Two-stage Hybrid Decision Feedback Equalization for DS-CDMA Systems

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***Abstract*— This letter proposes a hybrid decision-feedback equalizer (HDFE) for DS-CDMA systems. The proposed HDFE is carried out in two stages to improve the accuracy of the feedback signals by exploiting the spreading gain in the feedback filter. The spread signals for a symbol are buffered and then detected, after which re-spreading is applied to generate feedback signals at the chip level. Through the use of symbol detection for feedback filtering, more accurate feedback signals are achieved. Simulation results demonstrate the superior error performance of the proposed scheme over the frequency domain linear equalization (FD-LE) and RAKE schemes in highly dispersive channels.**

***Index Terms*— Equalization, hybrid decision feedback equalization, code-division multiple access.**

I. Introduction

THE achievable data rates over wireless channels are often limited by inter-symbol interference (ISI) due to the multipath propagation. In direct-sequence code-division multiple access (DS-CDMA) systems, multipath fading channels result in the interchip interference (ICI). Conventional receivers use RAKE combining to exploit multipath diversity in multipath fading channels. This combining technique is efficient only if ICI is negligible [1]. For high-data-rate transmission, RAKE receivers cannot efficiently recover the transmitted signal since the channel-induced ICI could destroy the orthogonality among the spreading codes.

In order to combat multipath channel impairments, single-carrier modulations equipped with frequency domain equalization (SC-FDE) are promising solutions for uplink transmission. Generally, time domain equalization (TDE) has a high computational complexity to overcome the effects of a large delay spread. An efficient way to reduce the complexity of TDE is to transform its operations into the frequency domain (FD), which results in frequency domain equalization (FDE). The concept of FDE was first introduced almost three decades ago [2]. After that, significant efforts have been dedicated to improve its performance. Recently, SC-FDE has been recommended for wireless broadband standard IEEE 802.16 [3].

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FDE has been investigated as one of the promising approaches to evolve the third-generation (3G) DS-CDMA systems. A particular FDE technique called cyclic-prefix CDMA was proposed for DS-CDMA in [4]. In fact, it was based on the simple FD-LE technique. In time-domain, decision-feedback equalization (TD-DFE) outperforms linear equalization (TD-LE). However, for channels having a long impulse response, TD-DFE becomes highly complex. Thus, there is a need to implement DFE in the frequency domain. For FD-DFE, a special hybrid structure of DFE (HDFE) was proposed, in which, the feedforward (FF) filter is implemented in the frequency domain, while the feedback process is performed in the time domain [5], [6]. This structure has advantages of reduced complexity by FF filtering in frequency-domain, and improved performance by feedback processing in time domain.

For DS-CDMA systems, most of related works on FDE are based on FD-LE [4], [7], or iterative block decision feedback equalization (IB-DFE) [8], [9]. In IB-DFE, the feedback filtering is performed in frequency domain and it has the inherent delay due to FFT processing since the symbol detection must be carried out in time domain [5]. Moreover, the performance of IB-DFE relies on the number of iterations. With two iterations, the performance of IB-DFE is relatively similar to HDFE. In this letter, we focus on the application of HDFE and propose a new HDFE scheme suitable for DS-CDMA systems. In the proposed scheme, spreading gain is exploited through feedback operation in the time domain to reduce the error-propagation effect. Both despreading and symbol detection are involved in generating the feedback signal. We recognize that interchip interference (ICI) must be eliminated at the chip level since the spread signals are distorted by the channel. As a result, re-spreading is applied to obtain the feedback signal. To take advantage of the spreading gain, the proposed HDFE is performed in two stages: the first stage makes use of linear equalization, and the second stage employs HDFE.

The rest of this letter is organized as follows. Section II introduces the data transmission formats for FDE. Section III presents the details of the proposed HDFE and its benefits are discussed. Finally, Section IV provides numerical results and discussions, followed by concluding remarks in Section V.

II. Data Transmission Formats for DS-CDMA Systems with FDE

FDE requires the data to be transmitted in blocks. Suppose that a data block containing M successive modulated symbols is transmitted in one time slot. The information data are spread with a spreading code having a spreading factor SF. The DS-

CDMA signals (chips) can be expressed as

$$s_{nSF+m} = c_m d_n \tag{1}$$

where s_k is defined for $0 \leq k \leq P\ (= M \times SF - 1)$, d_n is the n-th quadrature phase shift keying (QPSK) modulated symbol, and $c_m, (m = 0, 1, \ldots, SF - 1)$ is the complex spreading code.

