Implementation of Link Quality Evaluation Unit for HF Modems

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Abstract

Link quality evaluation unit is very essential in recent digital data receiver especially in time vary channel such as HF and mobile radio. It provides the adaptive detector with the required information or parameters optimize its performance from the power and rate points of views. These information parameters include the signal-to-noise ratio, Doppler spread, channel severity and the estimate of the sampled impulse response of the channel.

Efficient schemes for the estimation of HF channel were presented. The chosen parameters are the Doppler spread, channel severity and the signal-to-noise ratio which are the dominant ionosheric propagation parameters.

Computer simulation tests have been carried out to evaluate these estimators over HF channel that has been simulated based on Waterson model. Results of computer simulation tests have shown the ability of these estimators to track the variation of the HF channel accurately and efficiently.

الخلاصة

ان وحدة تقويم جودة القناة ضروريه في اجهزه الاستقبال الرقميه الحديثه و خاصة في القنوات المتغيره مع الزمن مثل و القنوات لراديويه المتنقله.توفر هذه الوحده المعلومات او العناصر الضروريه للكواشف الرقميه المتكيفه لتحسين اداؤها من حيث القدره و سرعة الارسال. تشمل هذه المعلومات او العناصر انشار دوبلر وسوء القناة و نسبة الاشارة الى الضوضاء و الاستجابة المتقطعه القناة الرديويه. يعرض البحث طرائق كفؤه لتخمين عناصرالقنوات الراديويه نوع بالاضافه الى تخمين الاستجابة المتقطعة للقناة الراديويه ذات التردد عالي النطاق للنبضه. العناصر المختاره هي انتشار دوبلر وسوء القناة و نسبة الاشارة الى الضوضاء و التي تمثل العناصر الغالبه في عناصر بث القناةالراديويه.تم اجراء اختيارات في الحاسوب لتقيم هذه المخمنات على القناة المنمذجه باسلوب واترسون. اثبتت نتائج الفحص قابلية المخمنات معرفة و متابعة التغيير في هذه العوامل بدقه و كفاءه.

Keywords Estimation techniques, HF channel, Adaptive digital data receivers, Link quality evaluation.

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1. Introduction

The difficulties of data (and voice) communication over HF channels are due to the high interference levels, fading and multipath phenomena on these channels. The problem is that one can not design an optimized modem for an HF channel since the channel parameters are time varying. Thus, an adaptive approach is required that will either select a new HF frequency if the present channel deteriorates, or adapt the modem and/or the communication protocol to the new conditions [1]. In both cases, a qualitative assessment of the channel quality is required. If this link quality evaluation (LQE) is to be effective, it has to be performed in real-time, in order to sense any variation in the channel parameters and subsequently activate the adaptation algorithm. These ideas are certainly not new and have been applied in numerous communication systems [2].

One of the difficulties in these adaptive HF systems is the implementation of the LQE unit. This unit will be responsible in vary the power and rate of the system. The SNR, frequency spread, channel sampled impulse response and the channel are the main parameters necessary in the adjusting the LQE unit.

2. Model of the System

Figure (1) shows the general block diagram for the receiver. It consists, mainly, from the LQE unit and the adaptive detector.

The communication system assumed is adaptive to the channel condition. It has full-duplex communication. A 4-QAM signal is transmitted over the HF channel. The HF channel introduces fading and both time and frequency spreading to the transmitted signal. The received noisy signal (r_i) is then used by the LQE to extract (estimate) values for these channel parameters for channel qualification. According to the quality (condition) of the channel, the communication system parameters are adapted such as power, data rate and complexity. For the receiver, the detection process may be switched among many detection algorithms. That is for good channel conditions, a simple detector may be sufficient to provide satisfactory operation. For more severe channel conditions, more complex detectors may be used like linear, nonlinear or decision feedback equalizers. For bad channel conditions, more sophisticated detection processes like the Viterbi algorithm or any of its derivatives may be used to cope with the severe channel distortion. Information about channel quality (condition), is also feedback to the transmitter, so that, the transmitter may change some communication system parameters to suit the current channel condition [3]. These parameters include the transmitted signal power and the baud rate. That is, when the channel deteriorates, the power of the transmitted signal may be increased and/or the baud rate may be decreased, and for good channel conditions, the baud rate may be increased for a given signal power.

The estimate of the signal-to-noise ratio at the input of the receiver and the frequency spread f_{sp} are made by the SNR and f_{sp} estimators, respectively. These estimates are derived from the received data symbols and knowledge about the modulation technique used. These information are then fed to the channel condition selector, that determines the channel condition according to a look-up table. It translates the estimated channel parameters to predefined channel conditions depending on the International Radio Consultative Committee CCIR classification.

