A Synchronization Algorithm for Burst-Mode MR-OFDM System of 802.15.4g

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Abstract. This paper presents a refined coarse frame synchronization algorithm for long repeated-sequence based preamble STF (short training field) in burst mode MR-OFDM system of 802.15.4g which is of low decision delay and resistant to timing uncertainty due to metric plateau. The precision of coarse timing and frequency offset estimation is improved. A fine symbol timing and frequency offset estimation algorithm based on LTF (long training field) is also proposed which provides an overall solution for the synchronization of burst mode MR-OFDM system. Simulation results are given to provide reference for implementation.

1. Introduction

Synchronization for OFDM system has been widely discussed in the literature[1][2][3]. Generally some specially designed preamble/training data is inserted within continuous or burst OFDM transmission system[4][5], which are usually repeated symbol sections of certain length to make it robust to initial synchronization for large frequency offset between transmitter and receiver. Available solutions are usually suitable for initial acquisition for moderate SNR range[1][2][3][6], which use one OFDM symbol’s training data.

If the acquisition SNR range is expected to be wide from below-zero dB to as high as 30dB with required timing and frequency synchronization accuracy, the preamble/training data should be longer to be several OFDM symbols as in MR-OFDM of 802.15.4g[4]. The burst transmission mode makes a more demanding task than continuous mode transmission since it’s not tolerable for large delay which means large storage requirement for the synchronization decision and it’s mandatory for one-shot acquisition with enough accuracy and acquisition ratio while continuous transmission mode usually relies on more times of acquisition to make a final expected acquisition ratio and accuracy. Unfortunately, this burst mode acquisition and synchronization targeting for wide SNR range has hardly been addressed.

This paper addresses the above issue under the framework of MR-OFDM of 1.2M bandwidth in 802.15.4g. A compound coarse timing metric is presented to accomplish the refined coarse synchronization with improved accuracy for wide SNR range which is of low decision delay and resistant to timing uncertainty due to metric plateau, and fine synchronization is also provided to make a complete solution for the MR-OFDM like burst mode OFDM system.
Section 2 describes the system model, section 3 presents the refined coarse synchronization algorithm and procedures, fine synchronization is presented in section 4, simulation and evaluation of the performance is provided in section 5. Finally, conclusion is provided in section 6.

2. Signal Model and Available Synchronization Methods
The burst mode OFDM signal for MR-OFDM of 1.2M bandwidth consists of SHR (synchronization header), PHR (physical header) and PHY payload, where the SHR consists of STF and LTF which is the preamble for packet acquisition, coarse and fine synchronization as well as channel estimation.

The STF consists of 4 OFDM symbols (5 times FFT size=5*128 in baseband samples, or 5*96us) which are 40 repeated subsections with one subsection of 1/8 FFT size and the last 4 subsections are negative basic subsection signals. The LTF consists of 2.5 repeated signal sections with one section of FFT size and the first 0.5 section serving as the CP[4]. Fig.1 shows the STF and LTF detail for MR-OFDM of 1.2M bandwidth.

Figure 1. STF and LTF structure in MR-OFDM of 1.2MHz bandwidth

The time domain STF signal is generated as in [4] by IFFT from the frequency domain STF BPSK modulation, with N of the FFT size. The time domain LTF signal is generated from frequency domain LTF BPSK signals. The time domain STF signal satisfies the following:

\[ x(n-l) = x(n-l-N_D) = x(n-l-2N_D) = ... = x(n-l-35N_D), l = 0,1,...,N_D-1 \] (1)

and

\[ x(n-l) = -x(n+l) = -x(n+l+N_D) = -x(n+l+2N_D) = -x(n+l+3N_D) \] (2)

for \( n \) at the last sample of the non-negated STF subsections, and \( N_D = N/8 \), with \( N \) of the FFT size.

