A Novel Method On Robust Timing Recovery

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Abstract:

This paper presents a timing recovery scheme that can tolerate channel interference and non-recovered carrier phase as well as the channel noise. The main feature of the scheme is that it is independent of the equalizer, AGC and phase recovery that may themselves depend on timing recovery. Furthermore there is no need to use training sequence. This feature of the proposed timing recovery is due to the designed timing error estimator and the proper closed loop control system.

I-Introduction:

Many researches are performed to improve the communication system and achieve near Shannon limit performance. Turbo coding invented by Berrou in 1993 [1] is an important attempt in this domain. But for a receiver system to be robust against noise, it is not sufficient that only error correcting part to be robust. This paper examines the design of a robust timing recovery process. In a timing recovery process, an estimate of the timing error is obtained by which VCO is derived. And VCO controls A/D or sampling instant of the match filter output. Timing recovery robustness depends on the design of the timing error estimation. Various types of the timing error estimation such as Spectral line or Decision Directed methods have been discussed [2]. In design of the system perfect conditions are usually assumed: ideal channel, recovered carrier phase, known receiver signal power, and reliability of the decisions or possibility of using a training sequence at start. On the other hand when we deal with equalizer and the other process, perfect timing recovery is assumed. This is a dead lock at start, and is a motivation to design a system that does not depend on the equalizer, AGC, and carrier phase recovery. Also it doesn't require any training sequence at start and even doesn't use the detected symbols. The structure of the timing recovery system depends on realization of the receiver: what parts of the receiver are analog and what parts are digital. In this work, an all-digital receiver is assumed, which means that the input signal is sampled by an A/D and then, match filtering, shift to baseband, AGC, and equalization are performed in the discrete time domain.

II- Receiver Timing Loop:

Consider the receiver timing loop as in Fig. 1. Assuming the ideal channel and using the raised cosine function q(t) to avoid the inter-symbol interference (ISI).

$$q(t)=\frac{\sin(\pi t/T)\cos(\alpha\pi t/T)}{(\pi t/T)(1-4\alpha^2t^2/T^2)} \qquad (1)$$

Where T is the symbol duration. The signal pulse shape and the match filter impulse response are the squared root raised cosine function, p(t) so that: $p(t)*p(t)=Tq(t)$. With $\alpha=0.2$, we have [4]:

$$p(t)\approx\frac{\sin(0.8\pi t/T)+0.8(t/T)\cos(1.2\pi t/T)}{(\pi t/T)(1-0.64t^2/T^2)} \qquad (2)$$

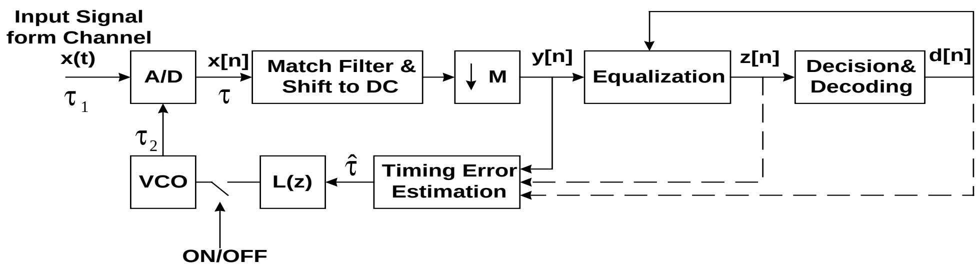

Fig. 1: Receiver Timing Loop

The input signal x(t) received from channel will be:

$$x(t) = \mathrm{Re}\left\{\sum_i I_i p(t - iT - \tau_1)\, e^{j(\omega_c t + \phi)}\right\} \qquad (3)$$

Where I_i are independent and identically distributed (IID) complex information or coded symbols with zero mean and τ_1 is the timing offset between the optimum sampling instant of x(t) and the receiver time base. ϕ is the phase shift between the signal carrier and the receiver local oscillator. Assume the receiver samples the signal at 2M times of the symbol rate to satisfy the Nyquist rate and the sampling instant specified by VCO is $t=nT/(2M)+\tau_2$. Thus the sampled signal will be:

$$x[n] = \mathrm{Re}\left\{\sum_i I_i p\left(\frac{nT}{2M} - iT - \tau\right) \exp\left(j\left(\omega_c \frac{nT}{2M} + \phi\right)\right)\right\} \qquad (4)$$

