Soft Timing Recovery Framework for Cellular Receivers

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***Abstract*—The increased demand for bandwidth and power efficient timing recovery in cellular systems has triggered the need for new timing recovery methods. Classical timing recovery methods rely on transmission of training symbol sequence for the timing information. The resulting timing information synchronizes the receiver through traditional phase-locked loops. Such timing recovery methods are however, bandwidth and power inefficient. In extremely lossy channels, the exploitation of code properties to derive soft timing signals is very crucial. In this paper we present a fast converging timing recovery algorithm for cellular mobile receivers in low signal to noise ratios. In the proposed method, the receiver exploits the soft decisions computed at each turbo decoding iteration to provide reliable estimates of a soft timing signal which in turn improves the decoding time. The derived method based on sequential minimization techniques, approaches the theoretical Cramer-Rao bound with unbiased estimates within a few iterations. Though an 8-PSK baseband equivalent communication system is simulated, it was found that this model worked well with most modulation schemes. The proposed scheme is also insensitive to carrier offsets recovery. Simulation showed that the proposed method outperforms conventional timing extraction methods with respect to jitter performance.**

***Index Terms*—a posteriori means, matched filter output, turbo codes, soft information exchanges.**

I. INTRODUCTION

Soft timing signals are estimated based on the received signal statistics rather than on actual signal values or derivatives. This is desirable in analyzing stochastic signals arriving at the receiver [8]. Such estimators however, need to be bounded and tested for unbiased performance. A benchmark for practical estimators is the Cramer-Rao Bound (CRB). The CRB is a lower bound on the error variance of any unbiased estimate [23, 24]. The CRB is known to be asymptotically achievable for a large enough number of observations, under mild regularity conditions [15]. In many cases, the statistics of the observation depend not only on the vector parameter to be estimated, but also on a nuisance vector parameter we do not want to estimate [25]. The CRB synchronizers have been applied in linearly modulated signals [25]. However, timing recovery in multilevel wireless cellular e.g. GSM modulation scheme with unknown data symbols as nuisance parameters is still a challenging task. This is due to computational complexities involved in low SNR scenarios and mobility of the receivers. Fortunately, Berrou *et al.* developed the revolutionary iterative "turbo" receiver for decoding two dimensional product-like concatenated codes [1]. The impressive performance of turbo codes has triggered the application of this powerful coding technique to digital communications in low SNR environments [2, 3]. Several receiver functions such as the signal detection, equalization, demodulation and timing recovery, are now possible with a combined turbo decoding algorithm[4]-[6].

In most classical timing phase estimations, the timing recovery and decoding process have been separated with little penalty: Timing recovery uses an instantaneous decision device to provide tentative decisions that are adequately reliable to estimate the timing phase error [7]. In such situations, however, the timing recovery process assumes that the neighboring symbols are mutually independent at high SNR and the theoretical framework is normally based on least mean square (LMS) and traditional phase-locked loops (PLL) [8].

Such a framework is susceptible to local minima and often presents additional block processing complexities which fail in low SNR. Due to operation in low SNR environments, combined timing recovery and turbo decoding algorithms are unavoidable in future wireless system. Refs. [9, 10] have shown that classical soft-input/soft-output (SISO) iterative detection/decoding algorithm embed timing parameter estimation in the decoding process. For instance in [10], combined iterative decoding, equalization and timing error estimation is performed with modified forward and backward recursions in the SISO decoders using a per-survivor processing algorithm [21]. Such methods are reliable but increase the receiver's design complexity with vast memory requirement. In order to reduce the complexity involved in designing the decoder structure, soft information provided at each iteration by a conventional turbo decoder can be used to derive reliable information on timing error estimation. This is the essence of turbo principle synchronization technique [6] [22]. Though recent research attention is focused on turbo synchronization method [11, 12], less effort has been directed towards achieving a fast converging timing recovery process.

