A Novel Spectrum Re-Planning Technique for Wideband Single Carrier Transmission

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***Abstract*— The flexible spectrum access and arrangement is a key requirement for future software defined radio (SDR). The capability in spectrum planning not only allows the efficient usage of the white space but can also enhance the system performance. This paper presents the example of using spectrum fragmenting and defragmenting technique to cope with multipath channels for a wideband single carrier (SC) signal. The single carrier modulation often requires both linear and non-linear equalization, i.e., linear equalizer (LE) and decision feedback equalizer (DFE), in the presence of multipath channel. The cost of the LE and DFE limits the usage of SC system as compared to multi-carrier (MC) system where DFE is not needed and LE can be achieved efficiently in the frequency domain. We show that by properly arranging the transmitted signal spectrum one can enhance the SC system performance and may possibly avoid building DFE.**

***Keywords*—perfect reconstruction channelizer; non-maximally decimated filter bank; wideband signal processing; channel equalization; single carrier.**

I. INTRODUCTION

The modem architecture has been dramatically evolved during the past decades driven by the increasing demand for high throughput and enhanced robustness. The MC transmission, e.g., OFDM [1], has captured major attention due to its simple equalization scheme and its low processing speed feature, i.e., serial to parallel conversion. These important characteristics allow one to build very broadband systems in a cost effective manner. However, as pointed out by [2], the OFDM based MC systems suffer from a number of drawbacks, such as CP overhead, high PAPR, sensitivity to carrier frequency offset; and these facts limit the system's throughput, power consumption, and BER performances. On the other hand, the legacy SC signal, i.e., square root raised cosine (SRRC) shaped QAM waveform, has these problems well controlled. Yet, legacy waveforms require very long equalizers, i.e., the ATSC-8VSB [3] has 6 MHz wide signal bandwidth; and requires an equalizer with total length exceeding 400 taps, the longest equalizer ever deployed. Authors in [4] have proposed joint diversity combining and filter bank (FB) transformed domain linear equalization technique for SC transmission. The implementation of [4] is based on non-maximally decimated filter bank (NMDFB) processing as detailed in [5]. It is shown in [4] that the NMDFB processing not only offers reduced power consumption as compared to legacy SC system but also allows low speed processing. Moreover, the simulation results also reflected significant superiority in bit error rate (BER) over uncoded OFDM system.

It is shown in [4] that, for a wideband SC system, one can avoid building non-linear equalizer (NLE), i.e., DFE, if supported by proper diversity combining methods. In this paper, we shall demonstrate another case that NLE can be avoided if provided with proper prior channel knowledge and extra bandwidth. Fig. 1 shows the conventional SC receiver with LE and NLE. It is well known that [6, 7] both minimum mean square error (MMSE) and zero forcing (ZF) LE cannot handle channels with deep notches especially in the low signal to noise ratio (SNR) region, i.e., low E_b/N_0. This is because LE significantly amplifies the noise floor while trying to invert the channel distortion.

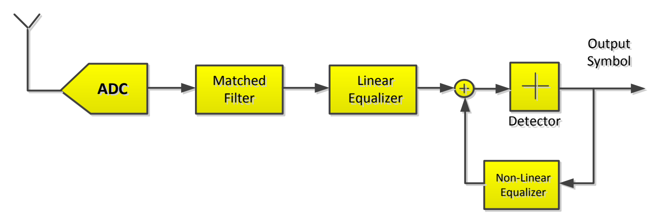

Fig. 1 Conventional SC Equalization

Besides LE and NLE, the conventional signal processing approach, when applied to wideband single carrier system, can hardly protect the signal from severe channel distortion. In this paper we propose using NMDFB processing to re-arrange the transmitted signal spectrum, or spectrum fragmentation at the transmitter end such that the wideband SC signal can "by-pass" the channel notches, Fig. 2. At the receiver end, a dual spectrum de-fragmentation process is performed to undo the fragmentation process done at the transmitter, Fig. 3. The carefully planned fragmenting and defragmenting process avoids channel notches and enhances the overall system performance. The block diagram of the proposed NMDFB transmitter and receiver is shown in Fig. 2 and Fig. 3 respectively.

The organization of this paper is: section 2 introduces the NMDFB signal model; section 3 demonstrates the spectral fragmenting and defragmenting approach; section 4 shows the simulation result; and the section 5 draw the conclusion.

