Project	**IEEE 802.16 Broadband Wireless Access Working Group**
Title	**Equalization Requirements for Millimeter-Wave Fixed Broadband Wireless Access Systems**
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Re:	Response to the PHY group Chair's oral call for contributions at meeting #1 for models with which to evaluate candidate PHY proposals, and subsequent call for contributions from the IEEE 802.16 PHY Task Group, on Sept. 22, for a proposed PHY solution for broadband wireless systems.
Abstract	Wideband impulse response measurements have been made in residential and light industrial areas of Ottawa, at 29 GHz. Based upon these measured data, we have estimated the parameters of adaptive equalizers that would be required for LMCS/LMDS systems receiver over a range of bit rates. The results should be useful as a guide to specifying modulation schemes and a channel model for evaluation of candidate PHY layers for the 802.16 interoperability standard.
Purpose	For 802.16.1 to consider the input to form a standardized channel model for evaluation of candidate PHY layers for the 802.16 interoperability standard.
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Equalization Requirements for Millimeter-Wave Fixed Broadband Wireless Access Systems[1]

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I. Introduction

One of the impediments in achieving high data rate (10'to 100's Mbps) communication in an outdoor wireless environment is the multipath channel between the transmitter (Tx) and the receiver (Rx) causing intersymbol interference (ISI), which can limit the maximum achievable data rate. Given a linear serial digital modulation scheme, a receiver can use an adaptive equalizer to compensate for the average range of expected channel amplitude and delay characteristics and thus combat the detrimental effects caused by ISI. The analysis for equalization requirements in this research is based on a large number of measured multipath profiles of the broadband wireless channel in the Parkwood Hills and Kanata, residential and light industrial neighborhoods of Ottawa, Canada.

II. System model description

We consider input data at various data rates of 10 M Symbols per second (Msps), 25 Msps and 50 Msps respectively using QPSK modulation format, thus corresponding to bit rates of 20, 50 and 100 Mb/s, respectively. The Tx and Rx filters are square-root raised cosine filters with roll-off factor of 0.3. The channel is the impulse response (IR) data obtained from the outdoor measurements. It is known that a DFE (decision feedback equalizer) with feedback (FB) taps and with a symbol-spaced feedforward (FF) filter is sensitive to the timing error in sampling at the output of the equalizer. A fractionally spaced (FS) equalizer can alleviate this problem [3]. In our analysis, we have used both symbol-spaced and FS (T/2) DFE and compared their performance. We use the mean-square-error (MSE) as the performance index at the output of the slicer, to adjust the tap weights of both filters of DFE to minimize MSE of the estimation error $e(n)$; i.e. we minimize [3]

$$MSE = J = E\left|e(n)\right|^2 = E\left|d(n-D) - y(n)\right|^2$$

where $y(n)$ is the input to the detector at time n, $d(n-D)$ is the desired signal and D is the delay of the system. For an optimum set of taps in FF and feedback (FB) filters, the MSE, J, is minimized to the minimum mean square error (MMSE), J_{min}, given by:

$$MMSE = J_{\min} = s_d^2 - \mathbf{P}^H \mathbf{R}^{-1} \mathbf{P} = s_d^2 - \mathbf{P}^H \mathbf{W}_{opt}$$

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where s_d^2 is the variance of the desired response (assumed to be 1 in our analysis) and $\mathbf{W}_{opt}$ is the set of optimum tap weights for FF and FB filters. Here, vector P is the cross correlation of the tap inputs of forward and feedback filters and the desired signal, and R is a correlation matrix of tap inputs of FF and FB filters.