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In the recent past, wireless cellular transmission systems have over-relied on traditional forward error correction (FEC) codes to either save bandwidth or reduce power requirements [20]. However, classical FEC coding schemes have limited coding gain. Data-aided synchronizers employed are faster but bandwidth inefficient. In cellular receivers, bandwidth conservation, convergence and jitter variation in steady state synchronization is very critical. Thus, to recover the majority of transmitted symbols and frames with minimum error, timing recovery speed and channel impairments become fundamental questions. The objective of this paper is therefore to develop a mathematical framework for a turbo-based timing recovery with a focus on faster convergence and jitter reduction. This goal is achieved by combining maximum likelihood estimation, modified Newton-Raphson minimization model and matched filtering. The performance bound of the estimator tends towards the Cramer-Rao bound.

This paper is organized as follows. Section I provided a broad overview of the problem area, related work and results achieved by other researchers. In section II, the turbo system model is presented. In section III, an improved soft timing framework is proposed. Simulation tests and results are given in sections IV and V. Conclusions are drawn in section VI.

II. SYSTEM MODEL

The baseband-equivalent of a turbo-coded communication encoder and decoder structure is depicted in Figure 1. It consists of two recursive systematic convolutional encoders (RSC) which are separated by a pseudo-random L-bits interleaver (INT), puncturing the output of the encoders increases the transmitted code rate from 1/3 to 1/2. On the other hand the receiver consists of modified turbo decoder with two separate Soft-In/Soft-Out maximum likelihood a posteriori (MAP) decoders which are connected with an interleaver (INT) and deinterleaver (DEINT). The most widely used MAP algorithm is the recursive Bahl-Cocke-Jelinek-Raviv (BCJR) algorithm [4]-[6], which permits easy calculation of the log likelihood ratio (LLR) on block by block basis. BCJR algorithm, however, requires priori knowledge of the channel parameters. Fortunately, the feedback loop provides prior soft bit information from one decoder to the other hence the name 'turbo code'. It works iteratively, i.e. the decoding improves with consecutive iterations upto a certain threshold limit when performance degrades. Implementing combined timing recovery and decoding helps with convergence.

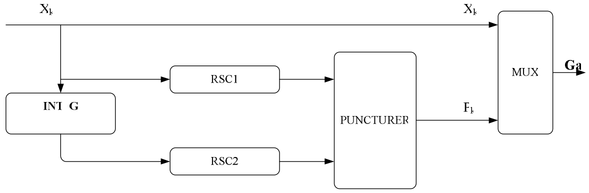

Figure 1a. Structure of a turbo encoder

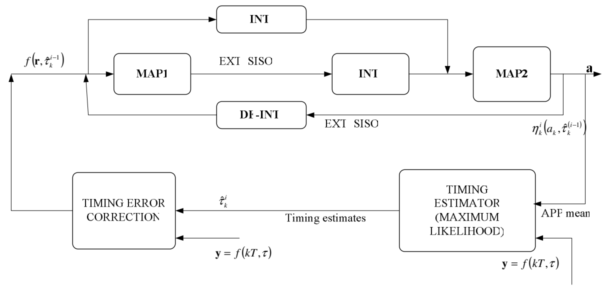

Figure 1b. Structure of modified turbo decoder

III. THEORETICAL CONSIDERATIONS

The received signal is

$$\mathbf{r} = \mathbf{HGa} + \mathbf{n}\,, \qquad (1)$$

where, after sampling, the received vector becomes $\mathbf{r} = [r_{-M} \ldots r_0 \ldots r_{N-1+M-1}]^T$. Here, we have collected $N + 2M$ samples from the received pulses, N is the number of transmitted symbols, $2M+1$ is the samples of the observation interval (so-called pulse truncation), and $M \to \infty$. for best noise performance.

$$\mathbf{H} = \begin{pmatrix} h(\tau) & & \\ \vdots & \ddots & \\ h(N_h T + \tau) & & h(\tau) \\ & \ddots & \vdots \\ & & h(N_h T + \tau) \end{pmatrix}, \qquad (2)$$

is a convolution matrix created from delayed samples of pulse shape $\mathrm{h(t)}$ by τ. $\mathbf{G}$ is an interleaving matrix (obtained by permuting the columns of an identity matrix of size L-bits), $\mathbf{a} = (a_1, \ldots, a_N)$ are the transmitted symbols with consecutive symbol duration, T and $\mathbf{n}$ is independently and identically distributed (i.i.d.) Gaussian noise with variance $N_0 / 2$.

