antenna at the receiver as function $g[n]$.

If the FFT block size of the OFDM system is N_B and, assuming proper synchronization of symbol blocks, we can describe the system as:

$$G[k] = \left(\sum_{i=1}^{M} H_i[k]C_i[k]\right)S[k] + W[k] \qquad (1)$$

where

$$G[k] = \text{FFT}(g[n], N_B)$$
$$H_i[k] = \text{FFT}(h_i[n], N_B) \text{ with } i \in \{1..M\}$$
$$W[k] = \text{FFT}(w[n], N_B)$$

Alternatively, we can describe the system in the time domain as:

$$g[n] = \left(\sum_{i=1}^{M} h_i[n] * c_i[n]\right) * s[n] + w[n] \qquad (2)$$

where

$$c_i[n] = \text{IFFT}(C_i[k], N_B) \text{ with } i \in \{1..M\}$$
$$s[n] = \text{IFFT}(S[k], N_B)$$

and $*$ is the circular convolution operator over length N_B.

B. Filter Set Design Process

The time domain sequence sets $\{c_1[n], c_2[n], \ldots, c_M[n]\}$ described in section III-A have an associated peak sidelobe correlation metric:

$$\text{PSL} = \max_{i \neq j}\left(c_i[n] * c_j[n]\right) \text{ over all } i, j \in \{1..M\} \qquad (3)$$

where the convolution operator is circular with respect to N_B.

The main objective in designing a good system is to minimize this PSL metric. This can be done through an iterative process similar to techniques that have been used to create unimodular sequences for radar sensing [14], but with an entirely different set of design constraints that are more relevant for the SFN application.

TABLE I
TDCFS ALGORITHM DEPICTING THE BASIC ITERATIVE REFINEMENT OF FILTERS WHILE KEEPING THE NECESSARY CONSTRAINTS IN PLACE

Step 1:
Initialize Code Filters: $\{c_1[n], c_2[n], \ldots, c_M[n]\}_{curr}$
Step 2:
Normalize Code Filters to have unity magnitude in the frequency domain:

$$\{c_1[n], c_2[n], \ldots, c_M[n]\}_{curr,norm}$$

Step 3:
Calculate Cross Correlation Peak Sidelobe PSL_{curr} over all filter pairs. If this is the first iteration or $\text{PSL}_{curr} < \text{PSL}_{opt}$, skip to Step 5; otherwise, continue to Step 4.
Step 4:
Adjust optimum filters in predetermined manner:

$$\{c_1[n], c_2[n], \ldots, c_M[n]\}_{curr} = \{c_1[n], c_2[n], \ldots, c_M[n]\}_{opt,adj}$$

and return to Step 2.
Step 5:
Set

$$\text{PSL}_{opt} = \text{PSL}_{curr}$$
$$\{c_1[n], c_2[n], \ldots, c_M[n]\}_{opt} = \{c_1[n], c_2[n], \ldots, c_M[n]\}_{curr,norm}$$

If stopping criteria is not satisfied, return to Step 5; otherwise continue to Step 6.
Step 6:
Set Transmit Diversity Code Filters:

$$\{c_1[n], c_2[n], \ldots, c_M[n]\}_{opt}$$

Table I shows the algorithm for the filter set design process, with particular parameters defined by the number of transmitters M and the filter length L. In order to provide a good starting point, the initial filter set can be a known set of sequences with desirable properties (e.g., Gold, Kasami) even if they do not meet the eventually imposed constraints. The adjustment in Step 4 can be accomplished in several ways, such as random perturbation or methods derived in [14] and related works, and switching between multiple methods during the iterative process may avoid local minima that a single method may have.

An extension of this process is to use a different cost function than the simple PSL when using the flowchart of Table I, such as using a weighted PSL:

$$\text{PSL}_{weighted} = \max_{i \neq j}\left[\left(c_i[n] * c_j[n]\right)m[n]\right] \text{ over all } i, j \in \{1..M\} \qquad (4)$$