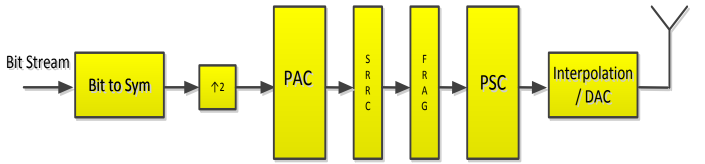

Fig.2 NMDFB based SC Transmitter with Spectral Fragmentation

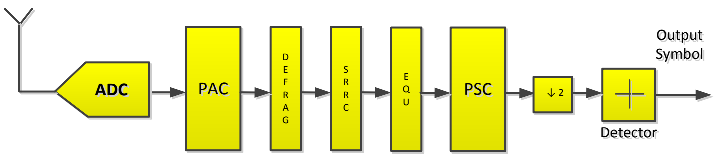

Fig.3 NMDFB based SC Receiver with Spectral Defragmentation

II. Background on PR-NMDFB

In this section we review the signal model of PR-NMDFB. The PR-NMDFB is composed by a pair of analysis filter bank (AFB) and synthesis filter bank (SFB) shown in Fig 4, whose composite response satisfies the PR property, i.e, the composite AFB and SFB only introduces delay in the absence of the intermediate processing stage . Historically, the PR filter bank has been proposed in many contexts, among which maximally decimated cosine modulated filter banks [8, 9] are the most popular choice in image coding communities. Yet, in the waveform level signal processing, we would particularly in favor of the NMDFB structure. The arguments are: 1) The aliasing management of NMDFB is easy [10, 11]. 2) The prototype filters are easy to design in terms of large system dynamic range. 3) It supports time varying update feature.

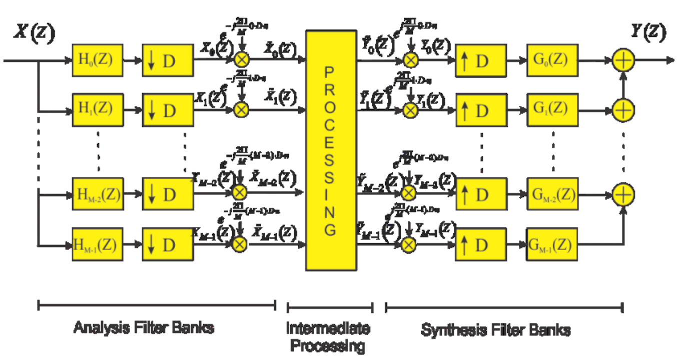

Figure 4. NMDFB Processing

The AFB contain M band pass filters (BPFs), whose Z-transforms are denoted as $H_m(Z)$, $m = 0,1,\dots,M-1$. Those filters have equal bandwidth, each centered on digital frequency $\theta_m = \frac{2\pi}{M}m$, $m = 0,1,\dots,M-1$. Let $h(n)$ be the impulse response of the low pass prototype filter (LPPF), whose Z-transform is $H(Z)$. The m^{th} BPF can then be represented as $h_m(n) = h(n)e^{j\frac{2\pi}{M}mn}$, whose Z-transform is $H_m(Z) = H\left(e^{-j\frac{2\pi}{M}m}Z\right) = H(W_M^m Z)$, and $W_M \triangleq e^{-j\frac{2\pi}{M}}$. A down sampling by D operation follows each BPF, where D is an integer that divides M. After the down sampling, we require the output signal on each channel to be centered on base-band. This task is accomplished by using a set of complex rotators $e^{-j\frac{2\pi}{M}mnD}$. The Z-transform of signal $\tilde{X}_m(Z)$ is:

$$\tilde{X}_m(z) = \frac{1}{D}\sum_{d=0}^{D-1} X\left(z^{1/D}W_D^d W_M^{-m}\right)H\left(z^{1/D}W_D^d\right) \qquad (1)$$

