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Joint frequency-phase estimation for pilot-limited communication systems: a novel method based on length-variable auto-correlation operator

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Dear editor,

Wireless communication systems, such as the fifth-generation (5G), unmanned aircraft vehicle, have widespread applications in our modern society [1–4]. This case will significantly decline available spectrum resources, thereby requiring us to provide some solutions. This study proposes a novel method for providing joint frequency-phase estimation for pilot-limited communication systems.

Traditional frequency-phase estimation (TFPE) method often first estimates the frequency offset, then estimates the phase offset post frequency offset compensation. So this TFPE method can be regarded as a "serial" estimation mode. However, this mode will suffer from the following fact: the performance of the frequency offset estimation directly influences that of the subsequent phase offset estimation [5,6]. Generally, the usage of a pilot sequence will directly determine the overall performance of the TFPE method, particularly for pilot-limited communication systems. For the TFPE method, the pilot-aided frequency offset estimation cannot achieve a good performance, thereby resulting in a serious impact on the performance of the coming phase offset estimation [7]. So far, no effective solution to the aforementioned problem has yet been provided. Thus, we propose a new frequency-phase estimation (NFPE) method to implement the decoupling of the frequency and phase offset estimations such that this method can be viewed as a "paralle" estimation mode.

System model. Figure 1 depicts the system model used in this study. First, a data stream along with a pilot stream is converted by a modulator (MOD) into two groups of complex baseband signals d with M and p with L symbols, respectively. Next, the signals d and p are spliced together via a multiplexer (MUX) to generate a popular "preamble" frame structure [8]. The multiplexed signals are then rotated by Doppler shift f_{d} and phase offset $\theta \in [-\pi, \pi)$, further passing through the additional white Gaussian noise (AWGN) channel. Single carrier transmission with an ideal symbol timing based on the classical interpolation idea [9] is assumed, such that the k-th received signal can be expressed as

$$r(k) = s(k) \exp\left[j\left(2\pi f_{\mathrm{d}} T_{\mathrm{s}} k + \theta\right)\right] + n(k), \quad (1)$$

where $k \in \{1, 2, \ldots, L, L+1, L+2, \ldots, L+M\}$, $s(k)$ is the normalized-energy signal; T_{s} is the symbol duration; and $n(k)$ is the circular symmetric complex Gaussian random variable, whose real and imaginary parts have variances $N_0/2$, $\kappa_p = \{1, 2, \ldots, L\}$ denotes the index set consisting of the whole pilot symbols.

The received pilot signals at the receiver are extracted via a de-multiplexer from the received

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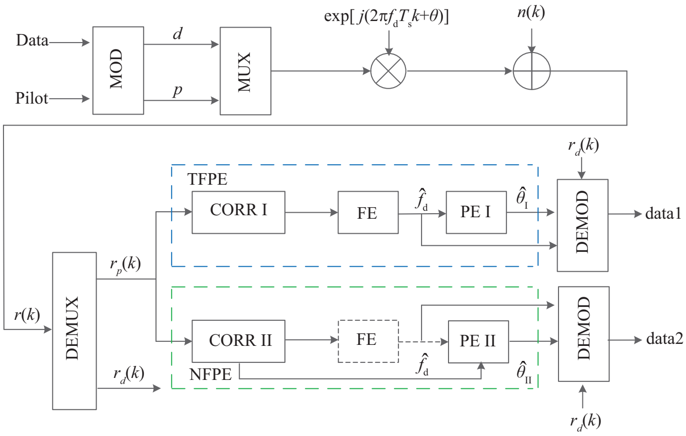

Figure 1 (Color online) System model.

signals to perform the TFPE and NFPE. Both methods comprise an auto-correlator (CORR), frequency estimator (FE), and phase estimator (PE) denoted by CORR I/II, FE, and PE I/II, respectively. In the end, the received signals are compensated and sent into a demodulator to recover the original data information.