with a pre-defined weight profile $m[n]$ that is based on the likelihood of such a corresponding lag at the receiver. For example, $m[n]$ might be defined as a simple triangular function over a set span, and this weighting would emphasize decorrelation of short lags over long lags in a linear fashion. Experimentation with different weight profiles for specific network topologies is a potential subject for further study.
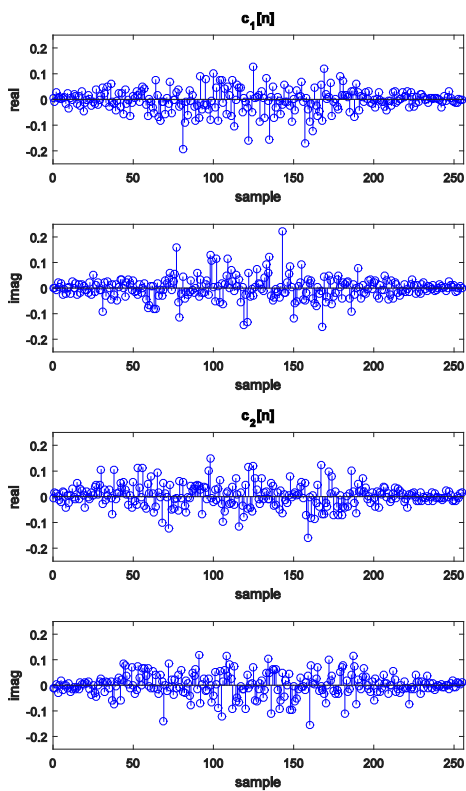
Fig. 4. The real and imaginary components of the filter impulse responses for a two transmitter system with filter lengths of $L = 256$ indicates the channel spreading that will occur to increase diversity.
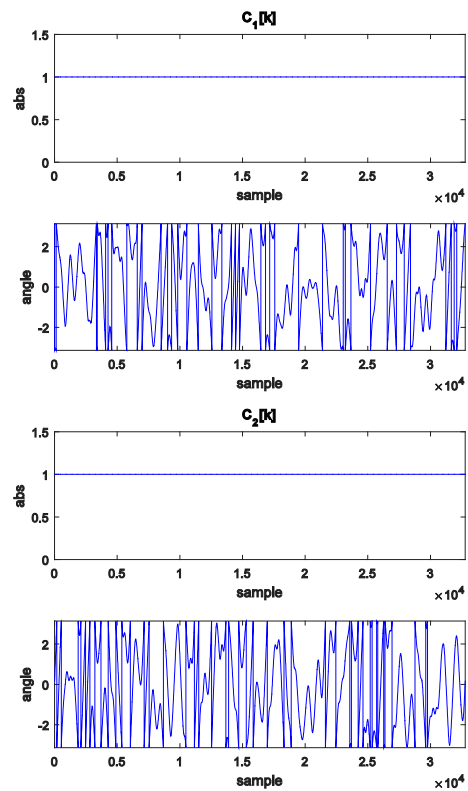

Fig. 5. The amplitudes and phases of the filter frequency responses in the two transmitter system show that there is no amplitude deviation to the original signals and that the perturbation can be simply implemented as a carrier dependent phase rotation.
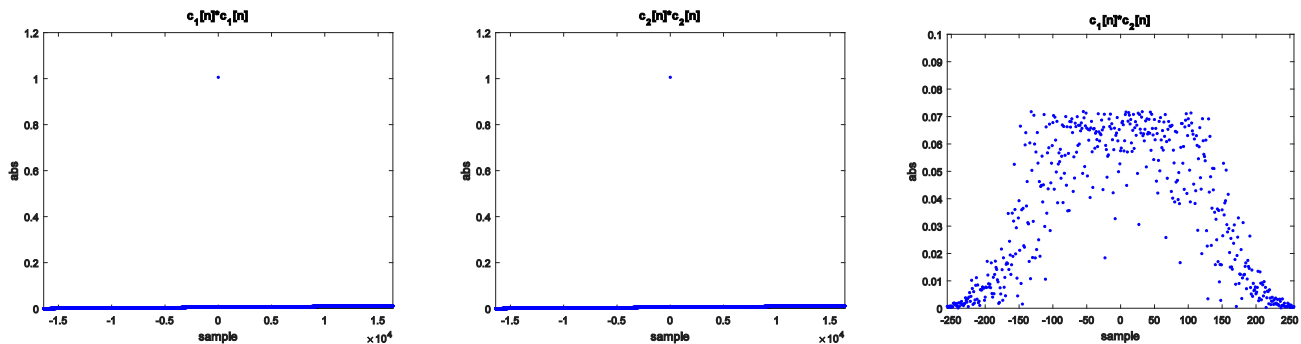

Fig. 6. The autocorrelations of the filters in the two transmitter system are a delta function (left, middle), a characteristic resulting from their flat amplitude spectrum, and the iterative process has spread the cross-correlation impulse response which also has unity power, thereby reducing the peak value (right).

Fig. 4 shows a sample set of filters in a finite impulse response representation that have been designed using the process depicted in Table I with the parameters of $M = 2$ transmitters, $L = 256$ samples, and with a continuous filter application.

Fig. 5 shows the same set of filters, but as a set of frequency responses. Note that because of the frequency response normalization step, these have nearly constant magnitude of 1, though with varied phase. This ensures that the code filters do not affect the spectrum of the transmitted signal.