FDE techniques are based on the *circular convolution* between the transmitted signal and channel impulse response (CIR) [6]. Thus, a particular transmission format at the transmitter is required to make the spread data s_n in (1) periodic. There are two common block transmission formats for FDE: *cyclic prefix* (CP) extension [5], and *unique word* (UW) extension [6]. In CP format, which is commonly used in OFDM systems, the last fragment of the data is repeated at the beginning of each transmission block. At the receiver, the CP part will be discarded.

In UW transmission format, a unique word (a training sequence [10]) is added to the end of each data block. Note that UW is not removed at the receiver and DFT is applied to the total received signals including the UW. Unlike CP, UW is not random; both the transmitter and the receiver know the values of the UW part. For feedback filtering, the interference caused by the UW can be reconstructed and completely removed in the initialization stage. This property enables feedback equalization process to start properly, mitigating the effect of error propagation. We thus choose UW-extension as a suitable transmission format for HDFE in this letter.

This transmission format is also known as 'PN-extension' [6], since a PN sequence is chosen as the unique word. It is worth of noting that the length of the cyclic prefix or the PN-extension must be longer than the maximum excess delay of the channel to achieve the circular convolution property between the CIR and transmitted signal.

PN extension: We denote h_l and τ_l, $l = 0, 1, \ldots, \rho - 1$, as the CIR (including the transmit filter and receive filter) and time delay of the $(l+1)$-th propagation path, respectively. A PN sequence of length $L \geq \tau_{\rho-1}$, say $\{p_n\}$, $n = 0, 1, ..., L-1$, is added to the spread transmit signal, where $\tau_{\rho-1}$ is the maximum excess delay of the channel. For efficient use of FFT and IFFT blocks, M and L should be selected to be such that $P = M \times SF + L$ becomes a power of 2. The transmitted signals (chips) $\mathbf{s} = [s_0, s_1, \ldots, s_{M\times SF-1}, p_0, p_1, \ldots, p_{L-1}]^T$ are circular on a block of size P, i.e., $s_n = s_{n+P}, \forall n = 0, 1, \ldots, L-1$.

III. Proposed Hybrid Decision-Feedback Equalizer for DS-CDMA Systems

The spread signals propagate through a multipath fading channel, and the received signals are written by

$$r_n = \sum_{l=0}^{\rho-1} h_l s_{n-\tau_l} + \eta_n \tag{2}$$

where η_n is the zero-mean complex Gaussian white noise samples with variance σ_n^2. In (2), the channel path gains $\{h_l\}$, $l = 0, 1, \ldots, \rho - 1$, are assumed to be quasi-static and mutually independent. By applying the circular property of transmitted signals, we transform (2) into the frequency domain, which is expressed as

$$R_k = H_k S_k + N_k, \ k = 0, 1, \ldots, P - 1 \tag{3}$$

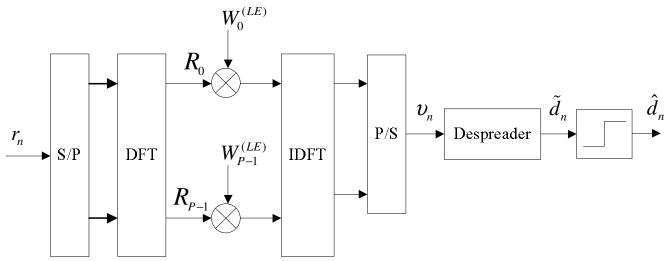

Fig. 1. Frequency-domain linear equalization structure.

where R_k, S_k and N_k are the P-point DFT of the received signal, the spread signal, and the white noise, respectively. We first examine the typical FD-LE for the DS-CDMA systems, and then describe our proposed HDFE approach.

