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Title	**Channel Quality Measurement Method for 802.16ab PHY Layers**
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Re:	Call for Comments: Measurement of RSSI and CCI in the 802.16ab PHY; Source Document: IEEE 802.16ab-01/09; Date: 2001-09-25
Abstract	Frequency domain method for channel quality measurement (RSSI , CCI-plus-noise and SINR) is described. Method can be used for single-carrier, OFDM, or OFDMA PHY layers.
Purpose	Provide 802.16ab PHY layer measurement method which provides channel quality measurements to the MAC layer.
Notice	
Release	
Patent Policy and Procedures	

Channel Quality Measurement Method for 802.16ab PHY Layers

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Background and Introduction

Due to the high cost of licensed frequency bands mass deployments of WPANs, WLANs and WMANs in license-exempt frequency bands will occur in the near future. Devices and systems such as cordless phones, amateur radios, satellite systems, synthetic aperture radars, road transport and traffic telematics systems, dedicated short-range systems, etc., will also operate in the same license-exempt bands. To operate reliably and to minimize interference devices operating in license-exempt bands should adapt to their operating environment. Specifically, by detecting signals transmitted within frequency bands available to a device, selecting the best available frequency band, and operating in the selected band with minimal power in order to minimize interference. That is, a wireless device must adaptively detect all available channels, select and connect to the optimal channel available, and operate within the selected channel at minimal power required for a desired quality of service.

Two adaptive methods which help mitigate interference and coexistence problems are dynamic frequency selection (DFS) and transmit power control (TPC). DFS is used for selecting the best frequency band available and TPC for interference control within the frequency band selected. Together DFS and TPC adaptively improve the reliability and quality of a link for a device and minimize interference to other operating devices. Both the IEEE 802.11h and IEEE 802.16b task groups are working towards incorporation of DFS and TPC within their PHY and MAC layers. IEEE 802.15 is incorporating adaptive frequency hopping to address interference and coexistence problems.

Channel quality measurements are crucial in the implementation of DFS and TPC methods. Obtaining accurate channel quality measurements can be difficult because of interference-plus-noise signals which are nonstationary. The statistical characteristics of interference-plus-noise signals are difficult to describe. Channel quality measurements can be easily and accurately computed in the frequency domain without statistical knowledge using a nonparametric signal processing method. This is especially true for IEEE 802.16ab PHY layers which require the use of a fast Fourier transform processor. In addition to accuracy and stability, a nonparametric frequency domain method allows channel quality measurements to be computed for specific sub-carriers within an OFDM symbol. In fact, channel quality measurements can be computed on frame-by-frame basis for specific sub-carriers or sub-channels within an OFDMA symbol.

In this document a frequency domain method for channel quality measurement is described. The method allows a subscriber station (SS) or base station (BS) to compute channel quality measurements on a frame-by-frame basis. Time-averaged channel quality estimates are readily available and can be provided for channel quality updates whenever needed by a DFS or TPC algorithm. Quantized channel quality measurements can be easily included in the TLV (Type-Length-Value) field of a channel state feedback message.

Description of the Measurement Method

Each frame transmitted on the downlink or uplink in IEEE 802.16ab compliant system contains a preamble. An IEEE 801.16ab compliant receiver uses the preamble for signal detection, AGC convergence, carrier recovery, symbol timing recovery and equalization. As described below the preamble can be used for another signal processing task: channel quality measurement. In the description and simulations which follow we employ the OFDM PHY. However, we emphasize that the method can be used for any of the IEEE 801.16ab PHY layers.

Measurement of Mean CCI-plus-Noise Power

Assume that the guard interval of an OFDM preamble has been removed. The *N* consecutive complex-valued samples of the preamble of the *n*th received frame can then be described as

$$y_k[n] = h_k[n]x_k[n] + v_k[n], \quad k = 0,1,\ldots,N-1$$

Here *k* denotes sub-carrier number, $h_k[n]$ denotes a sample of a time-varying complex-valued channel process, $x_k[n]$ a sample of a known preamble and $v_k[n]$ a sample of complex-valued nonstationary process which models CCI-plus-noise. The time-varying probability distribution of the CCI-plus-noise can be any continuous distribution since the proposed method is nonparametric.