After match filtering and shift to baseband, it is possible to down sampling the signal. With applying the Fractionally Spaced Equalizer (FSE) that uses the signal at two times of the symbol rate, the down sampling rate is equal to M. The match filter output signal after shift to baseband and down sampling, y[n] will be:

$$y[n] = \sum_i I_i q\left(\frac{nT}{2} - iT - \tau\right) e^{j\phi} = \sum_i I_i q_2[n-2i]\, e^{j\phi} \qquad (5)$$

Where $q_2[n]=q(nT/2-\tau)$. The timing loop is used to estimate the timing error τ from the signal y[n], or the equalizer output z[n], or even the decisions d[n], and then to prepare a timing command τ_2 so that settles the timing error τ to zero.

III- Timing Error Estimation:

The signal y[n] is composed of pulses q(nT/2) shifted in time by $iT-\tau$ and weighted by I_i. It is possible to extract one pulse $q_2[n]$ and then from it, prepare a timing error estimation $\hat{\tau}$. Extraction of one pulse is possible by getting correlation of y[n] and the symbols I_m:

$$E\{I_m^* y[n]\} = \sum_i E\{I_m^* I_i\} q_2[n-2i]\, e^{j\phi}$$
$$= \sum_i P_I \delta[m-i] q_2[n-2i] = P_I q_2[n-2m] e^{j\phi} \qquad (6)$$

Where P_I is the power of the IID symbols I_m. For robustness purpose, instead of the decisions, the values of y[n] before equalization are used.

$$E\{y^*[2m] y[n]\} = P_I \sum_i q_2[n-2i] q_2[2m-2i] \qquad (7)$$

Let us define the soft timing error function as:

$$\varepsilon(\tau) = \frac{\psi[1]-\psi[-1]}{\psi[1]+\psi[-1]}$$
$$= \frac{\left|E\{y^*[2m]y[2m+1]\}\right| - \left|E\{y^*[2m]y[2m-1]\}\right|}{\left|E\{y^*[2m]y[2m+1]\}\right| + \left|E\{y^*[2m]y[2m-1]\}\right|} \qquad (8)$$

Where:

$$\psi[n] = \sum_{i=-\infty}^{\infty} q_2[2i+n] q_2[2i] \qquad (9)$$

Although $E\{y^*[2m]y[2m \pm 1]\}$ is a real positive value and it is free of the phase shift ϕ, but we still use absolute value operation, because the actual average is not an exact real value. The error function $\varepsilon(\tau)$ is free of the symbols scale, the carrier phase shift and reliability of the decisions. As this error function doesn't require any training sequence, the timing recovery is blind. Fig.2 shows the timing error function $\varepsilon(\tau)$.

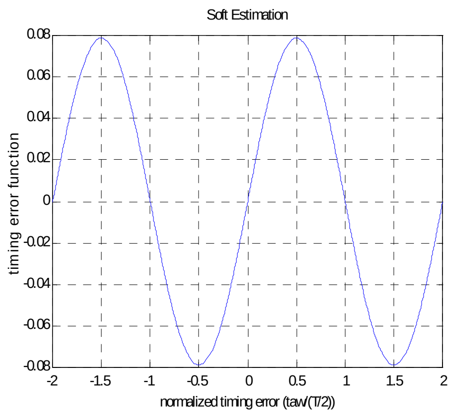

Fig. 2: Timing Error Function $\varepsilon(\tau)$

Implementation of the timing error estimation requires the averaging operation $E\{y^*[2m]y[2m \pm 1]\}$ in (8). It can be done by a first order recursive low pass filter, or by averaging over a block of data. The timing loop depicted in Fig. 1 has a considerable delay between A/D and timing estimation, produced by buffer for A/D output and the match filter. This delay causes instability in the timing control loop system. So to avoid this problem, it is decided to take the averages required in (8) over a block of data that prepared with a fixed timing command τ_2. The length of the data block for y[n], N is much larger than the delay, but small enough so that the timing error caused by the local oscillator frequency drift is negligible. Values of about 600 are suitable for N. Therefore in this technique the timing estimation and timing command are prepared each, for every N samples of y[n]. Thus, as seen in Fig.1 the VCO input command is switched on and off. It is switched on, one time for one block of y[n].

