A Novel Hybrid CFO Estimation Scheme for UFMC-Based Systems

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***Abstract*— In this letter, a robust hybrid carrier frequency offset (CFO) estimation scheme based on training symbols is devised for universal-filtered multi-carrier (UFMC) based systems. In this scheme, the CFO estimator based on a least square (LS) criterion is first derived to acquire a coarse frequency offset estimation with the use of a large frequency searching step. And then, the residual frequency offset is accurately estimated through the auto-correlation operation of received signals, which have been compensated. This algorithm has the features of larger estimation range as well as higher estimation precision. Through numerical simulations, it is validated that the performance of the proposed algorithm is better than that of existing algorithms.**

***Index Terms*— Universal-filtered multi-carrier, carrier frequency offset, least square, synchronization, 5G.**

I. Introduction

As A new air interface based on non-orthogonal waveform for 5G systems, universal-filtered multi-carrier (UFMC) that was firstly proposed in [1] combines the simplicity of orthogonal frequency division multiplexing (OFDM) with robustness of filter bank multi-carrier (FBMC). Compared with FBMC, UFMC [2], [3] demands a shorter filter length that often equals the length of the cyclic prefix (CP) in CP-OFDM. Moreover, it is very suitable for short uplink burst communications. Distinct from OFDM, it is not necessary for UFMC to insert the CP and UFMC has a low spectral side-lobe level.

Both [4] and [5] have analyzed the impacts of carrier frequency offset (CFO) on UFMC systems. Furthermore, a comparison with other candidate multi-carrier waveforms such as generalized frequency division multiplexing (GFDM), FBMC and circular FBMC, was also provided in [5]. A leakage-based filter optimization algorithm was developed in [4] and [6] to optimize the finite impulse response (FIR) filter design, which enabled UFMC systems to be insensitive to residual CFOs and timing offsets (TOs). Considering the presence of both the TOs and CFOs, a concept called autonomous timing advance (ATA) was introduced in [7] to enhance the overall performance of UFMC systems. Instead of using Dolph-Chebyshev filter, a Bohman filter based pulse shaping is exploited to improve the robustness against CFOs for UFMC [8]. In above-mentioned approaches, it is only assumed that the user of interest (UoI) is perfectly synchronized and other users are impaired with the CFOs or TOs at system receivers. But they do not consider the scenario where all users are not aligned in terms of CFOs.

To the best of our knowledge, so far, there are no CFO estimation approaches developed for UFMC systems. Since UFMC is very close in nature to OFDM, the frequency offset may be estimated in UFMC systems by referring to CFO estimation approaches of OFDM systems. In general, the CFO estimators devised for OFDM systems mainly consist of two categories: the data-aided (DA) scheme [9]–[12] and non-data-aided (NDA) scheme [13], [14]. A prominent drawback of NDA estimation schemes is high computational complexity. For CFO estimators [9], [10] based on the pseudo-random noise (PN) sequence, the lower estimation accuracy is obtained at very low signal-to-noise ratios (SNRs). Using two training symbols, the CFO estimation methods [11], [12] are developed to only estimate the fractional CFOs. Besides, their complexity is largely dependent upon the frequency searching resolution.

Motivated by the aforementioned observations, an efficient hybrid CFO estimation algorithm is proposed for UFMC-based systems, in order to eliminate the frequency offsets accurately. Thanks to the absence of the CP, unlike OFDM symbols, each UFMC symbol cannot be individually handled through using the least square (LS) approach to recover system carrier frequency under the condition of the multipath fading channel. With the aid of two identical consecutive training symbols, the CFO estimator based on the LS criterion is developed in detail for UFMC systems. Meanwhile, this estimator with the large frequency searching step is also utilized to roughly estimate the frequency offsets. Finally, the residual CFO is removed through the auto-correlation calculations of the compensated signals. This method not only achieves good estimation accuracy, but also owns large estimation range. The following numerical simulations confirm the practicability of the proposed estimation algorithm.

II. System Model

A block diagram of a common UFMC-based system [6] is depicted in Fig. 1. In this figure, "Freq. Synch." denotes that both the CFO estimation and compensation are completed, and "FDE" stands for the frequency domain equalization. Besides, T_s and f_c are the system sampling period and carrier frequency, respectively. Obviously in Fig. 1, a sub-filtering operation is performed on a group of sub-carriers in UFMC systems.