Let $S_{xx}(f)$, $S_{yy}(f)$ and $S_{xy}(f)$ denote windowed (Hann, Hamming, Kaiser, etc.) FFT-computed power spectral density (PSD) estimates of the transmitted preamble, received preamble and cross-PSD estimate between the transmitted and received preambles. A measure of the correlation between sub-carriers of the transmitted and received preambles at frequencies f_k, $k = 0,1,\ldots,N-1$, is given by the squared-coherence [1]-[4] defined as

$$C_{xy}(f_k) = \frac{\left|S_{xy}(f_k)\right|^2}{S_{xx}(f_k)S_{yy}(f_k)}$$

The squared-coherence satisfies the inequality

$$0 \le C_{xy}(f_k) \le 1$$

and may be interpreted as a frequency domain correlation coefficient. As shown in [1], [2] and [4] a nonparametric estimate of the CCI-plus-noise PSD can be written as

$$S_{vv}(f_k) = (1 - C_{xy}(f_k))S_{yy}(f_k)$$

Summing the received sub-carrier PSD estimates $S_{vv}(f_k)$ and applying a logarithmic operator gives the received CCI-plus-noise power estimate

$$P_{CCI}[n] = 10\log_{10}\left(\sum_{k=0}^{N-1} S_{vv}(f_k)\right)$$

Note that logarithmic operation stabilizes the variance of the estimates [4]. For $n \ge 1$ time-averaged CCI-plus-noise power measurements can be provided using the recursion

$$\overline{P}_{CCI}[n] = \overline{P}_{CCI}[n-1] + \alpha(P_{CCI}[n] - \overline{P}_{CCI}[n-1])$$

where $\overline{P}_{CCI}[0] = P_{CCI}[0]$ and parameter $0 < \alpha < 1$ controls the degree of smoothing.

Measurements of Mean Received Signal Power

For the *n*th received frame an estimate of the *k*th component of the channel transfer function can be written as

$$H(f_k) = \frac{S^*_{xy}(f_k)}{S_{xx}(f_k)} = \sum_{k=0}^{N-1} h_k[n] e^{-ik2\pi/N}$$

where the asterisk denotes complex conjugation. Assume a preamble with statistical properties similar to uncorrelated white noise with unit variance. A received signal power estimate $P_{RS}[n]$ can be computed in the time domain by cross correlation or in the frequency domain by an FFT operation. Specifically from the Parseval equality we have

$$P_{RS}[n] = 10\log_{10}\left(\sum_{k=0}^{N-1} \left|H(f_k)\right|^2\right) = 10\log_{10}\left(\sum_{k=0}^{N-1} \left|h_k[n]\right|^2\right)$$

For $n \geq 1$ time-averaged RS power measurements can be computed using the recursion

$$\overline{P}_{RS}[n] = \overline{P}_{RS}[n-1] + \beta(P_{RS}[n] - \overline{P}_{RS}[n-1])$$

where $\overline{P}_{RS}[0] = P_{RS}[0]$ and parameter $0 < \beta < 1$ controls the degree of smoothing.

Note that during a network access procedure *total* RS power in a particular frequency band can also be measured passively by computing the PSD of a received signal. Let $S_{rr}(f)$ denote this PSD estimate. A simple statistical hypothesis test based on the Rayleigh distributed components of $S_{rr}(f)$ (see [2] and [4]) can then be used as a channel selection criterion.

Quantized SINR Measurements for DFS and TPC Algorithms

An estimate of the received SINR for the *n*th received frame is then

$$\overline{\rho}[n] = Q\left(\overline{P}_{RS}[n] - \overline{P}_{CCI}[n]\right)$$

where Q is an appropriately chosen quantization function.

Quantized SINR Measurements for OFDMA Sub-Channels

For an OFDMA PHY layer the proposed method can be used to measure SINR associated with subsets of sub-carriers (OFDMA sub-channels). Specifically, using the above we can write

$$P_{CCISubSet}[n] = 10\log_{10}\left(\sum_{k=K_1}^{K_2} S_{vv}(f_k)\right)$$

$$\overline{P}_{CCISubSet}[n] = \overline{P}_{CCISubSet}[n-1] + \alpha(P_{CCISubSet}[n] - \overline{P}_{CCISubSet}[n-1])$$

$$P_{RSSubSet}[n] = 10\log_{10}\left(\sum_{k=K_1}^{K_2} \left|H(f_k)\right|^2\right)$$

$$\overline{P}_{RSSubSet}[n] = \overline{P}_{RSSubSet}[n-1] + \beta(P_{RSSubSet}[n] - \overline{P}_{RSSubSet}[n-1])$$

where integers K_1 and K_2 denote the start and end points of a subset of sub-carrier frequencies. Then

$$\overline{\rho}_{SubSet}[n] = Q\left(\overline{P}_{RSSubSet}[n] - \overline{P}_{CCISubSet}[n]\right)$$