III. Data evaluated and evaluation criteria

The impulse responses used for this evaluation had been measured by the Communications Research Center (CRC) and Carleton University, at 29.5 GHz frequency in residential and industrial areas of Kanata and Parkwood Hills between a typical cell site transmitter and many receive locations through the use of directional antennas at both ends[2]. For wideband (WB) measurements in Kanata, the Tx antenna, mounted at the rooftop of a high rise building, was a vertically polarized horn antenna with gain of 18 dBi, elevation beamwidth of +/- 11.5 deg., azimuth beamwidth of +/- 8 deg. and the Rx antenna was a vertically polarized cassegrain reflector antenna with beamwidth of +/- 1 deg. in both elevation and azimuth, with gain of 36 dBi. In Parkwood Hills, the Tx antenna, mounted on the rooftop of a high rise apartment building, was the same vertically polarized horn antenna, while the receiver antenna was a vertically polarized horn with +/- 6 degree beamwidth in both azimuth and elevation, with a 23 dBi gain. During the WB probe, the measurement method involved transmitting a BPSK signal at 29.5 GHz modulated by a 63 bit long PN sequence generated at 40 Mbps and then receiving the signal by employing a quadrature (I and Q) receiver. The demodulated quadrature signals were sampled at a rate of 100 M samples/sec. Following the I-Q demodulation process at Rx, the resultant complex envelope of the received signals were sampled and stored. Subsequently, sampled complex envelope was crosscorrelated with a replica of the transmitted sequence, resulting in an estimate of the low pass equivalent of the radio channel complex IR. A thresholding technique, described in [4] was applied to the recorded IRs to reduce the impact of noise on measured data and extract the valid IR echoes.

A typical delay profile i.e. magnitude of the received complex impulse response obtained in such a manner from the Kanata data, with +/- 1 degree receive antenna, is shown in Fig. 1. Fig. 2 shows the CDF (cumulative distribution function) of the multipath spread in measured Parkwood Hills data for the cases of receive antenna placements of 0 and +/- 6 deg. The 90th percentile for the multipath spread equals 90 ns in this plot. The vertical axis in Fig. 2 could be interpreted as the fraction of locations that would have acceptable multipath spread, given an equalizer capable of handling the delay spread given in the horizontal axis. Small multipath spread implies that for some of the lower transmission rates, the system may not require any equalizer at all. It should be pointed out that the measured delay profiles only cover the propagation multipath, and that indoor-to-outdoor triple-transit cabling reflections could add to the overall multipath delay spread.

[2] We are grateful to Dr. Robert Bultitude and his colleagues at CRC for their efforts in obtaining and providing the data.

IV. Impulse response analysis results

To explain our results regarding equalization requirements at different symbol rates, we consider the effect of varying the number of FF and FB filter taps on the output S/N ratio which is related to the MMSE by $-10\log_{10}(MMSE)$, where received baseband average signal power is assumed to be unity. We assume an input SNR of 20 dB, and determine the optimum combination of the FF and FB taps which could give us an output SNR close to this. We represent the variation of the output SNR with respect to FF and FB taps through contour plots, which are level lines above the plane of FF and FB taps in the units of the output SNR. All the following results are outcomes of the treatment of the example IR shown in Fig. 1. This 90 ns multipath spread is amongst the worst observed multipath spreads.

A) 10 Mega Symbol Per Second

The contour plots of Fig. 3 are for fractionally-spaced (FS) DFE case with transmission rate of 10 Msps. Here, we notice that the LMCS system receiver may not require an equalizer as we can achieve an output SNR of over 19 dB (this corresponds to the lower left corner of Fig. 3) with matched filter detection; but we see that there is an improvement in the performance by using an equalizer over no equalizer (a (1,0) DFE). Here, the notation (a,b) represents a FF filter of length ‘a’ symbols, and a FB filter with ‘b’ taps. We notice, using one FB tap, the system performance improves to 19.86 dB. The horizontal axis here is the length of FF filter in symbol durations and since there are 2 taps per symbol period, number of taps is the horizontal axis value multiplied by 2.

As expected [3], the FS DFE is insensitive to the choice of timing phase; the T-spaced DFE with optimal symbol timing, achieves nearly similar performance to the FS DFE, although it is more sensitive to a non-optimum timing phase.

