Symbol and Frame Timing Estimation for Multicarrier HF Packet Radio

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Abstract **— High frequency (HF) channel is a difficult medium for data transmission. Multi-carrier transmission over HF channel is used to avoid effects of frequency-selective fading. In this paper, we present algorithms for estimation of symbol and frame timing of particular HF packet-radio signals. In particular, we analyze the proposed data-aided frame timing estimation method.**

Key words **— estimation, high frequency, packet radio, timing.**

I. Introduction

PROPAGATION of radio signals in high frequency (HF) band suffers significant degradation. Multipath propagation and various ionosphere effects cause time dispersion, Doppler shift, and frequency selective fading. To increase reliability of transmission and simplify channel estimation, multi-carrier signals with differential phase modulation are used. This approach is exploited in many commercial HF modems such as SCS Pactor IIex [1] and Codan 3012 [2]. Typically, signals on all of the sub-carriers (tones) start with the known preamble - header. Using known headers, data-aided methods for frame synchronization are possible. We focus on symbol and frame synchronization for Pactor-like signals [1].

A. Architecture of Pactor-like Packet Radio Signal Demodulator

HF radio signal of the packet modem is first demodulated in radio receiver and than fed in software modem receiver as an audio signal in 3 kHz bandwidth.

Demodulation of the modem signal starts when signal is detected in the band of interest (Figure 1). Pass-band signal is translated into the base-band and then decimated and filtered with matched filter. Coarse frequency estimation follows and demodulator frequency is adjusted accordingly. Then, symbol timing is estimated and sampled symbols are fed into fine frequency and phase offset estimator. Finally, residual frequency and phase offset are estimated and corrected in phase estimator.

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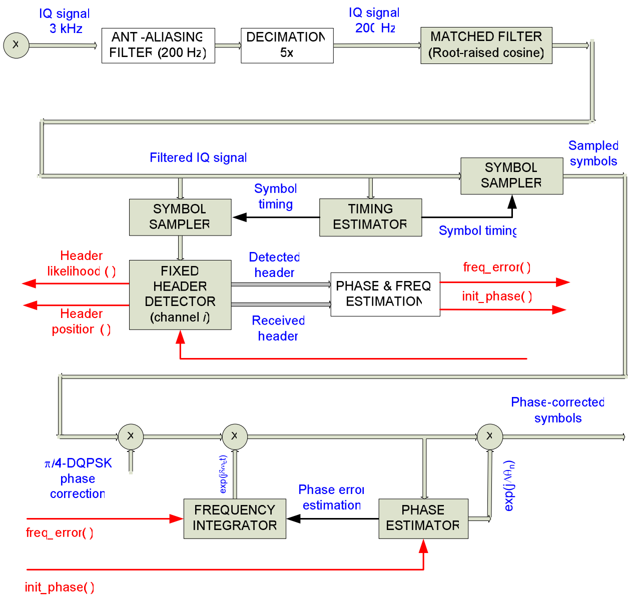

Figure 1. Block-diagram of single tone demodulation.

B. Pactor-III Signal Properties and Packet Format

The PACTOR-III signal [1] occupies spectrum from 400Hz to 2600Hz (measured between -40dB signal levels). There are 18 sub-carriers (tones). Separation between any two neighboring carriers is 120Hz. Symbol rate at each carrier is 100 symbols/s, and symbol duration is T = 10ms.

Each data packet begins with the reference pulse, followed by 8-symbol long header, a variable-length data field, status byte, and CRC field. Packets at odd tones are always aligned in time, and so are packets at even tones. Packets at odd tones are time-shifted by $T/2$ with respect to even tones – time shift alternates in each successive transmission between $+T/2$ and $-T/2$. The headers are always transmitted by using $\pi/4$-DQPSK modulation. The rest of the packet is modulated using either BPSK or QPSK depending on the speed level (SL). There are 6 SLs. At SL1, SL2 and SL3, active tones use BPSK modulation, and at SL4, SL5 and SL6, use QPSK. Active tones at SL1 are 5 and 12, at 1080Hz and 1920Hz, respectively. At SL2, 6 tones are active (3, 5, 7, 10, 12, 14). At SL3 and SL4, 14 tones are active (2 to 15). At SL5 and SL6, 16 (1 to 16) and 18 (0 to 17) sub-carriers are active.