Computational Issues

Let **x** and **y** denote transmitted and received preamble sequences. Using the FFT the following PSD are computed

$$S_{xx}(f) = \left|FFT(\mathbf{x})\right|^2, \quad S_{yy}(f) = \left|FFT(\mathbf{y})\right|^2, \quad S_{xy}(f) = \left(FFT(\mathbf{x})\right)^*\left(FFT(\mathbf{y})\right)$$

Hence

$$C_{xy}(f) = \frac{\left|S_{xy}(f)\right|^2}{S_{xx}(f)S_{yy}(f)} = \frac{\left|\left(FFT(\mathbf{x})\right)^*\left(FFT(\mathbf{y})\right)\right|^2}{\left|FFT(\mathbf{x})\right|^2\left|FFT(\mathbf{y})\right|^2} \equiv 1$$

To avoid this problem the coherence must be computed by segmenting and windowing operations. See [1] or the MATLAB implementation of cohere() for more information.

Note that in an implementation $S_{xx}(f)$ can be computed off-line, only $S_{yy}(f)$ needs to be computed for each received frame. To see this recall that the cross-spectral density can be computed as

$$S_{xy}(f) = S_{xx}^*(f)S_{yy}(f) = \left(FFT(\mathbf{x})\right)^*\left(FFT(\mathbf{y})\right)$$

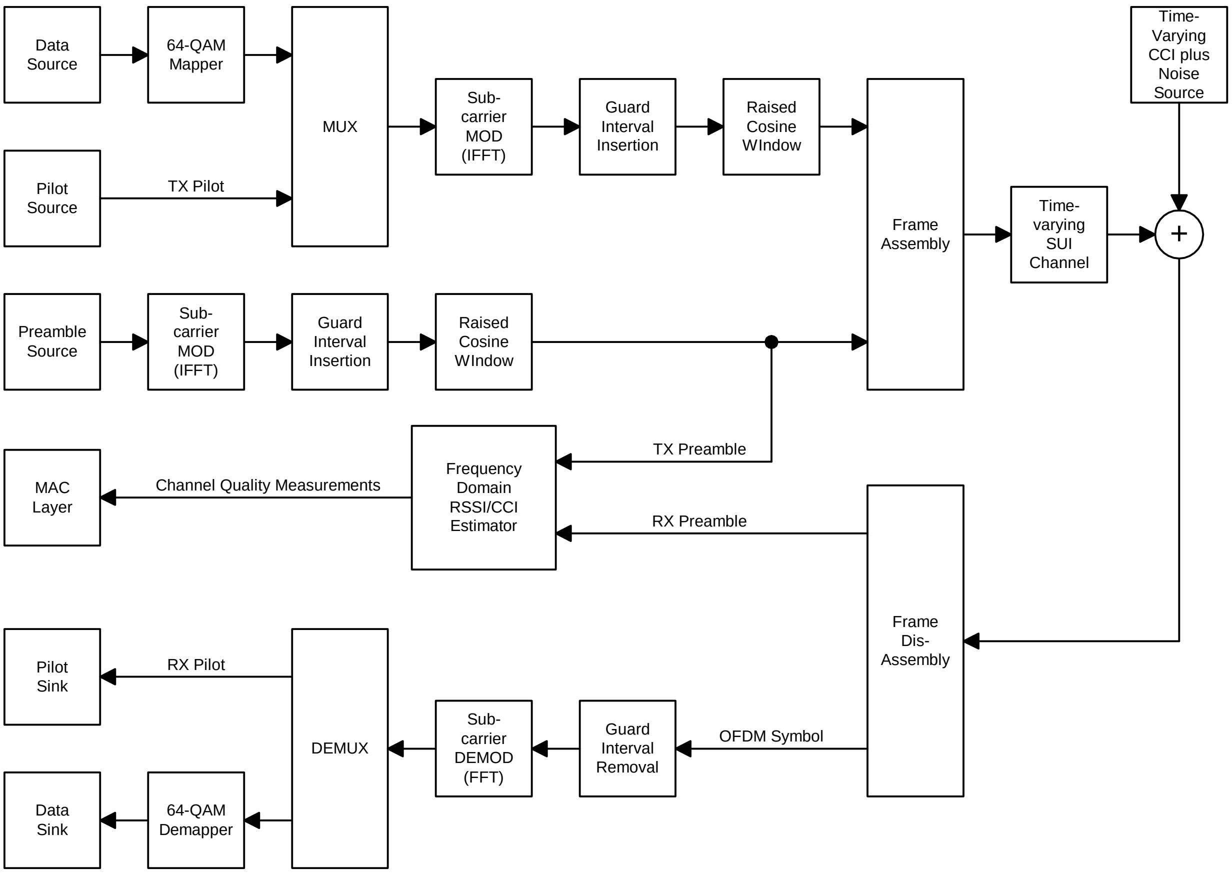
Thus if $S_{xx}^*(f)$ is stored in a look-up table only one extra FFT for $S_{yy}(f)$ needs to be computed in an implementation. Similarly