A. Frequency domain MMSE linear equalization

The FD-LE is shown in Fig. 1. The DFT output of the received signals $\{R_k\}$ is first weighted with linear equalizer coefficients $\{W_k\}$. The resulting signals are then transformed back into the time domain by an IDFT block. Finally, UW is discarded, and the information-bearing data are fed into the despreader, which is followed by a threshold detector to recover the transmitted data. The linear equalizer filter coefficients can be expressed as [4]

$$W_k^{(LE)} = \frac{H_k^*}{|H_k|^2 + \sigma_n^2}, \ k = 0, 1, \ldots, P - 1. \tag{4}$$

In highly frequency-selective fading channels, the residual ICI could drastically degrade the BER performance of FD-LE. Some ICI cancellation schemes combined with FD-LE have been introduced in [11] to further improve the performance of FD-LE. But, in this letter, we focus on ICI cancellation by feedback filtering.

B. The proposed HDFE for DS-CDMA systems

While the FD-LE is widely used for the reception of DS-CDMA signals due to its low complexity, the application of DFE for DS-CDMA systems received less attention. In this section, we present a novel HDFE, particularly designed for DS-CDMA systems. The proposed HDFE architecture is depicted in Fig. 2. In the proposed scheme, detection for the feedback operation is made at the symbol level to take advantage of the processing gain. Then, re-spreading is applied to the detected symbol to obtain the feedback signal at chip level. However, the feedback signals are not available before re-spreading. Hence, the proposed equalization consists of two stages. In the first stage, the FD-LE is carried out because the feedback signals are not available. The main purpose of the first stage is to provide tentatively detected chips for the feedback operations of the second stage. In the second stage, feedback filtering is carried out based on the re-spreading process, and the feedforward filter coefficients are updated based on the DFE. In the first stage, the FF filter coefficients are obtained from frequency domain linear equalization since the feedback signal is not available, and the FB filter is omitted. The frequency domain signals $\{R_k\}$

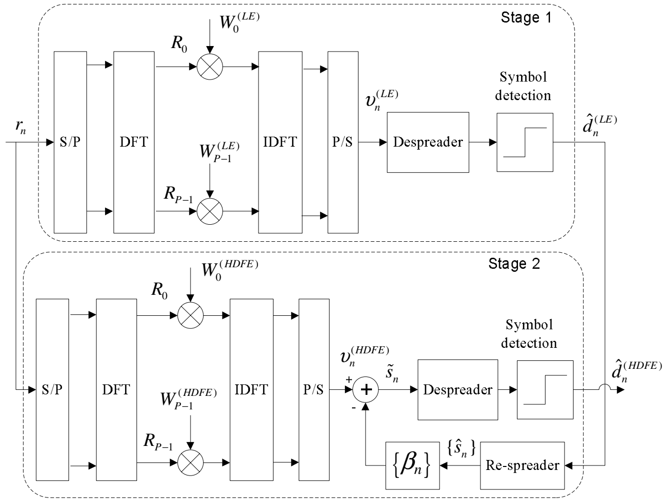

Fig. 2. The proposed HDFE structure for DS-CDMA systems. The outputs of linear equalization in the first stage is used as feedback singnals for feedback filtering in the second stage.

in (3) are weighted with the FD-LE coefficients $\{W_k^{(LE)}\}$, $k = 0, 1, \ldots, P-1$, and an IDFT is applied to the resulting signal to yield the time-domain signal $\upsilon_n^{(LE)}$ as

$$\upsilon_n^{(LE)} = (1/P)\sum_{k=0}^{P-1} R_k W_k^{(LE)} e^{j2\pi kn/P}, \; n = 0, 1, ..., P-L-1. \quad (5)$$

In (5), the last L values of $\{\upsilon_n^{(LE)}\}$ are discarded since they are distorted versions of the PN sequence. Since feedback filtering is not included in the first stage, the sequence $\{\upsilon_n^{(LE)}\}$ are directly fed into the despreader. Symbol detection and re-spreading are subsequently performed to generate the feedback signal, which is needed for the second stage. In the second stage, $\upsilon_n^{(LE)}$ in (5) is recalculated with the feedback signal from the first stage, which is denoted by $\upsilon_n^{(HDFE)}$ in Fig. 2. The FF filter coefficients of linear equalization $\{W_k^{(LE)}\}$ in (5) are adjusted to the FF filter coefficients of HDFE, i.e., $\{W_k^{(LE)}\}$ is replaced by $\{W_k^{(HDFE)}\}$. The feedback signals (chips) $\hat{s}_n$ are generated from the first stage, which are the re-spread version of the detected symbol. The FB filter is now taken into account, and the soft estimates of chips fed into the despreader are given by