Denote the intermediate processing transfer function for the m^{th} filter in the bank to be $K_m(Z)$, for $m = 0,1,\dots,M-1$. Thus the input to the m^{th} filter in the SFBs is $\tilde{Y}_m(Z) = K_m(Z)\tilde{X}_m(Z)$. Again, complex heterodyne $\tilde{Y}_m(Z)$ by $e^{j\frac{2\pi}{M}mnD}$, one obtains $Y_m(Z) = \tilde{Y}_m(ZW_M^{mD})$. The final output of the SFBs $Y(Z)$ is:

$$Y(z) = \frac{1}{D}\sum_{d=0}^{D-1} X\left(zW_D^d\right)\sum_{m=0}^{M-1} K\left(z^D W_M^{mD}\right) H\left(zW_D^d W_M^m\right) G\left(zW_M^m\right) \qquad (2)$$

Eq. (2) can be expressed compactly in the matrix form. Let us define the following matrices and column vectors.

$$\boldsymbol{G}(\boldsymbol{Z}) = \left[G\left(ZW_M^0\right)\dots G\left(ZW_M^{M-1}\right)\right]^T$$

$$\boldsymbol{H}(\boldsymbol{Z}) = \left[H\left(ZW_M^0 W_D^0\right)\dots H\left(ZW_M^{M-1}W_D^0\right)\right]^T$$

$$\bar{\boldsymbol{X}}(\boldsymbol{Z}) = \left[X\left(ZW_D^1\right)\dots X\left(ZW_D^{D-1}\right)\right]^T, \boldsymbol{X}(\boldsymbol{Z}) = \left[X\left(ZW_D^0\right), \bar{\boldsymbol{X}}(\boldsymbol{Z})\right]^T$$

$$\mathbb{H}(\boldsymbol{Z}) = \begin{bmatrix} H(ZW_M^0 W_D^0) & \dots & H(ZW_M^0 W_D^{D-1}) \\ \vdots & \ddots & \vdots \\ H(ZW_M^{M-1}W_D^0) & \dots & H(ZW_M^{M-1}W_D^{D-1}) \end{bmatrix}_{M\times D}$$

$$= \left[\boldsymbol{H}_{M\times 1} \mid \bar{\mathbb{H}}_{M\times(D-1)}\right]_{M\times D}$$

$$\mathbb{K}(\boldsymbol{Z}) = diag\left(K(Z^D W_M^{0D}),\dots,K(Z^D W_M^{(M-1)D})\right)$$

$$= \begin{bmatrix} K(Z^D W_M^{0D}) & \cdots & 0 \\ \vdots & \ddots & \vdots \\ 0 & \cdots & K(Z^D W_M^{(M-1)D}) \end{bmatrix}_{M\times M}$$

Then, the matrix representation of Eq. (2) can be written as:

$$\mathrm{Y(Z)} = \frac{1}{\mathrm{D}}\mathrm{G}_{1\times\mathrm{M}}^{\mathrm{T}}(\mathrm{Z})\mathbb{K}_{\mathrm{M}\times\mathrm{M}}(\mathrm{Z})\mathbb{H}_{\mathrm{M}\times\mathrm{D}}(\mathrm{Z})\mathrm{X}_{\mathrm{D}\times 1}(\mathrm{Z})$$

$$= \frac{1}{D}\boldsymbol{T}^{\mathbb{K}}{}_{1\times D}(\boldsymbol{Z})\boldsymbol{X}_{D\times 1}(\boldsymbol{Z}) \qquad (3)$$

where $\boldsymbol{T}^{\mathbb{K}}{}_{1\times D}(\boldsymbol{Z}) \stackrel{\text{def}}{=} \boldsymbol{G}_{1\times M}^{T}(\boldsymbol{Z})\mathbb{K}_{M\times M}(\boldsymbol{Z})\mathbb{H}_{M\times D}(\boldsymbol{Z}) = \left[T_s^{\mathbb{K}}(Z)\ \boldsymbol{T}_A^{\mathbb{K}}(Z)\right]$ is the total transfer function for the M-path, decimate by D, AFBs and SFBs, $T_s^{\mathbb{K}}(Z) \stackrel{\text{def}}{=} \boldsymbol{G}_{1\times M}^{T}(\boldsymbol{Z})\mathbb{K}_{M\times M}(\boldsymbol{Z})\boldsymbol{H}_{M\times 1}(\boldsymbol{Z})$ is the desired signal transfer function, whereas $\boldsymbol{T}_A^{\mathbb{K}}(Z) \stackrel{\text{def}}{=} \boldsymbol{G}_{1\times M}^{T}(\boldsymbol{Z})\mathbb{K}_{M\times M}(\boldsymbol{Z})\overline{\mathbb{H}}_{M\times(D-1)}(\boldsymbol{Z})$ is the undesired aliasing transfer function. We then rewrite Eq. (3) as:

$$Y(Z) = \frac{1}{D}T_s^{\mathbb{K}}(Z)X(Z) + \frac{1}{D}\boldsymbol{T}_A^{\mathbb{K}}(Z)\overline{\boldsymbol{X}}(Z) \tag{4}$$