$$H(f) = \frac{S_{xy}^*(f)}{S_{xx}(f)} = \frac{\left(FFT(\mathbf{x})\right)^*\left(FFT(\mathbf{y})\right)}{\left(FFT(\mathbf{x})\right)^*\left(FFT(\mathbf{x})\right)} = \frac{FFT(\mathbf{y})}{FFT(\mathbf{x})}$$

so only $S_{yy}(f)$ needs to computed to estimate $H(f)$ for each received frame. Computations can be further reduced by computing $S_{yy}(f)$ in periodic manner rather than for every received frame.

Simulation Model, Procedure and Results

The proposed method can be incorporated into any of the 802.16ab PHY layers: Single-carrier, OFDM, or OFDMA. To verify the above described method simulations were run using the OFMA PHY layer. The following figure shows a functional block diagram of the baseband simulation model. With reference to the figure we now describe the simulation procedure.



Transmit Operations

A uniform random number generator produced data source bits which were mapped to a 64-QAM signal constellation via Gray coding. The pilot source produced a BPSK signal using the characteristic polynomial and initial state vector specified in Clause 8.3.5.3.2.2.2 of the standard. Symbols of the 64-QAM constellation signal and the BPSK pilot signal were combined by the pilot inserter in order to construct complex-valued OFDM symbols of length 256. The sub-carrier mapper mapped each length-256 vector input to 256 sub-carriers by an IFFT transformation. The resulting 256 samples of an OFDM symbol were comprised of 192 data sub-carriers, 8 pilot sub-carriers, 28 left-guard carriers and 27 right-guard carriers (see Table 214 of standard). A guard interval was pre-pended to OFMD symbol using the last 64 samples of the 256 sample OFDM symbol. The resulting 320 sample OFDM symbol was spectrally shaped by a raised-cosine window with a rolloff parameter of .025 and a cutoff of .975. Within the frame assembler a packet was constructed by concatenating four successive length-320 OFDM symbols.

The preamble consisted of one OFDM symbol pre-pended by a cyclic prefix equal in length to the cyclic prefix associated with a data symbol, i.e. 64 samples. The preamble source produced a BPSK modulated pseudo-random noise sequence. The sequence was produced using the characteristic polynomial and initial state vector specified in Clause 8.3.5.3.2.2.2 of the standard. Note that the described method allows any type of preamble source to be used. A total of 256 symbols of the BPSK preamble signal were collected and mapped to sub-carriers by the an IFFT operation. A guard interval of length 64 was pre-pended to the 256 samples giving a length 320 preamble symbol (equal in length to data symbol). The preamble symbol was spectrally shaped using the same raised-cosine window described above. Within the frame assembler a length 320 preamble was pre-pended to each OFDM packet to form an OFDM frame.

Channel Operations

Channel operations did not include a radio impairment model such as Rapp or Saleh power amplifier (PA) model. In general, a nonlinear PA will produce spectral dispersion or regrowth without input power back-off or the incorporation of a PA predistortion method. It was assumed that spectral dispersion was controlled.

Each frame output by the frame assembler was corrupted by one of the Stanford University channel models: SUI-1 through SUI-6. Channel tap power, propagation delay, Doppler frequency spread and Ricean 90% K-factor for an omni-directional antenna configuration were used. The values used for these parameters are listed in [5]. For each frame a different set of channel taps was used. Thus the channel model varied at the frame rate. The taps were produced using the algorithms described in [5].

CCI and complex-valued Gaussian noise were added to each channel model corrupted frame. The CCI signal was generated according to the interference model specified in Annex A.2.5.3 of the standard. Note that the annex describes a two-state Markov on/off model for the interference source model in order to generate interference bursts have specified arrival time and duration. The interference model for the simulations did not use the two-state Markov model. Rather a continuous interference source was used. The interference signal was produced in accordance with clauses A2.5.3.2 and A2.5.3.3.