With the wireless channel modelled as discrete quasi-static multipath Rayleigh faded FIR in which the max delay corresponding to the max channel taps number \( L \), the received BB signal can be expressed as following with relative frequency offset \( \varepsilon \) (relative to the subcarrier spacing \( \Delta f \) ) between Tx and Rx and corrupted by white complex Gaussian noise \( w(n)\sim\mathcal{CN}(0,\sigma^2) \):

\[ y(n) = \exp\left(j2\pi\varepsilon n/N\right)\sum_{l=0}^{L-1}h(l)x(n-l)+w(n) \] (3)

It’s obvious that the repeated structure keeps between subsections of length \( N_D = N/8 \) before the last 4 subsections of total length \( N/2 \) even the multipath length greater than \( N_D \). This repeated structure of STF signal can be used for combined frame timing and frequency offset as follows[1][2][3]:

\[ A(n) = \sum_{l=0}^{N_D-1} y(n-l)y^\ast(n-l-N_D), n = n_0+2N_D-1, n_0+2N_D-1+1,... \] (4)

\[ B(n) = \sum_{l=0}^{N_D-1} |y(n-l)|^2, n = n_0+2N_D-1, n_0+2N_D-1+1,... \] (5)

where \( n_0 \) is the first sample index of the searching processing. Of course the above calculation can be done in an iterative manner to save hardware and power.

And for improving acquisition/synchronization performance, the auto-correlation and subsections’ energy are combined on M subsections where the total used STF samples is \((M+1)N_D\) as following:

\[ A_0(n) = \sum_{m=0}^{M} A(mN_D+n), n = n_0+2N_D+(M-1)N_D-1, n_0+2N_D+(M-1)N_D-1+1,... \] (6)

\[ B_0(n) = \sum_{m=0}^{M} B(mN_D+n), n = n_0+2N_D+(M-1)N_D-1, n_0+2N_D+(M-1)N_D-1+1,... \] (7)

where \( M \) is determined from the available repeated subsections before the beginning of the negated STF field with the coarse frame timing target set to the beginning of the negated part. And a synchronization metric can be calculated as follows[1][2][3] for every sliding sample index \( n \) satisfying \( n \geq n_0+2N_D+(M-1)N_D-1 \).
\[ \rho(n) = |A_s(n)|^2 / (B_s(n) \star B_s(n)) \]  

(8)

For burst mode transmission, the synchronization metric is usually compared to a predefined threshold \( \rho_t \) [7][8] which is related to the expected minimum SNR during the sliding procedure as following:

if \( \rho(n) > \rho_t \),

frame detected and \( n_f = n \) is the coarse frame timing

(9)

And coarse relative frequency offset \( \varepsilon \) (to the subcarrier spacing \( \Delta f \)) can be estimated based on the coarse frame timing as

\[ \hat{\varepsilon} = 8 \text{arg}(A_s(n_f)) / (2\pi) \]  

(10)

The frequency acquisition range is \([-4\Delta f, 4\Delta f]\), where \( \Delta f \) is the subcarrier spacing.

A more complex timing metric can be calculated as following[2]:

\[ A_s(n) = \sum_{l=0}^{N_s-1} y(n-l) y^*(n-l-2N_s), n = n_0 + 3N_D - 1, n_0 + 3N_D - 1 + 1,... \]  

(11)

\[ A_{31}(n) = \sum_{m=0}^{M-3} A_s(m \star N_D + n), n = n_0 + 2N_D + (M-1)N_D - 1, n_0 + 2N_D + (M-1)N_D - 1 + 1,... \]  

(12)

\[ \rho_t(n) = (M / (2M - 1)) \left| |A_s(n)|^2 + |A_{31}(n)|^2 \right| / (B_s(n) \star B_s(n)) \]  

(13)

The coarse frame timing can be determined similarly as (9) and frequency offset estimation same as (10).

The coarse frame timing from (9) works if the expected minimum working SNR is high and the SNR range is narrow, but if it’s expected the minimum SNR is below zero dB and the SNR range is e.g., [-3, 30]dB, the coarse timing accuracy is not satisfied for especially at high SNR because the threshold \( \rho_t \) is expected to be chosen according to the minimum SNR value and obviously the coarse timing error is very large when the used frame synchronization samples is long, while the coarse frequency offset estimation accuracy depends obviously on the timing accuracy, SNR and the accumulation samples number[2]. Fig.2 and Fig.3 are the one-shot timing metric of (17) with \( M=31 \) where the first half-FFT size’s STF samples is assumed to be used for AGC stabilization for wide SNR range and the timing offset of zero means the timing target which is 1 baseband sample before the beginning of the negated STF field. It’s obvious that the timing error for coarse frame synchronization is larger for even high SNR when the expected acquisition SNR range is wide. Also we can’t rely on a max metric search after the sliding metric bigger than a threshold, which is usually done in continuous transmission system[1][2][3][9] since in this case the reasonable search range is as large as at least the used STF samples number. The used STF samples is several OFDM FFT size meaning a large delay for the synchronization decision, and the max metric search also suffers from timing uncertainty in the metric plateau as shown in Fig.3.

