High Accuracy Data-Aided Frequency Offset Tracking for Burst Mode Mobile Radio

G. Narendra Varma and V.Umapathi Reddy
Hellosoft (India) Pvt. Ltd.
8-2-703, Road No. 12, Banjara Hills, Hyderabad-500 034
varma@hyd.hellosoft.com,vur@hyd.hellosoft.com

ABSTRACT

Carrier synchronization and tracking are crucial for satisfactory functioning of burst mode mobile communication, and implementation of the corresponding algorithms becomes challenging in the mobile receiver. The problem can be addressed in two ways: i) Estimate the frequency offset and correct the local oscillator frequency at the receiver using phase locked loop. ii) Estimate the frequency offset and incorporate the correction in software. Our technique belongs to the second category. The estimation and tracking can be performed at IF or at the baseband. In this paper, we propose a technique in the baseband. We use an IIR adaptive line enhancer for initial acquisition of the frequency offset and data-aided correlation based method for tracking the offset. We provide some simulation results to illustrate the effectiveness of the proposed method.

I. INTRODUCTION

In a wireless receiver such as the mobile handset for GSM (Global System for Mobile Communications) and GPRS (General Packet Radio Service), the RF section delivers I-Q (in-phase and quadrature) signals to the baseband processor. In GSM/GPRS, the receiver processing is done on a burst basis, and hence, synchronization is performed every burst. A mismatch between the transmitter frequency and that of the local oscillator used at the receiver gives rise to frequency offset, and the residual frequency offset as specified by ETSI for GSM/GPRS is within 0.1 ppm of the transmitted carrier . A straightforward solution is to use a local oscillator with 0.1 ppm accuracy. This is done at the base station. But it will be a very expensive solution for a mobile handset. An economical solution is to use a cheaper oscillator with 20 ppm accuracy and bring its accuracy to 0.1 ppm through a frequency offset estimation and correction algorithm. Of course, a software solution is preferred to a hardware solution for reasons such as upgradeability.

Several researchers have addressed the problem of symbol timing and frequency offset estimation. In [3], Huang and Fan considered the problem of joint symbol timing and frequency offset estimation for a GMSK (Gaussian Minimum Shift Keying) modulated receiver. Their method employs a frequency discriminator followed by FFT analysis. They do not, however, consider multipath channel. D'andrea et al. discuss carrier phase and clock recovery problem for continuous phase modulated signals in [4] and MSK (Minimum Shift Keying) modulated signal in [5]. They suggest Costas loop in [4] and apply non-linearity of certain order to the baseband signal followed by differentiation in [5]. Mehlan et al. [6] also addressed the joint estimation of symbol timing and frequency offset for MSK modulated signals and applied their method to multipath environment with only flat fading and two-ray frequency selective channels. Their differential phase technique is non-data aided and also it ignores channel effects.

Our approach is entirely different from those of [3-6] and the estimate is not sensitive to the channel profile. In our approach, we estimate the offset and apply the correction to nullify the effect of the offset. Estimation of the offset is carried out in two phases, acquisition initially and tracking on every data burst. The acquisition is done using a pilot sequence supplied by the base station. For example, the FCH (Frequency Correction Channel) burst on BCCH (Broadcast Control Channel) of GSM system contains 148 zeros. This burst, when modulated with GMSK, takes the form of a complex sinusoid of frequency $R_b/4$ (where R_b =1625/6 KHz is the bit rate) in the baseband. The nominal value of the sinusoid is thus 67.7083 KHz. The actual frequency, however, will be the sum of this frequency and the frequency offset of the oscillator, and our objective is to track the offset. We use an IIR adaptive line enhancer (ALE) for initial acquisition of the offset and data-aided correlation based method for tracking the offset on each data burst.

We describe the IIR ALE approach for initial acquisition in Section II, and data-aided tracking algorithm in Section III. Section IV discusses some simulation results and Section V concludes the paper.

II. IIR ALE FOR INITIAL ACQUISITION

Figure 1 shows an IIR ALE where W(z) is the IIR filter. When the input contains a strong narrowband signal embedded in wide-band noise, and the center frequency of the filter coincides with that of the narrowband signal, the filter output $z(n)$ contains the narrowband signal almost free of wide-band noise. The center frequency of the converged IIR filter gives an estimate of the signal frequency which is the sum of the offset and the nominal value of $R_b/4$. The offset frequency is computed by subtracting the latter value.