There are 16 distinct headers transmitted at tone 5, and 16 distinct headers transmitted at tone 12, which define whether the packet is retransmission of the previous packet, speed level, whether one-to-one communication or broadcast traffic pattern is used, and whether the packet is short or long. Headers at tones 5 and 12 always go in pairs. Other tones (0-4, 6-11, 13-17) have unique headers each. In other words, there are 16 “variable” header pairs at tones 5 and 12, and 16 “constant” headers at all other tones. SL5 and SL6 have the same encoding in „variable“ headers as SL1 and SL2, respectively, so that analysis of „constant“ headers determines if SL5 or SL6 is encoded.

Therefore, Pactor-III signal demodulator consists of at least 2 and at most 18 independent branches of the demodulator (Fig. 1), depending on the SL. The branches are identical, but each one has distinct tone frequency. Frequency offsets from nominal carrier frequencies are unknown. These offsets are estimated and corrected [3], so that we assume that residual frequency offset at any tone is limited to $|\Delta f| \le 1Hz$ (see performance in [3]).

II. Symbol Timing Estimation

A. Estimation Principle

Robust extraction of symbol timing is essential for proper functioning of the demodulator because other demodulator blocks rely on correctly sampled symbols. Symbol timing estimation in Pactor-III demodulator uses non-data aided (NDA) algorithm based on the search for maximum sum of magnitudes of sampled symbols [4].

Timing Estimator (TE) operates at the symbol rate. It takes signal sequence with 16 base-band samples per symbol. Assuming that every symbol should be sampled at sample *j* within the symbol (*j* = 1, 2, … 16), TE evaluates sum of magnitudes of these samples for the whole sequence or its part. This step is performed over all possible values of sampling position, *j,* within a symbol. Optimal sampling position yields maximum sum of magnitudes of data samples.

B. Timing Drift and Cycle Slipping

In previous subsection, it was assumed that the symbol period is exactly *T*. However, in practice, there is always difference between clocks in the receiver and transmitter, so that timing estimate will drift over time [4, 5]. TE splits data into non-overlapping segments and performs timing estimation on each segment independently of other segments, so that it is possible to adjust timing in the presence of drift.

Such drifting can result in “cycle slipping” when either symbol sample need not be taken in a symbol interval, or two symbols must be sampled in a symbol interval [4]. TE includes cycle slipping control.

At speed levels 3 through 6, joint processing of all active channels is performed in order to obtain more robust and reliable timing estimation.

III. Header detection and frame synchronization

Proper header detection is necessary to determine packet length, SL (modulation), and packet (frame) synchronization.

Header Detector (HD) employs data-aided algorithm to decide whether a vector of sampled symbols at its input is a valid packet header. The algorithm searches for one of 16 possible header pairs on tones 5 and 12. If a header pair or a single header at either tone 5 or 12 is reliably detected, IDs of detected headers, packet length, and employed speed level are provided to Demodulator control (DC) block. SL1 and SL5 have the same header pair on tones 5 and 12, and so do SL2 and SL6. SL5 is distinguished from SL1 by presence of constant headers on tones 1-4, 6-11, and 13-16, whereas SL6 is distinguished from SL2 by presence of constant headers on all tones not active at SL2.

The length of the input vector to the HD block - header detection window - is 12 symbols to account for possible error in positioning of the reference pulse. HD first differentially decodes samples of input signal by calculating phase difference between successive symbols, which removes information on the initial phase from the signal. Then HD de-rotates π/4-DQPSK constellation.

The next step is computation of phase differences between the received differential phases and the differential phases of all known headers. The 12-sample HD window contains the reference pulse and 8 differentially modulated header symbols. Therefore, there are 4 different time offsets in the 12-symbol HD window.

Let $h^{(k)} = (h_1^{(k)}, h_2^{(k)}, \dots h_8^{(k)})$ be the vector of *k*-th header’s 8 differential phases, for *k* = 1, 2,…, 32, which denotes all possible headers at tones 5 and 12.

Let $\varphi^{(i)} = (\varphi_1^{(i)}, \varphi_2^{(i)}, \dots \varphi_8^{(i)})$ be the vector of received signal 8 differential phases, when header *i* is transmitted, for *i* = 1, 2,…, 32, which includes all possible headers at tones 5 and 12.

Let $w = (w_1, w_2, \dots w_8)$ be the vector of noise realizations corrupting $\varphi^{(i)}$.

