FREQUENCY OFFSET ESTIMATION ALGORITHM FOR π/4-DQPSK DEMODULATORS

S. Y. Kruglikov,
Rybinsk State Academy of Aviation Technology

I. INTRODUCTION

The Japanese personal digital cellular (PDC) and North American TDMA cellular systems employ receivers that use either differential detection or coherent detection along with a sequence estimator or decision feedback equalizer (DFE). Time-division multiple-access (TDMA) receivers must have the ability to synchronize accurately on each received burst. Synchronization is difficult to achieve in the presence of signal impairments such as carrier frequency offset, Doppler spread, cochannel interference, multipath fading, thermal noise, and nonlinear disunion [1]. Of these impairments, carrier frequency offset is one of the main causes for the failure of a maximum likelihood sequence estimation (MLSE) receiver. Frequency offset rotates the overall discrete-time channel impulse response so that the adaptive channel estimator in the MLSE receiver cannot track the channel variations [1]. Frequency offset also degrades the performance of differential detectors operating under conditions of high Doppler shift and/or low signal-to-noise ratio (SNR). In this case, the received signal points are likely to cross decision boundaries because the phase of the received signal point is shifted in an amount proportional to the Doppler shift. For these reasons, it is important to correct frequency offset before any further processing on the received signal [1].

Synchronization is difficult to achieve in the presence of high-level noise in satellite communication systems too.

Thus, it is necessary to develop a robust frequency offset estimators that provides satisfactory performance over a wide range of channel characteristics. In this connection, a novel frequency offset estimation algorithm is proposed for π/4-DQPSK on multipath channels of cellular communications and on AWGN channel of satellite communication.

II. SYSTEM AND CHANNEL MODEL

With π/4-DQPSK, the phase difference between successive baud intervals has one of four possible values: π/4, 3π/4, -π/4, and -3π/4. The absolute carrier phase at epoch k is

$$\phi_k = \phi_{k-1} + \Delta\phi_k \tag{1}$$

where $\Delta\phi_k = x_k \times \pi/4$ is the differential carrier phase and $x_k \in \{\pm1. \pm3\}$ is the information symbol at epoch k. Assuming zero initial phase, the transmitted phase at epoch k belongs either to the set (π/4, 3π/4, -π/4, and -3π/4} or to the set {0, π/2, π, 3π/2} when k is odd and even, respectively. The transmitted signal point at epoch k. denoted by

$$\tilde{x}_k = e^{j\phi_k} \tag{2}$$

is shaped by transmit filter $p_t(t)$ and transmitted over the channel. The waveform at the output of transmit filter is

$$s(t) = \sum_k \tilde{x}_k p_t(t - kT) \tag{3}$$

The receiver filter is matched to the transmitted pulse, yielding an overall pulse response $p(t) = p_t(t) \otimes p_t(-t)$ of unit energy, where $\otimes$ denotes convolution. For a AWGN channel, the signal at the output of the receiver filter is

$$r(t) = \left(\sum_k \tilde{x}_k p_t(t - kT) + n(t) \right) \cdot e^{j2\pi f_e t} \tag{4}$$

where $n(t)$ is zero-mean complex Gaussian noise with a power spectral density

$$S_{nn}(f) = |P_t(f)|^2 N_0 \tag{5}$$

where

$P_t(f)$ Fourier transform of $p_t(t)$;

N_0 power spectral density of the zero-mean complex additive white Gaussian noise;

f_e carrier frequency offset.

Under the condition of perfect symbol synchronization, the sampled output of the receiver filter at epoch k is