Let $x(n)$ be a complex narrowband signal (complex sinusoid in our case) corrupted with a complex wide-band noise. W(z), the transfer function of the IIR filter, is given by [1]

$$W(z) = \frac{(1-s)(w - z^{-1})}{1-(1+s)w z^{-1} + s z^{-2}} \qquad (1)$$

where s is the bandwidth controlling factor in the range 0 to1, and w is a real parameter which is related to the center frequency of the filter f (rad/sample) as

$$w = \cos f \qquad (2)$$

The IIR LMS algorithm [1] for adapting the parameters w and s (after modifying the update equations given in [1] for the complex case) is given by

$$w(n+1) = w(n) + 2\mu_w \, real\,(e(n)\alpha(n)^*) \qquad (3)$$

and

$$s(n+1) = s(n) + 2\mu_s real\left[\frac{1}{(1-s(n))^2} y^2(n) + \frac{1+s(n)}{1-s(n)} y(n)\beta((n)\right] \qquad (4)$$

where $\alpha(n)$ and $\beta(n)$ are complex, and defined as

$$\alpha(n) = \frac{\partial y(n)}{\partial w(n)} = (1+s(n))w(n)\alpha(n-1) \qquad (5)$$

$$-s(n)\alpha(n-2) + (1+s(n))y(n-1) + (1-s(n))x(n)$$

$$\beta(n) = \frac{\partial y(n)}{\partial s(n)} = (1+s(n))w(n)\beta(n-1) - s(n)\beta(n-2) - \qquad (6)$$

$$(w(n)e(n-1) - e(n-2))$$

The superscript $*$ denotes complex conjugate. Here, the index n denotes n^{th} update. We chose (after some trials) the step-size parameters μ_s and μ_w as

$$\mu_s = .0001, \; \mu_w = 0.1(1-s(n))^3 \qquad (7)$$

The baseband FCH correction burst on the BCCH carrier in GSM/GPRS contains a complex sinusoid of 67.7083 KHz. At the sampling rate of 1 sample/symbol (i.e., 1625/6 KHz), the discrete angular frequency of the sinusoid is $\pi/2$. This suggests an initialization of $w(n)$ as $w(0) = \cos(\pi/2) = 0$. Note from (7) that the convergence rate is high when s(n) is small. For a 20 ppm local oscillator at 900 MHz, the ambiguity at the baseband is +/- 18 KHz. To accelerate the initial convergence rate, we initialized $s(n)$ as $s(0) = 0.65$. With the step-size parameters and initializations as above, the ALE achieved acquisition within 0.1 ppm for SNRs as low as 5 dB.

III. TRACKING OF CARRIER OFFSET

After initial acquisition of the frequency offset during the FCH burst, the receiver has to track the drifts in the frequencies of the transmitter and the local oscillator, which are caused due to temperature changes, aging etc., during the data bursts. Tracking has to be performed on the modulated data. The complex baseband signal at the receiver suffers from phase variations due to GMSK modulation, fading channel, carrier offset and white noise floor. Tracking requires estimation of the offset in the presence of these distortions.We use the training sequence provided in wireless systems, for example, the 26-bit training sequence in GSM/GPRS. The motivation for our tracking algorithm is as follows.

The baseband effects of all the phenomena occurring between the symbol source in the transmitter and the symbol sink in the receiver are modeled by a linear channel, and we refer to this as composite channel. The doppler effect due to mobile motion and the local oscillator offset introduce time-varying nature into the composite channel thereby making it a linear time-varying channel. In GSM/GPRS, we estimate the channel once in each data burst using the transmitted bits provided in the middle of the burst, following a linear time invariant (LTI) system identification approach. But, the composite channel will be time varying as stated above. This suggests that we remove the effects of both the doppler and carrier offset before the channel is estimated. Note that the symbol time in GSM/GPRS is such that the channel can be assumed to be stationary over each data burst for the maximum possible doppler. This means, if we remove the effect of frequency offset on the received complex signal, we can proceed with LTI model assumption for the composite channel. But, the frequency offset is unknown. In the following, we show that when we de-rotate the received complex signal by an angle determined by the frequency offset and assume the knowledge of the composite channel, certain correlation measure assumes real value. We now describe our proposed tracking algorithm.