In order to have sufficient statistics in the decoding-timing recovery module, (1) must be sampled at $T_S \le T/(1+\alpha)$ intervals, where α is the roll-off of the transmitting square root raised cosine waveform.

The received sampled vector is passed through discrete matched filters whose outputs take the form

$$\mathbf{y} = f(kT+\tau) = \sum_{k=1}^{k=N} a_k x(s - kT - \tau) + w(s) \qquad (3)$$

Computing the output of (3) at correct instants of the argument, $\{kT + \tau\}$ yields the solution to the problem of time recovery.

The solution to the problem of timing recovery is two fold; estimating the timing phase signal τ and determining the steady state location of the timing instants through suitable update mechanisms. Traditionally this has been achieved using a combination of timing error detections and phase locked-loops implemented without utilizing the decoder soft information or on the other hand using highly complex decoders [7,8], [10]-[12] and [14].

In the next section we show how decoder functions can be improved with timing recovery with little modifications. We further introduce the concepts of a low variance design for timing recovery in digital mobile receivers.

A. Estimating timing information

The problem addressed in this section is the estimation of τ subject to a trial value $\tilde{\tau}$. This estimate may be seen as the solution of the maximization problem

$$\hat{\tau} = \underset{\tilde{\tau}}{\operatorname{argmax}}\, \Lambda(\tilde{\tau}). \quad (4)$$

Here,

$$\Lambda(\tilde{\tau}) = \ln p(\mathbf{r} \mid \tilde{\tau}) \quad (5)$$

and

$$p(\mathbf{r} \mid \tilde{\tau}) = \int_{\mathbf{a}} p(\mathbf{a})(\mathbf{r} \mid \mathbf{a}, \tilde{\tau})d\mathbf{a}, \quad (6)$$

where, $p(\mathbf{a})$ is a prior probability mass function. The logarithmic function of second factor of the integrand in (6) is defined as

$$\ln p(\mathbf{r} \mid \mathbf{a}, \tilde{\tau}) = \Re\left\{\sum_{k=0}^{N-1} a_k^* y(kT + \tilde{\tau})\right\}, \quad (7)$$

where $y(kT + \tilde{\tau})$ corresponds to the matched filter output evaluated at $kT + \tilde{\tau}$.
In order to solve for (4), we take the derivative of (5) with respect to $\tilde{\tau}$ and we equate to zero, that is,

$$\begin{aligned} &\frac{\partial}{\partial \tilde{\tau}} \ln p(\mathbf{r} \mid \tilde{\tau}) \\ &= \int_{\mathbf{a}} \frac{p(\mathbf{a})p(\mathbf{r} \mid \mathbf{a}, \tilde{\tau})}{p(\mathbf{r} \mid \tilde{\tau})} \frac{\partial}{\partial \tilde{\tau}} \ln p(\mathbf{r} \mid \mathbf{a}, \tilde{\tau})d\mathbf{a} \\ &= \mathbf{0} \end{aligned} \quad (8)$$

We notice that the evaluation of (8) requires the knowledge of the priori probabilities, $p(\mathbf{a})$ of the transmitted symbols at the receiver. However, in this problem, we assume that such information can only be derived from posteriori information. If we invoke Baye's rule in the first factor of the integrand in (8), we have a posteriori conditional probability density function (PDF) of the transmitted vector $\mathbf{a}$. We can then represent it as

$$\frac{p(\mathbf{a})p(\mathbf{r} \mid \mathbf{a}, \tilde{\tau})}{p(\mathbf{r} \mid \tilde{\tau})} = p(\mathbf{a} \mid \mathbf{r}, \tilde{\tau}). \quad (9)$$