3. Parameters Estimation

The heavy early emphasis on received signal power measurements was a direct reflection of the fact that the essential unknown parameter in early communication systems was the received SNR. While, the received SNR will always be important, other parameters related to the selective fading of the channel have become important. In particular, it is clear that the multipath spread and Doppler spread in addition to SNR are basic parameters needed to ascertain the performance of digital data modems [4]. In the literature, there are two different approaches to the “instantaneous” real-time measurement of Doppler and multipath spreads. One approach, called the complex envelope approach, involves the in-phase and quadrature components of the received carrier, while the other approach, called the envelope approach, deals only with the envelope of the received carrier. The former procedure is more difficult, but, in principle, more accurate [5]. The latter approach is quite simple and inexpensive to implement but involves the assumption that the received carrier is a narrow-band Gaussian process. This assumption appears to be sufficiently accurate for HF channels. Therefore, this latter approach is of considerable help in any large-scale testing of the HF media [6].

3.1 The SNR Estimator

This part estimates the signal-to-noise ratio at the input of the receiver, from a predefined length block of received data symbols. The estimation procedure is based on the important fact that the variation of the mean square value (m.s.v.) of the received data block with SNR is very closely similar to the variation of the noise power for the same block length. This fact was reached after extensive study for the variation of all available (known) parameters at the input of the receiver with SNR. The SNR estimator is derived as follows:

Let the ith received data symbol be given by

$$r_i = \sum_{j=0}^{g} s_{i-j} y_{i,j} + w_i \qquad ..(1)$$

where s_i is the ith transmitted symbol, y_j is the jth component of the channel sampled impulse response of the linear baseband channel and w_i is the additive white Gaussian noise. Since, the variation of both the received block m.s.v and the actual noise power with signal-to-noise ratio are similar, we may define

$$SNR' = \frac{transmitted\ block\ energy}{received\ block\ energy} \qquad ..(2)$$

On the other hand, the estimated SNR in terms of the channel sampled impulse response is

$$SNR' = \frac{1}{\frac{1}{2N}\sum_{i=0}^{N-1}|r_i|^2 - |Y|^2} \qquad ..(3)$$

where N is the length of the received block used in the estimation Further details of the derivation of equ. 3 are given in [5].

3.2 Frequency Spread Estimation

The frequency (Doppler) spread introduced by an HF channel, may be estimated by measuring the r.m.s. bandwidth of the channel. This is achieved by transmitting a carrier and then measuring the r.m.s. bandwidth of the envelope of the received waveform. The frequency spread, usually, is a linear function of the r.m.s. bandwidth of the channel [7,8].

There are more than one method to detect the envelope of the received waveform. One of these is by multiplying the received carrier by both a local carrier and a 90^o shifted local carrier at the same frequency as the received carrier (or as near as possible) and then extracting the low-frequency components by lowpass filtering the resultant waveform. Strictly speaking, f_{sp} is independent of the mean Doppler shift, and thus precise knowledge of the received carrier frequency is not necessary [3,7,8].

Having the envelope being detected, we have time samples separated in time by the reciprocal of the baud rate, for a predefined time interval. The R.M.S. bandwidth of this envelope may be measured by performing frequency analysis on these samples. This is done by transforming these samples to the frequency domain using the Discrete Fourier Transform (DFT), and then taking the –3dB frequency to be the R.M.S. bandwidth ($f_{r.m.s.}$). The frequency (Doppler) spread is equal to $f_{r.m.s.}$ Multiplied by some constant. The value of this constant could be found by making a statistical study for the values of $f_{r.m.s.}$ for different HF channel conditions (different time and frequency spreads). From which, a general formula for the frequency spread estimation may be developed[7].

4. HF Channel Sampled Impulse Response Estimators

In the adaptive adjustment of the detector, the receiver must continuously estimate the sampled impulse response of the channel and appropriately adjust (update) the stored estimate that is used by the detector. In order to estimate the sampled impulse response, the receiver uses the detected values of the received data symbols (the detected signal elements) and assumes that these are all correct. When the signal distortion is severe, a small error in the stored estimate of the sampled impulse response of the channel can introduce a large reduction in tolerance to noise, therefore accurate estimation of the channel has to be maintained.

The detector in Figure (1) is a near-maximum likelihood detector and the delay in detection is (n-1) sampling intervals. Thus, following the receipt of ri+n-1 at time t = (i+n-1)T, the detected data symbol s'_i is detected. Since we are here concerned with the operation of the channel estimator not the detector, the correct detection of all data symbols is assumed, so that s'_i = si for all i. Tests have indicated that the performance of the estimator is only seriously affected by errors in the $\{s'_i\}$ at the higher error rates [1-6,].