Figure 2 shows a bank of 2M+1 rotors and analyzers and Figure 3 gives the details of the analyzer. The rotor minimizes the time varying nature of the composite channel introduced by the frequency offset. Of course, only one of the rotors meets this condition. M is chosen such that M times the required accuracy $\delta\omega$ is greater than the possible drift in the carrier frequency. In GSM, $\delta\omega$ should be 0.1 ppm of the 900 MHz (0.0021 rad/sample) at the sampling rate of 1 sample/symbol. If we assume that the worst case temperature and aging effects cause a drift equivalent to 0.4 ppm in one frame duration (8 slots), then M = 4 is adequate to estimate in one recursion. Let $r_{tr}(n)$ be the received complex signal when a training sequence is transmitted. The output of mth rotor is given by

$$r^m{}_{rot}(n) = r_{tr}(n) e^{-jm\delta\omega n} \qquad (8)$$

If mth branch in the bank gives the best-fit solution, then $m\delta\omega$ is the frequency offset estimate.

Now consider Figure 3. Let $I_{tr}(n)$ denote the complex symbol equivalent of the training bits and h(n) denote the baseband equivalent impulse response of the composite channel made up of Gaussian filter and multipath channel.

To keep the development of the proposed algorithm simple, we assume the noise to be absent. From the earlier argument on LTI nature of the composite channel, we can express

$$r_{tr}(n) = (h(n) * I_{tr}(n))e^{j\,\delta\omega_0 n} \quad ,n = 0 \text{ to } L_{tr} - 1 \tag{9}$$

where $\delta\omega_0$ is the difference between the frequency of the local oscillator and that of transmitter, L_{tr} is the length of the training sequence and * denotes convolution operation. n=0 corresponds to the start of the training sequence. The output of the channel matched filter with a sampling rate of 1 sample/symbol, which is the case in GSM, is

$$y(n) = r^{m}(n) * \hat{h}^{*}(-n) \tag{10}$$

Here, $\hat{h}(n)$ denotes the estimate of the impulse response of the composite channel. When $\delta\omega_0 = 0$ and $\hat{h}(n) = h(n)$, (9) and (10) (with m =0) can be combined to give

$$y(n) = I_{tr}(n) * \mathrm{R_{hh}}(n) \tag{11}$$

where $\mathrm{R_{hh}}(n)$ is the channel auto-correlation function. We can re-write (11) as

$$y(n) = \sum_{m=n-L}^{n+L} I_{tr}(m)\,\mathrm{R_{hh}}(n-m) \tag{12}$$

where L is the assumed delay spread of the composite channel. The terms in the summation for $m \neq n$ represent the ISI (Inter Symbol Interference). Using the reference training sequence and the received signal corresponding to this sequence, $\mathrm{R_{hh}}(n)$ can be computed and the ISI can be subtracted from (12). We will then have

$$y_{canc}(n) = I_{tr}(n)\,\mathrm{R_{hh}}(0)\,, n = L \text{ to } L_{tr} + L - 1 \tag{13}$$

We restricted the range of values for n as above, because for these values we can compute L significant ISI terms and subtract their contribution from (12).

We define the correlation metric as

$$CM(n) = I_{tr}^{*}(n) Y_{canc}(n) \tag{14}$$

which in view of (13) simplifies to

$$CM(n) = I_{tr}(n)\, I_{tr}^{*}(n)\mathrm{R_{hh}}(0) = |I_{tr}(n)|^{2}\, \mathrm{R_{hh}}(0) \tag{15}$$

From (15), it is clear that the metric is purely real when the carrier offset and noise are absent, and the channel estimate is accurate. In general, however, $CM(n)$ is not purely real. In the presence of noise, which is assumed to be Gaussian, $CM(n)$ is a Gaussian random process with mean equal to $|I_{tr}(n)|^{2}\, \mathrm{R_{hh}}(0)$. With inaccurate channel estimate, $CM(n)$ will be complex and it will have a non-zero imaginary component because of insufficient ISI cancellation. We define a measure J as

$$J(m) \equiv J(CM_m(n)) = \sum_{n=L}^{L_{tr}+L-1} \left\{ \mathrm{Im}\left[CM_m(n) \big/ |CM_m(n)| \right] \right\}^{2}$$