Let Δf be residual frequency offset between the receiver carrier frequency and the transmitter carrier frequency, after frequency estimation and correction [3]. Δf is assumed constant in header detection window.

$$\varphi^{(i)} = (h^{(i)} + 2\pi\Delta f T + w) \bmod 2\pi \tag{1}$$

Then, we can model difference between the received differential phases and differential phases of a header as:

$$\Delta\varphi^{(i,k)} = (\varphi^{(i)} - h^{(k)}) \bmod 2\pi \tag{2}$$

$$\Delta\varphi^{(i,k)} = ((h^{(i)} - h^{(k)}) + 2\pi\Delta f T + w) \bmod 2\pi \tag{3}$$

It follows that at symbol *m*, for *m* = 1,2,…,8, we have

$$\Delta\varphi_m^{(i,k)} = ((h_m^{(i)} - h_m^{(k)}) + 2\pi\Delta f T + w_m) \bmod 2\pi \tag{4}$$

The HD finally computes sum of squares of phase differences and chooses the header with minimum sum.

$$k_t = \arg\min_{k,position} \sum_{m=1}^{8} (\Delta\varphi_m^{(t,k)})^2 \tag{5}$$

The main idea behind this approach is that the phase difference between the received signal and the true header, appropriately positioned within the 12-sample window, should be smaller than a difference between the received signal and any other header starting at the same position as the true header. The metric should capture this idea. From the model for the received signal phase differences in (4), it follows that (5) is a function of the true header, small frequency offset Δf, and AWGN in the channel. In the next section, we approximate the joint probability distribution function (pdf) of the differences between the received signal and true header differential phases. We show that the sum of squares of differences between the received signal and a header, evaluated at every position within the 12-sample HD window, and across all headers, is suboptimal yet appropriate metric.

The largest metric, provided that it is above a threshold, yields the true header. Threshold selection is a trade-off between minimization of probability of missing the true header and minimization of probability of detecting a false header.

The decision-making step also considers the possibility of one carrier being in a deep fade, while the other carrier has good conditions. The correct header pair can be recovered by making a decision based on the metric of the carrier which is not in fading.

IV. Analysis of the Header Detector

A. Joint pdf of phase differences

Note that $\Delta h_m^{(i,k)} = h_m^{(i)} - h_m^{(k)}$ is the difference between differential phases of header *i* and header *k* at symbol *m*. By construction of the headers, this difference can take values in $\{-\pi/2,\ 0,\ \pi/2,\ \pi\}$, and any two headers, *i* and *k*, $i \neq k$, differ in at least 3 symbols by at least $\pm\pi/2$.

$$\sum_{m=1}^{8} \left| (h_m^{(i)} - h_m^{(k)}) \bmod 2\pi \right| \geq 3\pi/2 \tag{6}$$

We will approximate joint pdf of random vector $\Delta\varphi^{(i,k)}$. First, let us consider marginal pdf of $\Delta\varphi_m^{(i,k)}$, and correlation between it and $\Delta\varphi_l^{(i,k)}$, for $m \neq l$, and then approximate the joint pdf.

Note that in $\Delta\varphi_m^{(i,k)}$, $2\pi\Delta f T$ is a constant, unknown term. $\Delta h_m^{(i,k)}$ is a deterministic unknown parameter because *i* is unknown at the receiver before header detection. By knowing all the header differential phases, one can construct a deterministic search algorithm, which selects the most likely header, subject to the presence of frequency error drift and noise. We studied differential header phases across all 8 symbols in each header, and found that their distribution resembles Uniform descrete distribution on support set $\{-\pi/2,\ 0,\ \pi/2,\ \pi\}$ with true fraction of realizations being {65/256, 65/256, 31/128, 1/4} respectively. Therefore, to make our analysis tractable, we treat header differential phases $h_m^{(i,k)}$ as random variables with uniform discrete distribution of on support set $\{-\pi/2,\ 0,\ \pi/2,\ \pi\}$, where each point has probability 1/4. It follows that $\Delta h_m^{(i,k)} (\bmod 2\pi)$ has the same approximate distribution – uniform over $\{-\pi/2,\ 0,\ \pi/2,\ \pi\}$.

The third term, w_m, is noise in differential phase. We derive pdf of w_m for M-PSK modulation over AWGN channel. First, we observe that the differential phase noise term is a difference of two successive absolute phase noise terms: $w_m = n_{m+1} - n_m$. In AWGN channel, n_m, for all *m*, are independent, identically distributed (i.i.d.) random variables. The pdf of n_m follows from the following M-PSK model. Without loss of generality, let the correct phase of a symbol be 0. The point in constellation diagram corresponding to this phase is $(x, y) = (1,0)$. In AWGN channel, in-phase and quadrature components of the signal with zero phase, X and Y respectively, have independent distributions: X - Normal (Gaussian) distribution with mean 1 and variance $\sigma^2/2$, $N(1, \sigma^2/2)$, and Y - Normal distribution with zero mean and variance $\sigma^2/2$, $N(0, \sigma^2/2)$. Their joint pdf is:

$$f_{X,Y}(x, y) = \frac{1}{\pi\sigma^2} \exp\left(-\frac{1}{\sigma^2} \left((x-1)^2 + y^2 \right) \right) \tag{7}$$