A directly related property is clear in the autocorrelation plots of Fig. 6, which shows the nearly perfect autocorrelations for these two filters. Fig. 6 also shows the cross-correlation plot of these two filters. It is seen

that $PSL_{opt} < 0.08$, and in fact the sidelobe metric for many of the possible lags is notably smaller.

C. Example Using an Idealized Receiver

An analysis of the idealized equalizer performance for the SFNs in Fig. 3 in a simplified environment explains why minimizing the peak sidelobe is the criterion for improving performance. A general measure of an optimal receiver filter for an ISI channel with AWGN is the matched filter bound, which maximizes the SNR with respect to the channel impulse response. It can be shown that the matched filter response for the systems as described by the equations above is dependent on the cross-correlation at lags corresponding to the multipath and network delays, so minimizing the peak cross-correlation across the likely range of delays will indicate an improvement in overall performance.
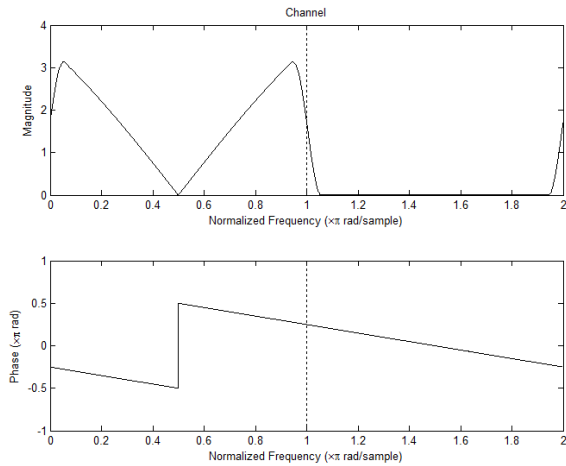
Fig. 7. The frequency spectrum in this example of a 0 dB echo with short delay created by two overlapping transmitters shows the deep fading across a broad span of carriers that can occur in certain locations of an SFN.
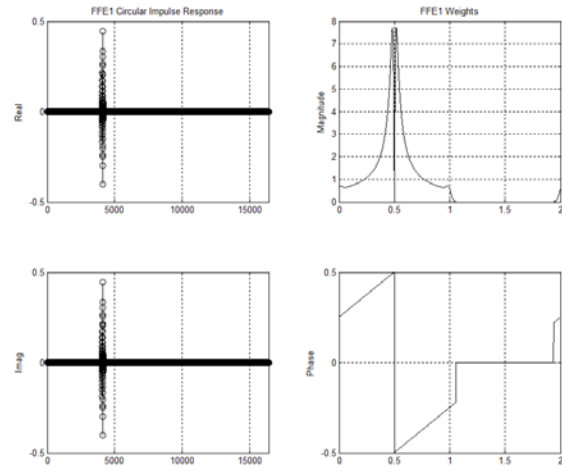

Fig. 8. Classical equalization techniques applied to the example of a 0 dB echo with short delay may create large amounts of noise enhancement to correct spans of deep fading.
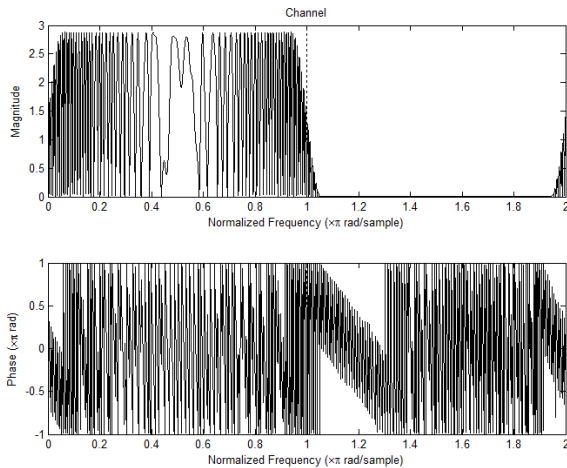

Fig. 9. The application of TDCFS to the 0 dB short echo delay example creates a more diverse channel frequency response.
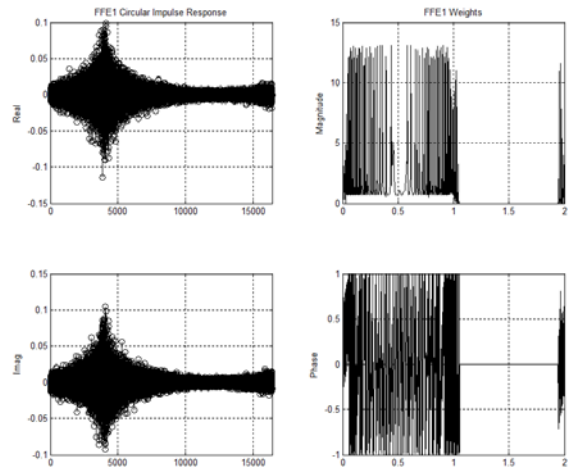

Fig. 10. The more complex equalizer response for the resulting application of TDCFS to the 0 dB short echo delay example also must create noise enhancement, but there is overall SINR gain of ~2 dB.