B) 25 Mega Symbol Per Second

Fig. 4 corresponds to the 25 Msps case for FS DFE, and we found that for this transmission rate the system would require an equalizer. For a T-spaced equalizer case, 2 to 3 FF and 1 FB taps seemed an optimum choice which gives an output SNR of better than 18.0 dB at the best sampling phase. Here, we found that a FF of minimum 2 symbol durations was required to achieve independence from the sampling phase and that (4,0) or (4,1) tap combination resulted in acceptable performance. Also, on comparing the results for symbol-spaced DFE and FS-DFE, we found the latter superior in performance over the other for same number of taps.

C) 50 Mega Symbol Per Second

At 50 Msps, for the symbol-spaced case, we find that DFE needs to be (5,1) to achieve an output SNR of more than 18 dB at the best sampling phase. For FS-DFE, (6,1) or (8,1) taps is found to be optimum resulting in output SNR of more than 18 dB (see Fig. 5).

V. MLSE using Viterbi Algorithm (VA)

MLSE (Maximum Likelihood Sequence Estimation) using the Viterbi Algorithm [5] as an alternative to DFE is evaluated here. The limitations to the applicability of MLSE to practical receivers is computational complexity associated with VA for large delay spreads. This computational load would be substantial at symbol rates, such as 50 Msps or higher when the channel dispersion can be as high as 4-5 symbol durations or more. Here we tried to find out the complexity along with the performance of the VA if suboptimal techniques of using a shorter impulse response are employed [6][7] i.e. by partially equalizing the channel so that the overall sampled response seen by the VA is significantly different from zero over only a small number of samples and any remaining ISI is considered to be noise. The linear portion of DFE (FF filter) can be assumed to work as a prefilter and in that case the initial sample of the overall desired IR (DIR) is rendered as large as possible compared to the additive noise. The overall sampled IR seen by the VA is of the type $\{F_k\}$, k=0,1,..,FB, where FB are the total number of the feedback taps in the DFE and F_0 ~1 [7]. The VA makes decisions on the assumption that the DIR is the actual overall channel IR, which is much shorter than the actual channel IR.

If the DFE has only one feedback tap, then the overall impulse response presented to VA has only two samples, thus has much reduced complexity and much simplified processing. For QPSK signalling, it results in a trellis with 4 states, or 16 states if we assume 2 FB taps in DFE. Hence, we can think of the cascade of the Tx filter, the channel, the Rx filter and FF filter of the DFE optimally conditioning the IR of the channel to approximate an IR of limited duration for which MLSE of the data is implementable in practice though this approximation does lead to some performance degradation. Using this we got the bounds for the probability of error (*Pe*) when MLSE is employed with shortened impulse response [6]. The upper bound on performance is computed by using the Pe equation suggested by Forney [5], by computing MMSE and the minimum coding distance of DIR. The MMSE includes the additive noise plus any residual ISI and is effective noise seen by VA. The *Pe* in this case is approximated by:

$$Pe = KQ(d_{\min} / 2s\,) = KQ(\sqrt{d_{\min}^2 / 4MMSE})$$

where d_{min} , the minimum energy of any non zero signal, s^2, the noise variance, and K - a small constant are defined in [5]. Using the optimum structures of the DFE, which minimize the MSE at the equalizer output, estimates of the error rate performance of DFE have been made. In the case of DFE, the MMSE may be roughly related to the performance in terms of the probability of symbol error for QPSK signalling via:

$$Pe \approx 2Q(\sqrt{|F_0|^2 / 2MMSE})(1 - \frac{1}{2}Q(\sqrt{|F_0|^2 / 2MMSE})$$

The probability of symbol error bounds for MLSE and DFE through above equations have been plotted in Fig. 6, for the IR of one of the channels, and using the MMSE calculated at 50 Msps with a FS-DFE of (8,1). For comparison, we have also plotted the matched filter bound of the probability of symbol error for QPSK modulation. We see from this plot that the MLSE has a performance edge over the DFE. This performance edge is at the cost of added complexity of MLSE over DFE which is equal to the complexity of DFE (as it includes a DFE (front end filter)) plus number of states times computations per state.