Phase of a point in the constellation diagram is $arg(X+jY) = arctg(Y/X)$. Pdf of an M-PSK phase in AWGN is obtained by transformation of random variables:

$$\rho = \sqrt{X^2 + Y^2}, \phi = arctg(Y/X),$$

which yields the following joint pdf of ρ, ϕ:

$$f_{\rho,\phi}(r, p) = \frac{r}{\pi\sigma^2} \exp\left(-\frac{1}{\sigma^2} (r^2 - 2r\cos p + 1) \right) \tag{8}$$

$$r \geq 0, p \in [-\pi, \pi].$$

Phase pdf is a marginal distribution obtained by integrating $f_{\rho,\phi}(r, p)$ over *r*:

$$f_\phi(p) = \int_{r=0}^{\infty} f_{\rho,\phi}(r, p) dr \tag{9}$$

$$= \frac{1}{\pi} \left[\exp(-\frac{1}{\sigma^2}) \right.$$

$$\left. 2\sqrt{\pi} \frac{\cos p}{\sigma} \exp\left(-\left(\frac{\sin p}{\sigma} \right)^2 \right) \left(1 - F_N\left(-\sqrt{2} \frac{\cos p}{\sigma} \right) \right) \right]$$

where $F_N(x)$ denotes the cumulative distribution function of Normal distribution. For small *p*, $f_\phi(p)$ is:

$$f_\phi(p) \approx \frac{1}{\pi} \exp\left(-\frac{1}{\sigma^2} \right) +$$

$$\frac{\sqrt{2}}{\sqrt{\pi}(\sigma/\sqrt{2})} \exp\left(-\frac{1}{2} \frac{p^2}{(\sigma/\sqrt{2})^2} \right) \left(1 - F_N\left(-\frac{1}{\sigma/\sqrt{2}} \right) \right)$$

in which the second term is scaled Normal distribution.

Pdf of w_m, $f_w(x)$, is equal to the convolution

$$f_w(x) = \int f_\phi(l) f_\phi(x - l) dl \tag{10}$$

$f_w(x)$ was computed numerically and compared with Normal pdf. It can be shown that for SNR = $1/\sigma^2 \geq 6$ dB, $f_w(x)$ is closely approximated by $N(0, (\sigma')^2)$, where SNR' = $1/(\sigma')^2$ = SNR – $(3+\varepsilon)$dB, where ε is approximately 0.4 at SNR = 6dB, 0.1 at SNR = 10dB, and decreases toward 0 as SNR increases.

It follows that the pdf of $\Delta\varphi_m^{(i,k)}$, for $i \neq k$, is a combined discrete-continuous pdf with peaks at $\{-\pi/2, \ 0, \ \pi/2, \ \pi\}$

$$f_{\Delta\varphi,m}^{(i,k)}(x) = \frac{1}{4}\left[f_w\left(\left(x - 2\pi\Delta fT + \frac{\pi}{2}\right)\mathrm{mod}\, 2\pi\right) + f_w\left((x - 2\pi\Delta fT)\mathrm{mod}\, 2\pi\right) + f_w\left(\left(x - 2\pi\Delta fT - \frac{\pi}{2}\right)\mathrm{mod}\, 2\pi\right) + f_w\left((x - 2\pi\Delta fT - \pi)\mathrm{mod}\, 2\pi\right)\right]. \quad (11)$$

On the other hand, pdf of $\Delta\varphi_m^{(i,i)}$ is:

$$f_{\Delta\varphi,m}^{(i,i)}(x) = f_w\left((x - 2\pi\Delta fT)\mathrm{mod}\, 2\pi\right). \quad (12)$$

It can be shown that for $i \neq k$, $\mathrm{var}(\Delta\varphi_m) \approx 3\pi^2/8 + \mathrm{var}(w_m)$, whereas for $i=k$, $\mathrm{var}(\Delta\varphi_m) = \mathrm{var}(w_m) = (\sigma')^2$, independent of m.