Let N denote the total number of all sub-carriers in this system, N_{SC} the number of all used sub-carriers and N_Q the number of consecutive sub-carriers contained in each sub-band. As seen from Fig. 1, by summing all filtered sub-band signals, the baseband transmitted signal on the nth sample of

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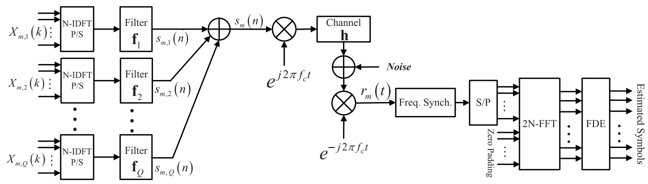

Fig. 1. Block diagram of a common UFMC-based system.

the mth UFMC symbol in this system is expressed as

$$s_m(n) = \sum_{i=1}^{Q} s_{m,i}(n) = \sum_{i=1}^{Q} \sum_{l=0}^{L_F-1} f_i(l) x_{m,i}(n-l) \tag{1}$$

where $\{f_i(0), f_i(1), \cdots, f_i(L_F-1)\}$ denote tap coefficients of the ith sub-band filter, L_F is the length of this filter, and Q is the number of the divided sub-bands. Furthermore, we assume that $\sum_{l=0}^{L_F-1} |f_i(l)|^2 = 1$.

In (1), $x_{m,i}(n)$ can be further written as

$$x_{m,i}(n) = \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} X_{m,i}(k) \cdot \exp\left(\frac{j2\pi kn}{N}\right) \tag{2}$$

where $X_{m,i}(k) = X_m(k)$ if $k \in \mathbb{Q}_i$; otherwise $X_{m,i}(k) = 0$. Here, $\mathbb{Q}_i$ is an integer set consisting of all sub-carrier indexes that belong to the ith sub-band and $X_m(k)$ is the modulated symbol on the kth sub-carrier of the mth UFMC symbol.

After passing through a multipath Rayleigh fading channel, the corresponding discrete-time baseband received signal in this system can be given by

$$r_m(n) = \exp\left[\frac{j2\pi f_d(mN_t+n)}{N}\right] \cdot [h(n) \otimes s_m(n)] + \omega_m(n), \quad n = 0, 1, \ldots, N_t - 1 \tag{3}$$

where $\otimes$ represents the linear convolution, $N_t = N + L_F - 1$, f_d denotes a normalized CFO which is the ratio of the CFO to the sub-carrier spacing of this system, $\omega_m(n)$ is a complex-valued additive Gaussian white noise (AWGN) with zero mean and a variance of σ_ω^2, $\{h(0), h(1), \cdots, h(L_h-1)\}$ represent the coefficients of the channel impulse response (CIR), and L_h is the length of the CIR. Additionally, the real part of $\omega_m(n)$ is completely independent from its imaginary part.

III. Proposed CFO Synchronization Scheme

Owing to the absence of the CP, these exists an unremovable inter block interference (IBI) between two adjacent UFMC symbols under the multipath fading channel. Consequently, the LS approach is not directly applied to each UFMC symbol for estimating the frequency offsets or channel impulse responses. To avoid the effects of the IBI on the frequency offset estimation, a robust CFO estimation scheme depending on two identical consecutive training symbols is proposed for UFMC systems in this section. Moreover, its principle is also analyzed in detail as follows.

Assume that the mth modulated symbol vector is denoted as $\mathbf{X}_m$, i.e., $\mathbf{X}_m = [X_m(0), X_m(1), \cdots, X_m(N_{SC}-1)]^T$. Furthermore, both $\mathbf{X}_0$ and $\mathbf{X}_1$ are two known training symbols and $\mathbf{X}_0 = \mathbf{X}_1$. Depending on (1) and (2), we can obtain $s_0(n) = s_1(n)$, $n = 0, 1, \cdots, N_t - 1$. Therefore, the equivalent replacement of the training sequences is shown as

$$\begin{gathered}\{s_0(N_t-L), \cdots, s_0(N_t-1), s_1(0), \cdots, s_1(N_t-1)\} \\ \Updownarrow \\ \{s_1(N_t-L), \cdots, s_1(N_t-1), s_1(0), \cdots, s_1(N_t-1)\}\end{gathered} \tag{4}$$

where L is a positive integer. When L is more than L_h, the effect of the 0th UFMC symbol on the 1st UFMC symbol is converted into the self-interference of the 1st UFMC symbol.