As an example of how the receiver can improve using a TDCFS scheme, consider a two transmitter simple SFN where the receiver is located such that it receives the signal from the second transmitter with the same strength as the first transmitter but with 1 sample of delay and opposite phase, giving a channel frequency response as shown in Fig. 7.

When the AWGN at the receiver is 25 dB, the MMSE (minimum mean square error) solution for the receiver equalizer provides a Signal-to-Interference Noise Ratio (SINR) of ~12.5 dB because of the large null created in the middle of the band due to the multipath. This equalizer response is shown in Fig. 8.

If a code filter set is used for the two transmitters under the same multipath conditions, the frequency response at the receiver is as shown in Fig. 9, which shows that the null in the middle of the band has been mitigated. The MMSE equalizer response, shown in Fig. 10, is clearly more complex, but the SINR out of the equalizer is now ~14.5 dB, a noticeable improvement over the simple SFN. Furthermore, it can be shown that due to the decorrelation effect of the code filter set over a large span of lags, the performance gain is even more significant when compiled across the wide range of multipath possibilities.
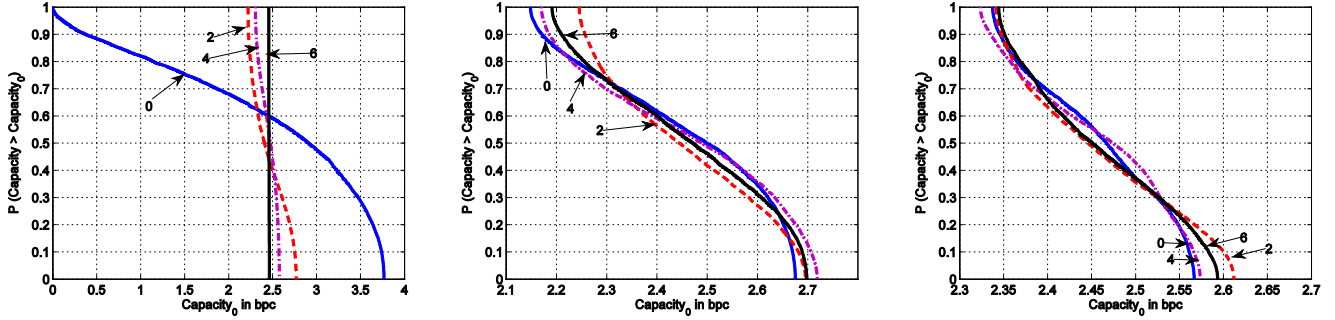
Fig. 11. CCDF of the channel capacity values in an SFN environment with two transmitter sites, received CNR of 8 dB, fixed delays (delay values of 0, 2, 4 and 6 samples) without pre-distortion (left), and with TDCFS pre-distortion with filter length 64 (middle), and with TDCFS pre-distortion with filter length 256 (right). The zero sample delay case is the worst-case scenario for SFN.

IV. PERFORMANCE SIMULATION RESULTS

In this section, we present capacity analysis and bit-error-rate (BER) simulation results to study the performance of TDCFS in SFN environments. We assume an SFN network with single antenna transmitters and with the same transmitted power. The transmitted signals can arrive at the receiver with different power levels, delays and with a phase that is a random variable uniformly distributed between 0 and 2π. We compare the performance of TDCFS with the conventional SFN setup where all the sites within the network transmit the same signal. We use the following signal parameters: FFT size of 8K carriers, 1/8 guard interval (GI), and 6 MHz bandwidth. In the BER simulations, we employ baseline ATSC 3.0 LDPC codes with a code-rate of 9/15, bit-interleavers, and 16-point two-dimensional non-uniform constellations. Since we assume static receivers, the random phases are kept constant within an OFDM symbol (for the capacity analysis) or time-interleaving block (for the BER simulations), but changed to a new random state every OFDM symbol or time-interleaving block.

A. Capacity Analysis with Perfect Channel Estimation

In Fig. 11, we present complementary cumulative distribution functions (CCDF) of the channel capacity values per OFDM symbol in an SFN environment with two transmitter sites. The signals from the two transmitters are received at the terminal with the same strength (0 dB echo) and different fixed delays. The delay between the first and second received signal takes the values of 0, 2, 4 and 6 samples, where the sampling time for a 6 MHz bandwidth system corresponds to 7/48 μs. We also assume the receiver has perfect knowledge of the channel realizations.

