Transmission Signal Quality Metrics for FM IBOC Signals

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Table of Contents

Contents

List of Figures

List of Tables

1 Abstract

This technical report describes the measurement of several transmission signal quality metrics specified in reference [1] for FM In-Band-On-Channel (IBOC) signals defined in reference [2]. In particular, the metrics are intended to assess signal distortion due to both nonlinear high-power amplifier (HPA) characteristics as well as linear filtering. The effects of the Peak-to-Average Power Ratio (PAPR) reduction algorithm on the reference subcarriers and the data (non-reference) subcarriers are measured with different metrics, as appropriate. Group delay and gain variation across subcarriers may also be measured with the techniques described.

2 Acknowledgements

The selection of Modulation Error Ratio (MER) as the HD Radio transmission signal quality metric for the FM IBOC signal and the method of measuring MER on the FM IBOC signal was developed by a working group of technologists representing iBiquity Digital, Broadcast Electronics, Continental Electronics, Harris Broadcast, Nautel Ltd., and other interested participants from the US radio broadcast industry. This group of technologists reached full consensus on the standardized method for FM IBOC signal MER measurement described in this document.

The list of regular participants included:

Geoffrey Mendenhall, Working Group Chairman - Harris Broadcast
Kevin Berndsen – Harris Broadcast
Harvey Chalmers – iBiquity Digital
Jeff Detweiler – iBiquity Digital
David Gates – Cesium Communications
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A special acknowledgement is extended to Kevin Berndsen, Harvey Chalmers, Dave Hershberger, Brian Kroeger, and Phillip Schmid for doing the majority of the work to develop, test, and verify the HD Radio MER measurement method described in this document.

3 References

[1] NRSC-5-B, “HD Radio™ FM Transmission System Specifications.” Document Number: SY_SSS_1026s.

[2] NRSC-5-B, “HD Radio™ HD Radio™ Air Interface Design Description Layer 1 FM.” Document Number: SY_SSS_1011s.

[3] NRSC-G201, “NRSC-5 RF Mask Compliance: Measurement Methods and Practice.”

4 Abbreviations, Symbols, and Acronyms

4.1 Abbreviations, Symbols, Acronyms

BPSK	Binary Phase Shift Keying
dB	decibel
EVM	Error Vector Magnitude
FFT	Fast Fourier Transform
FM	Frequency Modulation
HPA	High Power Amplifier
Hz	Hertz
I	In-Phase
IBOC	In-Band-On-Channel
L1	Layer One
MER	Modulation Error Ratio
NRSC	National Radio Systems Committee
OFDM	Orthogonal Frequency Division Multiplexing
PAPR	Peak-to-Average Power Ratio
PLL	Phase-Locked Loop
PPM	parts per million
Q	Quadrature
QPSK	Quadrature Phase Shift Keying
RF	Radio Frequency
rms	Root Mean Square

5 Measurement Conditions

Prior to OFDM demodulation, it is first necessary to perform frequency and symbol synchronization commonly across all subcarriers. A suggested method to accomplish this is described in Appendix A.

The sample rate (nominally 744187.5 complex samples per second) should be locked to 2160 times the symbol rate (nominally 344.53125 Hz). It is important that the symbol synchronization is accurate, since any symbol timing error will cause a linear phase shift across the subcarriers. However, group delay variations across the subcarriers will prevent perfect timing for every subcarrier. Further timing correction for individual subcarriers should not be implemented since this is an error we want to measure. Of course, both frequency and symbol synchronization algorithms already exist for receivers. For the purposes of these static measurements in non-real-time, a block of at least 128 OFDM symbols can be used to estimate the frequency and symbol timing parameters as a block (instead of PLLs, etc.), simplifying the synchronization process.

The first step is to capture N contiguous symbols of 2160 complex baseband samples each, where N is the block size. For good results, N should be 128 or greater. A value of 512 symbols corresponds to a total time span of one L1 Frame period (1.486 seconds).

A cyclic folding with root-Nyquist pulse shape weighting operation should be applied to each OFDM symbol, converting the 2160 samples into 2048. Next, a 2048-point FFT should be computed for each of the N symbols. The bins of the FFT correspond to the OFDM subcarriers.

Some of the transmission signal quality measurements specified in reference [1] may be conveniently computed using the reference subcarriers. These measurements include gain variation, group delay variation and error vector magnitude (EVM). However, in place of EVM, a slight variant called modulation error ratio (MER) is proposed in order to be in better alignment with standard practices within the broadcast industry.

The subcarrier index m corresponds to the active reference subcarriers, where $m \in \{-546, -527, -508, -489, -470. -451, -432, -413, -394, -375, -356, -337, -318, -299, -280, 280, 299, 318, 337, 356, 375, 394, 413, 432, 451, 470, 489, 508, 527, 546\}$. The active subcarriers per service mode are given in Table 5-1. Refer to [2] for further details.

Table 5-1: Active Subcarriers per Service Mode

Service Mode	Active Reference Subcarriers	
	Lower Sideband	**Upper Sideband**
MP1	−546, −527, −508, −489, −470. −451, −432, −413, −394, −375, −356	356, 375, 394, 413, 432, 451, 470, 489, 508, 527, 546
MP2	−546, −527, −508, −489, −470. −451, −432, −413, −394, −375, −356, −337	337, 356, 375, 394, 413, 432, 451, 470, 489, 508, 527, 546
MP3	−546, −527, −508, −489, −470. −451, −432, −413, −394, −375, −356, −337, −318	318, 337, 356, 375, 394, 413, 432, 451, 470, 489, 508, 527, 546
MP5, MP6, MP11	−546, −527, −508, −489, −470. −451, −432, −413, −394, −375, −356, −337, −318, −299, −280	280, 299, 318, 337, 356, 375, 394, 413, 432, 451, 470, 489, 508, 527, 546

6 Phase Adjustment for each Reference Subcarrier

The average phase φ in radians (with ambiguity of π radians) of each reference subcarrier is computed as the argument (-π to π) of the sum of the squared reference symbol samples (where the reference symbol samples are denoted by r), divided by 2. The division by 2 is necessary because the squaring of the samples doubles the phase. This phase represents the phase of the BPSK symbol constellation point.

$$\varphi_m = \frac{1}{2} arg\left\{ \sum_{n=0}^{N-1} r_{n,m}^2 \right\}; -\frac{\pi}{2} \leq \varphi_m < \frac{\pi}{2}.$$

A sample rate (frequency) error between the exciter clock and the measurement equipment will result in a linear phase ramp over the time span for any one subcarrier. The span of this ramp is proportional to the sample error drift over the signal duration. The slope of the ramp is computed for each subcarrier index m.

$$slope_m = \frac{d\varphi_{n,m}}{dn} = \frac{1}{2} \cdot arg\left\{ \sum_{n=1}^{N-1} \left(r_{n,m} \cdot r_{n-1,m}^{*} \right)^2 \right\} \; ; \; -\frac{\pi}{2} \leq slope_m < \frac{\pi}{2}$$

The phase can now be estimated for each OFDM symbol, and for each reference subcarrier.

$$\varphi_{n,m} = \varphi_m + slope_m \cdot \left(n - \frac{N-1}{2} \right)$$

For subsequent magnitude and noise computations, the BPSK symbols of each reference subcarrier are phase-adjusted to place the nominal BPSK constellation points on the real axis.

$$u_{n,m} = r_{n,m} \cdot e^{-j \cdot \varphi_{n,m}}$$

For these cases it is assumed that the reference subcarriers are not processed by the PAPR reduction algorithm, and the distortion of I and Q is small at the exciter output. However, some versions of the PAPR reduction algorithm allow some amplitude variation of the reference subcarriers, although well within acceptable noise-level limits. This is primarily a measure of the signal distortion caused by the path (e.g., HPA and filters) between the exciter output and the antenna.

The above algorithms should work well for a sample clock frequency error of approximately ±2.5 PPM for a block size of N=512. The tolerance is inversely proportional to the block size so that if N=128, a tolerance of ±10 PPM is acceptable.

7 Signal Magnitudes and Gain Deviation

The mean of the magnitude of each BPSK reference subcarrier can now be conveniently computed as the mean of the absolute value of the real part of the phase-adjusted symbols.

$$smag_m = \frac{1}{N} \cdot \sum_{n=0}^{N-1} \left| \mathrm{Re}\{u_{n,m}\} \right| \; ;$$

where N = the total number of OFDM symbols.

The gain variation across the subcarriers is determined by

$$\Delta GdB = 20 \cdot \log\left(\frac{\max(smag)}{\min(smag)} \right)$$

The above computation is a convenient way to verify the NRSC-5-B gain flatness specification [1], Subsection 4.9.

8 Group Delay

The local group delay can be approximated over a finite frequency difference between any two adjacent reference subcarriers indexed *m* and *m*+19 (19 subcarrier spacing). The average phase φ_m for each reference subcarrier is used for this computation.

$$\tau_{m,m+19} = \frac{\varphi_{m+19} - \varphi_m}{2 \cdot \pi \cdot 19 \cdot 344.53125 \cdot 135/128} \quad seconds,$$

or equivalently,

$$\tau_{m,m+19} = 23052 \cdot \left(\varphi_{m+19} - \varphi_m\right) \; nanoseconds$$

The maximum group delay variation is estimated by

$$\Delta\tau_{\max} = \max(\tau) - \min(\tau)$$

The above computation is a convenient way to verify the NRSC-5-B group delay flatness specification [1], Subsection 4.10.

Any additional group delay caused by symbol synchronization timing error will not affect this difference. Note that the specification is equivalent to 0.75 degrees over a frequency span between the two reference subcarriers with the minimum and maximum group delay.

9 Reference Subcarrier MER

The rms noise may be computed for each (phase-adjusted) reference subcarrier. This noise is normalized to the subcarrier nominal magnitude.

$$N(ref)_m = \frac{1}{smag_m} \cdot \sqrt{\frac{1}{N} \cdot \sum_{n=0}^{N-1} \left[\left(\left| \mathrm{Re}\{u_{n,m}\} \right| - smag_m \right)^2 + \mathrm{Im}\{u_{n,m}\}^2 \right]}$$

The normalized rms noise is then converted to MER:

$$MER(ref)_m = -20 \cdot \log\{\ N(ref)_m\}, \text{ or equivalently}$$

Equation 1: MER per Reference Subcarrier

$$\boxed{MER(ref)_m = -10 \cdot \log_{10}\left(\frac{1}{N \cdot smag_m^2} \cdot \sum_{n=0}^{N-1} \left[\left(\left| \mathrm{Re}\{u_{n,m}\} \right| - smag_m \right)^2 + \mathrm{Im}\{u_{n,m}\}^2 \right] \right)}$$

Equation 1 computes the MER for each individual reference subcarrier. To get a single signal quality metric for all reference subcarriers, the MER values can be averaged according to Equation 2.

Equation 2: Composite MER Measurement for all Reference Subcarriers

$$\boxed{MER(ref)_{Avg} = 10 \cdot \log_{10} \sum_m \frac{1}{M} 10^{MER(ref)_m/10}}$$

In Equation 2, the sum is computed for active reference subcarriers and M equals the total number of active reference subcarriers.

It should be noted that the composite MER measurement computed by Equation 2 excludes any group delay or amplitude variations across the channel bandwidth. This is because each reference subcarrier has been separately adjusted for amplitude and phase variations as described in Sections 6 and 7. This is appropriate since the receiver will also process each reference subcarrier separately. Amplitude and group delay distortions are specified separately in [2] and do not need to be included in the MER.

In addition, the reference subcarriers do not have the PAPR reduction algorithm applied. So the reference subcarrier MER is a good measure of the signal constellation spread caused by noise and distortion introduced solely by the transmission signal path after the digital to analog conversion in the exgine/exciter.

10 PAPR Reduction Noise on Data Subcarriers

The PAPR reduction algorithm introduces noise onto the data-bearing, non-reference subcarriers. Since these noise samples are intentionally biased in a way that mitigates errors, a simple rms or MER computation is not appropriate here. The measurement strategy is to first adjust the symbol constellations of the data subcarriers such that the constellation points are centered in the quadrants, and the in-phase (I) and quadrature (Q) components of each point is nominally ±1. The metric in this case would penalize values that are closer to the decision region axes.

11 Equalization of Data Subcarriers

The data subcarriers within each partition of 18 carriers bounded by reference subcarriers on either side must be equalized to perform subsequent measurements. A linear interpolation of the outer reference subcarriers is used to equalize each data subcarrier across each OFDM symbol. The reference phases are first adjusted to avoid complications due to modulo π wrapping.

$$\text{If } \left|\varphi_{n,m} - \varphi_{n,m+19}\right| > \frac{\pi}{2} \text{ , then let } \varphi_{n,m+19} = \varphi_{n,m+19} + \pi \ ;$$

where n is the OFDM symbol index, m is the index of the Active Reference Subcarrier to the left (lower index number) of the partition, $\varphi_{n,m}$ and $\varphi_{n,m+19}$ are the phases of the corresponding reference subcarriers that bound the partition, and $\varphi_{n,m+19}$ may be adjusted for the modulo π wrap case.