According to (3) and (4), the received samples of the 1st UFMC symbol are rewritten as in a compact matrix format

$$\mathbf{R}_1 = \exp\left(\frac{j2\pi f_d N_t}{N}\right) \cdot \mathbf{EH} \underbrace{\mathbf{F}^H \mathbf{D} \mathbf{X}_1}_{\mathbf{S}_1} + \mathbf{W}_1 \tag{5}$$

where $\mathbf{S}_1 = [s_1(0), s_1(1), s_1(2), \cdots, s_1(N_t-1)]^T$, $\mathbf{W}_1 = [\omega_1(0), \omega_1(1), \cdots, \omega_1(N_t-1)]^T$, $\mathbf{E}$ is a diagonal matrix, i.e., $\mathbf{E} = \text{diag}\left\{1, \exp\left(\frac{j2\pi f_d}{N}\right), \cdots, \exp\left(\frac{j2\pi f_d(N_t-1)}{N}\right)\right\}$, $\mathbf{H}$ is an $N_t \times N_t$ circular shift matrix with the diagonal value being $h(0)$ and the elements in the first row of $\mathbf{H}$ are equal to $[h(0), 0, \cdots, h(L_h-1), \cdots, h(1)]$, $\mathbf{D}$ and $\mathbf{F}$ are the Fourier transform matrix and the FIR filter matrix, respectively, i.e.,

$$\mathbf{D} = \text{diag}\left\{\widetilde{\mathbf{D}}_1, \widetilde{\mathbf{D}}_2, \cdots, \widetilde{\mathbf{D}}_Q\right\} \tag{6}$$

$$\mathbf{F} = \left[\overline{\mathbf{F}}_1^H, \overline{\mathbf{F}}_2^H, \cdots, \overline{\mathbf{F}}_Q^H\right]^H. \tag{7}$$

In (6), $\widetilde{\mathbf{D}}_m = \left[\mathbf{d}_{(m-1)N_Q+1}, \mathbf{d}_{(m-1)N_Q+2}, \cdots, \mathbf{d}_{mN_Q}\right]$, where $\mathbf{d}_k$ is the kth column vector of $\mathbf{D}_0$. Note that $\mathbf{D}_0$ is an $N \times N$ inverse discrete Fourier transform (IDFT) matrix, of which the (k,l)th entry is given by $\mathbf{D}_0(k,l) = \exp\left(j\frac{2\pi kl}{N}\right)$, $k, l = 0, 1, \cdots, N-1$. In (7), $\overline{\mathbf{F}}_m$ is an $N \times N_t$ Toeplitz matrix with the elements in the first row being $\left[f_m^*(0), f_m^*(1), \cdots, f_m^*(L_F-1), \cdots, 0\right]$.

A. Frequency Estimation Based on LS Criterion

In many practical scenarios, system receivers cannot commonly attain the perfect knowledge of wireless channel in advance, that is, the CIR $\mathbf{h}$ is unknown. Hence, different from (5), another matrix format of received signals is derived as

$$\mathbf{R}_1 = \overline{\mathbf{E}} \mathbf{S}^1 \mathbf{h} + \mathbf{W}_1 \tag{8}$$

where $\overline{\mathbf{E}} = \exp\left(\frac{j2\pi f_d N_t}{N}\right) \cdot \mathbf{E}$, $\mathbf{h} = [h(0), h(1), \cdots, h(L_h-1)]^T$, and $\mathbf{S}^1$ is further expressed as

$$\mathbf{S}^1 = \begin{bmatrix} s_1(0) & s_1(N_t-1) & \cdots & s_1(N_t-L_h+1) \\ s_1(1) & s_1(0) & \cdots & s_1(N_t-L_h+2) \\ \vdots & \vdots & \ddots & \vdots \\ s_1(N_t-1) & s_1(N_t-2) & \cdots & s_1(N_t-L_h) \end{bmatrix}. \tag{9}$$