The removed-modulation pilot signal for the pilot-aided communication systems must be obtained based on (1),

$$z(k) \triangleq r(k)s(k)^* = \exp[j(2\pi f_{\rm d} T_{\rm s} k+\theta)]+v(k), \quad (2)$$

where $v(k) \triangleq n(k)\tilde{s}(k)^*$ is also the Gaussian noise.

Detailed descriptions of TFPE and NFPE. Since residual frequency offset can be equivalent to being a time-varying phase post frequency offset compensation, the frequency offset estimation will directly influence the performance of the phase offset estimation in the TFPE method. Thus, we propose an efficient solution called the NFPE method. The implementation of both TFPE and NFPE methods is specifically presented as follows.

• The TFPE method. CORR I. Based on (2), the auto-correlation operator is performed as follows:

$$\begin{aligned} R_{\rm I}(\alpha) &\triangleq \frac{1}{L-\alpha}\sum_{k=1}^{L-\alpha} z(k)^* z(k+\alpha) \\ &= \exp(j2\pi f_{\rm d} T_{\rm s}\alpha)+V_{\rm I}(\alpha), \end{aligned} \quad (3)$$

where $\alpha \in [1, (L+1)/2]$ is the correlation delay length for adjusting the accuracy and range (as well as complexity) of the frequency offset estimation, and $V_{\rm I}(\alpha)$ is the cumulative noise.

FE. The frequency offset is estimated as follows from (3):

$$\hat{f}_{\rm d} = \frac{1}{2\pi\alpha T_{\rm s}}\arg\{R_{\rm I}(\alpha)\}, \quad |\hat{f}_{\rm d}| < 1/(2\alpha T_{\rm s}), \quad (4)$$

where $f_{\rm d,threshold} \triangleq \pm 1/(2\alpha T_{\rm s})$ is called the frequency threshold for the frequency offset estimation.

PE I. Post frequency offset compensation, the maximum likelihood (ML) phase offset estimation will be performed with the form:

$$\hat{\theta}_{\rm I} = \arg\left\{\sum_{k=1}^{L} z(k)\exp\left(j2\pi\hat{f}_{\rm d}T_{\rm s}k\right)^*\right\}. \quad (5)$$

The accuracy of the phase offset estimation depends on that of the frequency offset estimation in (4).

• The NFPE method. CORR II. Based on (2), the auto-correlation operator is utilized as follows:

$$\begin{aligned} R_{\rm II}(\alpha) &\triangleq \frac{1}{L'-\alpha}\sum_{k=1}^{L'-\alpha} z(k)^* z(k+\alpha) \\ &= \exp(j2\pi f_{\rm d} T_{\rm s}\alpha)+V_{\rm II}(\alpha), \end{aligned} \quad (6)$$

where $L' \in (L+1/2, L]$ is the variable correlation length for the adjustment of complexity and accuracy of the auto-correlation operator, and $V_{\rm II}(\alpha)$ is the cumulative noise.

FE. If necessary, the frequency offset estimation is executed according to (4) with the substitute of $R_{\rm I}(\alpha)$ by $R_{\rm II}(\alpha)$ with $L' = L$ in (6).

PE II. Without the frequency offset compensation and letting $\alpha = (L+1)/2$ (defined as α') in (6), the ML-type phase offset estimation is developed as follows:

$$\begin{cases} \hat{\theta}_{\rm II} = \arg\left\{\sum_{k=1}^{L} z(k)R_{\rm II}(\alpha')^*\right\}, \\ R_{\rm II}(\alpha') = \exp\left(j2\pi f_{\rm d}T_{\rm s}\cdot\dfrac{L+1}{2}\right)+V_{\rm II}\left(\dfrac{L+1}{2}\right). \end{cases} \quad (7)$$

Intuitively, the phase offset estimation in (7) is independent of the frequency offset estimation.