Notice that there is no partition corresponding to the largest value of m for each sideband. Next, compute a set of 18 equalizer coefficients to be applied to each of the 18 data subcarriers (k=1,2,…18). A phase shift of π/4 (equivalently a factor of 1+j) is also applied to center the QPSK constellations in the quadrants. The equalizer coefficients are computed as

$$C_{n,k} = \frac{19 \cdot (1+j)}{(19-k) \cdot smag_m \cdot e^{j \cdot \varphi_{n,m}} + (k) \cdot smag_{m+19} \cdot e^{j \cdot \varphi_{n,m+19}}} ; k = 1,2 \ldots 18$$

Then apply the equalizer coefficients to the data subcarrier symbols within each partition.

$$v_{n,m+k} = s_{n,m+k} \cdot C_{n,k} ; k = 1,2,\ldots 18$$

12 PAPR Reduction Algorithm Increases Subcarrier Power

The PAPR reduction (noise) causes a modest increase in digital signal power (e.g., 0.5 dB), which must be adjusted back to the target power level. This post-PAPR reduction adjustment should simply scale all subcarriers equally with a single gain factor. Also, the power increase of the reference subcarriers is generally less than the power increase of the data subcarriers. Although we know that the present PAPR reduction algorithm requires only a small gain adjustment (e.g., 0.5 dB), the change in ratio between the reference and data subcarriers power level should be limited. The average reference subcarrier power and the average overall power are computed. The average subcarrier power over all subcarriers is computed as

$$P_{avg} = \frac{1}{N \cdot SC} \sum_{sc=-546}^{546} \sum_{n=0}^{N-1} \left[\left(\mathrm{Re}\{v_{n,sc}\}\right)^2 + \left(\mathrm{Im}\{v_{n,sc}\}\right)^2 \right] \quad ;$$

where SC is the number of active subcarriers, and the outer summation is computed only for subcarrier indices pertaining to active subcarriers.

Note that -546 and 546 represent the minimum and maximum possible subcarrier indices.

The average power of the reference subcarriers is

$$P_{ref} = \frac{1}{M} \sum_{m} \sum_{n=0}^{N-1} \left[\left(\mathrm{Re}\{v_{n,m}\}\right)^2 + \left(\mathrm{Im}\{v_{n,m}\}\right)^2 \right] ;$$

where M is the number of active reference subcarriers, and m are the active reference subcarrier indices.

Since there is generally one reference subcarrier for every 18 data subcarriers, the voltage ratio R of the average data subcarrier to the reference subcarrier may be very closely approximated by:

$$R \approx \sqrt{\frac{19 \cdot P_{avg} - P_{ref}}{18 \cdot P_{ref}}}$$

Equation 3: Computation of Data Subcarrier to Reference Subcarrier Power in dB

$$\boxed{R_{dB} = 20 \cdot \log_{10}(R)}$$

This ratio is related to the expected loss due to the PAPR reduction power increase, and subsequent power-level adjustment. The computed value of R is used in the data subcarrier MER measurements.

Note: In the Data MER expressions (below), R restores the proper level of reference subcarriers for the computation and prevents the possibility of artificially lowering the reference subcarriers to improve MER results.

13 Data Subcarrier Partition MER

In this subsection, the noise metric computation for partition *m* is described, where *m* is the index of the active Reference Subcarrier to the left (lower index number) of the partition. Note that in this calculation, the reference subcarrier is not included.

$$N(dat)_m = \sqrt{\frac{1}{N \cdot 18} \cdot \sum_{k=1}^{18} \sum_{n=0}^{N-1} \left[\left(\max\left\{0, R - \left|\mathrm{Re}\left\{v_{n,m+k}\right\}\right|\right\}\right)^2 + \left(\max\left\{0, R - \left|\mathrm{Im}\left\{v_{n,m+k}\right\}\right|\right\}\right)^2 \right]}$$

$$MER(dat)_m = -20 \cdot \log_{10}\left\{ N(dat)_m \right\}$$, or equivalently

Equation 4: MER per 18-Data-Subcarrier Partition

$$\boxed{MER(dat)_m = 10 \cdot \log_{10}\left(\frac{N \cdot 18}{\sum_{k=1}^{18} \sum_{n=0}^{N-1} \left[\left(\max\left\{0, R - \left|\mathrm{Re}\left\{v_{n,m+k}\right\}\right|\right\}\right)^2 + \left(\max\left\{0, R - \left|\mathrm{Im}\left\{v_{n,m+k}\right\}\right|\right\}\right)^2 \right]} \right)}$$

Equation 4 computes the MER for a single partition. To get a single signal quality metric for all data subcarriers, the MER values can be averaged according to Equation 5.

Equation 5: Composite MER Measurement for all Data Subcarriers

$$\boxed{MER(dat)_{Avg} = 10 \cdot \log_{10} \sum_m \frac{1}{M} 10^{MER(dat)_m / 10}}$$

where the sum is computed for active partitions and M equals the total number of partitions

The above equations may be used to assess whether or not the digital transmission system is in compliance with the requirements stated in Subsection 4.8 of reference [1]. Refer to the next section of this document for proposed changes to these requirements. The MER is inclusive of any distortions caused by the PAPR reduction algorithm as well as any local variations in amplitude and group delay not removed by the equalizer described in Section 11. Thus it should be very indicative of actual receiver performance.

The intent of the specifications proposed in the next section of this document is to utilize MER to evaluate the performance of the transmission system equipment as measured at the RF output connection point to the antenna system. However, it must be noted that the same MER computations could be accomplished by a monitor receiver located at the broadcaster installation in order to evaluate the effects of the antenna system as well as any interference to the digital subcarriers caused by an overmodulated or otherwise distorted analog FM host signal. iBiquity will not impose specific requirements on this. But they may offer voluntary recommendations in the future, once the MER specifications have been fully characterized. For example, the broadcaster could utilize an MER monitor receiver to signal an alarm indicating a fault condition that may be compromising receiver performance in the field.

14 Proposed Changes to NRSC-5 Transmission Spec, Subsection 4.8

Reference [1], Subsection 4.8, specifies the EVM requirements for the HD Radio system. Both an average requirement as well as a worst-case requirement for a single subcarrier is listed:

Average EVM not to exceed 10% → Equivalent MER = 20 dB

Worst-Case individual subcarrier EVM not to exceed 20% → Equivalent MER = 14 dB

These figures included a reasonable amount of margin. However, the current HD Radio system has never been verified in this manner due to the lack of a unified test procedure. Given the proposed MER measurement techniques proposed in this paper, the following changes to NRSC-5, Subsection 4.8 are offered. Information within this section that is enclosed in brace brackets {} is subject to change.

4.8 Modulation Error Ratio

The following specifications shall be met, using the test configuration described in Subsection 4.2 of Reference [3].

4.8.1 Reference Subcarriers

1. The MER for each and every Binary Phase Shift Keying (BPSK) reference subcarrier, measured at the RF output of the transmission system at the connection point to the antenna system (including any RF filters), shall be greater than or equal to {11} dB, as computed by Equation 1. The parameter N in Equation 1, the total number of contiguous symbols used in the average, shall be set to {128}.
2. The average MER of the Binary Phase Shift Keying (BPSK) reference subcarriers, measured at the RF output of the transmission system at the connection point to the antenna system (including any RF filters), shall be greater than or equal to {14} dB, averaged across all reference subcarriers, as computed by Equation 2. This computation shall be based on a block of N = {128} contiguous symbols.

4.8.2 Data Subcarriers

1. The MER for each and every Quadrature Phase Shift Keying (QPSK) data subcarrier partition, measured at the RF output of the transmission system at the connection point to the antenna system (including any RF filters), shall be greater than or equal to {11} dB, as computed by Equation 4. The parameter N in Equation 4, the total number of contiguous symbols used in the average, shall be set to {128}.
2. The average MER of the Quadrature Phase Shift Keying (QPSK) data subcarriers, measured at the RF output of the transmission system at the connection point to the antenna system (including any RF filters), shall be greater than or equal to {14} dB, averaged across all data subcarrier partitions, as computed by Equation 5. This computation shall be based on a block of N = {128} contiguous symbols.

4.8.3 Data Subcarrier to Reference Subcarrier Power Ratio

In addition to the gain flatness specifications stated in Subsection 4.9, the ratio of the average data subcarrier power to the average reference subcarrier power, as computed by Equation 3, shall comply with the following limits:

$$-0.5 \le R_{dB} \le 1.0 \text{ dB}$$

Comments

- The language from the original NRSC-5 specifications regarding the measurement point has been retained.
- Note that the proposed data MER specifications are equivalent to the original EVM values. The reference subcarrier MER has been relaxed several dB, removing much of the original margin. This will still provide acceptable performance. The data subcarrier specifications are tighter primarily to account for the different measurement techniques.
- All of the specifications listed above need to be confirmed via actual system-level testing.
- For individual equipment specifications, such as for an exciter/exgine, all of the language above is still valid, except that the measurement point will be different, and it may be desirable to tighten the specifications somewhat.

Appendix A Frequency and Symbol Timing Acquisition from FM IBOC MER Measurement

This section provides a suggestion for performing synchronization of the carrier frequency and symbol timing.

The sample rate, fs=744,187.5 Hz. Symbol timing is less sensitive to sample rate frequency errors than is the carrier phase. So the sample rate frequency tolerances discussed in Section 6 apply here as well.

If the analog FM signal is present, it must first be removed in order for the techniques described here to work properly.

In order to process *N* "complete" symbols, each consisting of *K*=2160 complex signal samples, an additional symbol is buffered to accommodate the possibility of partial symbols at the boundaries (the minimum number of required additional samples is actually one sample less than a whole symbol). The total number of input complex signal samples is then $(N+1)\cdot K$.

The sequence of signal samples x_n , $n=0,\ldots,[(N+1)\cdot K]-1$, is complex-conjugate multiplied by the sequence delayed by 2048 samples (i.e., x_{n+2048}). The products are stored in a single-symbol, *K*-sample complex buffer by folding modulo *K*. Notice that only N symbols are processed (not N+1) to allow for the delay.

$$y_{\mathrm{mod}(n,K)} = y_{\mathrm{mod}(n,K)} + x_n \cdot x^{*}_{n+2048} \quad ; \text{ for } n=0,\ldots,N\cdot K-1$$

where * indicates complex conjugate, and *y* is previously initialized to zero because of the recursive operation. Next the vector *y* is match-filtered, using cyclic convolution, for the expected cyclic prefix autocorrelation pulse shape *h*, to produce vector *v*.

$$v_k = \sum_{m=1}^{111} h_m \cdot y_{\mathrm{mod}(k+m,K)} \quad ; \ k=0,\ldots,K-1$$

$$\text{where } h_m = \sin\left(\frac{m\cdot\pi}{112}\right) \quad ; \ m=0,\ldots 111$$

The sample offset error is simply computed as the index of the peak magnitude sample of the vector *v*. This corresponds to the starting (zero index) sample of the first complete symbol. Symbol synchronization to the nearest whole sample is sufficient, since the subsequent equalization process will remove the linear phase distortion resulting from residual symbol timing error.

$$samperr = \text{index } k \ ; \ \text{corresponding to } \max_{k} |v_k|$$

The estimated frequency error computed in Hz is

$$freqerr = \frac{-fs}{2\cdot\pi\cdot 2048}\cdot \arctan\left(\frac{\mathrm{Im}\{v_{samperr}\}}{\mathrm{Re}\{v_{samperr}\}}\right)$$

The sample and frequency error estimates are then used to adjust the original sequence, and output exactly N synchronized symbols.

$$xadj_n = x_{(n+samperr)} \cdot \exp\{-j\cdot 2\cdot\pi\cdot freqerr \cdot n / fs\} \quad ; \text{ for } n=0,\ldots,K\cdot N-1$$

OFDM Demodulation of the Block of N Symbols

Demodulation of the block of N OFDM symbols is next. Define the root-raised cosine Nyquist pulse shape window function for k=0,…,K-1, and K=2160.

$$\mathrm{win}_k = \begin{cases} \sin\left(\frac{k\cdot\pi}{224}\right) & ;\ \mathit{if}\ \ k < 112 \\ \sin\left(\frac{(K-k)\cdot\pi}{224}\right) & ;\ \mathit{if}\ \ k > 2048 \\ 1 & ;\ \ \mathit{otherwise} \end{cases}$$

The demodulation of a signal vector consisting of a block of N synchronized symbols can be computed as follows. First initialize the matrix *sig* for the appropriate size (2048 rows by N columns), and zero-fill to accommodate the initial condition of the recursive computation.

$$sig_{2047,N-1} = 0$$

Arrange the signal column vector into a matrix of N symbol column vectors. Then window and fold the ending cyclic suffix onto the prefix.

$$sig_{\mathrm{mod}\left[n-floor\left(\frac{n}{K}\right)\cdot K,2048\right],\,floor\left(\frac{n}{K}\right)} = sig_{\mathrm{mod}\left[n-floor\left(\frac{n}{K}\right)\cdot K,2048\right],\,floor\left(\frac{n}{K}\right)} + xadj_n \cdot win_{\mathrm{mod}[n,K]}\ ;$$
$$n = 0,...,K\cdot N-1 \text{ (i.e., } N \text{ symbols)}$$

Demodulate each column (OFDM symbol) of *sig* using complex FFT, and appropriate scaling such that constellation points are nominally (±1, ±1), with possible phase rotation.

$$OFDM\_VECTOR(\text{column } m) = CFFT(sig(\text{column } m))\cdot\sqrt{\frac{256\cdot 2\cdot 191}{135}}\quad ;\ m\text{=0,… N-1}$$

A symbol timing error (*samperr*) results in a phase rotation of the constellation across subcarriers (k). There is no phase rotation over time (not affected by m). The phase rotation is resolved by the equalizer.