The channel estimator is a linear feedforword filter with (g+1) taps which are equal to the number of components in the sampled impulse response of the channel and these tap gains are adjusted in such a way to minimize the mean square error between the actual received sample r_i and its estimate r'_i at the output of the estimator for time invariant channels. Under ideal conditions, the resulting values of the tap gains are the components of the sampled impulse response of the channel. On the receipt of r_{i+n} and before the detection of s_{i+1}, the estimator is fed with the received sample r_i and the detected data symbol s'_i. If Y'_{i-1} is the previously stored estimate of the Y_i, then an estimate r'_i of r_i at the output of the estimators is given by

$$r'_i = \sum_{h=0}^{g} s'_{i-h} y'_{i-1,h} \qquad ..(4)$$

the error in this estimate which is

$$e_i = r_i - r'_i \qquad ..(5)$$

is then scaled by a small positive quantity, Δ resulting in the signal Δe_i. Each signal s'_{i-h} for h = 0, 1,…, g is multiplied by Δe_i.and the products are added to the corresponding components of the previous estimates Y'_{i-1}, giving the newly stored estimates Y'_i, where the (h+1)th component of Y'_i is given by

$$y'_{i,h} = y'_{i-1,h} + \Delta e_i s_{i-h} \qquad ..(6)$$

Equation (6) is usually known as the stochastic gradient algorithm. The factor Δ is usually known as the step size of the estimator and need not necessarily be a constant. It is desirable to make Δ as small as possible so that the additive noise will have a small effect on Y'_i [7]. However, this results in the estimator having a slower rate of response to the change in Y. Clearly, the feedforward estimator can be implemented easily and it is also able to track slow variations in the channel response. In this method there is no prediction, so it uses Y'_i as an estimate of Yi+n in the detector. Hence,

$$Y'_{i+n,i} = Y'_i \qquad ..(7)$$

The adaptive channel estimators are a development of the conventional gradient estimator. The actual error in $Y'_{i,i-1}$ can, in principle, be derived from the fact that the prediction algorithm employs a degree-1 least square fading-memory polynomial filter [5]. The latter assumes that the rate of change of Yi with i is constant or only slowly varying with i. Thus, a significant source of error in a prediction $Y'_{i,i-1}$ is likely to be the acceleration (variation in rate of change) in Yi. If the only error in $Y'_{i,i-1}$ is due to the acceleration in Yi, then a function of the acceleration, which is measured by employing fading-memory or growing-memory averaging for the components of the previous estimates of the channel [7], may be inserted into the adaptation formula of eq. (6) as a variable step size. However instead of attempting to measure the acceleration in Yi directly, use can be made of the fact that the greater the maximum magnitude of any yi,h, the greater is likely to be its maximum acceleration and hence the greater the probable value of the largest error in the corresponding prediction $y'_{i,i-1,h}$. The gradient algorithm of equ. (6) is now replaced by

$$y'_{i,h}=y'_{i,i-1,h}+\Delta u_{i,h} e_i s^*_{i-h} \quad (8)$$

where:

$$u_{i,h}= k_1 \quad \text{for } x^2_{i,h} \le b,$$
$$= k_2 \quad \text{for } x^2_{i,h} > b.$$

The parameters k_1, k_2 and b are appropriate positive real valued constants [6] and

$$x^2_{i,h}= | y'_{i,i-1,h}|^2$$

5. Computer Simulation Tests

Extensive computer simulation tests have been carried out to compare the performance of the different estimator arrangements, over three different HF channel conditions, namely the moderate, poor and flutter HF channels (channels 1,2 and 3, respectively). Details of the development of the channel fading model, together with all computer simulation tests are given in [8,9] . Simulation tests have been carried out using the MatLab 6.1 software package.

5.1 Channel Sampled Impulse Response Estimator Evaluation

Channel sampled impulse response estimation have carried out to investigate the capability of the estimator to determine and track the sampled impulse response of the channel. The results of the tests are shown in Figures (3-7). The signal-to-noise ratio is Ψ dB, where

$$\psi = 10\log_{10}\left(1/\tfrac{1}{2}N_0\right) \quad ..(9)$$

The average transmitted and received energy per bit of information, at the input and the output, respectively, of the HF radio link is unity in each case, and $1/2N_o$ is the two-sided power spectral density of the additive white Gaussian noise at the output of the HF radio link.

Every individual measurement used in plotting a graph has involved the transmission of 24000 data symbols $\{s_i\}$ over the appropriate channel. Each of the three channels (1-3) has been represented by a particular sequence of 24000 vectors {Yi}. The parameter $\square_i$ is taken to be the square of the error in $Y'_{i,i-n}$, measured in dB relative to unity, and is given by

whereas, the parameter ξ_i is the mean square error in $Y'_{i,i-n}$, measured in dB relative to unity, which is given by

$$\xi = 10\log_{10}\left(\frac{1}{20000}\sum_{i=4001}^{24000}\left|Y_i - Y'_{i,i-n}\right|^2\right) \quad ..(10)$$

The first 4000 $\{Y'_{i,i-n}\}$ are ignored in order to eliminate any effect on ξ of the transient behavior of the estimators. Thus, ξ gives a measure of the steady-state performance of the channel estimator, which is here taken to be its performance during the prolonged and uninterrupted transmission of the data signal. In eqs. (7) and (8), $\left|Y_i - Y'_{i,i-n}\right|$ is the unitary length of the vector $Y_i - Y'_{i,i-n}$ and so is the unitary distance between the vectors Yi and $Y'_{i,i-n}$. In all tests (n = 17), where n sampling intervals is the delay in estimation, and the chosen value of n is typical of that likely to be used in practice [1-6].