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The actual timing error estimation becomes:

$$\hat{\varepsilon}(\tau) = \frac{\left|\sum_{m=D}^{N/2} y^*[2m]y[2m+1]\right| - \left|\sum_{m=D}^{N/2} y^*[2m]y[2m-1]\right|}{\left|\sum_{m=D}^{N/2} y^*[2m]y[2m+1]\right| + \left|\sum_{m=D}^{N/2} y^*[2m]y[2m-1]\right|} \qquad (10)$$

Where D represents the match filter delay. When using a buffer for A/D output, the produced delay is equal to N which can not be covered in the averaging process. So it must be considered in the timing control loop system. The estimated timing error is $\hat{\tau}=\hat{\varepsilon}(\tau)/\text{slope}$, where is normalized by T/2, slope=0.2474 as seen in Fig. 2.

IV- Timing Control Loop:

In the 'timing shift' domain, the timing control is a system as shown in Fig. 3, considering the VCO as a phase shift integrator. This discrete time system is clocked one time for every N-sample block of data y[n]. If a buffer of size N*M for A/D output is used, a delay element z^{-1} must be included before L(z). (Note that the data rate of A/D is M times of the match filter output.) It is desired that the system have a zero steady state error for a ramp input corresponding to deviation of the receiver VCO oscillator frequency. Thus the control system must be type II and L(z) must have at least one pole at z=1 so that the open loop system have a double pole in z=1. The closed loop control system will be:

$$H(z) = \frac{z^{-2}L(z)}{1-z^{-1}+z^{-2}L(z)} \qquad (11)$$

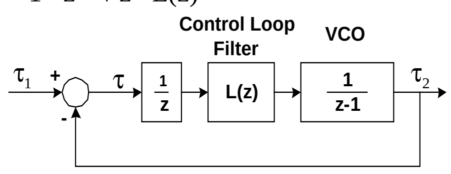

Fig. 3: Timing control system

There are 2 cases of system design which are (a) Fast convergence timing recovery (Acquisition), and (b) Low variance timing recovery (Tracking). In the acquisition mode, the system at the start is to recover the timing in a short time. In this case, the impulse response of the closed loop system, H(z) must be short, or equivalently poles of H(z) must have a small amplitude. But we must try to design H(z) not to be high pass. L(z) and H(z) are selected as:

$$L(z) = \frac{1.25-z^{-1}}{1-z^{-2}} \qquad (12)$$

$$H(z) = \frac{(1.25-z^{-1})z^{-2}}{1-z^{-1}+0.25z^{-2}} \qquad (13)$$

In the tracking mode it is desired to have a low pass system. L(z) and H(z) are designed as:

$$L(z) = \frac{0.125(1-0.9922z^{-1})}{1-z^{-1}} \qquad (14)$$

$$H(z) = \frac{0.125(1-0.9922z^{-1})z^{-2}}{1-2z^{-1}+1.125z^{-2}-0.12402z^{-3}} \qquad (15)$$

Error function non-linearity and channel interference:

As shown in Fig. 2, the error function $\varepsilon(\tau)$ is non-linear. But we don't have to apply its inverse function to prepare τ and can still use $\hat{\tau}=\hat{\varepsilon}(\tau)/\text{slope}$. Non-linearity of the error function decreases the gain of the control loop for $\tau \neq 0$. Decreasing the gain doesn't cause instability while the gain remains positive. But when the slope of the error function is negative for some τ, the control loop becomes unstable. This doesn't cause any problem. In this situation, the system will exit form the instability region and be locked in the stability region around τ=0.

Non-ideal transmission channel also decreases the closed loop gain and therefore makes the system a little more low pass and slow but not unstable. It also makes the error function $\varepsilon(\tau)$ not cross zero at τ=0. But this offset is small and can be ignored in the simulated situation.

V- Simulation Results and Conclusion:

Simulations have been performed for a twisted pair channel, used by xDSL modems. Fig. 4 shows the amplitude response of the 1 km line of type 'AWG24' [3]. For mentioned channel with severe interference, the parameters are considered as 2.7 Msymbol/sec rate, with 1.82 MHz carrier, and 64QAM modulation. In this work, a fractionally spaced DFE is used. The results for a low and high noise levels are presented in Fig. 5 and Fig. 6 respectively. The timing command is normalized by (T/2).