VI. Results summary

Similar evaluations were made for all the channels measured in Parkwood Hills and Kanata. The distributions of the output SNR for fractional spaced DFE's at 25 Msps ((4,1) DFE) and 50 Msps ((8,1) DFE) are summarized in Fig. 7 and Fig. 8, respectively. It is seen that the output SNR's cluster around or above 18 dB, except from some measured links with high path loss or unfavorable antenna positioning or pointing. Since at each location the antenna, the measurements were done at 4-5 closely located sub locations, so it was found that each of these abnormal cases could be dealt with by proper receive antenna positioning or placement.

VII. Conclusions

The study showed that for data rates of the order of 10 Msps using narrow beamwidth directive antennas, it is possible to avoid an equalizer. A similar conclusion was also drawn from measurements reported in [9]. For higher symbol rates of the order of 25 Msps, a (3,1) DFE seems effective for the symbol spaced case. The equalization requirements grow to (4,1) or (5,1) taps for still higher symbol rates of the order of 50 Msps for a symbol spaced DFE. The analysis with a fractionally-spaced (T/2) DFE showed requirement of (4,1) and (8,1) taps at 25 Msps and 50 Msps respectively for acceptable performance. Also, a MLSE can be employed with acceptable complexity when suboptimal measures such as approximating the overall system IR with its shorter version are employed and its performance has been shown to be better than that of DFE.

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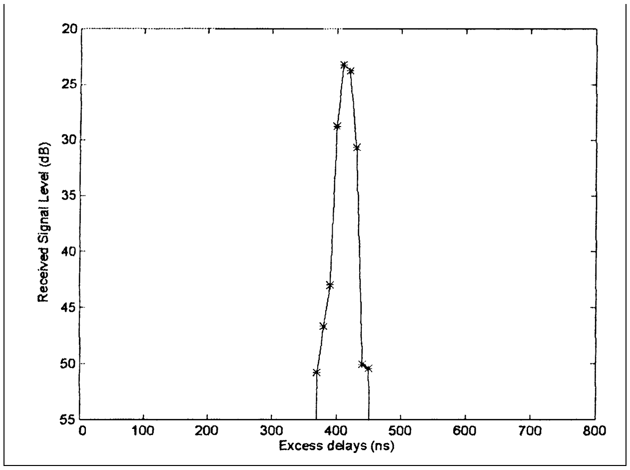

Fig. 1 Power delay profile for channel kab29044 at 0^o

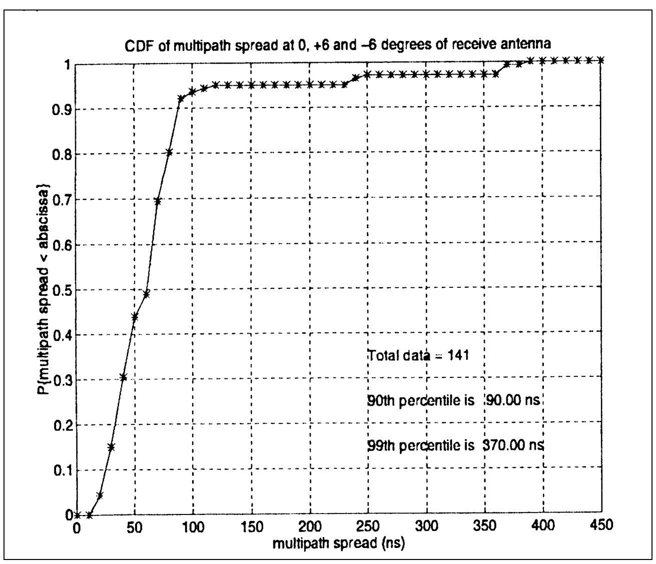

Fig. 2 CDF of multipath spread for receive directions of 0 and +/- 6 deg. for Parkwood Hills data