The performance of of the channel estimator is as shown in Table (1). The parameters, of the estimator have been adjusted as far as reasonably possible to minimize the ξ. However, in the time available, it has not been possible to carry out the complete optimization of every system, so that it may well be possible to achieve further small improvements in ξ for some systems. Three different values of Ψ (20, 30 and 60) have been used in the tests, where the values 20 and 30 are such that a significant number of errors in detection of the received data symbols are likely to be caused from time to time by the additive noise, whereas the value 60 represents a high signal-to-noise ratio, where the fading predominates over the noise.

Table 1 Performance of the channel estimator

Channel	ψ	ξ
1	20	-30.8
	30	-39.9
	60	-47.9
2	20	-29.6
	30	-34.8
	60	-43.7
3	20	-19.2
	30	-28.0
	60	-35.4

The good performance achieved by the channel estimator is due to the fading-memory averages that has been used in the acceleration. This types of adjustment has been compared with other form of adjustment such feedforword, feedforword with predection, and other form of acceleration but these results are beyond the scope of the paper. The results clearly demonstrate the good performance of the suggested estimator. Therefore, the latter is used in the computer simulation tests.

5.2. Channel Parameters Estimation Evaluation

Extensive computer simulation tests have confirmed the validity of this fact for all possible channel conditions. A channel condition here means having one relative time delay from the values 1, 2 and 3 ms and one frequency spread from the values 0.5, 1 and 2 Hz.

For the qualitative channel condition estimator, the channel severity for channels 1, 2, and 3 was measured for a period of 800 time instants ($800 \times \frac{1}{2400} = 0.333$ seconds). This parameter, d, the channel severity given by [3]

$$d = \frac{1}{b_g} \sum_{i=g+1}^{2g} |b_i| \qquad ..(11)$$

where

$$b_i = \sum_{h=-\infty}^{\infty} y_h y_{i-g+h} \qquad ..(12)$$

was calculated at every time instant (iT). It gives a clear idea about the severity of the amplitude distortion. The b_i is the ith component of the aperiodic auto-correlation function of the sequence Y.is plotted for these channels in Figure (2). For channel 1, it is obvious that the channel introduces a significant signal distortion (amplitude distortion), ½>d>¼, for most of the time and severe distortion, d>½, for relatively short time periods. For channel 2, the signal distortion is always severe, even more severe than the signal distortion introduced by channel 3. Physically, the signal distortion introduced by HF channels results from two main causes, the relative time delay between individual skywaves, τ, and the frequency spread due to the fading characteristics of the channel.

Next, consider the quantitative channel parameter estimators, proposed to estimate the signal-to-noise ratio. It is based on the close similarity between the variation of the received signal mean square value and the actual noise power.

Extensive computer simulation tests have confirmed the validity of this fact for all possible channel conditions. This result shows that the noise term in the equation of the ith received symbol, eq. (1), is dominant upon the signal term. Since the latter is equal to the transmitted signal elements weakened (faded) by the channel sampled impulse response, and it is almost constant for different values of signal-to-noise ratio. Therefore, the overall shape of the variation of the received signal mean square value with signal-to-noise ratio will, of course, follow the variation of the actual noise power given by

$$\sigma_{noise}^2 = \overline{s^2}\,10^{-\frac{SNR}{10}} \qquad ..(13)$$

where $\overline{s^2}$ is the transmitted signal mean square value (energy) and it is equal to 2 (4-QAM signalling). Depending on this fact, the estimated SNR has been proposed to be equal to the ratio of the transmitted signal energy to that of the received signal, with a correction factor calculated from the channel sampled impulse response to compensate the difference in amplitude between the received signal mean square value (m.s.v) and the actual noise power.

The m.s.v. of the received signal elements is taken to be the square of the length of the vector of the received signal elements R, given by eq. (2), divided by N. The

value of N is equal to the number of received signal elements needed by the estimator to estimate a single value of the signal-to-noise ratio at the receiver input.