A frequency error (*freqerr*) results in a phase rotation of the constellation for any one subcarrier over time (m). Even one Hz error causes more than one full rotation over 512 symbols. However, all subcarriers are in phase sync. This frequency estimate algorithm is very accurate, and should result in no significant phase rotation. The fixed phase offset is arbitrary, and is resolved by the equalizer (phase just happens to be zero in this simulation).

MathCad Example

Generate one more symbol than the required target to accommodate partial symbols at the boundaries:

Number of symbols $N := 64$ $\quad samperr := 0$ $\quad freqerr := 0$ $\quad x := FM\_dig\_sig(N+1, samperr, freqerr)$

Samples per symbol K: $K := 2160$ $\quad k := 0..K-1$

Process samples allowing for delay of 2048 samples (~1 symbol): $n := 0..N\cdot K - K - 1$

Complex multiply delayed samples and fold symbols: $y_{K-1} := 0$ $\quad y_{mod(n,K)} := y_{mod(n,K)} + x_n \cdot \overline{x_{n+2048}}$

Create matched filter for autocorrelation pulse shape: $m := 0..111$ $\quad h_m := \sin\left(\frac{m\cdot\pi}{112}\right)$

Matched filter for the resulting autocorrelation vector: $v_k := \sum_{m=1}^{111} \left(h_m \cdot y_{mod(k+m,K)}\right)$

Compute the sample and frequency errors from v:

$$
sampfreqerr(v) := \left|
\begin{array}{l}
\text{"find peak and its location"} \\
samperr \leftarrow 0 \\
pk \leftarrow \left|v_0\right| \\
\text{for } k \in 1..K-1 \\
\quad \text{if } \left|v_k\right| > pk \\
\qquad \left| \begin{array}{l} samperr \leftarrow k \\ pk \leftarrow \left|v_k\right| \end{array} \right. \\
\left(\begin{array}{c} samperr \\ freqerr \leftarrow -\frac{fs}{2\cdot\pi\cdot 2048}\cdot \operatorname{atan}\left(\frac{\operatorname{Im}\left(v_{samperr}\right)}{\operatorname{Re}\left(v_{samperr}\right)}\right) \end{array} \right)
\end{array}
\right.
$$

$$\begin{pmatrix} samperr \\ freqerr \end{pmatrix} := sampfreqerr(v)$$

$$\begin{pmatrix} samperr \\ freqerr \end{pmatrix} = \begin{pmatrix} 0 \\ 0 \end{pmatrix}$$

Adjust signal x for sample and freq errors, 64 symbols: $n := 0..N\cdot 2160 - 1$ $\quad xadj_n := x_{n+samperr} \cdot e^{-j\cdot 2\cdot\pi\cdot freqerr\cdot\frac{n}{fs}}$

$var(x) = 1$

Figure A-1: Symbol Synchronization and Frequency Acquisition

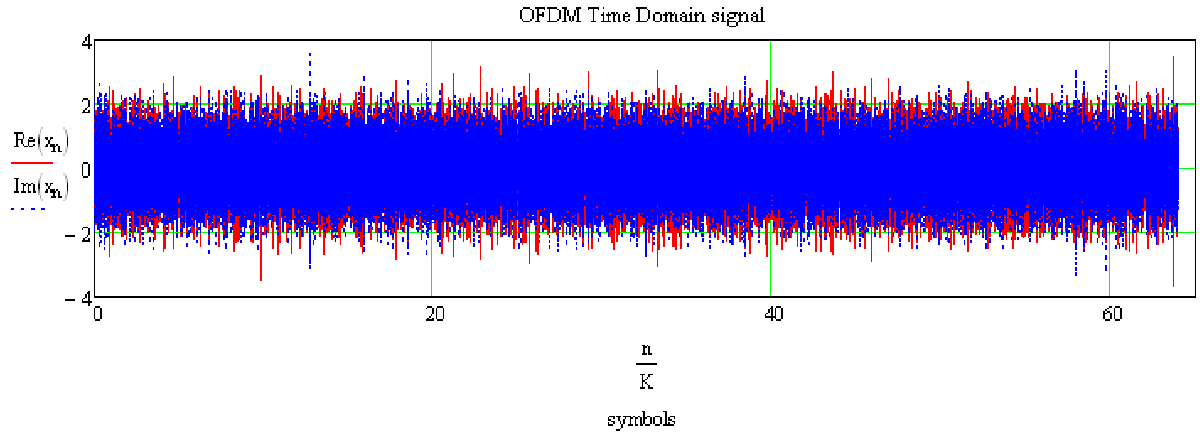


Figure A-2: OFDM Time Domain Signal

$SIG := CFFT(x)$

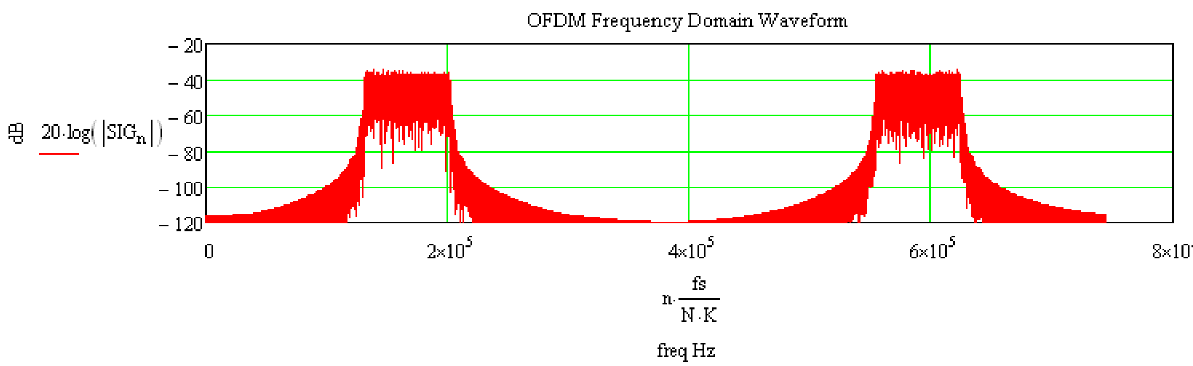


Figure A-3: OFDM Frequency Domain Waveform

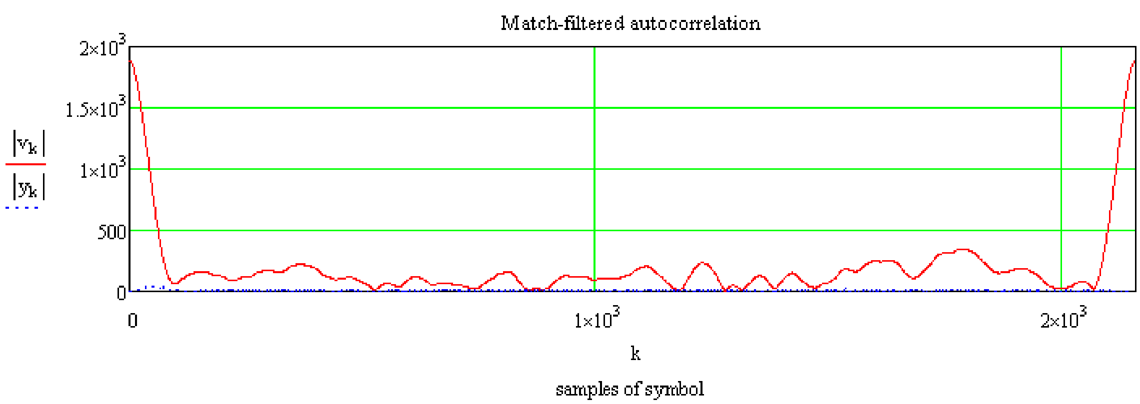


Figure A-4: Match-Filtered Autocorrelation

Demodulate synchronized OFDM signal:

Apply window to matrix:

$$
win := \left|
\begin{array}{l}
\text{"Generate 2160-sample window for pulse shape"} \\
Nw \leftarrow 2048 \\
TAPER \leftarrow 112 \\
\text{for } nt \in 0..Nw + TAPER - 1 \\
\quad w_{nt} \leftarrow \left|
\begin{array}{ll}
\sin\left(nt \cdot \dfrac{\pi}{2 \cdot TAPER}\right) & \text{if } nt < TAPER \\
\sin\left[(Nw + TAPER - nt) \cdot \dfrac{\pi}{2 \cdot TAPER}\right] & \text{if } nt > Nw \\
1 & \text{otherwise}
\end{array}\right. \\
w
\end{array}\right.
$$

Arrange signal vector xadj into matrix of symbol vector columns: Fold ending cyclic suffix onto prefix:

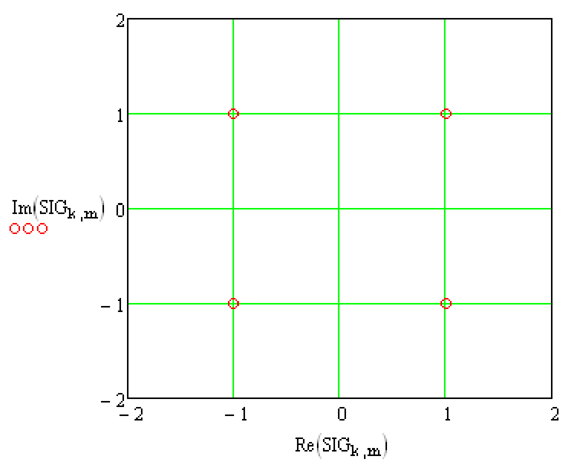
$$sig_{2047,63} := 0 \quad sig_{mod\left(n-floor\left(\frac{n}{2160}\right)\cdot 2160, 2048\right), floor\left(\frac{n}{K}\right)} := sig_{mod\left(n-floor\left(\frac{n}{2160}\right)\cdot 2160, 2048\right), floor\left(\frac{n}{K}\right)} + xadj_n \cdot win_{mod(n,K)}$$

$k := 1502$

$m := 0..63$

$k := 1692$

$$SIG^{\langle m \rangle} := CFFT\left(sig^{\langle m \rangle}\right) \cdot \sqrt{\frac{256 \cdot 2 \cdot 191}{135}} \qquad k := 546 \quad k := 356..546 \qquad \begin{pmatrix} samperr \\ freqerr \end{pmatrix} = \begin{pmatrix} 0 \\ 0 \end{pmatrix}$$



A symbol timing error (samperr) results in a phase rotation of the constellation across subcarriers (k). There is no phase rotation over time (not affected by m). The phase rotation is resolved by the equalizer.

A frequency error (freqerr) results in a phase rotation of the constellation for any one subcarrier over time (m). Even one Hz error causes more than 1 full rotation over 512 symbols. However, all subcarriers are in phase sync. This frequency estimate algorithm is very accurate, and should result in no significant phase rotation. The fixed phase offset is arbitrary, and is resolved by the equalizer (phase just happens to be zero in this simulation).

A sample rate error results in a phase rotation of the constellation for any one subcarrier over time (m). For example, if the sample rate error causes a drift of 1 sample over the signal time span, then this would result in a phase shift of 360*fc/fs degrees, or about 97 degrees for a subcarrier at 200 kHz from center frequency.

Figure A-5: Demodulated OFDM Symbol Constellation

Appendix B Computed MER Values and Measured Receiver Performance Using iBiquity Reference Test Data

Table B-1 shows the MER values computed from the reference data files supplied iBiquity Digital. The block size (N) is 512.