Figure 2. Timing Metric for low SNR in MR-OFDM of 1.2M bandwidth  
Figure 3. Timing Metric for high SNR in MR-OFDM of 1.2M bandwidth
3. Refined Coarse Frame Synchronization algorithm for Burst-Mode OFDM

The sliding timing metric is of monotonously non-decreasing property before the timing target as we can clearly see from Fig.3. Based on it, we propose a refined coarse timing estimation together with frame detection with the timing target set at just before the beginning of the negated STF part.

The timing metric calculation in (13) is sliding in every or several baseband samples. During the sliding procedure and at the first time the timing metric being greater than a predefined threshold value, the frame detection is decided as passed and the timing metric is recorded for a comparison with the timing metric later than it by $N_D$ samples; if the later metric is greater than the former metric multiplied by a coefficient, the timing metric is updated to the later one and a new timing metric calculation later than the updated timing metric by $N_D$ samples is done and comparison goes on until a later metric is detected as not greater than the former timing metric multiplied by a coefficient, at which time the detection is ended and this decision instant is expected to be of a max delay of $N_D$ baseband samples to the target timing, thus it is of low decision delay. The detailed sliding detection procedure is as following pseudo code:

Initialization:

```markdown
SyncFlag = 0;
SyncNo = -1;
MaxSyncTimes = C (> 1, decided according to used STF samples number divided by $N_B$)
```

Sliding:

```markdown
for n = $n_0 + 2N_D + (M - 1)N_D - 1$,
     $n_0 + 2N_D + (M - 1)N_D - 1 + 1$,
     $n_0 + 2N_D + (M - 1)N_D - 1 + 2$,...

if SyncNo >= MaxSyncTimes - 1{
    break;
}
if $\rho_r (n) > \rho_r$, {
    SyncNo = SyncNo + 1;
    $n_{STR}^{out} = n$;
    $\rho^{out} = \rho_r (n)$;
    SyncFlag = 1;
    continue;
}
if SyncFlag == 1{
    if $n = (n_{STR}^{out} + N_D)$
        continue;
}
```

Decision:

```markdown
if SyncFlag == 1{
    Frame Sync Detected
    $n_{STR}^{end} = n_{STR}^{out}$;
    $\hat{\epsilon} = 8 \cdot \text{arg}(A_r(n_{STR}^{out}))/\text{(2} \pi)$
}
else{
    Frame Sync Fail
}
```

where $r_c$ is a constant coefficient approaching 1.0, $n_{STR}^{out}$ and $\hat{\epsilon}$ is the refined coarse frame timing estimate (targeting on the end of STF part just before the negated part) and relative frequency offset estimate, $n_D$ is the decision instant. For reduce calculation speed for decision metric in (15) or (8), the sliding in (16) can be done every $N_D / G (G > 2, G < N_D)$ samples while not every sample like following:

```markdown
for n = $n_0 + 2N_D + (M - 1)N_D - 1, n_0 + 2N_D + (M - 1)N_D - 1 + N_D / G \times 1,
     n_0 + 2N_D + (M - 1)N_D - 1 + N_D / G \times 2$,...

(15)
```

The above refined frame synchronization algorithm makes the final refined frame synchronization timing error expectation within $[-N_D / 2, 0]$ baseband samples independent of used STF samples
number. And the frame synchronization timing decision instant \( n_D \) is \( N_D \) samples delay to the decided frame synchronization timing position thus is expected to be within \([N_D/2, N_D]\) from the targeted position which is related to the random beginning acquisition instant \( n_0 \) for the whole SNR range and independent of the used STF samples in the algorithm. The decision delay is low for the proposed coarse timing algorithm which is preferable but somewhat difficult for burst receiver. To accommodate the timing metric variance especially for low SNRs’ case (below zero dB), the error tolerance for refined timing is relaxed to be within offset of \([-N, N/2]\) when doing subsequent fine symbol timing, and we have \( N_{pT} = N/2 \) and \( N_{nT} = N \) for the positive timing tolerance and the negative timing tolerance.