Since from (1), vector $\mathbf{r}$, is a function of vector $\mathbf{n}$, but $\mathbf{n}$ is not a function of τ, we can ignore $\mathbf{n}$ in the following definition without loss of generality. We had indicated that the interleaving matrix is an identity matrix, hence (9) becomes

$$p(\mathbf{r} \mid \mathbf{a}, \tilde{\tau}) = f_{\mathbf{n}}(\mathbf{r} - \mathbf{H}(\tau)\mathbf{a}) = \exp\left(-\frac{\|\mathbf{r} - \mathbf{H}(\tau)\mathbf{a}\|^2}{N_0}\right). \quad (10)$$

The Maximum likelihood estimation problem in (4) now becomes an expectation problem given as follows

$$\begin{aligned} &\frac{\partial}{\partial \tilde{\tau}} \ln p(\mathbf{r} \mid \tilde{\tau}) \\ &= \int_{\mathbf{a}} p(\mathbf{a} \mid \mathbf{r}, \tilde{\tau}) \frac{\partial}{\partial \tilde{\tau}} \left\{\Re\left\{\sum_{k=0}^{N-1} a_k^* y(kT + \tilde{\tau})\right\}\right\}. \\ &= \mathrm{E}_{\mathbf{a}}\left\{\frac{\partial}{\partial \tilde{\tau}} \ln p(\mathbf{r} \mid \mathbf{a}, \tilde{\tau}) \mid \mathbf{r}, \tilde{\tau}\right\} \\ &= \mathbf{0} \end{aligned} \quad (11)$$

Since $\tilde{\tau}$ appears in both factors of the expectation problem in (10), the solution of (10) is non-trivial. According to [13], an iterative approach that generates a set of values for $\hat{\boldsymbol{\tau}} = \{\hat{\tau}^1 \ldots \hat{\tau}^i \ldots \hat{\tau}^n\}$ is a possible solution. Theoretically it is possible to prove that within the limit as $n \to \infty$ the sequence of timing estimate converges to a desired solution [15]. However, the proof is analytically complex. Fortunately, an iterative receiver employing turbo codes will help in achieving faster convergence [6]. The turbo principle in [1] provides soft information described by posteriori means, illustrated in Figure 1b, and given by

$$\begin{aligned} \eta_k^i\left(\mathbf{r}, \hat{\tau}^{(i-1)}\right) &\cong \int_{\mathbf{a}} a_k\, p\left(\mathbf{a} \mid \mathbf{r}, \hat{\tau}^{(i-1)}\right) d\mathbf{a} \\ &= \sum_{\alpha_m \in \mathrm{A}} \alpha_m\, p\left(a_k = \alpha_m \mid \mathbf{r}, \hat{\tau}^{(i-1)}\right) \end{aligned} \quad (12)$$

The second factor in (12) is the marginal a posteriori probabilities (APPs) computed from extrinsic SISO exchanges off log likelihood ratios (LLR) from the second maximum a posteriori (MAP) decoder to the first MAP decoder. Detailed derivation of (12) can be obtained from [2]. The log likelihood ratio as depicted in Figure 1b is thus given by

$$\begin{aligned} &\Lambda_{21}\left(a_k \mid \mathbf{r}, \hat{\tau}^i\right) \\ &= \log\left(\frac{p\left(a_k = +1 \mid \mathbf{r}, \hat{\tau}^i\right)}{p\left(a_k = -1 \mid \mathbf{r}, \hat{\tau}^i\right)}\right). \end{aligned} \quad (13)$$

Since a fast converging estimator is required for our problem, the Modified Newton-Raphson method in [13] is applied to (10) to give a numerical solution, that is

$$\hat{\tau}^{(i)} = \hat{\tau}^{(i-1)} - \left(\frac{\partial \tilde{\Lambda}}{\partial \tilde{\tau}}\right)_{\tilde{\tau} = \hat{\tau}^{(i-1)}} \left(\frac{\partial^2 \tilde{\Lambda}}{\partial \tilde{\tau}^2}\right)^{-1}_{\tilde{\tau} = \hat{\tau}^{(i-1)}}, \quad (14)$$

where, $\tilde{\Lambda} = f(\eta_k, \mathbf{y})$, (see (3) and (12)).