(16)

where $CM_m(n)$ is the $CM(n)$ evaluated in the analyzer of mth branch (see Figure 2). If mth branch yields minimum value of $J(.)$, then the offset estimate is given by

$$\hat{\omega}_{off} = m\,\delta\omega \tag{17}$$

To reduce the variance of $\hat{\omega}_{off}$ due to noise, we perform post-filtering as given below

$$\hat{\omega}_{filt}(k) = (1-\rho)\,\hat{\omega}_{filt}(k-1) + \rho\,\hat{\omega}_{off}(k) \tag{18}$$

where $\hat{\omega}_{off}(k)$ refers to the estimate of the frequency offset on k th burst. The offset estimate from the previous burst is used for de-rotation in the current burst and the tracking algorithm is invoked on the resultant burst. The recursion begins at k=1 with $\hat{\omega}_{filt}(0)$ set at zero value and offset estimate given by (17) as $\hat{\omega}_{off}(1)$. For small ρ, the variance due to noise is less but the tracking rate is also less. So, a trade-off is made between the tracking rate and the accuracy of the estimate.

IV. SIMULATION RESULTS

We have conducted simulations for the Equalizer Channel (this is one of the channel models specified by the ETSI standard 5.05 for testing GSM/GPRS), with and without noise, to verify the effectiveness of our correlation measure $J(m)$ and post-filtering. We used a grid of size 9, i.e., M=4, and assumed the impulse response duration of the composite channel as 4, i.e., L=4. The post filter ρ is fixed at 0.01. In the case without noise, we evaluated $J(m)$ for two values of frequency offset, 60 Hz and 150 Hz. This means, the ALE yielded estimates of frequency offset as above. The results of $J(m)$ are given in Table 1. Note that for offset values of 60 Hz and 150 Hz, the imaginary part of $J(m)$ is least for m=-1 and -2, respectively, as predicted by our algorithm. Figure 4 depicts the convergence behaviour of the post filter. Note that at higher SNR, the filtered estimate converges to around -150 Hz quickly, and the residual error of 30 Hz is well within 0.1 ppm accuracy.

V. CONCLUSIONS

We have presented a carrier tracking algorithm in the baseband for wireless applications such as burst mode mobile communication. The algorithm estimates the frequency offset and then applies a correction in software to minimize the effect of the offset. Some simulations are provided to demonstrate the effectiveness of the proposed method.

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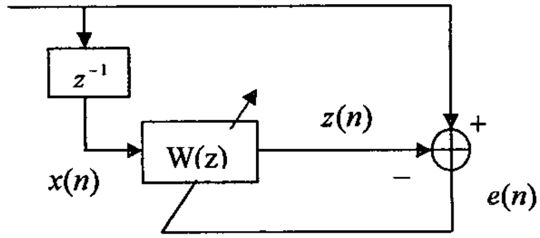

Figure 1 IIR Adaptive Line Enhancer

Table 1. Real and Imaginary values of correlation metric for different frequency offsets

	Re (J(m))		Im (J(m))	
m	60 Hz	150 Hz	60 Hz	150 Hz
4	15.9166	15.8837	0.0834	0.1163
3	15.9439	15.9168	0.0561	0.0832
2	15.9650	15.9440	0.0350	0.0560
1	15.9799	15.9651	0.0201	0.0349
0	15.9882	15.9799	0.0118	0.0201
-1	15.9898	15.9882	**0.0102**	0.0118
-2	15.9846	15.9898	0.0154	**0.0102**
-3	15.9727	15.9846	0.0273	0.0154
-4	15.9541	15.9726	0.0459	0.0274