Figure (3) shows plots for the estimated SNR versus the actual SNR, measured over channels 1, 2 and 3. The channel condition selector considers the SNR to be low, moderate or high according to the value of the estimated SNR. That is, an estimated SNR less than 20 dB is considered to be low, and to be moderate if it lies in the interval (20-40) dB. The SNR is high when the estimated SNR is greater than 40 dB. The margins between these signal-to-noise ratio conditions are, of course, not that sharp that a change of say 1 dB in the estimated SNR will drive the channel condition selector to switch to a new condition, but there is an uncertainty region between the different SNR conditions of a width not less than 5 dB. This makes the fluctuations in the estimates of the SNR of less than 5 dB, above or bellow the actual SNR, to be acceptable. Therefore, it may be concluded that the proposed SNR estimator is satisfactorily accurate, since all deviations in the estimated SNRs from their corresponding actual SNR values are sufficiently less than 5 dB as shown in Figure (3).

The SNR estimator must be able to operate correctly without knowing the channel sampled impulse response. Equation (2), was tested to estimate the SNR without utilizing the channel sampled impulse response. These tests were carried out over channels 1, 2 and 3. Plots of the estimated SNR versus the actual SNR for these tests are shown in Figure (4). From this figure, the efficiency of this estimator is clear for the case when channel 1 is used. The performance is degraded for channel 2 for SNRs greater than 50 dB and it is furthermore degraded for the case of channel 3, that is the estimator is no more able to estimate SNRs greater than 30 dB. In fact, this degradation is not important, since it is taking place for channel conditions other than that for which it is switched to the SNR estimator at hand. That is the performance of this estimator is only important when the HF channel is moderate, as shown in Figure (2a). Furthermore, the degradation in the performance of this estimator when tested over channels 2 and 3 emphasizes the importance of the correction factor in the denominator of eq. (3) calculated from the estimates of the channel sampled impulse response, in compensating the effect of the fading over these channels. The latter tests have been carried out but whose results do not presented since it show perfect estimation of the SNR for all channels tested.

The second quantitative channel parameter estimator is the frequency spread estimator. The frequency spread can be estimated by measuring the r.m.s. bandwidth (fading bandwidth) of the signal resulting from lowpass filtering (averaging) the received signal. This filtered signal is composed of the low frequency components (due to the fading characteristics) of the received faded signal. The estimation process is performed in three stages. Firstly, the envelope of the received signal (fading envelope) is extracted. Secondly, the r.m.s bandwidth for this envelope is measured. Finally, the frequency spread, f_{sp}, is related to the fading bandwidth as will be described. In the computer simulation tests, fading envelope extraction was achieved by performing decimation to the sequence of received data symbols, with a decimation ratio of 5 (i.e. taking a sample from each five successive samples), and then performing interpolation to the decimated sequence with an interpolation ratio of 5. This operation, forces the length of the received data sequence used in the estimation process to be relatively long, so that the shape of the low frequency components will appear clearly. A received data sequence of 300 samples ($300 \times \frac{1}{2400} = 0.125$ seconds) was used to give acceptable envelopes that are very close to the actual fading envelopes of the simulated HF channels. For the second stage, the r.m.s. frequency measurement, the following steps were performed on the extracted envelope.

The sequence of 300 samples, is padded by 724 zeros, so that a 1024-point FFT can be performed. Zero padding is used in order to enhance (smoothen) the shape of the spectrum, so that the r.m.s. frequency may be measured easily and accurately.The fading bandwidth is measured, the –3 dB frequency, for all channel conditions shown in Table (2). That is

$$f_{r.m.s.} = \frac{i}{N} f_s \qquad ..(14)$$

where i is the index of that component in the spectrum whose amplitude is equal to the r.m.s. value of the spectrum amplitude, N, here is the length of the data record, and f_s is the sampling frequency for the extracted envelope, and it is equal to the baud rate (2400 samples per second).

Having the fading bandwidth, $f_{r.m.s.}$, of the low frequency components of the received signal for the different channel conditions of Table (2), the individual factor relating the f_{sp} to the corresponding $f_{r.m.s.}$ for each channel condition may be calculated from

$$F_i = \frac{f_{sp}}{f_{r.m.s.}} \qquad ..(15)$$

where f_{sp} is the known frequency spread and $f_{r.m.s.}$ is the measured r.m.s. frequency. From these calculated factors, F_i, the general, unified factor, F, relating all these r.m.s. frequencies with their corresponding frequency spreads is calculated by taking the average value of the F_i's. It has been found from the data of Table (2), that the factor F is equal to 1.48. Therefore, the estimated frequency spread is given by

$$f'_{sp} = F\, f_{r.m.s.} = 1.48\, f_{r.m.s.} \qquad ..(16)$$

Next, the estimator of eq. (15) was tested on an HF channel with variable frequency spread to assess its trackability. This channel is a linear combination of channels 1, 2 and 3. Each channel is windowed to appear for a certain period of time whereas the other two channels are attenuated. The results of this test are plotted in Figure (5), show a good capability for the estimator to track the changing frequency spread over this channel. The values of the estimated frequency spreads, f'_{sp}, are given in Table (2) together with the absolute percentage errors between each f'_{sp} and the corresponding actual frequency spreads, f_{sp}, is given by

$$\varepsilon = \frac{\left| f_{sp} - f'_{sp} \right|}{f'_{sp}} \times 100 \qquad ..(17)$$

since the frequency spread values to be estimated are close to each other, a percentage error greater than 50% will mislead the channel condition selector. For the estimator at hand an average percentage error of about 11% is achieved, indicating a satisfactorily accurate performance.