Table B-1: Computed MER Values and Measured Receiver Performance

Test File Name	Service Mode	PAPR Reduction On/Off	Cd/No dB-Hz	MER(ref)$_{Avg}$	MER(ref)$_{Worst\ Case}$ / Subcarrier Index	MER(dat)$_{Avg}$	MER(dat)$_{Worst\ Case}$ / Partition Index	P1 Channel BER
MP1_PAROff_52dB	MP1	Off	52	1.5	1.0	4.8	4.7	7.80E-03
MP1_PAROff_54dB	MP1	Off	54	3.1	2.8	5.2	5.0	3.30E-05
MP1_PAROff_56dB	MP1	Off	56	5.0	4.7	6.0	5.8	1.10E-07
MP1_PAROff_58dB	MP1	Off	58	6.9	6.6	7.2	7.0	0
MP1_PAROff_60dB	MP1	Off	60	8.9	8.5	8.9	8.7	0
MP1_PAROff_62dB	MP1	Off	62	10.9	10.5	10.8	10.6	0
MP1_PAROff_64dB	MP1	Off	64	12.8	12.5	12.8	12.6	0
MP1_PAROff_66dB	MP1	Off	66	14.8	14.5	14.8	14.6	0
MP1_PAROff_68dB	MP1	Off	68	16.8	16.5	16.8	16.5	0
mp1_PARoff_noAnalog	MP1	Off	No noise	88.7	88.2	88.6	88.1	0
MP1_PAROn_52dB	MP1	On	52	1.0	0.54	4.6	4.4	1.10E-02
MP1_PAROn_54dB	MP1	On	54	2.6	2.2	4.8	4.7	6.80E-05
MP1_PAROn_56dB	MP1	On	56	4.4	4.0	5.5	5.3	5.70E-08
MP1_PAROn_58dB	MP1	On	58	6.3	5.8	6.6	6.4	0
MP1_PAROn_60dB	MP1	On	60	8.1	7.6	8.0	7.8	0
MP1_PAROn_62dB	MP1	On	62	10.0	9.5	9.6	9.3	0
MP1_PAROn_64dB	MP1	On	64	11.8	11.3	11.2	10.9	0
MP1_PAROn_66dB	MP1	On	66	13.6	13.1	12.6	12.3	0
MP1_PAROn_68dB	MP1	On	68	15.2	14.7	13.9	13.6	0
mp1_PARon_noAnalog	MP1	On	No noise	21.6	21.0	18.0	17.5	0

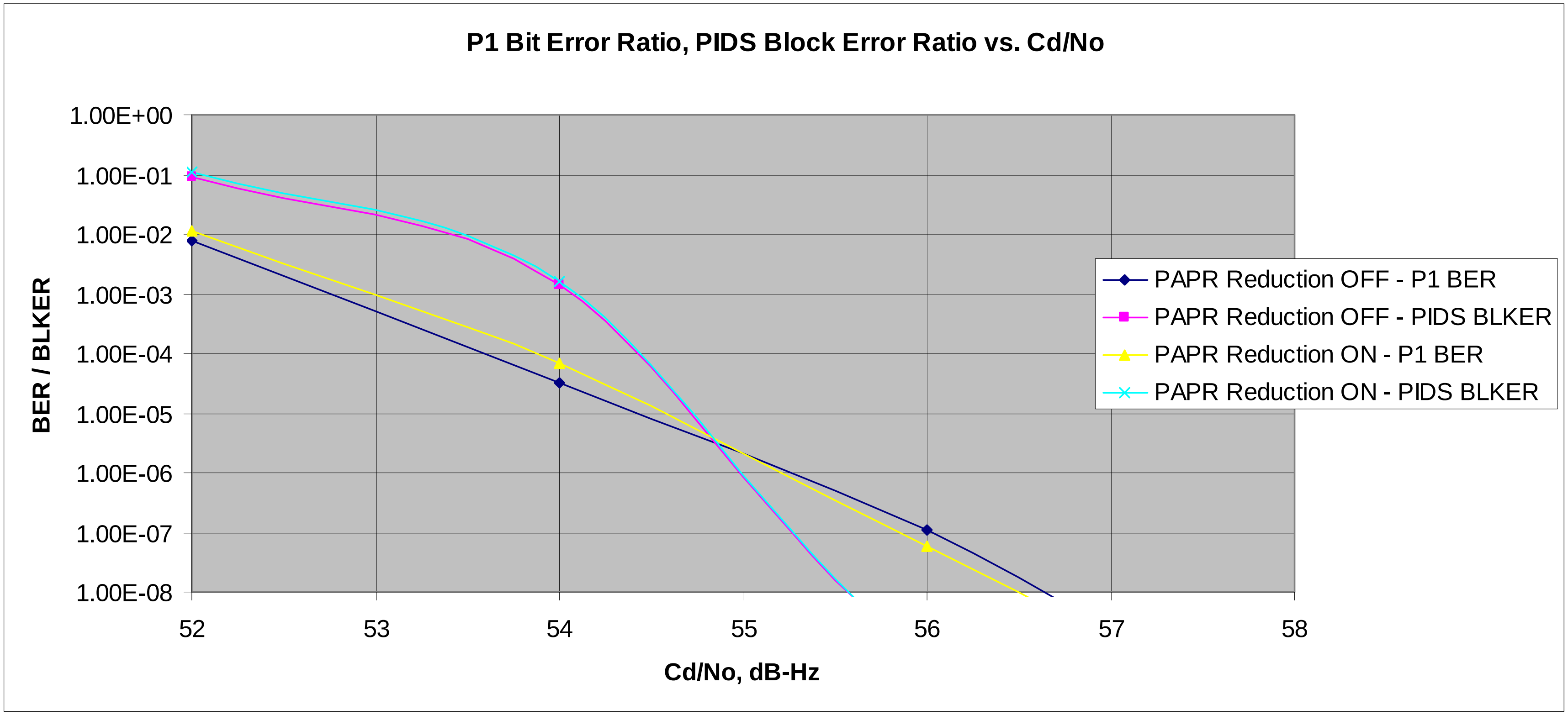


Figure B-1: P1 Bit Error Ratio and PIDS Block Error Ratio vs. Cd/No

Discussion

From the noise vectors analyzed here, it is apparent that as white Gaussian noise is added to the signal, the noise quickly dominates over the PAPR-reduction-induced constellation noise before bit errors become apparent. This statement is supported by comparing the MER numbers for the PAPR-reduction-enabled and PAPR-reduction-disabled cases versus the signal without added noise. The difference in MER numbers quickly diminishes with added noise, indicating that the introduced noise is dominating. Visual inspection of Figure B-4 and Figure B-9 shows that the constellations start to look quite similar when Cd/No=64 dB-Hz, where no bit errors are detected.

Bit errors are only detected when Cd/No=56 dB-Hz, where the difference in MER between the PAPR-reduction-enabled and PAPR-reduction-disabled cases is only about 0.5 dB.

The argument could be made that as long as the PAPR-reduction-induced constellation noise, as measured by the IBOC quality metric, is sufficiently better than the MER of the PAPR-reduction-disabled signal with added noise, then the impact of the PAPR-reduction-induced noise should be minimal to the received signal in an AWGN channel. So in the case where Cd/No=56 dB-Hz, the PAPR-reduction-disabled signal shows bit errors at a MER of 6.0 dB for data carriers and 5.0 dB for reference carriers. The PAPR-reduction-induced MER of 18.0 dB for data carriers and 21.6 dB for reference carriers provides a rather large margin before the system breaks down.

It is, however, prudent to add significant margin to account for multipath and mobile channel environments. The following analysis provides a method to derive a suitable MER specification point with a reasonable margin taken into account.

Consider a receiver that, due to channel noise, is operating approximately 2 dB above the blending point. According to the BER data, this is at Cd/No=56 dB-Hz. This is the point where the BER is around 10^{-7}; or almost error free. Assume for this analysis that the transmission system is perfect; that is, equivalent to the “PAPR Reduction Off – No Added Noise” reference file (mp1_PARoff_noAnalog, Figure B-7).

It is recommended that, when a realistic transmission system is introduced into the above scenario, the total Cd/No of the link shall not be degraded by more than 0.5 dB; that is, Cd/No(total) ≥ 55.5 dB-Hz.

Cd/No(total) = Combined Cd/No of channel and transmitter.

This is computed by combining the noise from the two sources: No(total) = No(channel) + No(transmitter).

This computation must be a linear addition; so the No values must be converted from dB to linear.

For 0.5-dB degradation:

Cd/No(total) = 55.5 dB-Hz

$$55.5 = -10*\log_{10}\{10^{-56/10} + 10^{-Cd/No(transmitter)}\}$$

Solving the above equation for Cd/No(transmitter), the following result is obtained:

Cd/No(transmitter) = 65 dB-Hz

Referring to Table B-1, for the PAPR Reduction Off case, Cd/No=65 dB-Hz produces an average MER value of 13.8 dB for both the data and reference subcarriers. This can be obtained through interpolation of the MER cases for Cd/No= 64 dB-Hz and Cd/No=66 dB-Hz.

Rounding this average MER value to the nearest dB produces a specification of 14 dB. Allowing for a 3-dB margin for the worst case, the specification would be 11 dB. See Section 14.

Constellation Plots

Figure B-2 through Figure B-6 show the signal constellations for various Cd/No values with PAPR-reduction-enabled; Figure B-7 through Figure B-11 show the equivalent cases with PAPR-reduction-disabled.

Data carriers are shown in blue, reference carriers are shown in red. The average power of the reference and data carriers is indicated by thicker blue and red dots with a green outline.