In multipath fading channels, the uncertainty of $\mathbf{h}$ has a great impact on the CFO estimation. Accordingly, the LS criterion is directly applied to (8) for estimating the residual CFO, i.e.,

$$\langle f_A, \hat{\mathbf{h}} \rangle = \arg\min_{f_d, \mathbf{h}} \Gamma(f_d, \mathbf{h}) = \arg\min_{f_d, \mathbf{h}} \| \mathbf{R}_1 - \overline{\mathbf{E}} \mathbf{S}^1 \mathbf{h} \|^2 \tag{10}$$

where f_A and $\hat{\mathbf{h}}$ are the estimations of f_d and $\mathbf{h}$, respectively.

From (10), it is observed that there are two unknown parameters in $\Gamma(f_d, \mathbf{h})$, namely f_d and $\mathbf{h}$. We can assume that f_d keeps unchanged and $\mathbf{h}$ is a variable vector firstly. Therefore, letting $\frac{\partial \Gamma(f_d, \mathbf{h})}{\partial \mathbf{h}} = 0$, $\hat{\mathbf{h}}$ can be derived as

$$\hat{\mathbf{h}} = \left[\left(\mathbf{S}^1\right)^H \mathbf{S}^1\right]^{-1} \left(\mathbf{S}^1\right)^H \overline{\mathbf{E}}^H \mathbf{R}_1 \tag{11}$$

By combining (10) with (11), $\Gamma(f_d, \hat{\mathbf{h}})$ can be expressed as

$$\Gamma(f_d, \hat{\mathbf{h}}) = \mathbf{R}_1^H\mathbf{R}_1 - \mathbf{R}_1^H\overline{\mathbf{E}}\mathbf{S}^1\left[\left(\mathbf{S}^1\right)^H\mathbf{S}^1\right]^{-1}\left(\mathbf{S}^1\right)^H\overline{\mathbf{E}}^H\mathbf{R}_1. \tag{12}$$

For each trial value of the CFO f_d in (12), since $\mathbf{R}_1^H\mathbf{R}_1$ is a constant, the CFO estimation f_A is obtained as

$$\begin{aligned} f_A &= \arg\min_{f_d}\Gamma(f_d, \hat{\mathbf{h}}) = \arg\max_{f_d}\Lambda(f_d) \\ &= \arg\max_{f_d}\mathbf{R}_1^H\overline{\mathbf{E}}\mathbf{S}^1\left[\left(\mathbf{S}^1\right)^H\mathbf{S}^1\right]^{-1}\left(\mathbf{S}^1\right)^H\overline{\mathbf{E}}^H\mathbf{R}_1, \end{aligned} \tag{13}$$

where $\Lambda(f_d)$ is only a function of f_d under the condition in which $\mathbf{S}^1$ is a known transmitted signal related to $\mathbf{S}_1$ in (9). Therefore, with using the frequency grid searching method, f_A can be solved by maximizing $\Lambda(f_d)$ in (13).

This LS based estimation algorithm has larger frequency estimation range compared with other algorithms [9]–[12]. Moreover, the smaller the frequency searching step is in (13), the higher the estimation accuracy will become. In such a case, however, the computational complexity of this algorithm will greatly increase. As a result, it needs a reasonable compromise for this approach between the computational complexity and the frequency estimation precision.

B. Fine Frequency Estimation

Since two identical consecutive training symbols are used in this system, the auto-correlation operation of received signals is performed to estimate the residual CFO precisely. Relying on (3), the fine frequency estimation is expressed as

$$f_B = \frac{N}{2\pi N_t}\arg\left[\sum_{k=L_1}^{N+L_F-2} r_0^*(k)\cdot r_1(k)\right]. \tag{14}$$

Because $s_0(n) = s_1(n)$, we have $\sum_{l=0}^{L_h-1} h(l)s_0(k-l) = \sum_{l=0}^{L_h-1} h(l)s_1(k-l)$ if $k > L_h$. Consequently in (14), when $L_1 > L_h$, the detailed expression of $\sum_{k=L_1}^{N+L_F-2}\left[r_0^*(k)\cdot r_1(k)\right]$ is shown in (15) and (16), as shown at the bottom of this page.