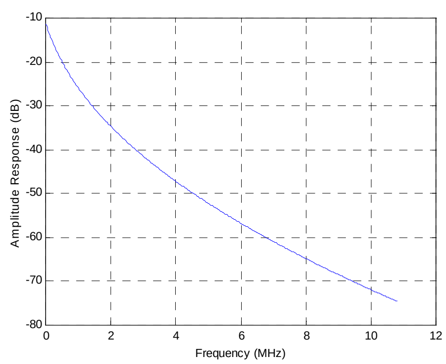

Fig. 4: 1 km 'AWG24' line frequency response

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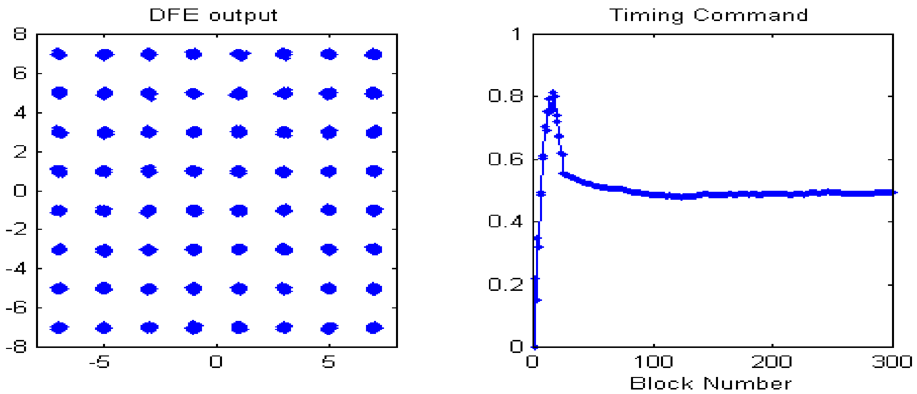

Fig. 5: Simulation for Eb/N0=40 dB

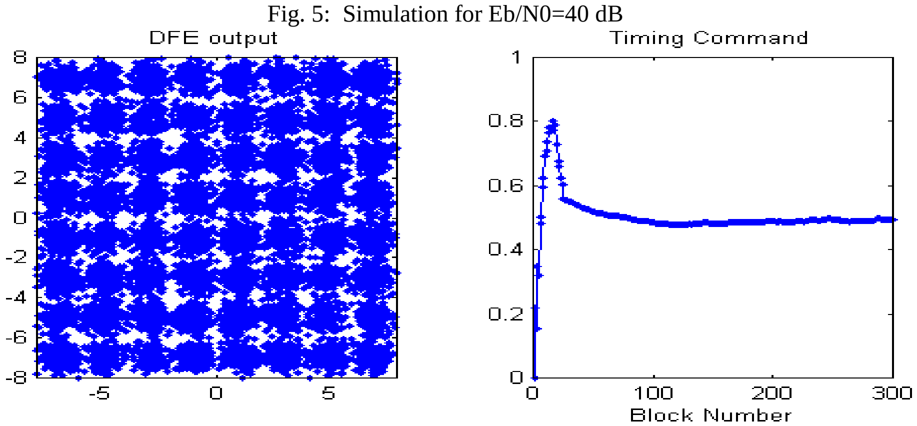

Fig. 6: Simulation for Eb/N0=20 dB

Robustness of the timing recovery can be seen by comparing Fig. 5 and Fig. 6. The timing command in these cases are alike and the channel noise has not affected the performance. For the first 25 data blocks, the acquisition mode and for the rest, the tracking mode is used. As seen, in the acquisition mode the timing command reaches near the correct value fast, but with a high variations, and in the tracking mode it smoothly converges to the correct value.

The new timing recovery scheme presented as said before is not depended AGC, equalizer, carrier phase recovery, and training sequence. It is shown to be robust, because of tolerating a powerful '*self noise*' (Large variance of the timing error estimation).

VII- References:

[1] C.Berrou, A.Glavieux, "Near Shannon limit error-correcting coding and decoding: turbo-codes", ICC93, May 1993, pp.1064-70

[2] E.A. Lee, D.G. Messerschmitt, "Digital communication", 2nd edition, Kluwer Academic Publishers

[3] 'Twisted-pair cable model parameters for 19,22,24, and 26 AWG', Committee T1-telecommunications working group T1E1.4 (DSL Access), November 5, 2001.

[4] "VDSL technical specification, part2: Technical specification for a single-carrier modulation (SCM) transceiver", Committee T1- Telecommunications working group T1 E1.4 (DSL Access), Savanah, GA, Nov. 13-17, 2000.

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