$$\tilde{s}_n = \upsilon_n^{(HDFE)} - \sum_{i=1}^{N_b} \beta_i \hat{s}_{n-i}, \; n = 0, 1, ..., P-L \quad (6)$$

where $\{\beta_i\}$, $(i = 1, \ldots, N_b)$ are the FB filter coefficients and $\{\hat{s}_n\}$ is the previously estimated chips obtained from the first stage. Again, the PN sequence in (6) is discarded, and the detector input is written by

$$\tilde{d}_n = (1/SF)\sum_{m=0}^{SF-1} \tilde{s}_{nSF+m} c_m^*, \; n = 0, 1, ..., (P-L)/SF, \quad (7)$$

where (*) denotes complex conjugate and c_m, $(m = 0, 1, \ldots, SF-1)$ is the complex spreading code. We derive the optimal solution for the feedforward and feedback filter coefficients under the assumption of correct decisions on $\{d_n\}$. Thus, the effect of error propagation is not evaluated mathematically. The filter coefficients are adjusted to minimize the mean-square error (MSE) at the output of the detector. The objective function is

$$J = E\left\{|e_n|^2\right\} = E\left\{\left|d_n - \hat{d}_n\right|^2\right\}. \quad (8)$$

Applying the orthogonality principle and after a few steps of algebra, the FF filter coefficients are calculated as

$$W_k = \frac{H_k^*(1 + \sum_{l=1}^{N_b} \beta_l e^{-j2\pi kl/P})}{|H_k|^2 + \sigma_n^2}, \; k = 0, 1, \ldots, P-1, \quad (9)$$

and the FB coefficients are determined by solving a set of linear equation, which is written in the vector-matrix form as

$$\mathbf{\Lambda\beta} = \boldsymbol{\lambda}, \quad (10)$$

where $\boldsymbol{\beta} = [\beta_1, \beta_2, \ldots, \beta_{N_b}]^T$ is a column vector of the FB coefficients; elements of $\boldsymbol{\lambda} = [\lambda_1, \lambda_2, \ldots, \lambda_{N_b}]^T$ are defined as

$$\lambda_m = \sum_{k=0}^{P-1} \frac{e^{j2\pi mk/P}}{|H_k|^2 + \sigma_n^2}, \; m = -N_b + 1, \ldots, 0, \ldots N_b - 1, \quad (11)$$

and $\mathbf{\Lambda}$ is an $N_b \times N_b$ matrix with $\mathbf{\Lambda}_{i,j} = \lambda_{i-j}$. We can view λ_m as the IDFT of $\{1/(|H_k|^2 + \sigma_n^2)\}$. Note that, $\mathbf{\Lambda}$ is a Toeplitz matrix; hence, efficient algorithms can be applied to reduce the complexity of solving (10). The linear equalization coefficients for the first stage can be obtained easily from (9) by eliminating the feedback part.

The significance of the proposed structure lies on the second stage; by exploiting the spreading gain in the feedback path, the quality of the feedback signal is significantly improved, resulting in a better performance with the proposed HDFE. In terms of hardware complexity, the two-stage equalization is identical to that of traditional HDFE since the equalization is just performed two times on the same hardware devices. Obviously, more computational complexity is needed to update the feedforward filter coefficients, but it is negligible since the first stage employs the frequency domain linear equalization.

The proposed two-stage equalization can be extended for more reliable reception of multicode DS-CDMA signals in highly dispersive channels. This can avoid the multi-level detection needed to get the feedback signals when HDFE is applied to multicode DS-CDMA systems. The two-stage equalization for multicode CDMA applications is processed in two steps as described above. The first stage is the FD-LE applied to multicode signals. The feedback signals are derived by applying the multicode re-spreading to the detected symbols from the first stage, and feedback equalization starts based on those despread multi-level signals. With the multicode despreading in the feedback filter, we can overcome the difficulty of the multi-level detection, concurrently increase the accuracy of feedback signals. Simulation results show that the two-stage HDFE greatly improves the performance of multicode DS-CDMA systems in frequency selective fading channels, compared with the FD-LE as presented in [7].