6. Conclusion

An efficient procedure has been presented for the estimation of certain important parameters of the HF channel. These parameteres will play a major role in the design and operation of LQE system employed in advanced and recent data communication that makes power and rate control in optimum state. Computer simulation tests carried out on these channel parameter estimators (the channel sampled impulse response, channel severity, SNR and frequency spread estimators) have shown the ability of these estimators to track the corresponding HF channel parameters efficiently under different HF channel conditions. It is believed that although these parameters do not completely characterize the HF channel, they provide enough useful information for the implementation of the various adaptive HF systems.

Finally, extra studies were also carried out to employ such parameter estimators for mobile channel. The results clearly show that the estimator correctly estimates the parameters required.

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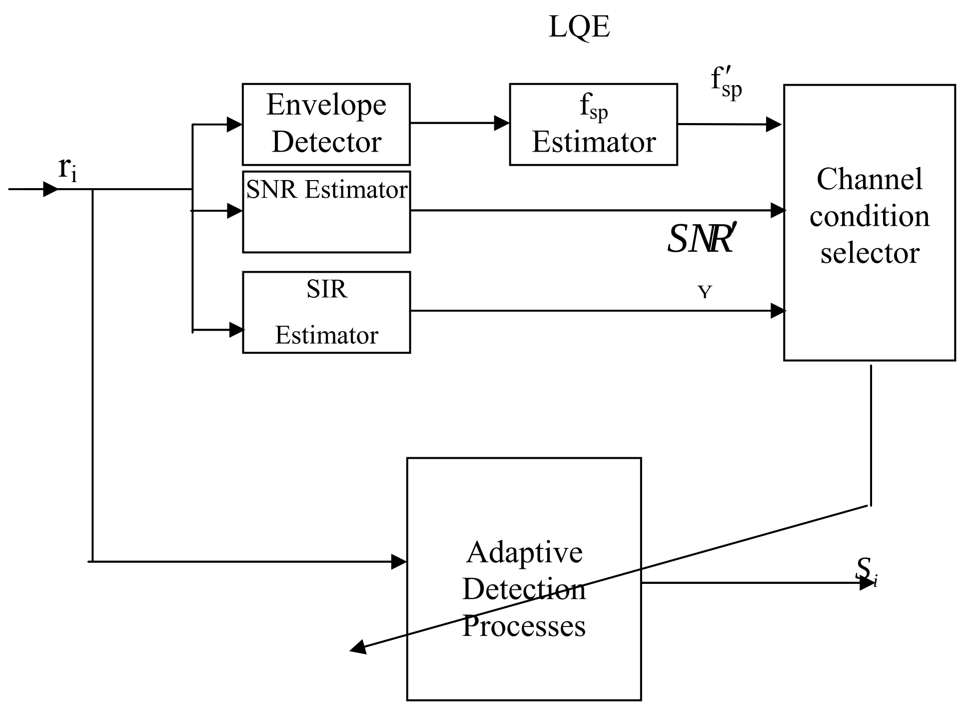