Here, i and $(i-1)$ denote the current and previous turbo iterations, respectively.

B. Updating timing phase estimates

We begin the iteration by assuming that the $(i-1)$th timing offset estimate is zero. The estimated timing offset finally updates the early and late samples of the discrete matched filter output and optimal synchronization is attained when the early and late samples become equal [17]. The new timing estimate will be based on the discrete matched filter output $y(s)|_{s=kT+\hat{\tau}^{i-1}}$ as shown in Figure 2 and the mean of posterior probabilities $\eta_k^{(i-1)}$ from the previous iteration as shown in Figure 1. This can be seen in the following expression

$$\left(\frac{\partial\tilde{\Lambda}(\tilde{\tau})}{\partial\tilde{\tau}}\right)_{\tilde{\tau}=\hat{\tau}^{(i-1)}} \approx \frac{\tilde{\Lambda}\left(\hat{\tau}^{(i-1)}+\Delta\hat{\tau}\right)-\tilde{\Lambda}\left(\hat{\tau}^{(i-1)}-\Delta\hat{\tau}\right)}{2\Delta\hat{\tau}} \quad . \tag{15}$$

Thus, at low SNR (15) is well approximated by

$$= 1/\Delta\hat{\tau}\sum_k \Re\left\{\eta_k^{*(i-1)}\begin{pmatrix} y\left(kT+\hat{\tau}^{(i-1)}+\Delta\hat{\tau}\right) \\ -\,y\left(kT+\hat{\tau}^{(i-1)}-\Delta\hat{\tau}\right)\end{pmatrix}\right\}, \tag{16}$$

and it easy to show that the second derivative in (14) is

$$= 1/\Delta\hat{\tau}^2\sum_k \Re\left\{\eta_k^{*(i-1)}\begin{bmatrix} y\left(kT+\hat{\tau}^{(i-1)}+\Delta\hat{\tau}\right) \\ +\,y\left(kT+\hat{\tau}^{(i-1)}-\Delta\hat{\tau}\right) \\ -2y\left(kT+\hat{\tau}^{(i-1)}\right)\end{bmatrix}\right\}, \tag{17}$$

where, $\Delta\hat{\tau}$ is an adjustable advance/delay parameter that satisfies $0<\Delta\hat{\tau}<T/2$.

C. Lower Bound on timing error variance

Our goal is to arrive at a lower bound on timing estimation error variance given a time-varying timing offset. We model the timing offset as a random walk, according to

$$\begin{aligned}\tau_{k+1} &= \tau_k+\omega_{k+1} = \tau_{-1}+\sum_{j=0}^{k+1}\omega_j \\ &= \tau_k+(k+1)\Delta T\end{aligned} \tag{18}$$

where, $\omega_k \subset \mathrm{N}(0,\sigma_\omega^2)$ are i. i. d. of kth symbol and σ_ω^2 determines the severity of the timing jitter. The random walk is chosen because of its simplicity and because of its ability to model a wide range of mobile channels. We assume a perfect acquisition by setting $\tau_{-1}=0$. In [14], the Cramer-Rao bound (CRB) on the timing estimation error variance for generic channel is presented. Timing offset is assumed constant over the duration of the packet length without loss of generality. Hence, the CRB gives a lower bound on the estimation error variance of unbiased estimators of deterministic parameters $\boldsymbol{\tau}=[\Delta T,\tau]^T$ In [15], the CRB is given by

$$\mathrm{E}_{\mathbf{r}}\left[(\hat{\tau}_i-\tau_i)^2\right] \geq CRB_i(\boldsymbol{\tau}) \tag{19}$$