If the aliasing transfer function $\boldsymbol{T}_A^{\mathbb{K}}(Z) = \boldsymbol{0}_{1\times(D-1)}$, the aliasing energy would be completely cancelled. It can be shown that the condition ensures aliasing cancellation is:

$$H(ZW_D^d)G(Z) = 0, \forall\ d = 1, \dots, D-1 \tag{5}$$

Examine the 1st term in Eq. (4), the condition for producing a distortionless response for the input signal $X(Z)$ is:

$$T_s^{\mathbb{K}}(Z) = \sum_{m=0}^{M-1} K(Z^D W_M^{mD})H(ZW_M^m)G(ZW_M^m) = Z^{-n_D}$$

where n_D is a positive integer representing the total delay introduced by the AFBs / SFBs, plus the intermediate processing. And, we do not require the intermediate processing matrix $\mathbb{K}_{M\times M}$ to participate in distortionless response, thus setting it to identity $\mathbb{K}_{M\times M} = \mathbb{I}_{M\times M}$. The distortion-less condition then becomes:

$$\sum_{m=0}^{M-1} H(ZW_M^m)G(ZW_M^m) = Z^{-n_D} \tag{6}$$

Therefore, in the absence of the intermediate processing matrix $\mathbb{K}_{M\times M}$, the PR condition for a NMDFB has to simultaneously satisfy Eq. (5) and (6). Seen from Eq. (5), it is clear that as long as $H(ZW_D^d)$ and $G(Z)$, $\forall\ d = 1 \dots D-1$ do not share common pass band and certain part of their transition bands, the aliasing energy can be made arbitrarily small by increasing the prototype filters' stop-band performance. And Eq. (6) is satisfied when the composite response of $H(Z)$ and $G(Z)$, the LPPF for AFB and SFB, forms a Nyquist channel. Let us define the Nyquist channel $H^{NYQ}(Z) \triangleq H(Z)\,G(Z)$, with integer delay n_D. The author in [10] suggested letting both $H(Z)$ and $G(Z)$ to be identical SRRC filters. In our observation, if the intermediate processing stage is emphasized, the analysis filter bank prototype filter, $H(Z)$, may be chosen to be any Nyquist pulse while the synthesis prototype filter, $G(Z)$, can be designed via Remez algorithm satisfying Eq. (5) and (6). The authors in [5, 11] proposed efficient polyphase implementation of NMDFB requiring an M-point FFT and one LPPF to achieve either AFB or SFB.

III. Spectral Fragmenting and Defragmenting

Examining Eq. (3, 4), by properly designing the intermediate processing matrix $\mathbb{K}_{M\times M}$, one can achieve NMDFB based wideband filtering to the input signal [5]. On top of this, we can introduce a permutation matrix $\mathrm{F}_{M\times M}$ followed by $\mathbb{K}_{M\times M}$ to rearrange the already shaped spectrum, Eq. (7).

$$\mathrm{Y}(\mathrm{Z}) = \frac{1}{\mathrm{D}}\mathrm{G}_{1\times\mathrm{M}}^{\mathrm{T}}(\mathrm{Z})\mathbb{K}_{\mathrm{M}\times\mathrm{M}}(\mathrm{Z})\mathrm{F}_{M\times M}\mathbb{H}_{\mathrm{M}\times\mathrm{D}}(\mathrm{Z})\mathrm{X}_{\mathrm{D}\times 1}(\mathrm{Z}) \tag{7}$$

Examining Eq. (7), the permutation matrix $\mathrm{F}_{M\times M}$ can be mathematically thought as altering the location of the AFB output channels; the SFB then synthesizes the spectrum location altered signal as modulated signal or signal to be transmitted. In practice, one can think it as taking a part of the signal spectrum and load it onto some other frequencies. The purpose of this work is to show by properly fragmenting the signal spectrum one can make the system more robust to severe channel distortions.