Receive Operations

Perfect time and frequency synchronization were assumed at the receiver. For each received frame the preamble was extracted and passed to the RS/CCI estimator. The OFDM symbol data component of a received frame is not required and will not be mentioned again. Given the known transmitted preamble and the received preamble, $S_{xx}(f)$, $S_{yy}(f)$, and $S_{xy}(f)$ were computed using Welch's PSD estimation method. Welch's method [1] used to compute two-sided PSD using a length 64 Hamming window. The windowed preambles were overlapped by 32 samples and a length 512 FFT was employed. Using the above PSD estimates the squared-coherence $C_{xy}(f)$, channel transfer function $H(f)$, and CCI-plus-noise PSD $S_{vv}(f)$ were computed.

Simulation Results

The results of the simulations or displayed in the six figures at the end of the document. For each plot the dotted line serves as a reference for the ideal estimator.

Conclusions

A frequency domain method has been described for channel quality estimation. The proposed method is ideally suited for IEEE 802.16ab compliant systems which require an FFT module. The proposed method allows channel quality estimates to be computed for specific sub-carrier frequencies or clusters of sub-carriers. The increased computational complexity per frame is approximately equal to one FFT per received preamble. Simulation results show the proposed method is very accurate for the types of channels over which an IEEE 802.16ab complaint system will operate.

References

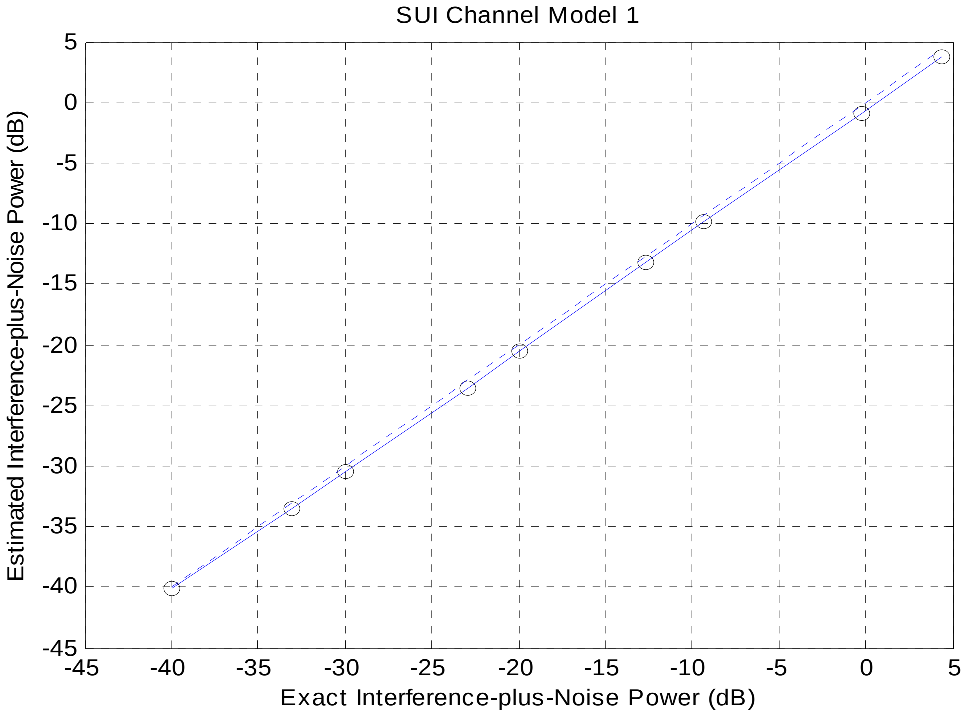
[1] Bendat, J. S., and A. G. Piersol, 1993, Engineering Applications of Correlation and Spectral Analysis, 2nd ed., John Wiley & Sons.