IV. Simulation Results

In this section, we evaluate the performance of the proposed HDFE scheme using computer simulation. We consider the channel B of the vehicular environment as described in

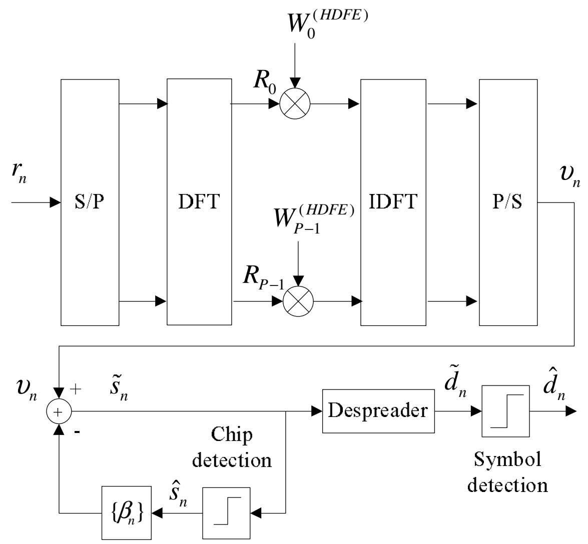

Fig. 3. HDFE for DS-CDMA systems with chip level detection.

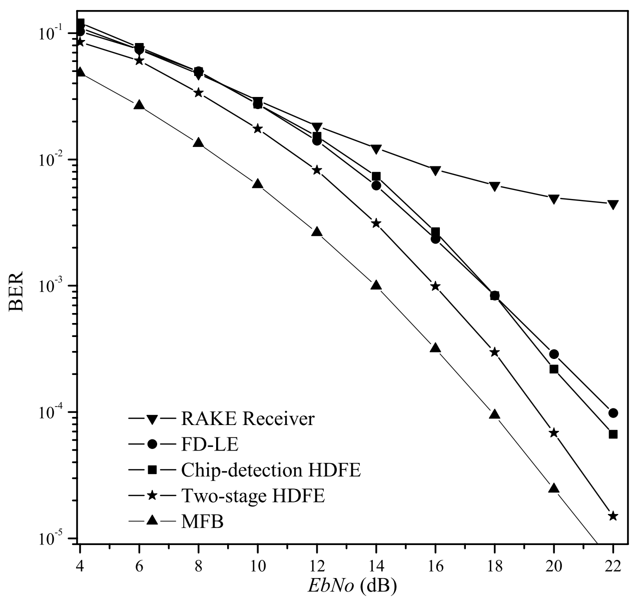

Fig. 4. Comparison of BER performances in quasi-static channels for single code DS-CDMA systems.

Recommendation ITU-R M.1225 [12]. This channel model produces an *rms* delay spread of 4023ns, and the channel excess delay extends to 74 chips. Therefore, a PN sequence of 80 chips is inserted into each data block. The chip rate is selected to be 3.6864 Mbps. We assume ideal channel estimation and consider the performance of an uncoded system using QPSK modulation. A 16-ary Walsh code is used as the spreading code. With an FFT size of 1024, the number of modulated symbols in one transmission block is 64 (single-code case). Theoretically, FDE is developed based on the assumption of block fading where the propagation gain is constant over one data block duration. In this section, the performance of FDE is also evaluated in time-varying channels. The maximum Doppler shift will be calculated assuming a carrier frequency of 2 GHz. We compare the performance of the proposed system with that of the following anti-multipath approaches:

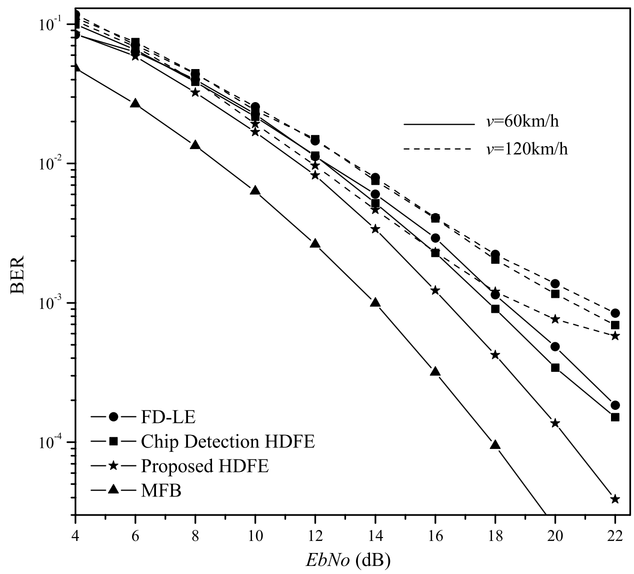

Fig. 5. BER performance of the proposed HFDE in time-varying fading environments.