Assuming the transmitter has prior channel knowledge, i.e., the channel gain accross the operating frequency band, then it is possible to avoid deep channel notches by using permutation transform matrix $\mathrm{F}_{M\times M}$. For instance, one can reload the signal spectrum portion that subjected to severe channel distortion to other available frequency bands with better channel conditions. This operation is common to OFDM based system, i.e., waterfilling technique, but can hardly be applied to SC system as of today. The beauty of using NMDFB based signal processing tool to perform spectrum planning is of course the efficiency. Eq. (7) suggests that we can perform the SC shaping and permutation at one shot with already significantly lowered power consumption [5].

At the receiver end, one needs to first perform the defragmenting process, i.e., to undo the fragmentation done at the tranmistter. Ideally, one can design the defragmenting permutation matrix such that the product of the two permutation matrices, i.e., one at the transmitter end; and the other one at the receiver end, produces identity matrix. This is true if and only if AFB channels do not overlap with each other. In practice, the defragmenting process needs also to include the edge bins as well so that PR can be achieved with reasonable amount of distortion.

IV. Simulation Results

This section presents the example of the proposed spectrum re-planning technique. We first show the fragmenting and defragmenting process based on a square root raise cosine (SRRC) filter shaped quadrature amplitude modulation (QAM) signal. We then show how fragmenting and defragmenting process improve the system perform in the presence of severe channel distortions.

Figure 5 below shows the fragmented and defragmented spectrum of a SRRC shaped QAM signal. As we described earlier, the fragmenting process performs the spectrum re-planning task. We can see from the 2nd subplot that part of the signal spectrum is transported to other frequency band. Subplot 3 shows the signal spectrum after the defragmenting process. To confirm the fragmenting / defragmenting process has very limited distortion to the original signal, we also plotted the matched filtered constellation and its dispersion in subplot 1. It is clear that the re-planning process did not introduce much distortion to the signal.

Let the multipath channel having impulse response $H(z) = 1 + (0.9\text{-}j0.45)z^{-2}$. Fig. 6 shows the transmitted SRRC QAM signal spectrum and the received spectrum (channel distorted) without using fragmenting / defragmenting transform. Fig. 7 shows the fragmented SRRC QAM signal spectrum and the received signal spectrum. Comparing Fig.6 and 7, it is clear that fragmenting / defragmenting transform is able to reassign the signal spectrum distribution to avoid channel distortion. The MMSE equalizer response is also plotted in Fig 6, and 7, where we can see the MMSE linear equalizer can hardly deal with deep channel notches.

Based on the fixed multipath channel $H(z) = 1 + (0.9\text{-}j0.45)z^{-2}$, we show in Fig. 8 the mean square error (MSE) observed at the detector after MMSE linear equalization. Clearly, the proposed spectral re-planning approach produces much lower MSE after linear equalization. This suggests that one can potentially avoid building NLE if the fragmenting / defragmenting transform can be employed. Of course, this also requires extra bandwidth and prior channel knowledge.

I. CONCLUSION

In this paper, we proposed the spectrum re-planning technique for a wideband SC communication system. The virtue of this technique is to enhance the overall system performance in the presence of severe channel distortions. The simulation results show the fragmenting / defragmenting process does not hurt PR constrain. Moreover, this process does improve the system performance in terms of MSE after equalization. This suggests the potential possibility of eliminating NLE in a wideband SC system.