Figure (1) Model of the adaptive digital data receiver

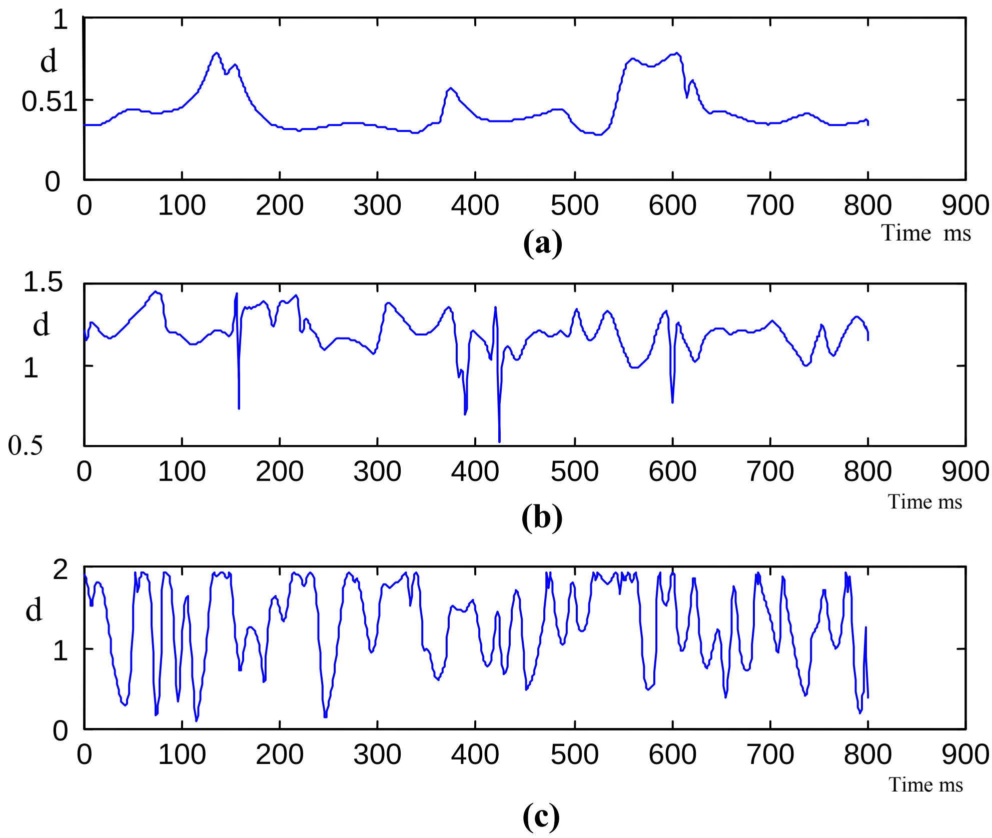

Figure (2) Severity for (a) channel 1 (b) channel 2 (c) channel 3.

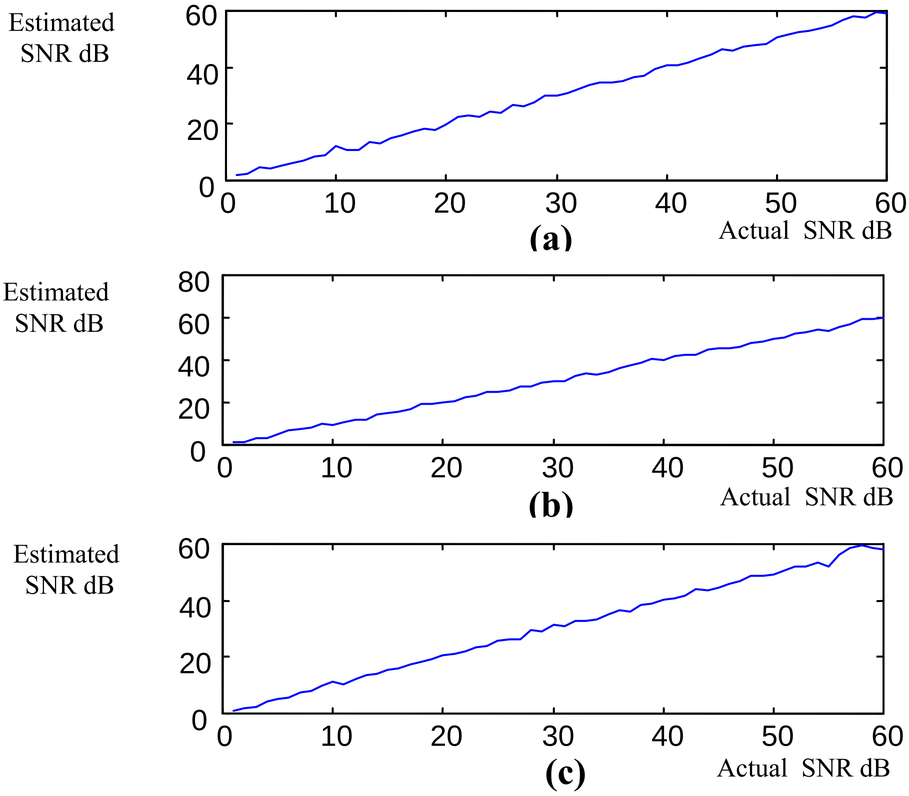

Figure (3) Estimated SNR versus actual SNR over (a) Channel 1 (b) Channel 2 (c) Channel 3.