On the left side of Fig. 11 we show the performance of an SFN scheme without pre-distortion. We can observe that in the case that the received signals reach the receiver at the same instant (0 samples delay case) the CCDF has a wide variance due to constructive (signals with same phase) and destructive (signals with opposite phase) interference. As the delay increases, the variance of the CCDF diminishes, and therefore low capacity values have low probability. The middle and right plots of Fig. 11 show the results for TDCFS schemes with filter lengths (cf. section III) equal to 64 and 256. TDCFS improves the performance of the SFN scheme worst-case
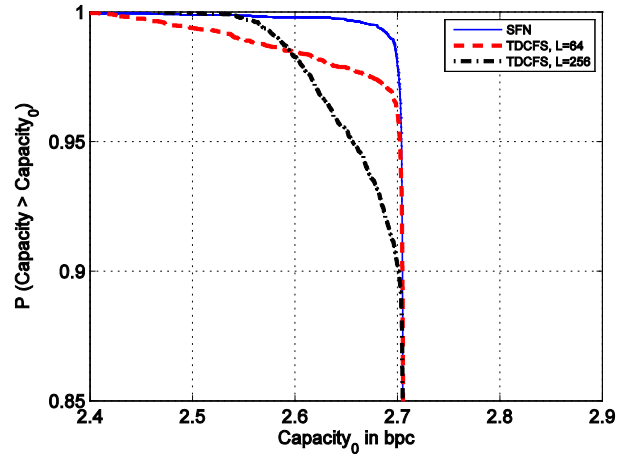

Fig. 12. CCDF of the channel capacity values in an SFN environment with two transmitter sites, received CNR of 9 dB, and random delays uniformly distributed in the $[0, N_{GI}]$ range, where N_{GI} is the number of samples in the GI. Performance of SFN (without pre-distortion), TDCFS with filter lengths of 64 and 256 included.

scenario, i.e., zero sample delay. The performance for the different delay values with TDCFS has an arbitrary order due to the interaction of the channel response and the TDCFS frequency pre-distortion. We also note that CCDF curves for TDCFS with filter length 256 have lower variance than the curves for the filter length 64.

In Fig. 12 we present a similar result as in Fig. 11 but here the delays between the two received signals are random variables uniformly distributed in the $[0, N_{GI}]$ range, where N_{GI} is the number of samples in the GI. Here, we can see that for CCDF values below 90%, SFN and TDCFS schemes provide similar performance. At higher CCDF values such as 98%, SFN provides slightly better performance than TDCFS schemes with both filter lengths. However, at CCDF values close to 100%, TDCFS improves the performance of SFN scheme due to the zero sample delay worst case for SFN (also shown in Fig. 11) which is solved with TDCFS.

B. Bit Error Rate Performance with Perfect Channel Estimation

In Fig. 15 we analyze the BER performance with two transmitter sites and fixed delay values of 0, 2, 4 and 6 samples (same scenario as in Fig. 11). Here, we can clearly
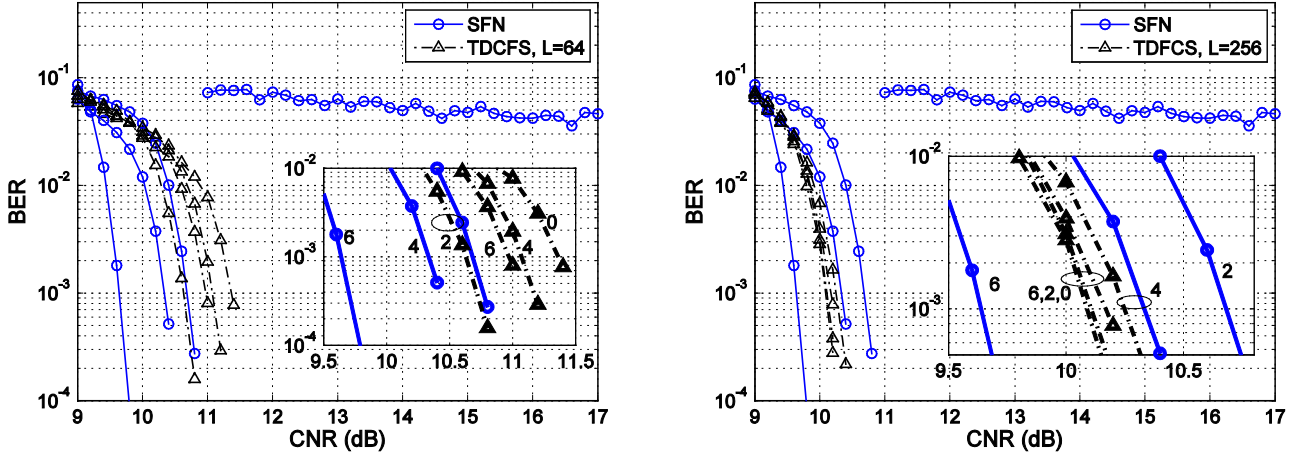
Fig. 13.   BER vs. CNR for two transmitter sites with fixed delay values of 0, 2, 4, and 6 samples. TDCFS with filter length of 64 (left) and 256 (right). Performance with ideal channel estimation. The figure shows how TDCFS can improve the performance of the worst-case SFN scheme.