PAPR-Reduction-Enabled Constellation Plots

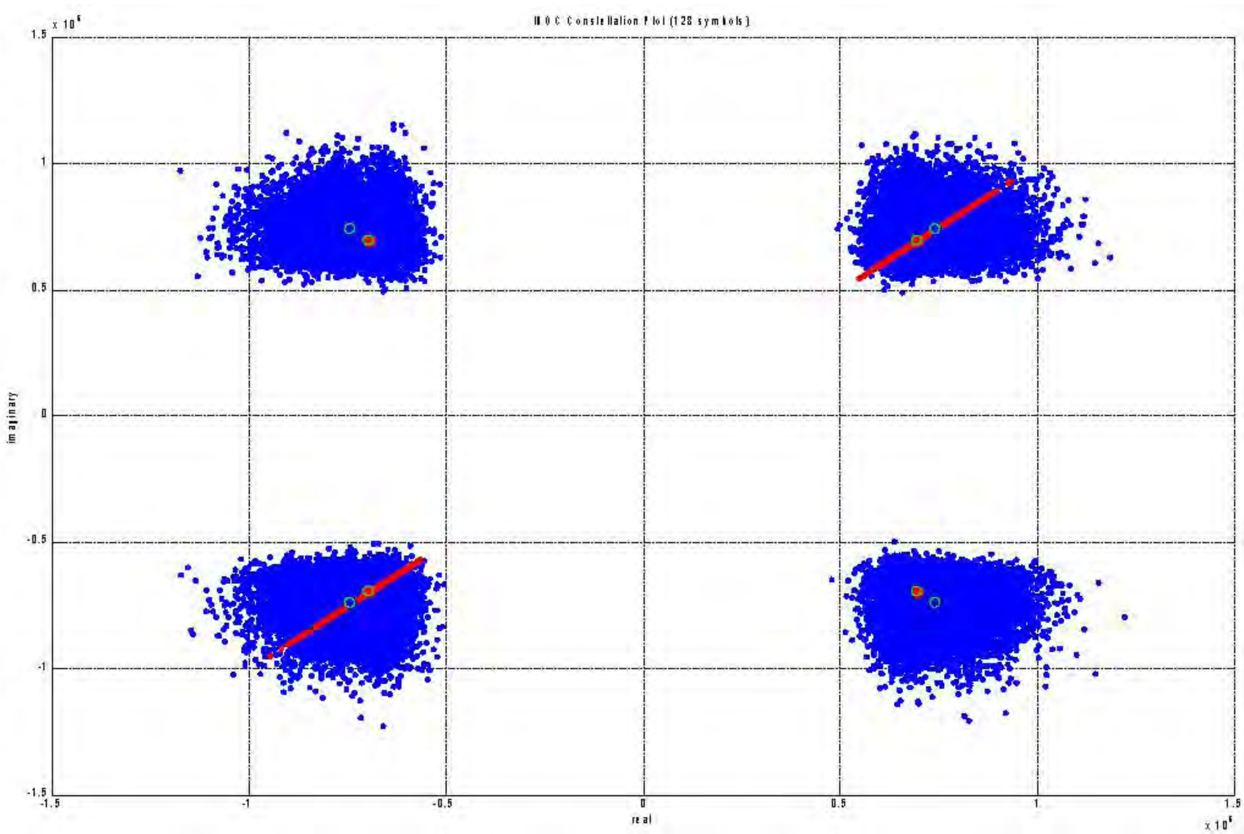


Figure B-2: IBOC Constellation with PAPR Reduction Enabled – No Added Noise

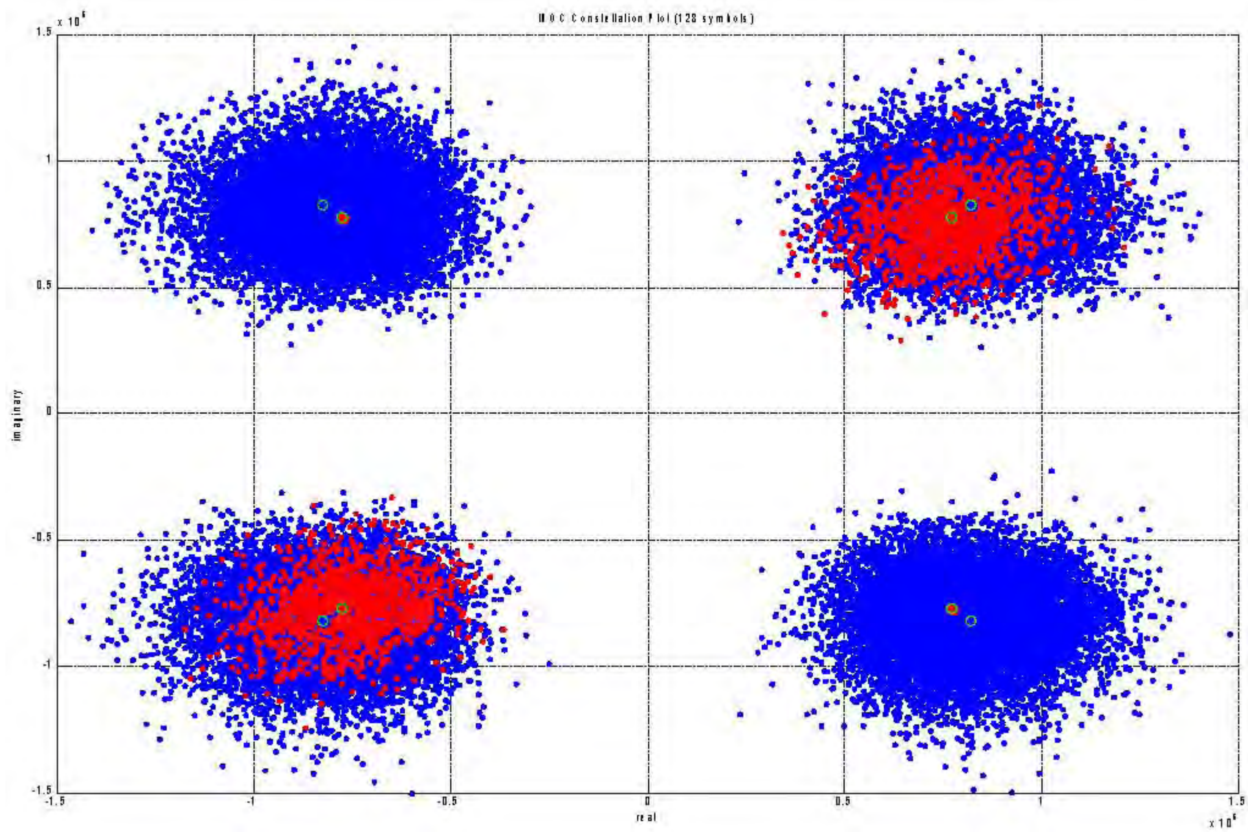


Figure B-3: IBOC Constellation with PAPR Reduction Enabled – Cd/No = 68 dB-Hz

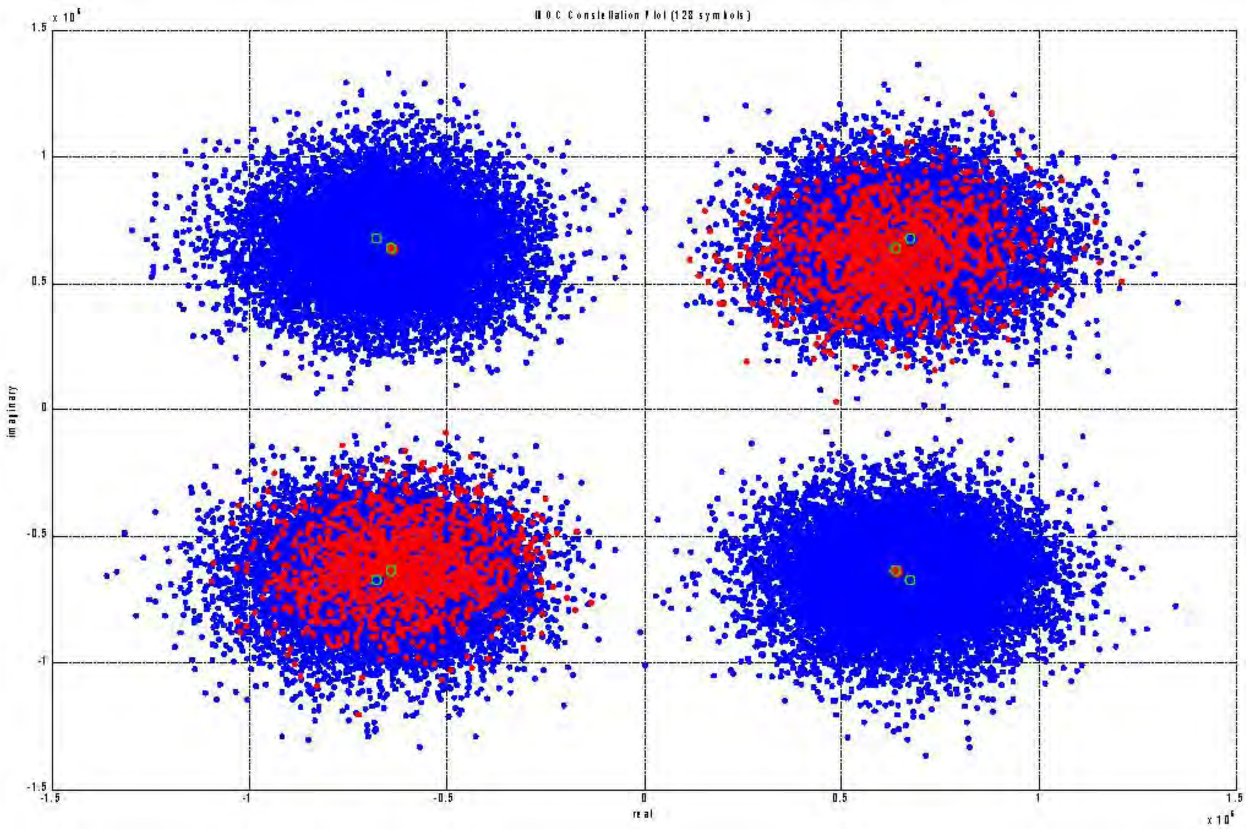


Figure B-4: IBOC Constellation with PAPR Reduction Enabled – Cd/No = 64 dB-Hz

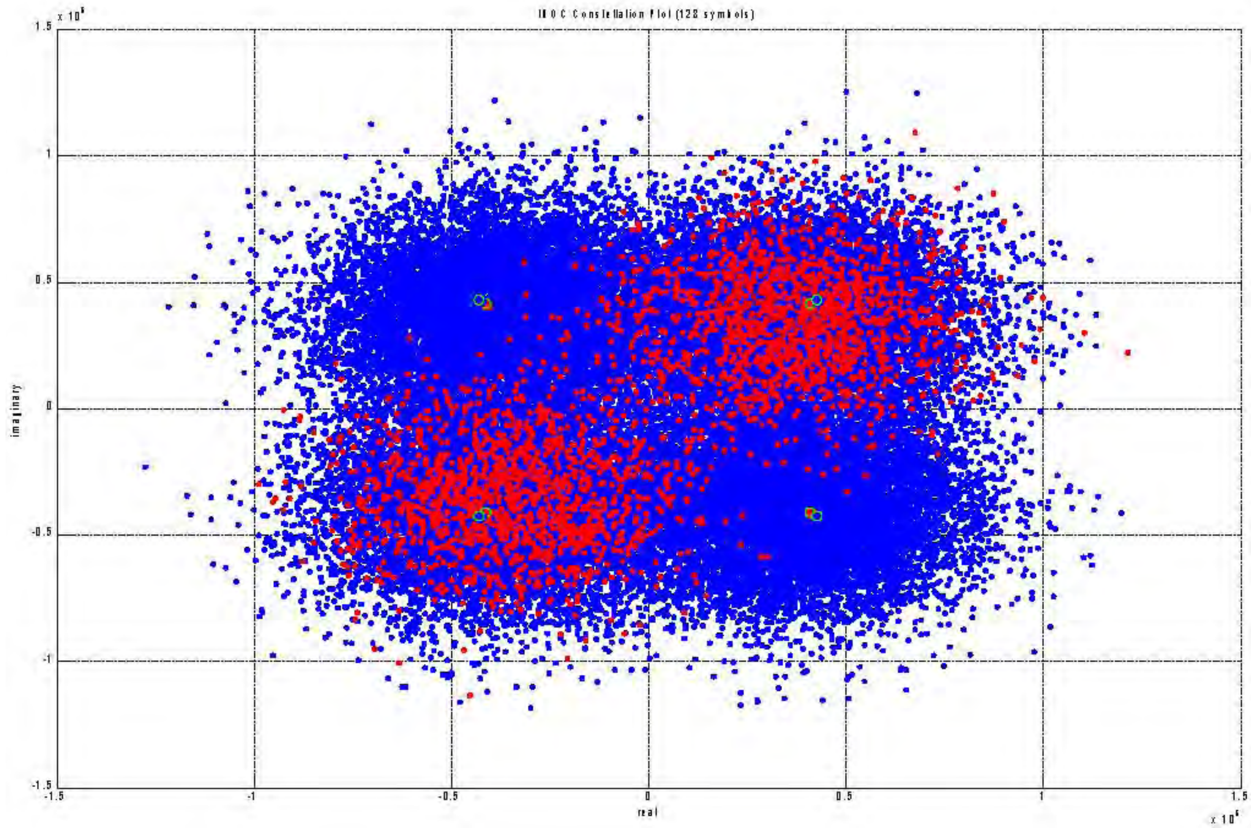


Figure B-5: IBOC Constellation with PAPR Reduction Enabled – Cd/No = 56 dB-Hz

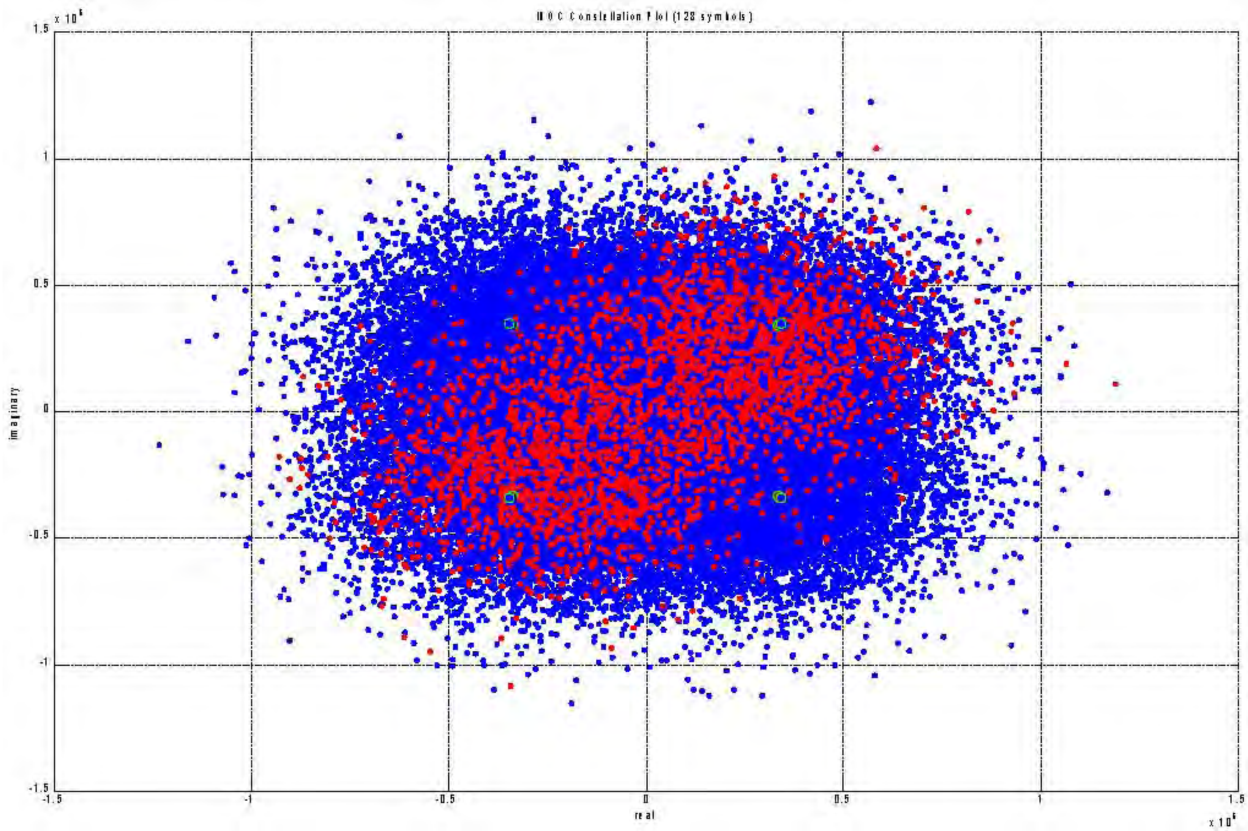


Figure B-6: IBOC Constellation with PAPR Reduction Enabled – Cd/No = 52 dB-Hz

PAPR-Reduction-Disabled Constellation Plots

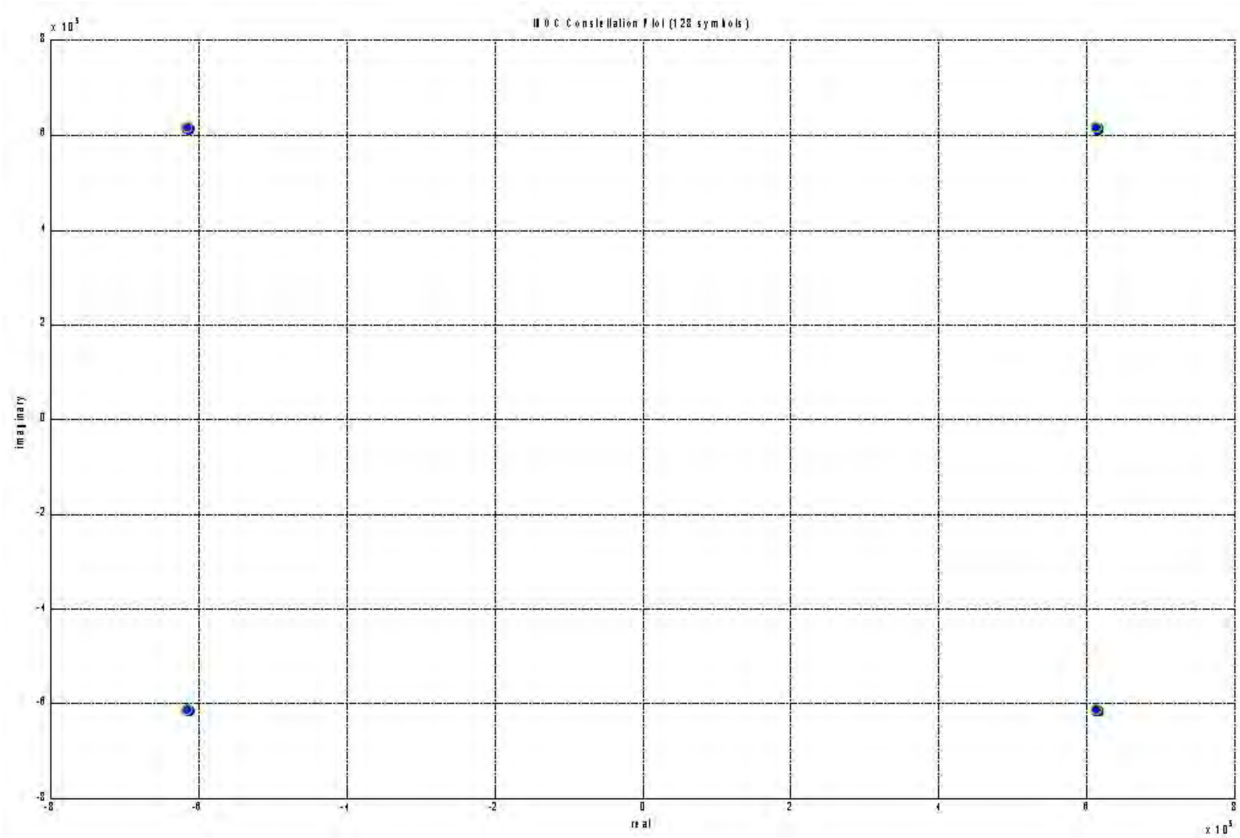


Figure B-7: IBOC Constellation with PAPR Reduction Disabled – No Added Noise

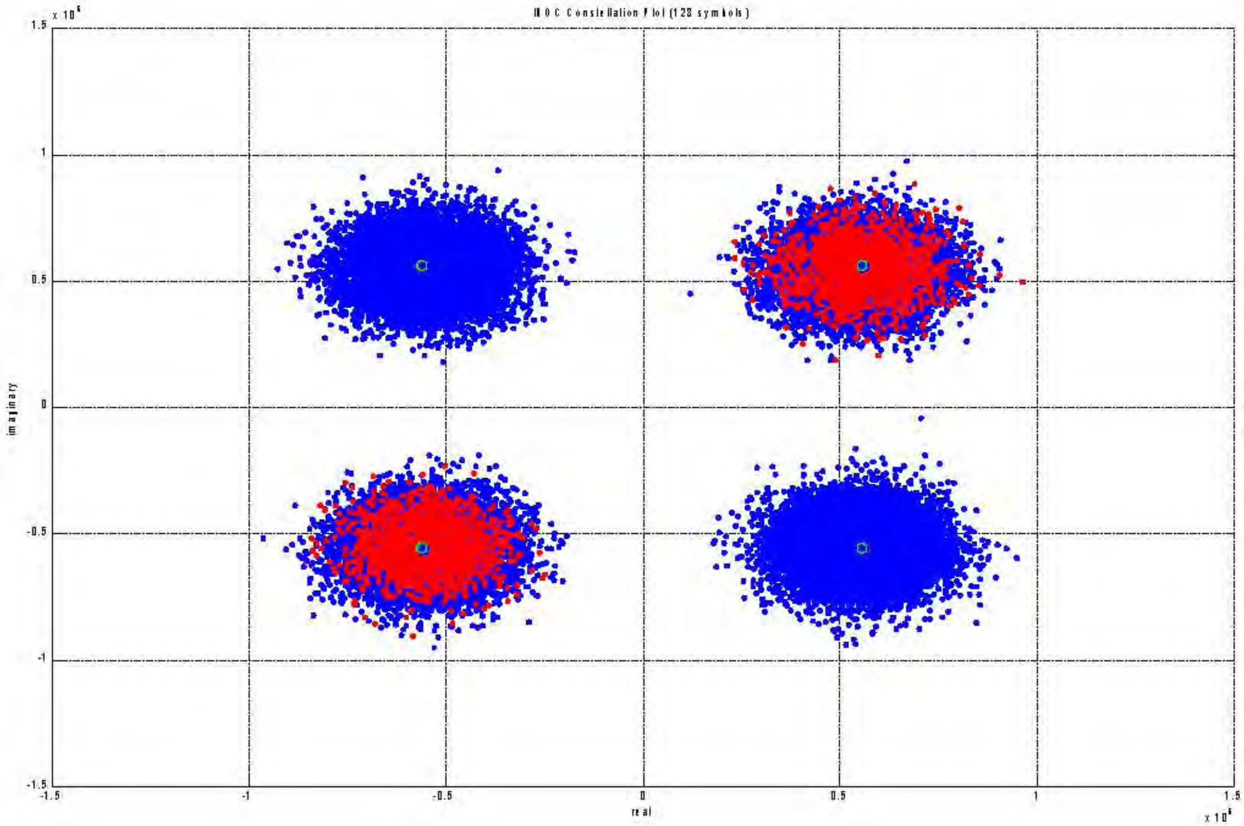


Figure B-8: IBOC Constellation with PAPR Reduction Disabled – Cd/No = 68 dB-Hz

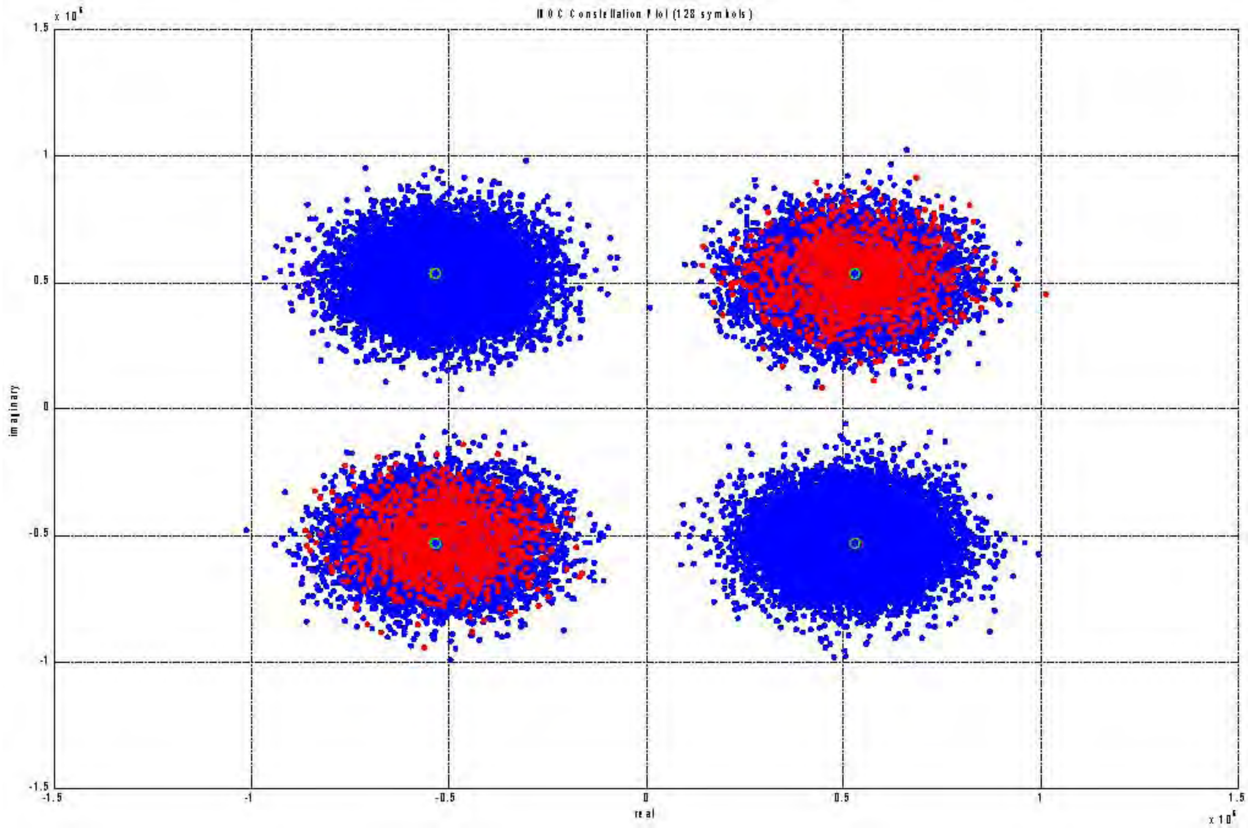


Figure B-9: IBOC Constellation with PAPR Reduction Disabled – Cd/No = 64 dB-Hz