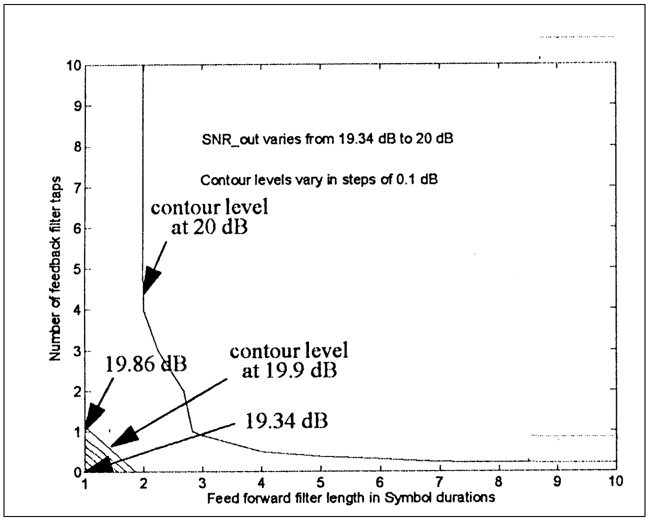

Fig. 3 Contour plots of SNR_out variation with FF filter and FB taps at 10 Ms/s (forward filter T/2-spaced)

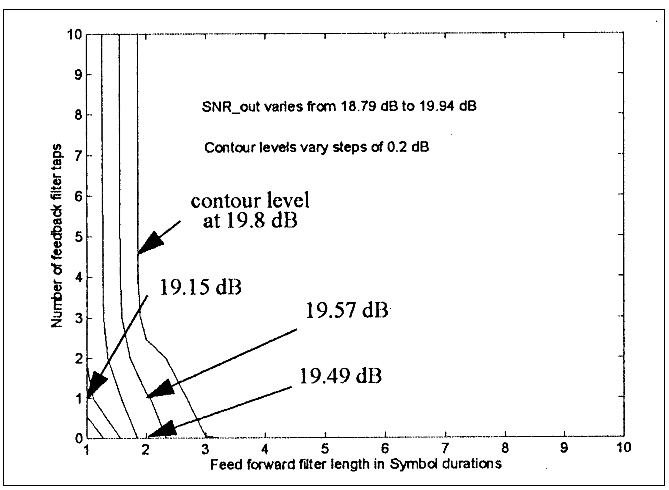

Fig. 4 Contour plots of SNR_out variation with FF filter and FB taps at 25 Ms/s (forward filter T/2-spaced)

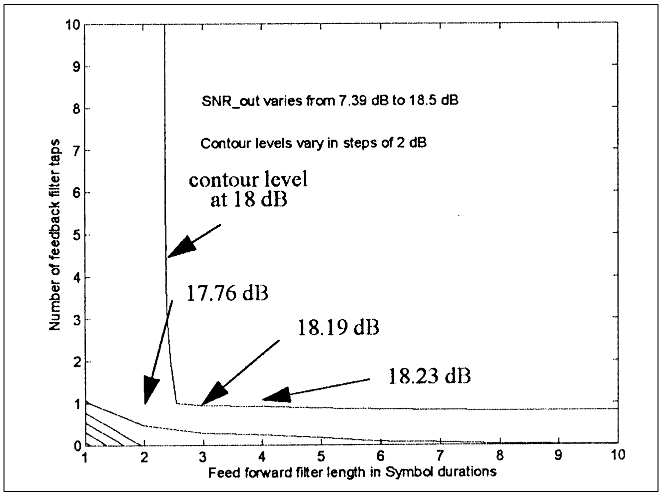

Fig. 5 Contour plots of SNR_out variation with FF filter and FB taps at 50 Ms/s (forward filter T/2-spaced)