observe how the worst-case for SFN scheme is the zero delay case that makes the service undetectable. On the other hand, performance of the SFN scheme improves for increasing delay. Similar to the CCDF capacity analysis, these results indicate that the implementation of TDCFS overcomes the SFN scheme worst-case scenario. We note that as the delay increases past a certain point (2 delay samples for $L = 64$ and 4 delay samples for $L = 256$), SFN scheme outperforms TDCFS as can be seen in Fig. 13. In the zoomed area of the plots we can observe that for TDCFS the performance order for the different delays takes (as in the capacity analysis of Fig. 11) arbitrary order. Comparing the critical 0 dB echo performance of TDCFS with respect to different filter lengths, filter length of 256 obtains better performance than filter length of 64 for short echo delays due to the higher diversity achieved for higher filter lengths. However, for cases of a 0 dB echo with delay exceeding 64 delay samples, it is expected that $L = 64$ will perform better than $L = 256$ due to greater decorrelation in the perceived channel impulse responses (similar to the above comparisons to SFN for echo delays beyond a certain point).

C. Impact of Channel Estimation Errors in System Performance

Next, we study the impact of channel estimation errors due to practical algorithms in performance of TDCFS scheme. At the transmitter the pilot insertion follows the patterns defined in ATSC 3.0 with $D_x = 6$ (one pilot every 6 OFDM carriers) and $D_y = 2$ (one pilot every 2 OFDM symbols). At the receiver, a cascade of two orthogonal one-dimensional filters in time and frequency domains realize the estimation of the channel realization in the data positions. As recommended in [14] the receiver performs temporal interpolation first, followed by the frequency interpolation. Regarding channel estimation algorithms at the receiver, we select linear interpolation in the time domain [14], although more sophisticated techniques may be required in time-varying channels common in mobile reception. In frequency domain we use linear interpolation or an estimation algorithm based on Wiener filtering [15]. Linear interpolation is a simple algorithm and provides a lower bound on the performance of practical estimation algorithms. The Wiener filtering is a more advanced algorithm that requires knowledge of the statistics of the channel and noise, which we
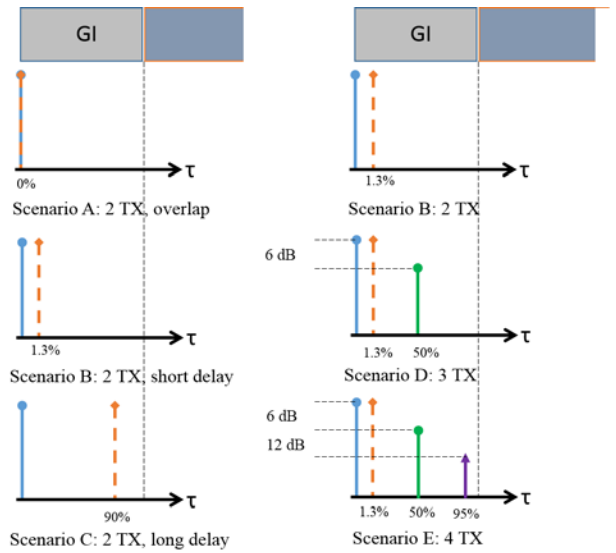

Fig. 14.   Diagram of SFN scenarios for the performance evaluation of SFN and TDCFS schemes with real channel estimation algorithms of Fig. 15.

assume in this analysis to be perfectly known at the receiver. The assumptions made for the Wiener filtering are clearly an idealization but provide an upper bound performance of a practical receiver using this type of algorithm.

The performance of TDCFS with real channel estimation algorithms is evaluated in the scenarios summarized in Fig. 14. Scenarios A, B and C in Fig. 14 (left) present the reception of two signals with the same strength at different delays relative to the GI duration. Specifically, scenario A, B and C present the case when signals arrive at the same time instant ("overlap"), at the 1.3% of the GI duration ("short delay"), and at the 90% of the GI duration ("long delay"), respectively. The short delay case is extracted from [13] and corresponds to 1.95 µs given the system sampling rate and FFT parameters. Scenarios B, D and E in Fig. 14 (right) present the reception of signals from two, three and four transmitters, respectively, at different delays. Specifically, for scenario B the delay of the second signal is at the 1.3% of GI with 0 dB of imbalance with respect to the first signal, scenario D includes scenario B and

an additional signal with a delay at 50% of the GI with 6 dB of power imbalance with respect to the first signal, and scenario E includes the scenario D and an additional signal with a delay at 95% of the GI and with a power imbalance of 12 dB with respect to the first signal.