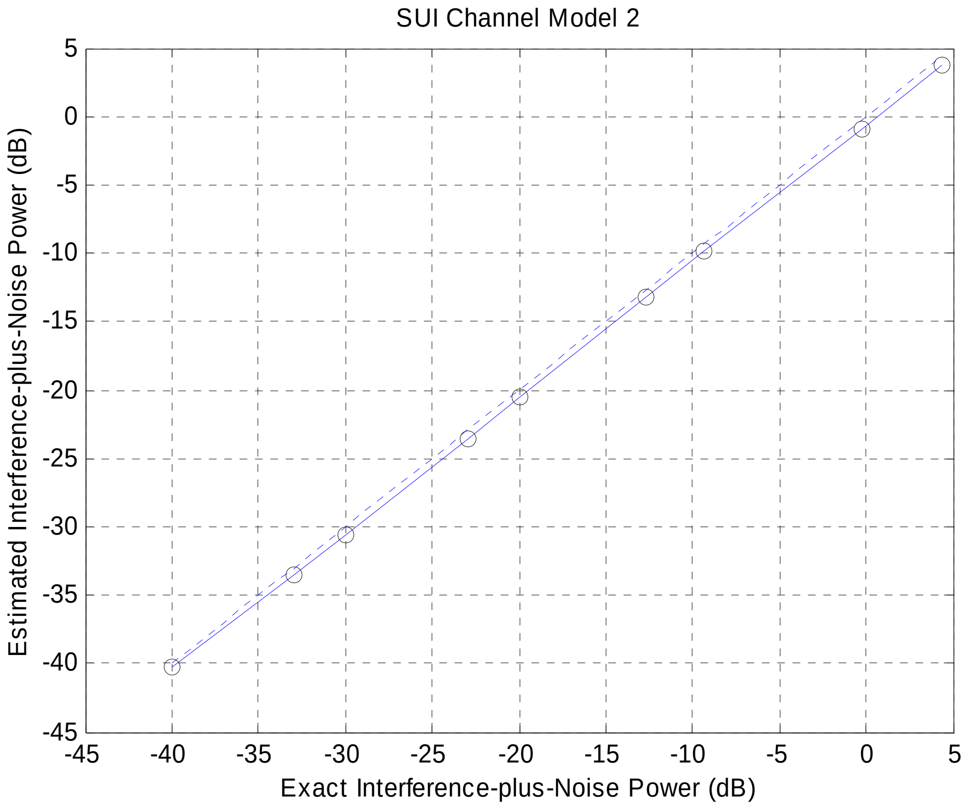
[2] P. J. Brockwell and R. A. Davis, Time Series: Theory and Methods, 2ed., New York: Springer-Verlag, 1991.

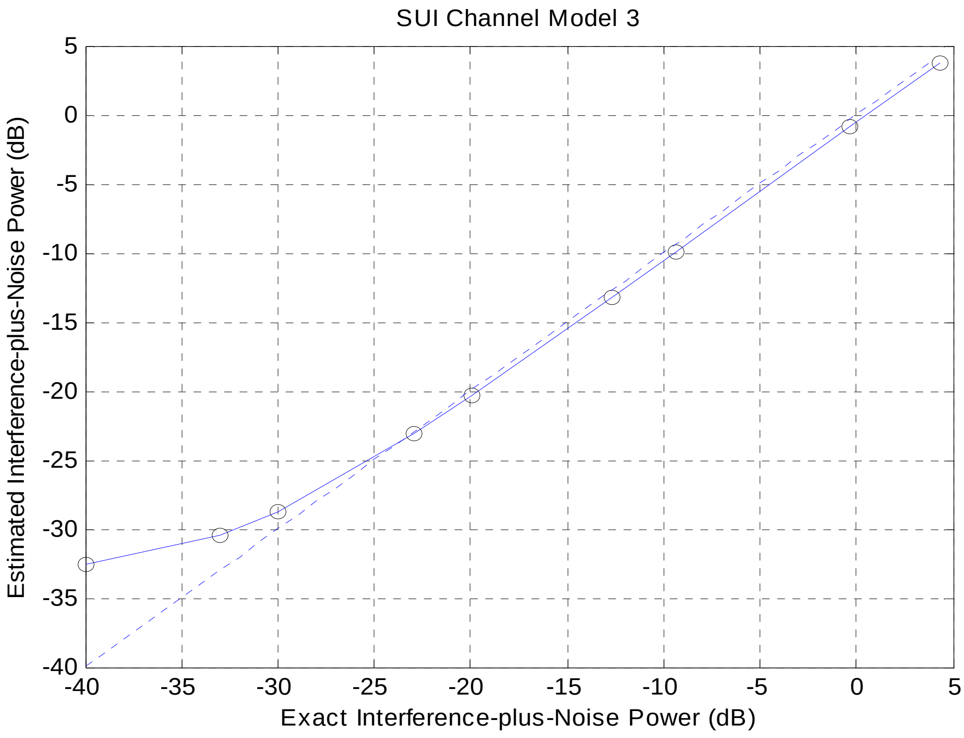
[3] A. Papoulis, Probability, Random Variables, and Stochastic Processes, McGraw-Hill, NY, 1984.