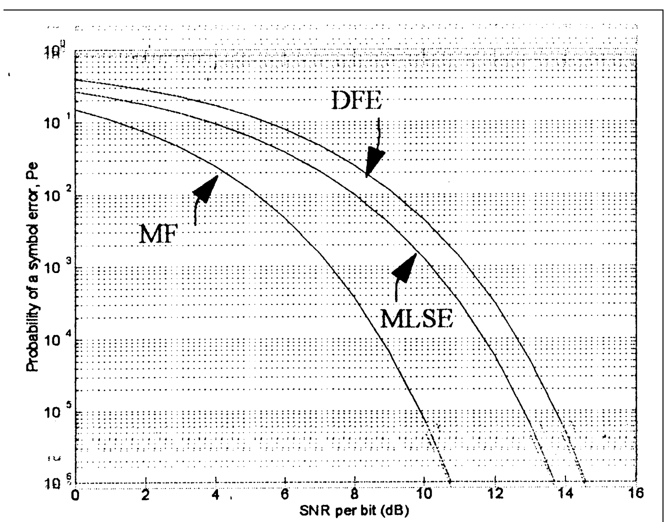

Fig. 6 Symbol error probability bounds for MLSE and DFE for QPSK, for one impulse response at 50 Ms/s, with MMSE calculated for (8,1) FS-DFE

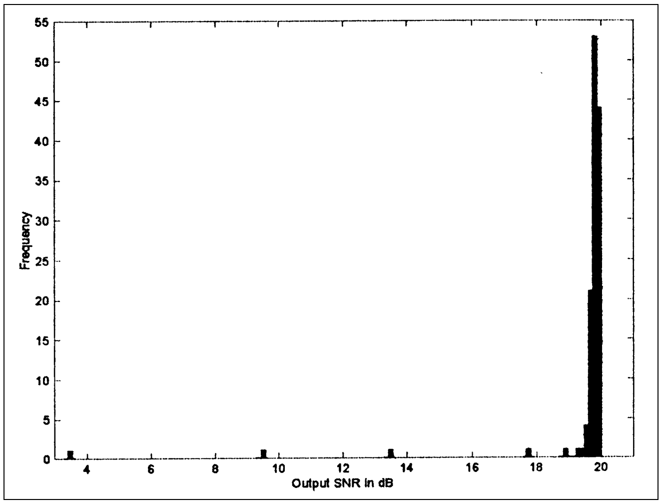

Fig. 7 Histogram of output SNR (dB) for the channels of Parkwood Hills and Kanata with a FS-DFE of (4,1) at 25 Ms/s

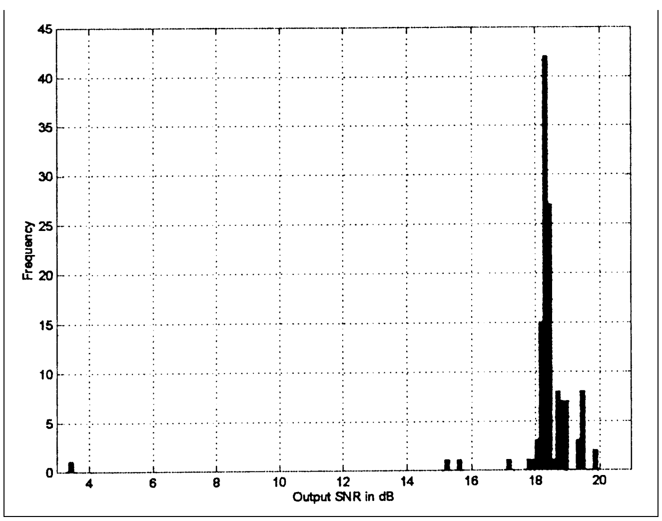

Fig. 8 Histogram of output SNR (dB) for the channels of Parkwood Hills and Kanata with a FS-DFE at of (8,1) at 50 Ms/s