4. Fine Symbol Timing and Frequency Offset Estimation

It’s important that the coarse frame synchronization timing decision delay is small for burst-mode initial frame synchronization. The designed total synchronization solution which consists of coarse frame synchronization and frequency offset estimation based on STF as well as fine symbol timing and frequency offset estimation based on LTF for MR-OFDM is as fig.4.

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**Figure 4.** Synchronization Procedure in MR-OFDM of 1.2MHz bandwidth

At the refined coarse frame synchronization decision instant \( n_D \), the refined coarse frame timing \( n_{STF}^{end} \) is decided and relative frequency offset is estimated as \( \hat{\epsilon} \) which is translated into crystal frequency correction and also sampling clock correction finished at a delay of \( n_{DF}^{stf} \) baseband samples. Now for the coarse frequency corrected and sampling clock corrected received baseband samples sequence \( \{z(n), n = n_{STF}^{end} + N - N_{pT} + 1, n_{STF}^{end} + N - N_{pT} + 2, \ldots, n_{STF}^{end} + N - N_{pT} + N_{nT}\} \) with an assumption/requirement that \( n_{STF}^{end} + N - N_{pT} \geq n_D + n_{DF}^{stf} \) we do following sliding cross-correlation processing to accomplish fine symbol timing and frequency offset estimation:

**Initialization:**

\( \text{fineSyncFlag} = 0; \)

**Sliding:**

\( n = n_{STF}^{end} + N - N_{pT} + 1, \)

\( n_{STF}^{end} + N - N_{pT} + 2, \ldots, \)

\( n_{STF}^{end} + N - N_{pT} + N_{nT}\}

\( c(n) = \sum_{i=0}^{N-1} z(n-i) z^*(n-l-N); \)

\( z_i(n-l) = z(n-l) + z(n-l-N) \)

\( \exp(j \cdot \arg(c(n)))\); \)

\( R(n) = \sum_{i=0}^{N-1} z_i(n-l) x_i^{stf}(N-l); \)

\( P(n) = \sum_{i=0}^{N-1} |z_i(n-l)|^2; \)

\( r(n) = |R(n)|^2 / \langle P(n)\rangle \cdot Q; \)

\( \text{if } r(n) > \rho_T \)

\( n_{STF}^{end} = n; \)

\( \text{fineSyncFlag} = 1; \)

\( n_{D,if} = n; \)

break;

\} \)

**Decision:**

\( \text{if } \text{fineSyncFlag} == 1 \)

\( \text{LTF Frame Sync Detected; } \)

\( n_{LTF}^{end} \) is the fine LTF end position;

\( \hat{\epsilon}_{if} = \arg(c(n_{LTF}^{end})) / (2\pi); \)

\} \)

\( \text{else} \)

\( \text{LTF Frame Sync Fail} \)

\} \)

(16)
where \( Q = \sum_{l=0}^{N-1} |x^f(l)|^2 \) is the sum of LTF time domain signal’s power of FFT size, \( \rho_f \) is a predefined constant threshold for LTF fine timing metric, \( n_{f_{\text{est}}} \) is the fine symbol timing decision instant and a refined relative frequency offset estimation is combined as

\[
\hat{\epsilon} = \hat{\epsilon} + \hat{\epsilon}_{f_{\text{est}}}
\]

(17)

It is obviously that this fine symbol timing decision instant based on LTF is of zero samples delay, which makes it of practical value to implementation since zero delay meaning no extra storage requirement.