Wre proceed with evalution of correlation coefficient of $\Delta\varphi_m^{(i,k)}$ and $\Delta\varphi_s^{(i,k)}$ for $m \neq s$. w_m is independent of all w_s except for $s = m$-1 and $s = m$+1, that is, $\Delta\varphi_m^{(i,k)}$ is correlated only with neighboring differential phases $\Delta\varphi_{m-1}^{(i,k)}$ and $\Delta\varphi_{m+1}^{(i,k)}$, which are equal by symmetry.

Then, for $i \neq k$,

$$\rho_{\Delta\varphi,m+1,m} = \frac{\mathrm{cov}(\Delta\varphi_{m+1}, \Delta\varphi_m)}{[\mathrm{var}(\Delta\varphi_{m+1})\mathrm{var}(\Delta\varphi_m)]^{1/2}} = \frac{\mathrm{cov}(\Delta\varphi_{m+1}, \Delta\varphi_m)}{\mathrm{var}(\Delta\varphi_m)}$$

It can be shown that

$$\mathrm{cov}(\Delta\varphi_{m+1}, \Delta\varphi_m) = \mathrm{cov}(w_{m+1}, w_m) + E(\Delta h_{m+1}\Delta h_m) + (2\pi\Delta fT)^2 \quad (13)$$

It is provable that $\mathrm{cov}(w_{m+1}, w_m) = -\mathrm{var}(w_m)/2$. For $|\Delta f| \le 1Hz$, we have $|2\pi\Delta fT| \le 2\pi/100 \approx 0.063$, and so $(2\pi\Delta fT)^2 \le 0.004$, which makes a negligible floor of $\mathrm{cov}(\Delta\varphi_{m+1}, \Delta\varphi_m)$. Strictly speaking, Δh_m is deterministic, although unknown at the receiver, and can be obtained from the table of headers for every pair of headers. Therefore, strictly speaking, $E(\Delta h_{m+1}\Delta h_m) = 0$. However, we abuse notation to emphasize that construction of headers, that is differential phases of the headers, can additionaly increase the floor level of $\mathrm{cov}(\Delta\varphi_{m+1}, \Delta\varphi_m)$. There remains the noise term:

$$\mathrm{cov}(w_{m+1}, w_m) \approx -\frac{1}{2} \cdot \frac{1}{\frac{3}{8}\cdot\left(\frac{\pi}{\sigma'}\right)^2 + 1} .$$

Ratio (σ')/π was numerically evaluated for various SNR values. One finds that at SNR = 6dB, $\rho_{\Delta\varphi,m+1,m} \le -0.038$; at SNR = 10dB, $\rho_{\Delta\varphi,m+1,m} \le -0.01$. As SNR increases, correlation coefficient decreases until it reaches a floor determined by the construction of headers and residual frequency error. We conclude that for all practical purposes one can consider two neighboring differential phases, $\Delta\varphi_m^{(i,k)}$ and $\Delta\varphi_{m+1}^{(i,k)}$, uncorrelated.

We have shown that the pdf of a vector of differences between the received differential phases and differential phases of a header, $\Delta\varphi^{(i,k)}$, for $i=k$, and SNR > 6dB, can be approximated by a truncated multivariate Normal distribution of uncorrelated random variables with zero mean and standard deviation σ'. Therefore, the joint pdf of random vector $\Delta\varphi^{(i,k)}$ is closely approximated by a product of 8 individual pdf's:

$$f_{\Delta\varphi}(x_1, x_2, ..., x_8) \approx \prod_{m=1}^{8} f_{\Delta\varphi,m}(x_m) .$$

By header design, for any other k, $i \neq k$, there are at least 3 symbols whose mean is ±π/2, that is, whose at least 3 marginal pdf's $\Delta\varphi_m^{(i,k)}$ fall into the tail of truncated Normal distribution. Hence, one can use the joint pdf of $\Delta\varphi^{(i,k)}$ and select the header which yields the largest likelihood $f_{\Delta\varphi}(x_1, x_2, \ldots, x_8)$. However, this can be approximated by a multivariate Normal distribution of i.i.d. random variables, so that it suffices to evaluate log-likelihood, which is the negative sum of squares of differences between the received differential phases and differential phases of a header:

$$\Lambda = -\sum_{m=1}^{8} (\Delta\varphi_m)^2 . \quad (14)$$

We conclude that *max* Λ over all header pairs is a suboptimal yet appropriate decision metric for header detection.

V. Conclusions

Algorithms presented in this paper provide robust solutions to symbol and frame timing estimation under adverse conditions of HF channel. They are developed for usage in software receiver for specific signal properties and packet format, but can be generalized for use in other packet radio software receivers.

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