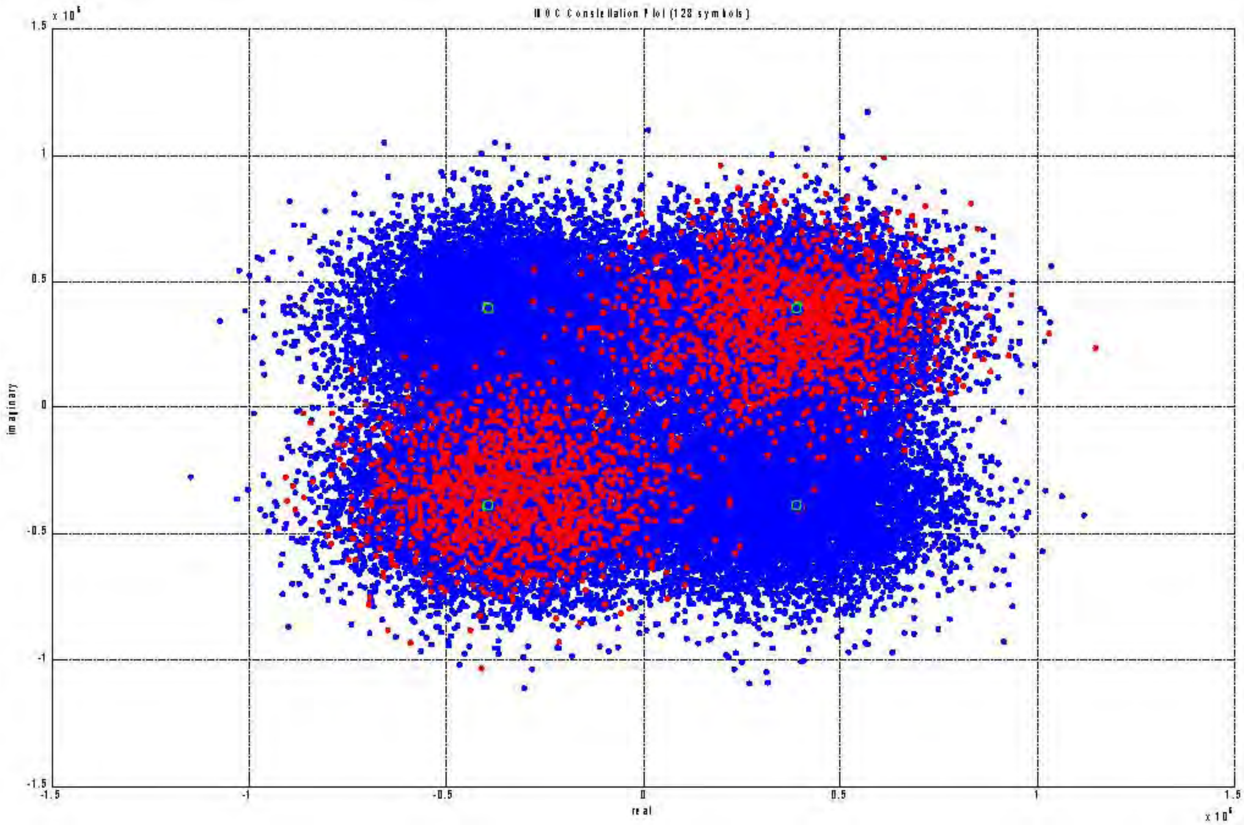


Figure B-10: IBOC Constellation with PAPR Reduction Disabled – Cd/No = 56 dB-Hz

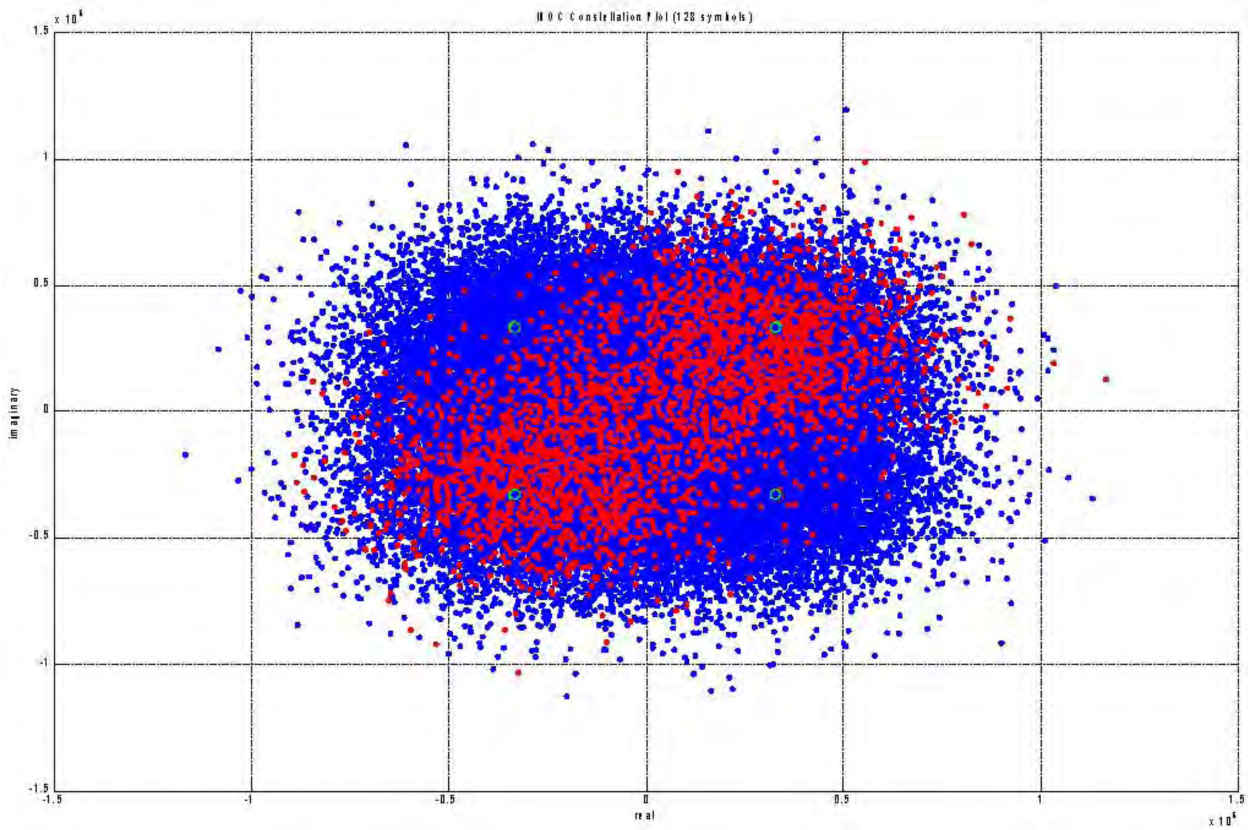


Figure B-11: IBOC Constellation with PAPR Reduction Disabled – Cd/No = 52 dB-Hz

Plots of MER versus Partition Index/Subcarrier Index

For all figures (Figure B-12 through Figure B-21), the MER versus subcarrier index are plotted. The lower sideband is shown on the left and the upper sideband is shown on the right; however, the values on the horizontal axis do not directly correspond to the subcarrier index.

Data MER versus Partition Index

Figure B-12 shows the MER for each individual data partition for the case of PAPR reduction enabled and no noise.

Figure B-13 through Figure B-16 show the MER for each individual data partition for the case of PAPR reduction disabled with various values of Cd/No. From these graphs, the worst case data MER values can be obtained. By comparing Figure B-12 to the other figures, the effect of PAPR reduction versus channel noise on the worst case MER can be compared.