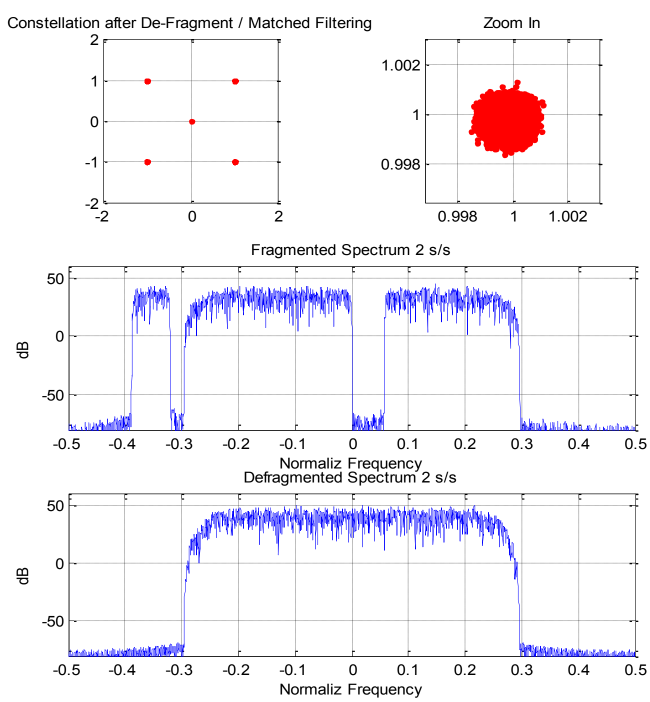

Figure 5. Fragmenting and Defragmenting of a SRRC Shaped QAM Signal

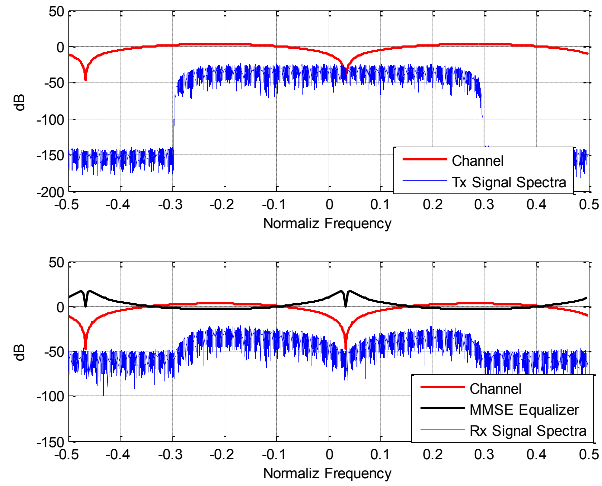

Figure 6. Conventional SRRC QAM Signal Spectrum and Multipath Channel, MMSE Equalizer Frequency Responses

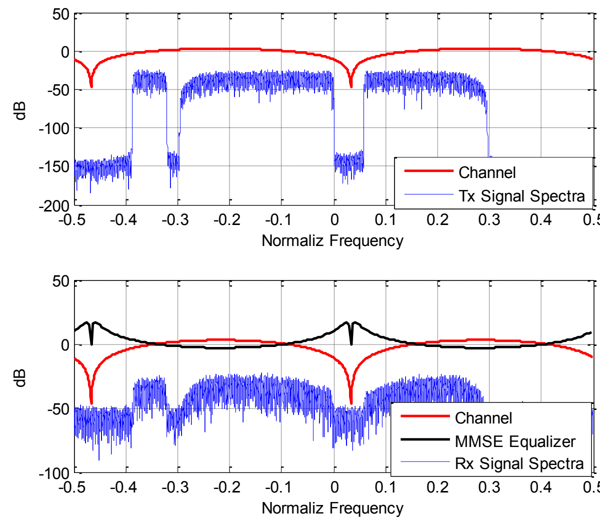

Figure 7. Fragmented SRRC QAM Signal Spectrum and Multipath Channel, MMSE Equalizer Frequency Responses

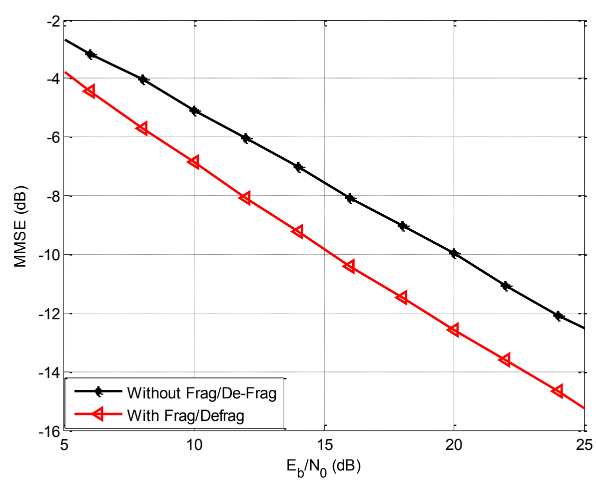

Figure 8. MSE vs. E_b/N_0

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