It’s to be pointed that the FFT windowing position for later channel estimation or data demodulation is usually delayed by a max delay corresponding samples number \( n_t \) as in [3]:

\[
h_{\text{FFTwindow start end}}^{\text{PHR LTF CP}} = n_{\text{FFT window start end}}^{\text{PHR LTF CP}} + n_t - n_r
\]

(18)

5. Performance Evaluation and Simulation

To evaluate the performance of the presented refined coarse frame synchronization algorithm and the fine symbol timing algorithm, simulation is set up with the 802.15.4g frame signal including the synchronization preamble of STF and LTF as fig. 1 as well as successive PHR and PHY payload samples. A white Gaussian noise sequence samples of 4 times the preamble length (STF+LTF, total 7.5 times of FFT size) is pre-inserted before the time domain frame signal. The beginning BB receiving sample \( n_0 \) is randomly selected within the pre-inserted white Gaussian noise sequence for every simulated frame to simulate the random starting position in receiver to the beginning of the frame. For the maximum frequency correction range of \([-4\Delta f, 4\Delta f]\), uniform frequency offset within this range is added on the received frame samples for every simulated frame. Performance is evaluated on AWGN and quasi-static multipath fading channel model EPA, ETU and EVA[10] where the quasi-static channel is implemented as: within every frame, the path coefficients are in-variant and every path coefficient is Rayleigh distributed in amplitude and uniform distribution within \([0, 2\pi]\) in phase and the path is independently faded.

The performance is evaluated in two aspects: detection ratio and false-alarm ratio.

The false-alarm ratio can be evaluated in two cases: no signal with only noise and OFDM signal with no synchronization preamble within the searching procedure.

When the receiver begins searching before a whole MR-OFDM frame including the synchronization preamble and PHY, the synchronization detection is expected to be passed with the coarse detection ratio defined as the coarse synchronization passed frame number divided by the total simulated whole-frame number. To evaluate the quality of the coarse synchronization algorithm, a coarse frame synchronization correct ratio is defined as the passed coarse synchronization frame number with timing error within \([-N, N/2]\) and relative frequency error within \([0.25, 0.25]\) divided by the total simulated whole-frame number. And a final coarse detected & correct ratio is defined as the passed coarse synchronization frame number with timing error within \([-0.25, 0.25]\) and relative frequency error within \([-0.25, 0.25]\) divided by the total simulated whole-frame number.

Furtherly, overall detection ratio is defined as the passed fine synchronization frame number divided by the total simulated whole-frame number, and correct fine synchronization ratio defined as the passed fine synchronization frame number with timing error within \([0, N_{mp}]\) and relative frequency error within \([-0.06, 0.06]\) divided by the total simulated whole-frame number. And a final fine synchronization detected & correct ratio is defined as the passed fine synchronization frame number with timing error within \([-N/32, N/32+N_{mp}]\) and relative frequency error within \([-0.06, 0.06]\) divided by the total simulated whole-frame number, where \(N_{mp} \) accounts for the max multipath delay samples number which is 6 for the longest ETU channel in MR-OFDM of 1.2M bandwidth. It’s apparent that the final fine synchronization detected & correct ratio is of essence to the final burst packet BLER performance, thus this ratio is compared to a usually agreed threshold ratio e.g. 90% to evaluate the workable SNR range for the receiver and system.
With appropriate chosen threshold constants $\rho = 0.06250$, $\rho_0 = 0.125$, we get zero false-alarm ratio for only noise case on different noise-levels for both coarse and fine synchronization. And for OFDM signal with no synchronization preamble within the searching frame case, under different $\text{Es/N0}$ settings, we also get zero false alarm ratio for coarse frame synchronization.

With same threshold constants, we get the coarse and overall performance as in Fig. 5 and Fig. 6 for AWGN and quasi-static EPA, ETU, EVA fading multipath. Table 1 is the summary of the required $\text{Es/N0}$ in dB for overall fine synchronization detected and correct ratio greater than 90%, it’s reasonable we get worse performance on Rayleigh faded single path channel such as EPA than Rayleigh faded multipath channel such as ETU and EVA.

Table 1. Min $\text{Es/No}$ in dB for overall fine detected & corret ratio greater than 0.9

| Channel Type     | AWGN | EPA fading | ETU fading | EVA fading |
|------------------|------|------------|------------|------------|
| Required min $\text{Es/N0}$ in dB | -3.0 | 5.0        | 1.0        | 1.2        |

6. Conclusion
Simulations show that satisfactory performance from both detection and detection quality as well as false alarm point of view is got for the proposed refined coarse detection algorithm and fine synchronization algorithm. This high performance together with the low detection delay makes it valuable in receiver implementation for general burst mode OFDM system’s initial synchronization with repeated preamble sequence.
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