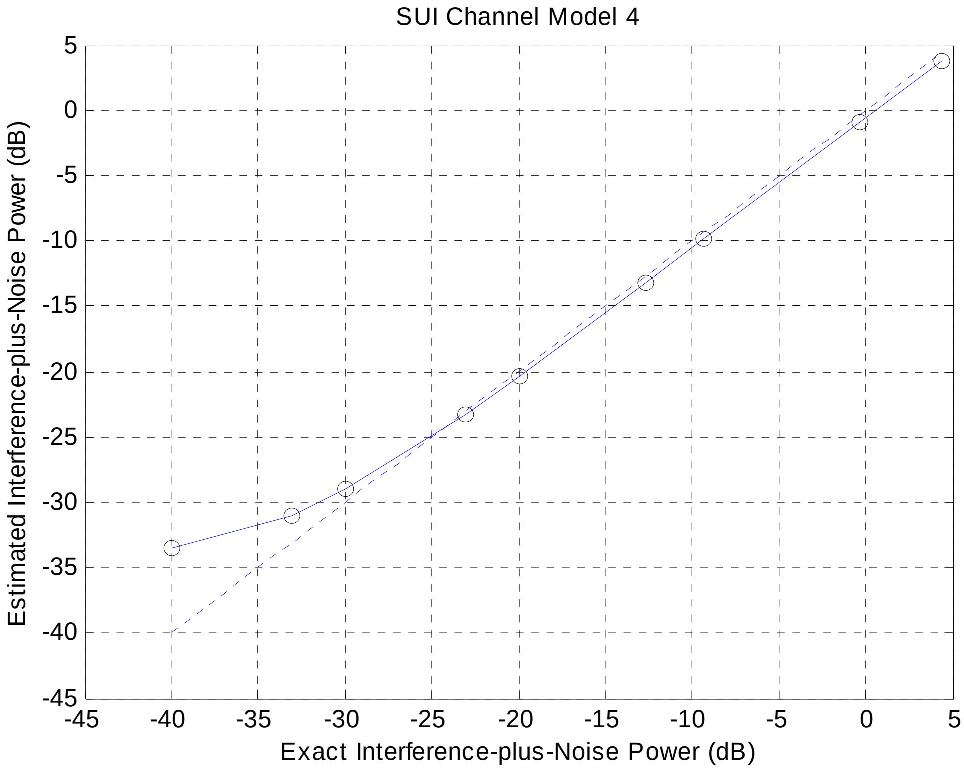
[4] Priestley, M.J., Spectral Analysis and Time Series, Academic Press, San Diego, 1996.

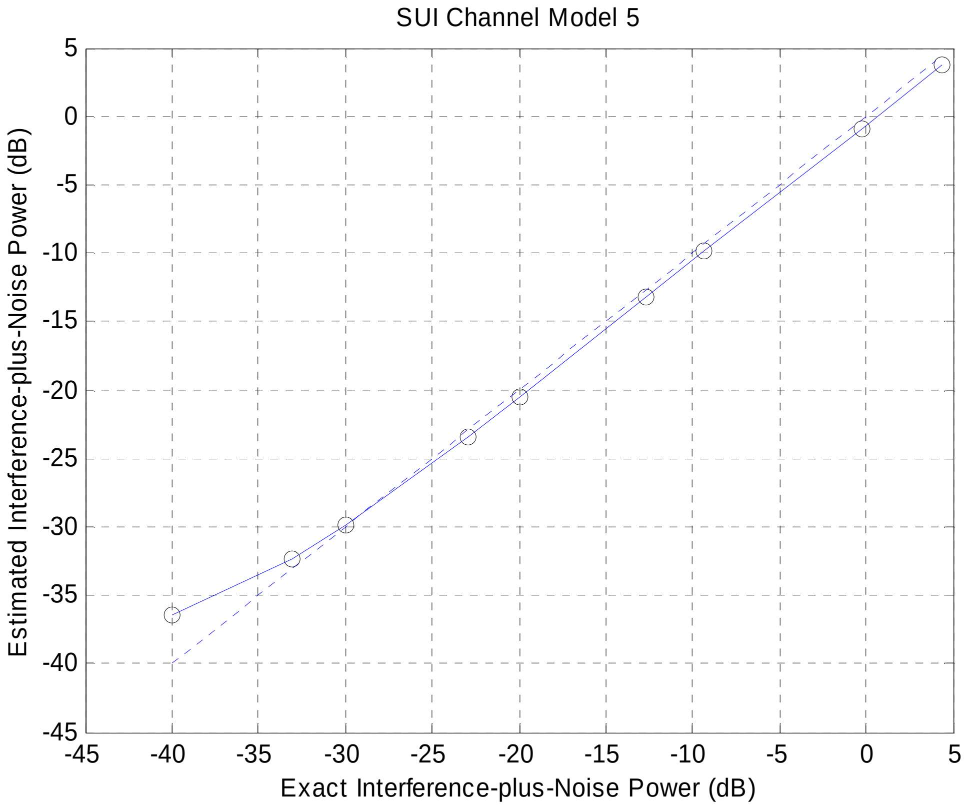
[5] V. Erceg, et al, "Channel Models for Fixed Wireless Applications," IEEE 802.16.3c-01/29r4.