1) Frequency domain MMSE linear equalization, 2) RAKE combining. In addition, the HDFE, in which the detection for feedback is carried out in chip level, is also considered. Such a HDFE structure is illustrated in Fig. 3. Its performance is labeled by '*chip-detection* HDFE' in Figs. 4 and 5.

Simulation results assuming a quasi-static fading model are shown in Fig. 4. The theoretical matched filter bound (MFB) derived from [13] is included as the baseline. The proposed method is found to outperform other frequency-domain equalization techniques. While the chip-detection HDFE provides a small gain over the FD-LE, the proposed HDFE provides about 2dB gain over the FD-LE at a BER of 10^{-3}. A critical assumption for the application of FDE is that the channel gains are constant within one block duration. The performance of FDE could be sensitive to the fading rate since even a small fluctuation of the channel coefficients whose effects can be neglected in the time-domain might destroy the equivalency between convolution in time domain and multiplication in frequency domain. Thus, the performance of the frequency domain approach might be severely degraded in time-varying channels. Fig. 5 compares the BER performance of FD-LE, chip-detection HDFE, and the proposed HDFE schemes under different mobile speeds. It is observed that the proposed two-stage HDFE outperforms other common FDE schemes; however, as the mobile speed increases to 120km/h, the performance gain diminishes. Fig. 6 shows the performance of multicode DS-CDMA systems combined with FDEs in time-varying channels. In pedestrian environment (user speed =3 km/h), the two-stage HDFE achieves a much better performance than other traditional FDE structures. As the speed increases, significant performance degradation is observed. Notably, the performance of RAKE receivers for multicode CDMA systems drastically degrades. This is due to the fact that the received signals in multicode-CDMA systems suffer from both inter-code interference and multipath interference. With FDE, both mulith-path interference and

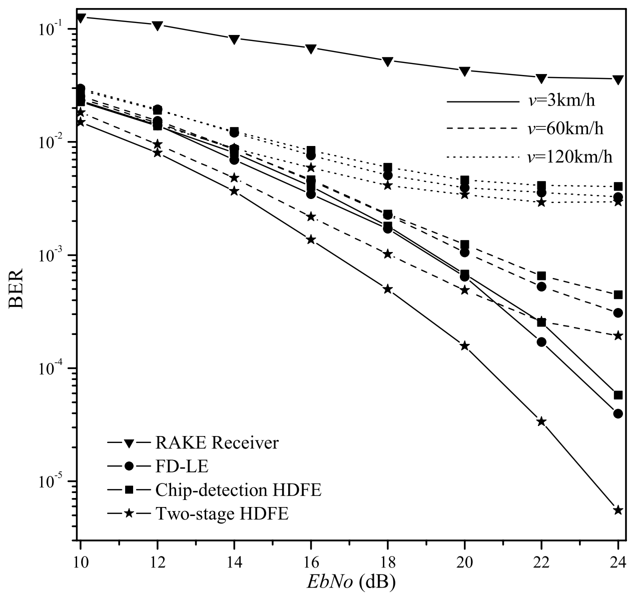

Fig. 6. BER performance of FDEs for multicode CDMA transmission over time-varying fading environments.

inter-code interference can be simultaneously reduced.

V. Conclusion

We have presented a two-stage HDFE scheme for DS-CDMA systems. The spreading gain is obtained through the feedback path, so that the reliability of the feedback signals is improved. The overall performance improvement lies on the improved accuracy of the detected symbols, which is due to two-stage equalization. In addition, the effect of error propagation is mitigated by the proposed scheme. Simulated BER performance of the proposed system shows significant gains over the conventional anti-multipath methods. The two-stage HDFE is then effectively extended for the reception of multicode DS-CDMA systems. The proposed equalization technique is suitable for high-date-rate wireless communications in highly dispersive channels.

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