where $CRB_i(\boldsymbol{\tau})$ is the ith diagonal element of the inverse of the *Fisher information matrix* $\mathbf{J}(\boldsymbol{\tau})$. The (i,j)th element of $\mathbf{J}(\boldsymbol{\tau})$ is given by

$$\mathbf{J}(\boldsymbol{\tau}) = \mathrm{E}_{\mathbf{r}}\left[-\frac{\partial^2}{\partial\tau_i\partial\tau_j}\ln\left(p(\mathbf{r}\mid\boldsymbol{\tau})\right)\right] \tag{20}$$

The probability density $p(\mathbf{r}\mid\boldsymbol{\tau})$ of $\mathbf{r}$, corresponding to a given value of $\boldsymbol{\tau}$, is called the likelihood function of $\boldsymbol{\tau}$. The expectation $\mathrm{E}_{\mathbf{r}}[.]$ is with respect to $p(\mathbf{r}\mid\boldsymbol{\tau})$. Equivalently (15) can be re-written as

$$\mathbf{J}(\boldsymbol{\tau}) = \mathrm{E}\left\{\left[\frac{\partial}{\partial\boldsymbol{\tau}}\ln p(\mathbf{r}\mid\boldsymbol{\tau})\right]\left[\frac{\partial}{\partial\tau}\ln p(\mathbf{r}\mid\boldsymbol{\tau})\right]^T\right\}. \tag{21}$$

From a detailed proof in [16], we obtain Cramer-Rao bound as

$$\frac{\mathrm{E}\left[(\tau-\hat{\tau})^2\right]}{T^2} \geq \frac{2\sigma^2(2N-1)}{\left(\frac{2\pi^2}{3}-1\right)N(N+1)}.$$

$$\frac{\mathrm{E}\left[\left(\Delta T-\hat{\Delta}T\right)^2\right]}{T^2} \geq \frac{2\sigma^2}{\left(\frac{2\pi^2}{3}-1\right)(N-1)N(N+1)} \tag{22}$$

where, σ is standard deviation of noise and other parameters retain their definitions as we have given earlier.

IV. SIMULATION TESTS

To verify the performance of our turbo aided timing recovery scheme, we simulated a baseband communication system transmitting 8-constellation alphabet for phase shift keying (8-PSK) symbols in MATLAB. We considered convolutional turbo code (031,027) with punctured net rate ½. The interleaver length was set to block sizes of 460 bits and square root raised cosine signaling pulse with roll-off of 0.25 was used. 1000 blocks were transmitted over a Rician distributed flat fading channel with additive white Gaussian noise (AWGN). The received signal was passed through anti-aliasing filter and then sampled at a rate higher than the baud rate. A Discrete matched filter was embedded at the input of the decoder as an early-late gate synchronizer (Refer to Figure 1b).

V. RESULTS

Figure 3 shows the estimator variance approaching the Cramer-Rao bound and indicating unbiased estimator in low SNR up-to 0.35 as iteration increases.

In Figure 4, the number of iterations was pre-set to 10. It is clear that for known normalized variances, the performance of the synchronizer improves with a decrease in variance. As expected data aided timing recovery performs better, but at the expense of extra bandwidth. This indicates that timing synchronizer uses soft information to get good estimates that converge to true channel offsets and on the other hand, the turbo iterative decoder uses this soft timing signal to converge faster. In Figure 5, it is seen that increasing the iterations improves performance, however whenever the number of iterations exceeds pre-set value, the performance degrades as seen in Figure 3. This is because the estimator becomes biased with further iterations.

Figure 6 gives a comparision between the early-gate synchronizer, the proposed method and data-aided GSM frame error rate performance. Though data-aided gives the best performance, both power and bandwidth requirements make it a prohibitive option.