From (14), it is clear that the fine frequency estimation range is $\left(-\frac{N}{2N_t}, \frac{N}{2N_t}\right]$. Since $N_t > N$, $\frac{N}{2N_t}$ is less than 0.5. Hence, the estimation range attained by using (14) is relatively small. But this estimation method can achieve high estimation precision.

C. Novel Hybrid Frequency Estimation Algorithm

According to the analysis of Section III-A, we know that the CFO estimator based on LS approach has large estimation range. Moreover, it also has better estimation accuracy when adopting a smaller frequency searching step, whereas its complexity dramatically increases. As seen from Section III-B, the estimator exploiting (14) is able to achieve better estimation performance. But its frequency estimation range is correspondingly small.

Based on the above analyses, a novel hybrid frequency synchronization scheme is proposed, which combines the utilization of (14) with the LS based estimation algorithm. Firstly, the coarse CFO estimation is acquired by employing the LS based estimator with a large frequency searching step. And then, (14) is used to obtain the fine CFO estimation after correcting received signals. This scheme owns the advantages of larger frequency estimation range, better estimation accuracy, and moderate complexity.

The specifical hybrid frequency offset estimation procedure is summarized in **Algorithm 1**. In this algorithm, note that λ is a large frequency searching step and $[F_{min}, F_{max}]$ denotes a frequency searching range. In addition, since the fine frequency estimation range is $\left(-\frac{N}{2N_t}, \frac{N}{2N_t}\right]$, the selection of λ must satisfy $0 < \lambda < \frac{N}{N_t}$.

D. Complexity Analysis

In this devised scheme, it needs to calculate the inverse matrix of $\left(\mathbf{S}^1\right)^H\mathbf{S}^1$ in (13). As seen from (9), it is evident that $\mathbf{S}^1$ is an $N_t \times L_h$ matrix. Hence, $\left(\mathbf{S}^1\right)^H\mathbf{S}^1$ is an $L_h \times L_h$ matrix, which is not related with N and L_F. Moreover, L_h is usually much less than N. The inverse matrix of $\left(\mathbf{S}^1\right)^H\mathbf{S}^1$ can be attained by using the LU decomposition approach. Accordingly, when L_h is small and a comparatively larger frequency searching step is used, the computational complexity of solving (13) for the coarse CFO estimation is acceptable. Besides, a relatively lower complexity is required to calculate (14). Therefore, the proposed algorithm has an acceptable complexity.

IV. Simulation Results

In this section, the CFO estimation mean square error (MSE) performance of the proposed method is evaluated with computer simulations. In UFMC-based systems, simulation conditions are set as follows: the QPSK modulation

$$\begin{aligned} \sum_{k=L_1}^{N+L_F-2} r_0^*(k)r_1(k) &= \sum_{k=L_1}^{N+L_F-2}\left[\sum_{l=0}^{L_h-1} h(l)s_0(k-l)\exp\left(j\frac{2\pi f_d k}{N}\right)\right]^*\cdot\left[\sum_{l=0}^{L_h-1} h(l)s_1(k-l)\exp\left(j\frac{2\pi f_d(N_t+k)}{N}\right)\right] + N_0 \\ &= \left(\sum_{k=L_1}^{N+L_F-2}\left|\sum_{l=0}^{L_h-1} h(l)s_0(k-l)\right|^2\right)\exp\left(j\frac{2\pi f_d N_t}{N}\right) + N_0 \end{aligned} \tag{15}$$

$$\begin{aligned} N_0 = &\sum_{k=L_1}^{N+L_F-2}\omega_1(k)\left[\sum_{l=0}^{L_h-1} h(l)s_0(k-l)\exp\left(j\frac{2\pi f_d k}{N}\right)\right]^* + \sum_{k=L_1}^{N+L_F-2}\omega_0^*(k)\left[\sum_{l=0}^{L_h-1} h(l)s_1(k-l)\exp\left(j\frac{2\pi f_d(N_t+k)}{N}\right)\right] \\ &+ \sum_{k=L_1}^{N+L_F-2}\omega_0^*(k)\omega_1(k) \end{aligned} \tag{16}$$

Algorithm 1 Novel Hybrid CFO Synchronization Scheme

Initialization:

Initialize $\lambda > 0$, F_{min}, F_{max}, and $L_1 > 0$;
Set $f_d^1 = F_{min}$, $f_d^{opt} = 0$, $f_A = 0$, $f_B = 0$ and $\Lambda_{max} = 0$;

Iteration:

while $f_d^1 \le F_{max}$ **do**
  Calculate the cost function $\Lambda(f_d^1)$ according to (13);
  if $\Lambda_{max} \ge \Lambda(f_d^1)$ **then**
    f_A remains unchanged;
  else
    $\Lambda_{max} = \Lambda(f_d^1)$;
    $f_A = f_d^1$;
  end if
  Let $f_d^1 = f_d^1 + \lambda$;
end while
Compensate the coarse CFO estimation f_A, i.e., $r_m^{'}(k) = r_m(k) \cdot \exp\left(-\frac{j2\pi f_A(mN_t+k)}{N}\right)$, $m = 0, 1$;
Calculate $\sum_{k=L_1}^{N+L_F-2} \left[r_0^{'}(k)\right]^* \cdot r_1^{'}(k)$ through (15) and (16);
Obtain the fine CFO estimation f_B by employing (14);
Calculate the final CFO estimation $f_d^{opt} = f_A + f_B$;

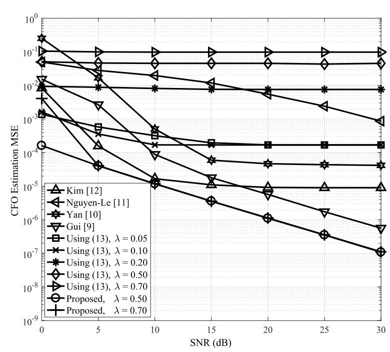

Fig. 2. Performance comparison of different CFO estimators versus SNR when $f_d = 0.213$.

scheme is used; the Dolph-Chebyshev filter is employed to filter each sub-band signal; $N = 128$, $N_{SC} = 108$, $N_Q = 12$, $L_F = 30$, $L_h = 8$, and $L_1 > L_h$. The modulated signals are transmitted through a multipath Rayleigh fading channel which has path delays of $l = 0, 1, 2, \cdots, L_h - 1$ samples and an exponential power delay profile written as $\frac{\exp(\frac{-l}{L_h})}{\sum_{p=0}^{L_h-1} \exp(\frac{-p}{L_h})}$ as used in [12].

With keeping the same transmitted data rate, Figs. 2 and 3 compare the performance of the proposed scheme to that of the existing schemes at $f_d = 0.213$ and $f_d = 2.33$, respectively. From Fig. 2, it is clear that this novel synchronization algorithm with $\lambda = 0.7$ achieves smaller MSE than available algorithms [9]–[12] when the SNR is larger than 1.5 dB or so under the condition of the small CFO. Evidently in the presence of large CFOs, Fig. 3 indicates that the performance of the proposed approach with $\lambda = 0.5$ or $\lambda = 0.7$ dramatically outperforms that of other existing approaches. As seen from both figures, there exists an estimation performance plateau for the LS-based CFO estimator (i.e., utilizing (13)) at high SNRs. Furthermore, its MSE performance gradually deteriorates with the increase of λ. These results illustrate the effectiveness of this devised scheme.

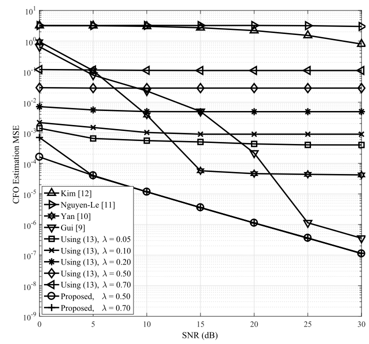

Fig. 3. Performance comparison of different CFO estimators versus SNR when $f_d = 2.330$.

V. CONCLUSIONS

For UFMC-based systems, a novel CFO estimation approach has been proposed over the multipath Rayleigh fading channel. Depending on two identical training symbols, a LS based frequency estimator is developed for estimating the frequency offset coarsely. After compensating this coarse CFO estimation, the auto-correlation calculation of corrected signals is performed to achieve the fine frequency offset estimation. This approach incorporates the advantages of both the LS-based frequency estimator and another frequency estimator only using (14). Simulations and comparisons are presented to verify the feasibility of the proposed approach.

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