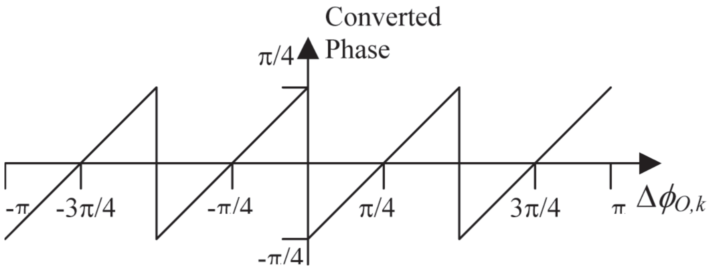

Fig. 1 Remove the phase shift due to the transmitted symbol

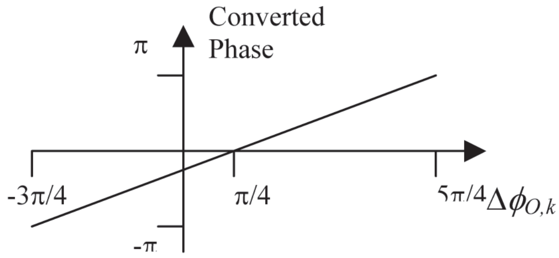

Fig. 2 Remove the phase shift due to a priori symbol

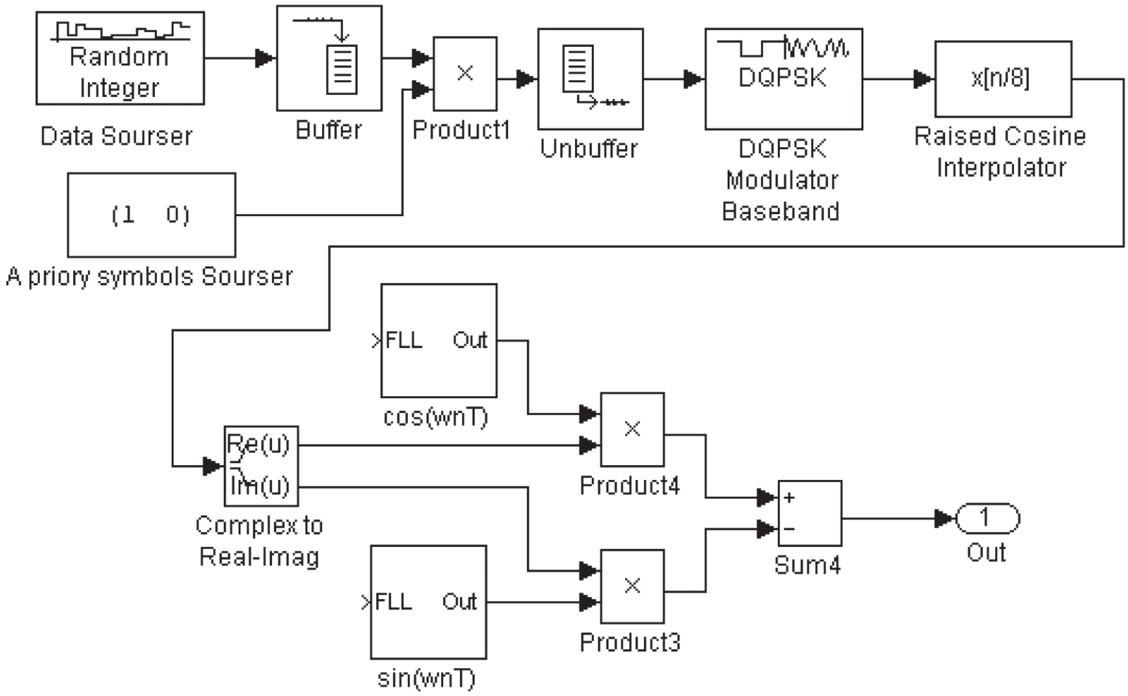

Fig. 3 Modulator simulation model with a priory symbols channel

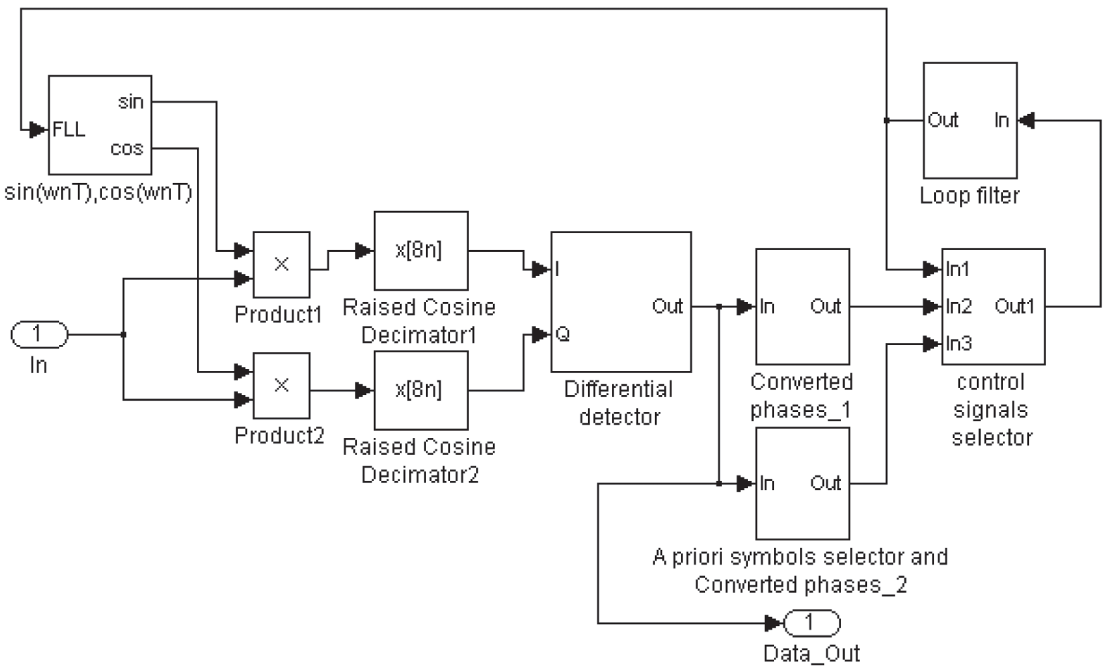

Fig. 4 Demodulator simulation model with FLL

$$r_k \equiv r(kT) =$$

$$= \left(\sum \tilde{x}_k\, p(0) + n(kT) \right) \cdot e^{j2\pi f_e kT} \quad (6)$$

The output of the differential detector is from (6)

$$y_k = r_k r_{k-1} =$$

$$= e^{j(\phi(kT) - \phi((k-1)T))} p^2(0) e^{j2\pi f_e T} +$$

$$+ n(kT)\, n^*((k-1)T) \cdot e^{j2\pi f_e T} \approx$$

$$\approx e^{j(\Delta\phi_k + \Delta\phi_{f,k} + 2\pi f_e T)} p^2(0) \quad (7)$$

The approximation in (7) is valid under conditions of high E_b / N_0 when the noise $n(t)$ are negligible. In the absence of noise, the phase at the output of the differential detector is

$$\Delta\phi_{O,k} = \Delta\phi_k + 2\pi f_e T \quad (8)$$

where $2\pi f_e T$ is the phase shift due to the frequency offset. If the frequency offset is small then the carrier frequency offset can be estimated by just averaging the differential detector output after removing the phase shift due to the transmitted symbols. In order to remove the phase shift due to the transmitted symbol, the output phases from the differential detector are first converted as shown in Fig. 1. Before phase conversion, the output phase $\Delta\phi_{O,k}$ can assume any value on the interval [-π, π]. The converted phases always lie on the interval [-π/4, π/4]. However, when the frequency offset is large, the output phases $\Delta\phi_{O,k}$ are likely to be located near the decision boundaries. As a result of additive noise, some of these output phases will cross the decision boundaries. In this case, a large positive frequency offset may be mistaken for a large negative frequency offset. To solve this problem, it is offered to introduce a priori symbols channel (Fig. 3).