A typical wireless cellular radio channel model performance is depicted in Figure 7. For simplicity in radio channel modeling, Rician multipath flat fading is considered with a Rician K-factor denoted by $K = \beta^2 / 2\sigma_0^2$. Here, β and σ_0^2 are the amplitude of the specular path component (dominant LOS) and variances of Gaussian random channel samples with zero-means.

For a smaller Rician factor K, the performance of synchronizer is worse than for large K. This contributes to the fact that, a Rician distribution is best modeled as a Rayleigh distribution when K is small. The resulting Rayleigh distribution provides a poor soft decoder outputs and consequently a poor soft timing signal for the next decoder iterations.

To explain our hypothesis of low jitter, the proposed method was compared with the well-known work of Mueller and Muller in [7]. In Figure 8, we notice that at start-up, the algorithm in [7] demonstrates some jittering effects and converges to a steady state after a few symbols. On the other hand the proposed algorithm starts up with minimal jittering effects and converges to steady state within 10 iterations.

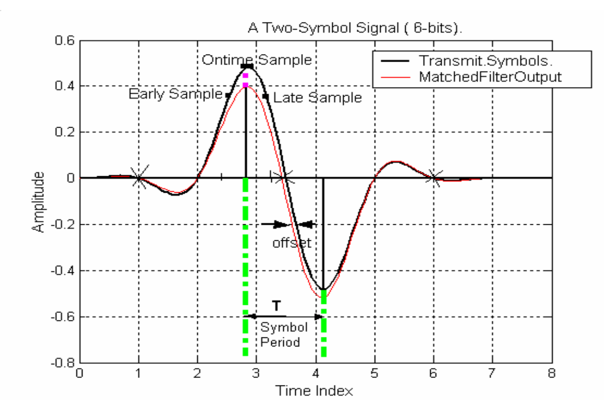

Fig. 2 demonstrates timing recovery problem

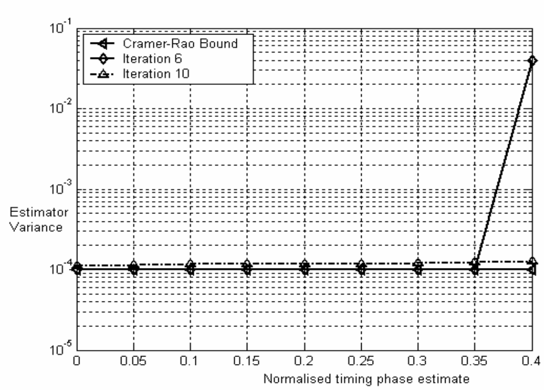

Fig. 3. Estimator variance performance

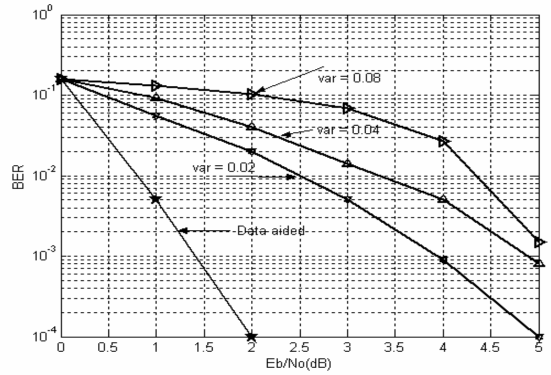

Fig. 4 BER Vs Eb/No at different normalized variances

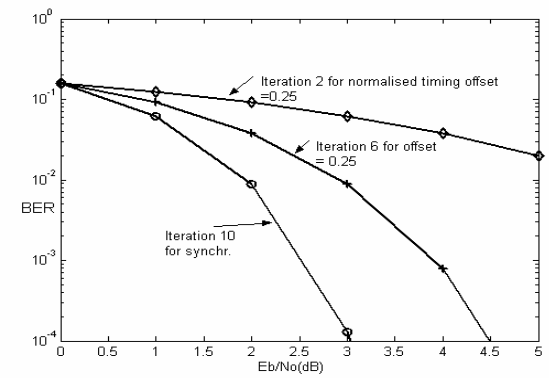

Fig. 5 BER Vs Eb/No at normalized timing offset 0.25

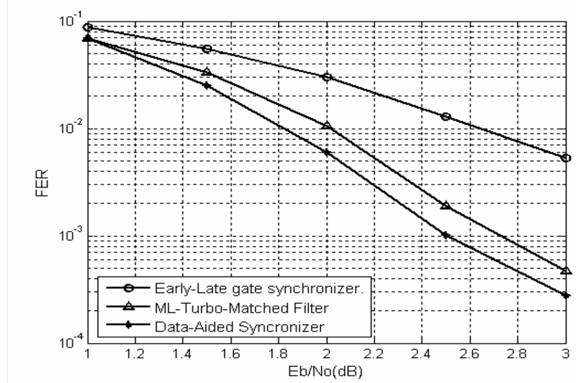

Fig. 6 compares performance of FER Vs Eb/No at normalized timing offset 0.25 and Iteration set to 10.

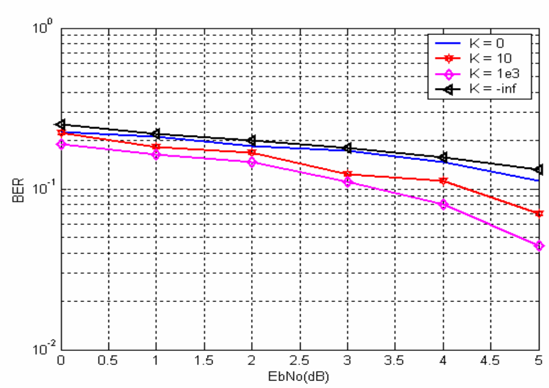

Fig. 7 Multipath fading iterative timing recovery performance for different Rician factor-K in dB at 8-PSK modulation GSM system.

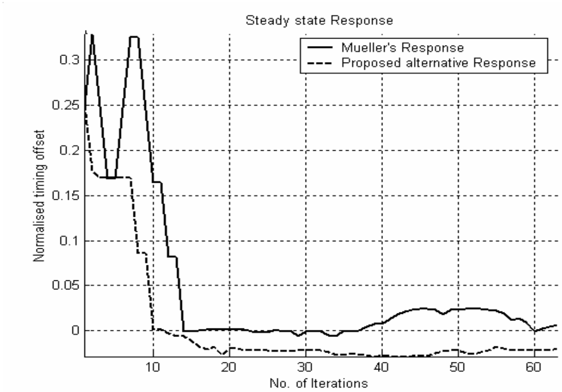