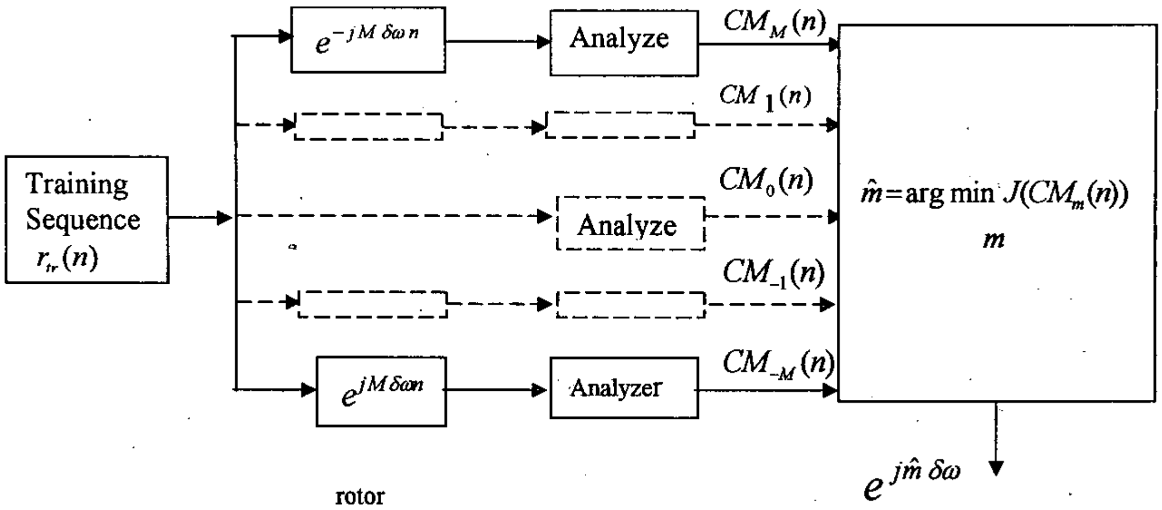

Figure 2 Implementation of the proposed tracking algorithm

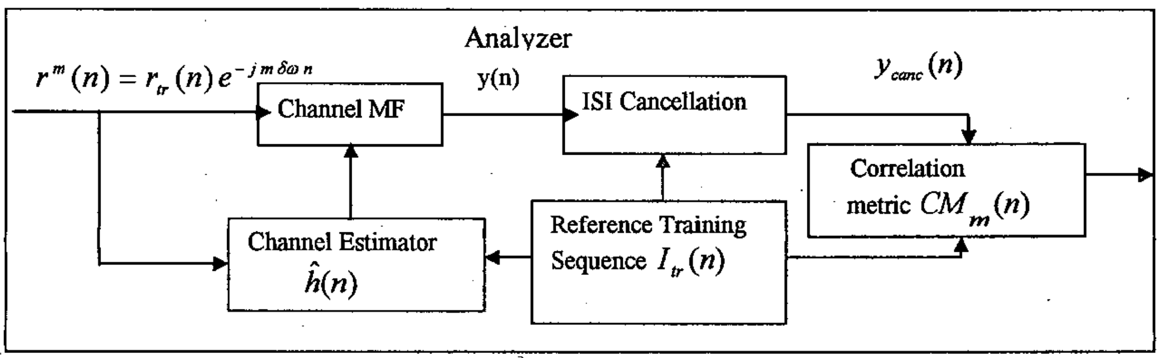

Figure 3 Sub-blocks of analyzer ($r^m(n)$ denotes the output of m[th] branch rotor)

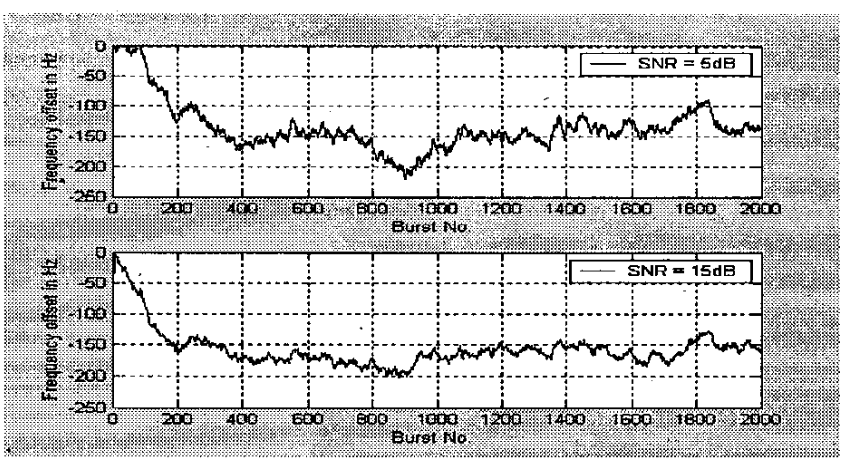

Figure 4. Convergence performance of the post filter (frequency offset was kept at -180 Hz)