Fig. 15 shows the performance for the five scenarios described in Fig. 14 and different channel estimation types. In the y-axes we plot required CNR to achieve a BER of 10^{-3} after BCH decoding. In the x-axes, we present the performance for SFN, TDCFS with filter lengths of 64 and 256 for 3 receiver types: a receiver with ideal channel estimation ("ideal"), a receiver using linear interpolation in time and Wiener interpolation in frequency ("Wiener"), and a receiver using linear interpolation in both time and frequency domains ("linear").

For scenarios A and B, the use of real channel estimation algorithms maintains similar CNR requirements as with ideal channel estimation. Furthermore, TDCFS maintains the performance improvement over the SFN worst-case scenario with all receiver types. For scenarios B and C, SFN requires lower CNR than TDCFS. For scenario C, the linear receiver type suffers a significant performance degradation for SFN and TDCFS due to high selectivity of the channel, and a more advanced receiver is required such as Wiener receiver, which performs close to ideal channel estimation. In these three scenarios for TDCFS, the filter length of 256 outperforms the filter length of 64.

In Scenarios B, D, and E we can observe that the CNR requirements increase with increasing number of transmitters. As expected, this CNR increase is higher for the linear receiver type. Although we can observe that SFN provides the best performance, TDCFS presents similar CNR requirements for all receiver types. Again, for TDCFS the filter length of 256 outperforms the filter length of 64.

V. IMPACT OF MAXIMUM DIVERSITY NETWORK ON THE ATSC 3.0 TRANSITION PLAN IN THE U.S.

For ATSC 3.0 to go forward, both economically and technologically, a feasible transition plan is necessary [17]. Without a mandate from the Federal Communications Commission (FCC), a transition period and additional bandwidth are required.

A. VHF Band

The current FCC plans indicate that additional bandwidth in the UHF band will not be available, so attention must be paid to the viability of the VHF band. With the Maximum Diversity Network, the high-band VHF band reception capability can be greatly improved. The band can now provide mobile and handheld services, in addition to fixed reception. With the improved interference characteristic of the Maximum Diversity Network, an adjacent co-channel cell (i.e., a neighboring cell using the same transmission frequency but from a different broadcaster) can be spaced with no guard distance between cells, as indicated in Fig. 16 (Left).

With some accommodations from the FCC, the low-band VHF can also provide additional channels. Transmitted power levels would need to be increased to overcome the additional
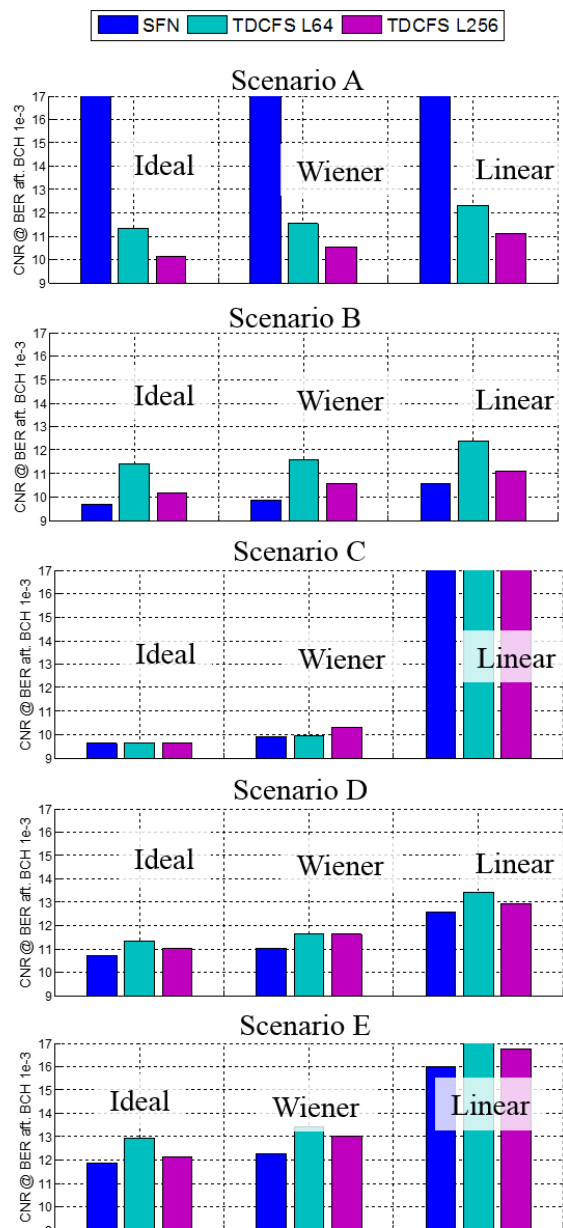