Fig. 8 Convergence criterion from normalized timing offset of 0.25.

VI. CONCLUSION

Combined synchronization and decoding of turbo codes is likely to give good results in low SNR environments. In such applications timing recovery is extremely difficult with traditional methods. Deriving good timing estimator function is crucial in both the decoding process and steady-state sampling phase. In Figure 4, a low variance timing offset estimator with promising results is reported. The variance is also bounded even when the number of iterations is increased beyond the pre-set value. The overall performance index in Figures 5 and 6 shows that, the bit error rate decreases with signal to noise ratio, and the performance degrades after some iterations. After some iterations the estimator becomes biased. The timing phase is therefore, tracked and locked during this period and optimum matched filter outputs are now generated to the APP decoder. Deep fading degrades iterative synchronizer performance as shown in figure 7. The most interesting results are depicted in Figure 8. In comparison of our contribution with the work proposed by Mueller and Muller, our proposed method exhibits a lower jitter at start-up time of synchronization. This is a good indication that the proposed method is a viable solution in cellular network (e.g. GSM) applications where high jitter performance is critical. The proposed solution also reaches steady-state earlier than Mueller and Muller's work hence providing a better solution for timing recovery problem in mobile receivers. Moreover, the proposed scheme is both bandwidth and power conservation efficient for use in cellular standards.

However, the proposed scheme is still complex and a scheme of training pilot sequences employed in GSM standards provides lower bit error-rate.

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