A priori symbols are sequence of repeating symbols appropriate to a phase shift π/4. Now, in order to remove this phase shift, it is converted as shown in Fig. 2. and the converted phases lie on the interval [-π, π]. As result, the frequency offset estimating interval is extended in 4 times.

III. IMPLEMENTATION

The approach to frequency offset estimate can be used for design of demodulator with FLL [2] and without it [1].

If the AWGN channel is used (satellite communication systems), it is possible to design of demodulator with FLL. In FLL the converted phase directly act on an input of a loopback filter (Fig. 4).

If the multipath channel is used (cellular communication systems), it is necessary to design of demodulator without FLL. Reason of this fact that the robust estimation of frequency offset is possible only at a significant signal delay [1].

Agrees [1], to define the deep fade, the mean magnitude $\overline{|r|}$ of the received signal vectors is required. The mean magnitude $\overline{|r|}$ is given by the average magnitude of the received signal vectors over the time-division multiple access (TDMA) burst, i.e.,

$$\overline{|r|} = \frac{1}{N} \sum_{k=1}^{N} |r_k|, \quad (9)$$

where N is the burst length (N = 162 in IS-54).

The frequency offset should be removed from temporary burst. Therefore, compensation of offset by means of r_k multiplying by signal $e^{-j2\pi\Delta f kT}$ is used. Last Δf is average frequency offset.

VI. CONCLUSIONS

A new frequency offset estimation algorithm with a priori symbols channel has been developed. Results of simulation (model – Fig. 3 and Fig.4) show that the frequency offset estimating interval is extended in 4 times.

REFERENCES

[1] Jinsoup Joung, Gordon L. “Stuber Frequency Offset Estimation Algorithm for π/4-DQPSK TDMA Mobile Radio”, IEEE Transaction on Veh. Technol., vol. 49, pp. 1885. 1892, Sept. 2000.

[2] Kruglokov S.Yu. Rybinsk, 2003. 36 p. VINITI no. 1891-B2003.