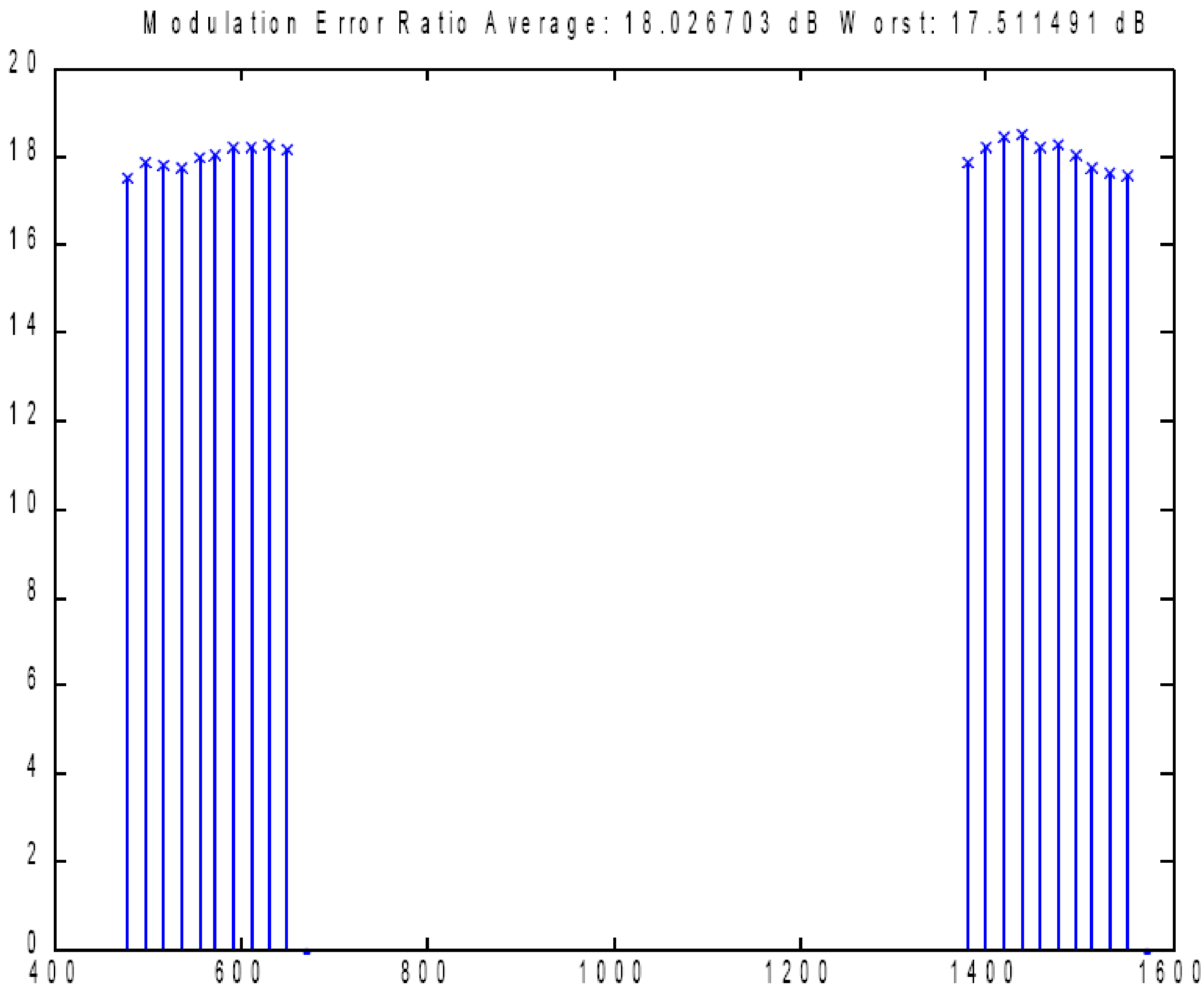


Figure B-12: Data MER with PAPR Reduction Enabled – no noise

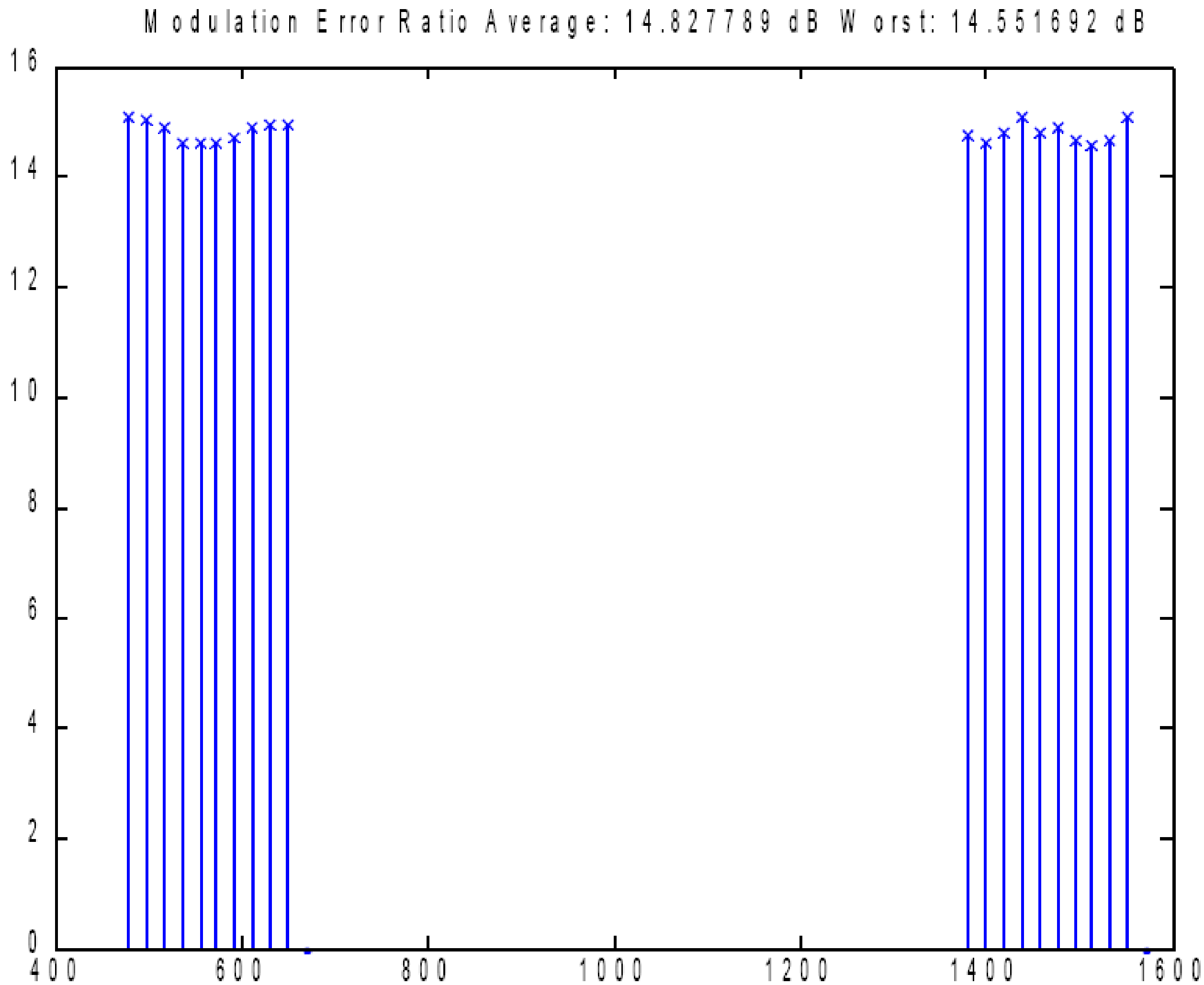


Figure B-13: Data MER with PAPR Reduction Disabled – Cd/No = 68 dB-Hz

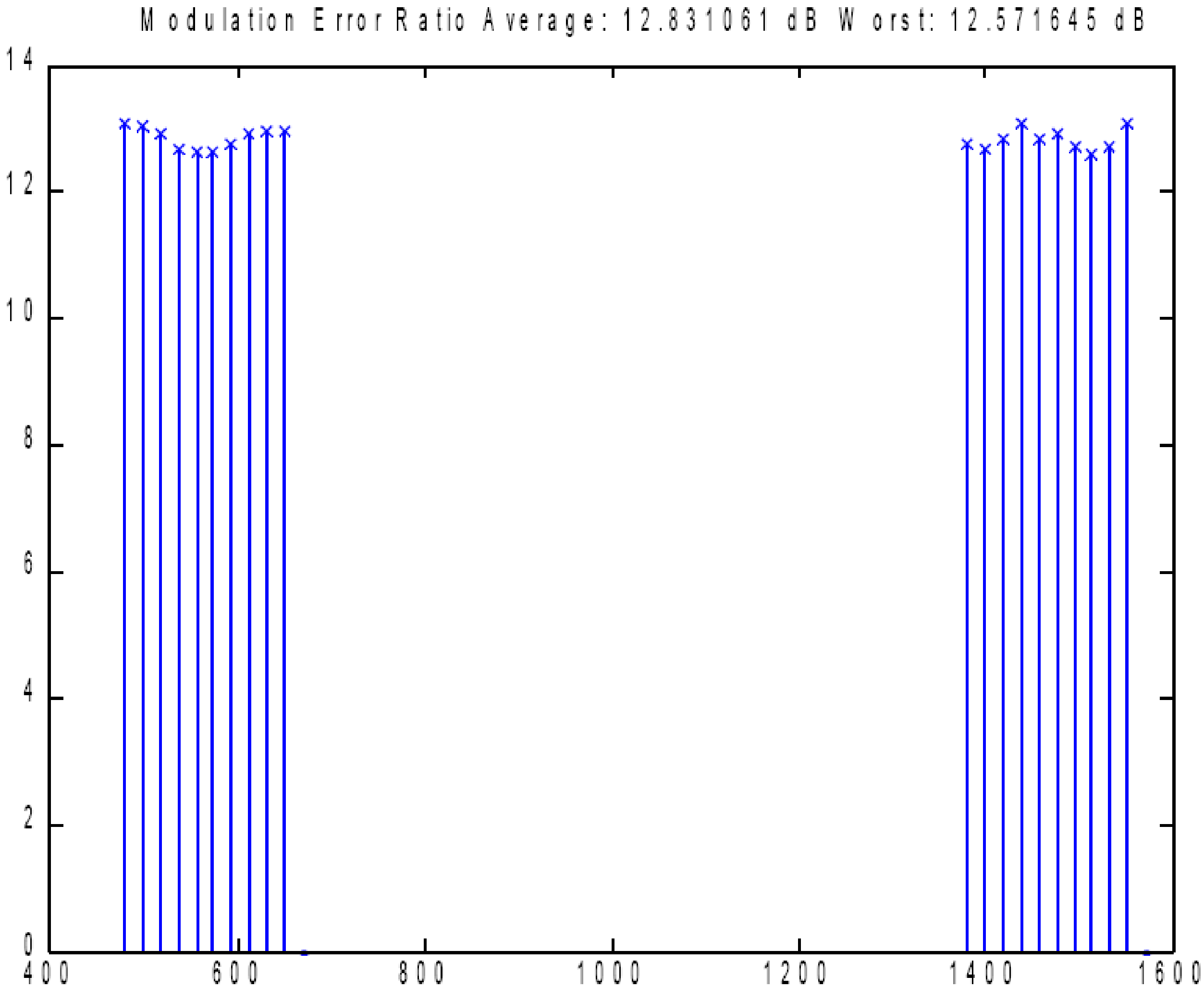


Figure B-14: Data MER with PAPR Reduction Disabled – Cd/No = 64 dB-Hz

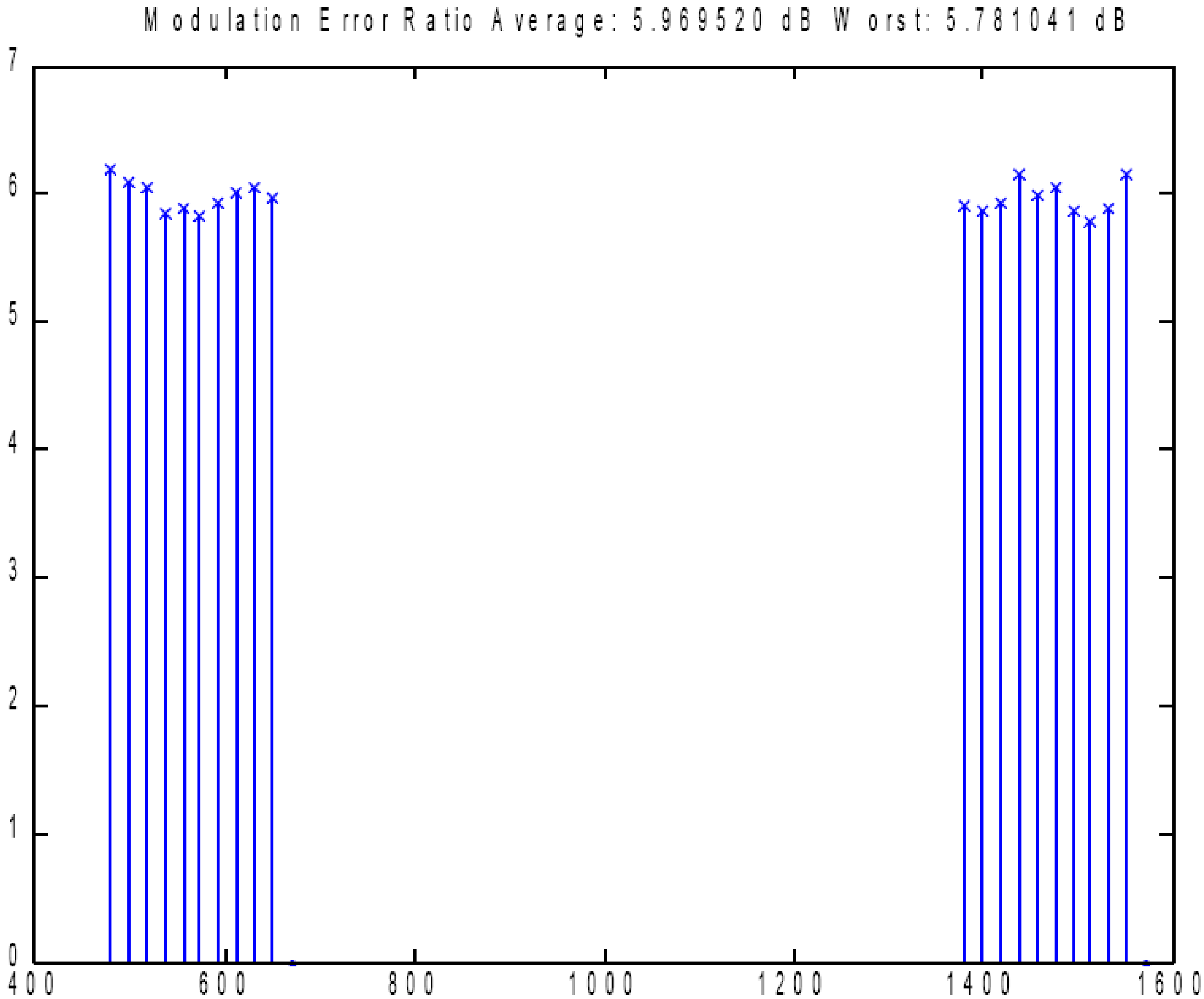


Figure B-15: Data MER with PAPR Reduction Disabled – Cd/No = 56 dB-Hz

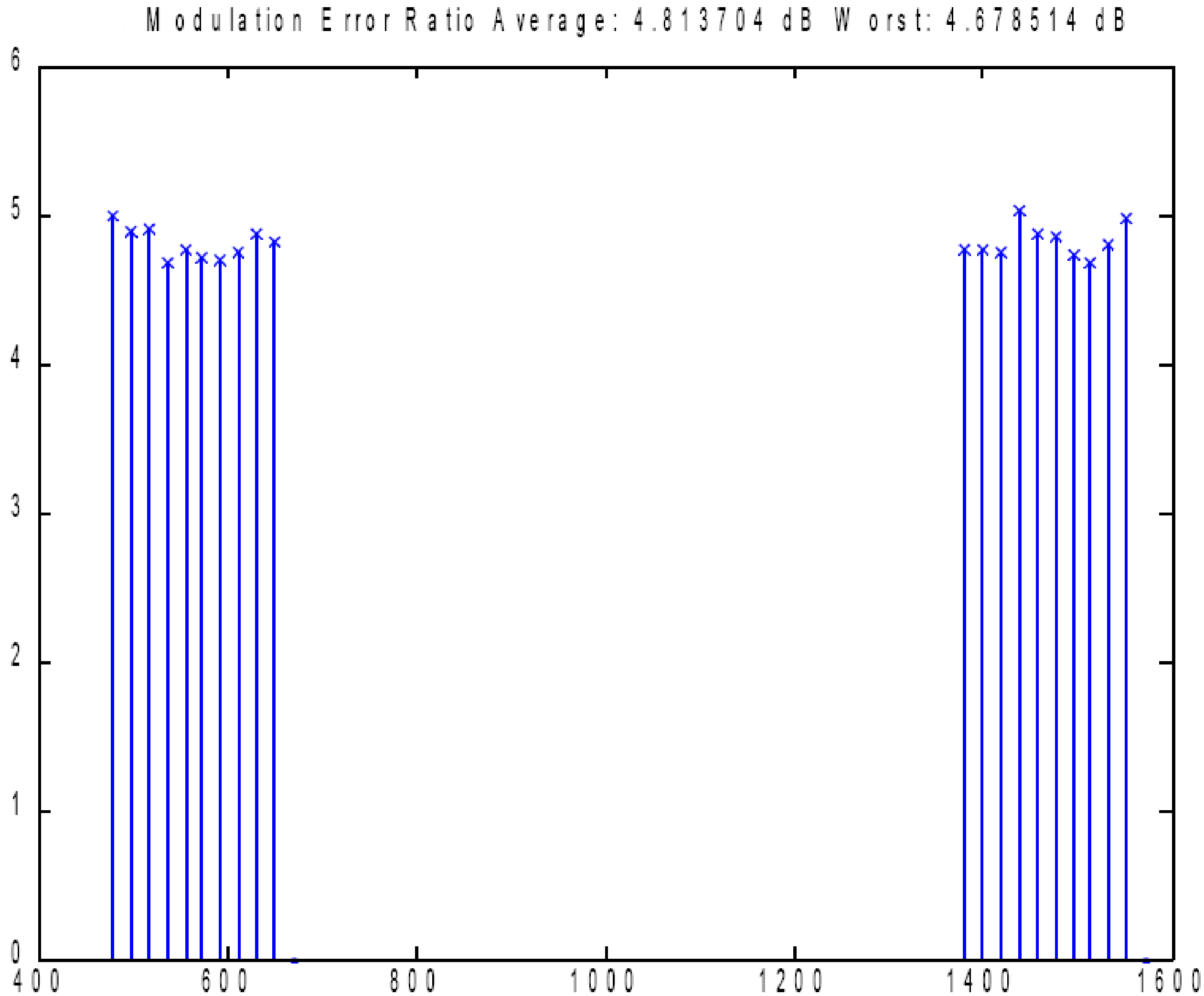


Figure B-16: Data MER with PAPR Reduction Disabled – Cd/No = 52 dB-Hz

Reference MER versus Reference Subcarrier Index