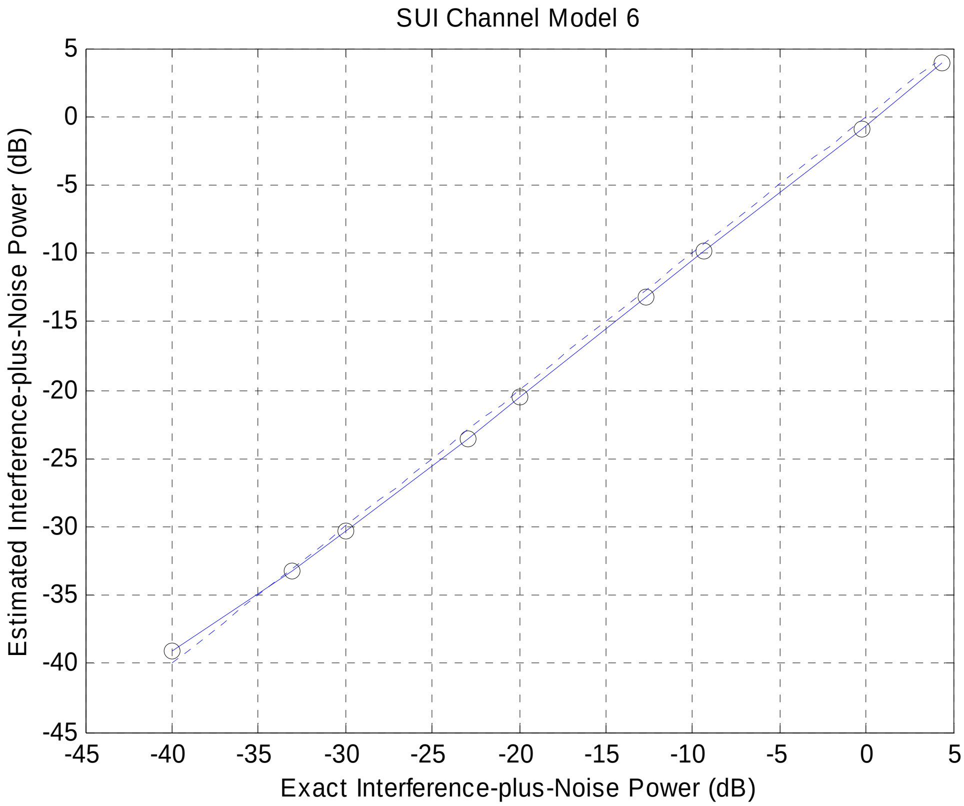



SUI Channel Model 2
Estimated Interference-plus-Noise Power (dB)
Exact Interference-plus-Noise Power (dB)
5
0
-5
-10
-15
-20
-25
-30
-35
-40
-45
-45
-40
-35
-30
-25
-20
-15
-10
-5
0
5


SUI Channel Model 3
Estimated Interference-plus-Noise Power (dB)
Exact Interference-plus-Noise Power (dB)
5
0
-5
-10
-15
-20
-25
-30
-35
-40
-40
-35
-30
-25
-20
-15
-10
-5
0
5


SUI Channel Model 4
Estimated Interference-plus-Noise Power (dB)
Exact Interference-plus-Noise Power (dB)
5
0
-5
-10
-15
-20
-25
-30
-35
-40
-45
-40
-35
-30
-25
-20
-15
-10
-5
0
5


SUI Channel Model 5
Estimated Interference-plus-Noise Power (dB)
Exact Interference-plus-Noise Power (dB)
5
0
-5
-10
-15
-20
-25
-30
-35
-40
-45
-40
-35
-30
-25
-20
-15
-10
-5
0
5


SUI Channel Model 6
Estimated Interference-plus-Noise Power (dB)
Exact Interference-plus-Noise Power (dB)
5
0
-5
-10
-15
-20
-25
-30
-35
-40
-45
-45
-40
-35
-30
-25
-20
-15
-10
-5
0
5