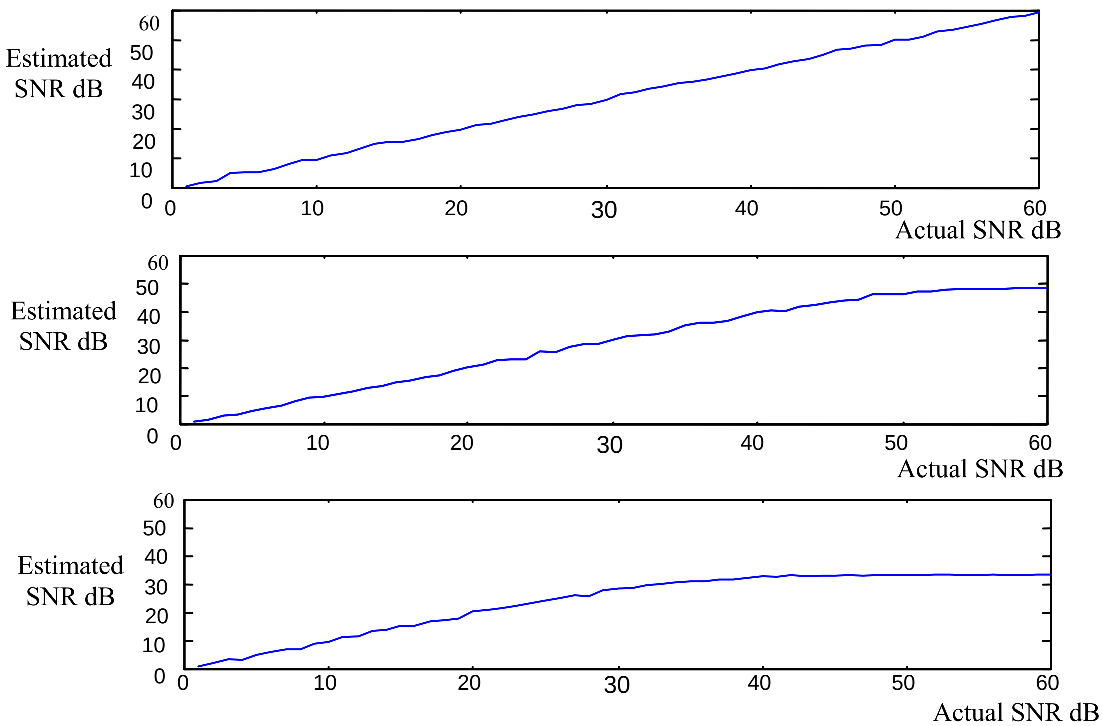

Figure (4) Estimated SNR versus actual SNR over (a) Channel 1 (b) Channel 2 (c) Channel 3.

Table 2 Data for f_{sp} estimation.

Channel condition		$f_{r.m.s.}$	F_i	f'_{sp}	%error ε
Delay	f_{sp}				
1	0.5	0.3492	1.4318	0.5166	3.2
	1	0.7614	1.3134	1.1263	11.2
	2	1.4308	1.3978	2.1167	5.5
2	0.5	0.3751	1.3329	0.5549	9.9
	1	0.5779	1.7305	0.8549	16.9
	2	1.4545	1.3751	2.1516	7.0
3	0.5	0.2552	1.9593	0.3775	32
	1	0.6895	1.4503	1.0200	1.9
	2	1.5112	1.3235	2.2356	10.5

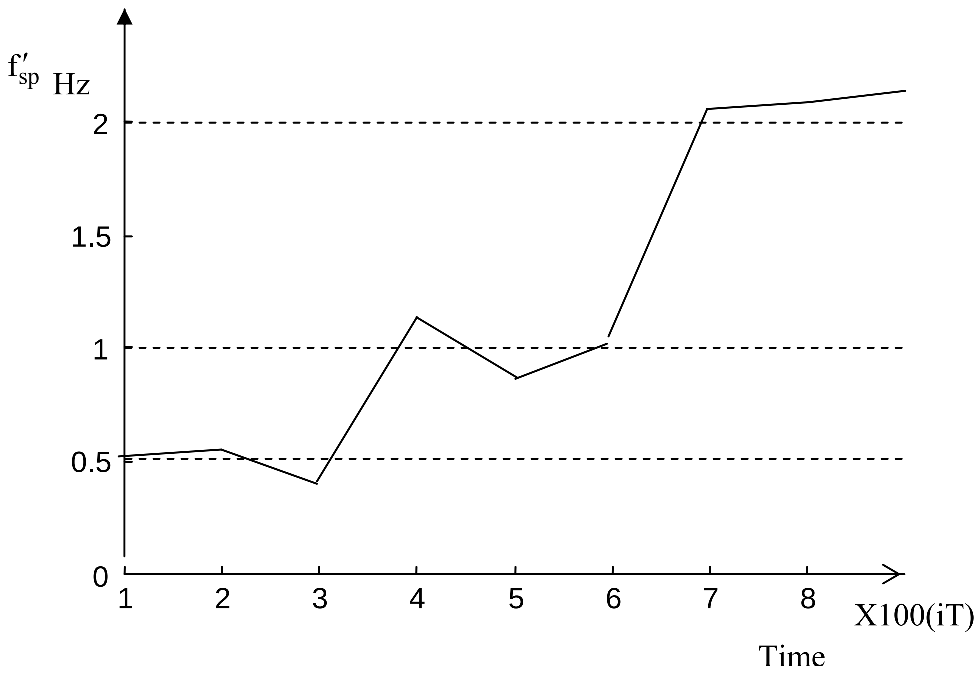

Figure (5) Estimated f_{sp} over the f_{sp}-varying channel.