Fig. 15. Performance of SFN and TDCFS (with $L = 64, 256$) with real channel estimation algorithms in five different scenarios. Scenarios A, B and C study the case with two transmitter sites and different delays. Scenarios D and E study the case with three and four transmitters. Delay and power imbalance values for the different scenarios are presented in Fig. 14.

man-made noise in this band. The low-band VHF with its limitations could be used in two ways:

1). To motivate a broadcaster to move to low band VHF, they could be assigned two 6 MHz channels (12 MHz total) [18]. A receiver already equipped with a dual frontend would change from antenna diversity to channel (frequency) diversity. This option would be particularly attractive to broadcasters prioritizing ultra-high-definition broadcast television over handheld devices.

2). Cross-band sharing, where two broadcasters share two assignment channels, one in low-band VHF and the other in the UHF band. Both broadcasters would transmit mobile and
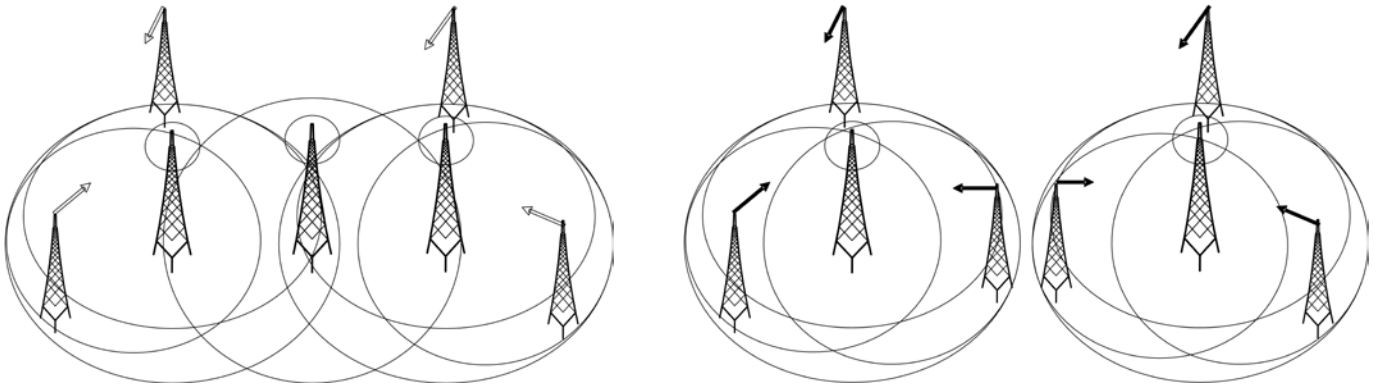
Fig. 16. Left: The use of networks with well-controlled secondary transmitters on the periphery allows closer geographical spacing of co-channel cells. Right: Merging smaller cells into regional network cells to achieve greater diversity can be accomplished with TDCFS even though the overlapping coverage has created more critical areas of potential interference.

handheld programs in the UHF band, and their high data rate programs would use low-band VHF.

B. Regional Network

Another feature of the Maximum Diversity Network allows two adjacent cells to be combined to form a regional network covering two metro areas that are currently separate under ATSC 1.0, as shown in Fig. 16 (Right). The network would use both of the original main towers and would add three to six additional towers. Sharing across metro areas would be especially beneficial to broadcasters in large and extended metro areas, such as New York, Miami, and Chicago, where a common broadcast service is appropriate over a larger geographical area.

VI. CONCLUSION

ATSC 3.0 has adopted a novel transmitter diversity technique known as Transmit Diversity Code Filter Sets (TDCFS) in which the same data is sent from multiple transmitters, but the signal waveforms are differentiated in such a way so that the likelihood of signal cancellation is minimized regardless of channel conditions. Due to the linearity of the transmission process, TDCFS requires no increase in receiver complexity which is a desirable feature for mobile devices with limited power resources, though minimal added complexity specific to the use of TDCFS may also improve performance.

TDCFS may be instrumental in deploying maximum diversity SFN to exploit multipath diversity gain, receiver diversity gain, and transmitter diversity gain to cover metropolitan areas in the U.S. Compared to the coverage of a single high-power high-tower transmitter, the average gain across an entire metropolitan area is expected to be about 10 dB, but more importantly, the diversity gain in the most difficult environments will be considerably larger.

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