For the TFPE method, substituting (2) into (5) can yield the following:

$$\begin{aligned}\hat{\theta}_{\mathrm{I}} &= \arg\left\{\exp(j\theta)\sum_{k=1}^{L}\exp\left[j2\pi(f_{\mathrm{d}}-\hat{f}_{\mathrm{d}})T_{\mathrm{s}}k\right]+V_{\mathrm{I}}\right\}\\ &\approx \arg\left\{\exp(j\theta)\sum_{k=1}^{L}\exp(j2\pi\Delta f_{\mathrm{d}}T_{\mathrm{s}}k)\right\}\\ &= \begin{cases}\theta, & |\Delta f_{\mathrm{d}}|=0,\\ \theta+\phi_{\mathrm{I}}(\Delta f_{\mathrm{d}},L), & |\Delta f_{\mathrm{d}}|\neq 0.\end{cases}\end{aligned} \tag{8}$$

For NFPE method, considering (2) into (7) can obtain the following:

$$\begin{aligned}\hat{\theta}_{\mathrm{II}} &= \arg\left\{\exp(j\theta)\sum_{k=-(L-1)/2}^{(L-1)/2}\exp(j2\pi f_{\mathrm{d}}T_{\mathrm{s}}k)+V_{\mathrm{II}}\right\}\\ &\approx \arg\left\{\exp(j\theta)\frac{\sin\left(\pi f_{\mathrm{d}}T_{\mathrm{s}}(L+1)\right)}{\sin(\pi f_{\mathrm{d}}T_{\mathrm{s}})}\right\}\\ &= \begin{cases}\theta, & |f_{\mathrm{d}}|<1/(L+1)\,T_{\mathrm{s}},\\ \theta+\phi_{\mathrm{II}}(f_{\mathrm{d}},L), & |f_{\mathrm{d}}|\geqslant 1/(L+1)\,T_{\mathrm{s}}.\end{cases}\end{aligned} \tag{9}$$

In Eqs. (8) and (9), $\Delta f_{\mathrm{d}} \triangleq f_{\mathrm{d}} - \hat{f}_{\mathrm{d}}$ is the residual frequency offset and $\phi_{\mathrm{I}}(\Delta f_{\mathrm{d}}, L)$ and $\phi_{\mathrm{II}}(f_{\mathrm{d}}, L)$ are the phase ambiguity functions of Δf_{d} or f_{d} and L. $f'_{\mathrm{d,threshold}} \triangleq \pm 1/(L+1)T_{\mathrm{s}}$ is also called the frequency threshold for the phase offset estimation. Note that $f'_{\mathrm{d,threshold}}$ is equal to $f_{\mathrm{d,threshold}}$ when $\alpha = (L+1)/2$ in (4), which contributes to the simulation analysis. Both V_{I} and V_{II} are cumulative noises.

Furthermore, on the comparison of (8) and (9), the former can achieve good performance if and only if the residual frequency offset is equal to or close to zero; otherwise, phase ambiguity phenomenon will occur. The latter can also obtain a good performance provided that the frequency offset is less than $|f'_{\mathrm{d,threshold}}|$. Frequency offset pre-estimation should be considered once beyond the frequency threshold.

So far, we can conclude that in our proposed NFPE method, the frequency-phase estimation can be decoupled, confirming that this method can be viewed as a "parallel" estimation.

Conclusion. We consider a pilot-limited communication system and propose a pilotless-aided NFPE method that includes a length-variable auto-correlation operator, alternative FE, and a PE performed by adding over all or partial results of the entire removed-modulation pilot signals multiplying by a conjugate form of the correlation value from the midpoint of sampling instants. The numerical simulation results provided in Appendix A show that, when the frequency offset is less than any one of two thresholds for the frequency-phase estimation, both the methods can achieve a good performance, but the NFPE method will not require the frequency offset estimation. Furthermore, the frequency offset estimation should be considered first when the existing frequency offset is more than the threshold for the phase estimation, and once the variable correlation lengths are more than the correlation delay lengths, the proposed NFPE method can achieve a better performance than the TFPE method. This implies that the proposed NFPE method will be more suitable for high-order modulation systems.

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Supporting information Appendix A. The supporting information is available online at info.scichina.com and link.springer.com. The supporting materials are published as submitted, without typesetting or editing. The responsibility for scientific accuracy and content remains entirely with the authors.

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