Figure B-17 shows the MER for each individual reference subcarrier for the case of PAPR reduction enabled and no noise.

Figure B-18 through Figure B-21 show the MER for each individual reference subcarrier for the case of PAPR reduction disabled with various values of Cd/No. From these graphs, the worst case reference subcarrier MER values can be obtained. By comparing Figure B-17 to the other figures, the effect of PAPR reduction versus channel noise on the worst case MER can be compared.

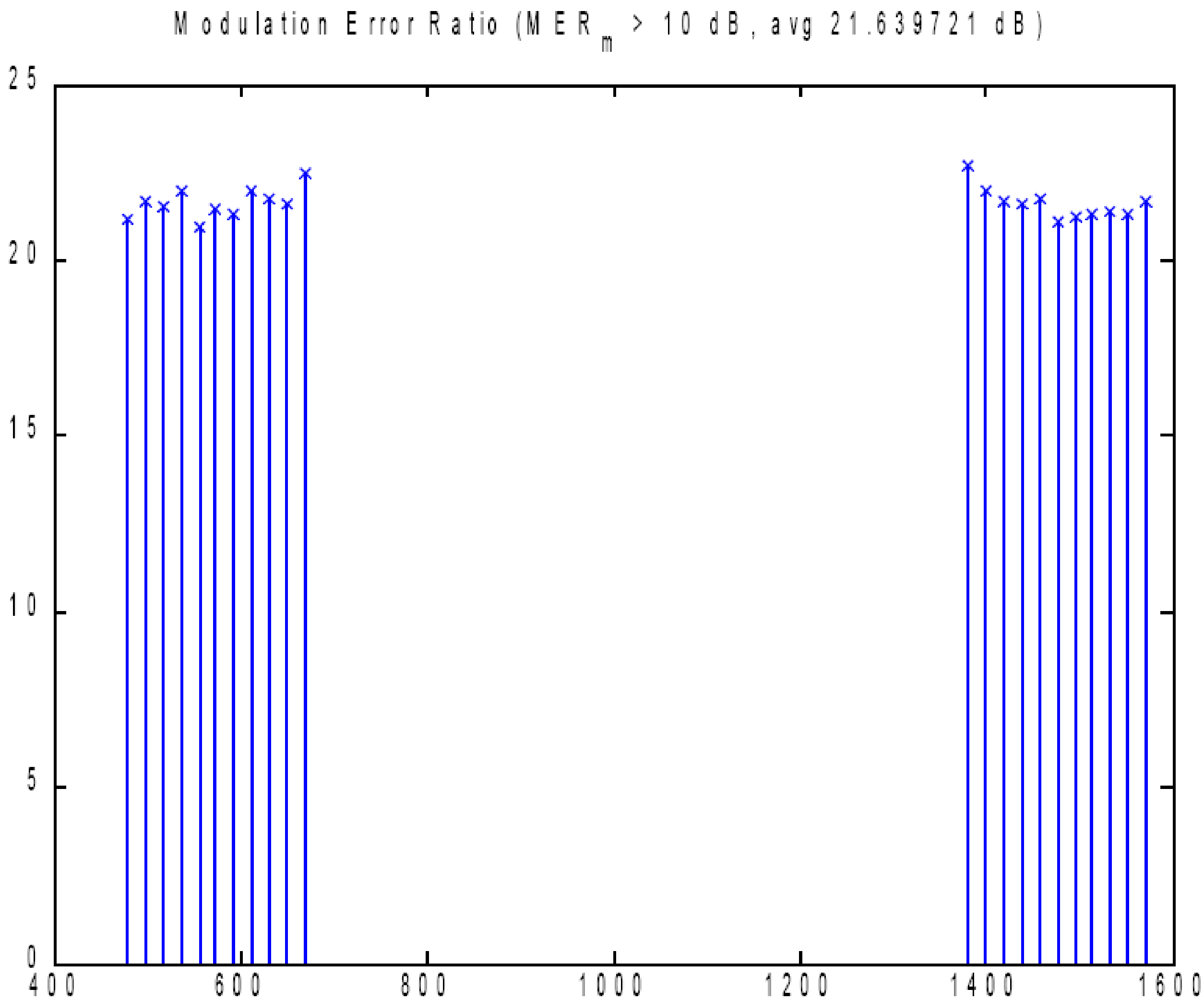


Figure B-17: Reference MER with PAPR Reduction Enabled – no noise

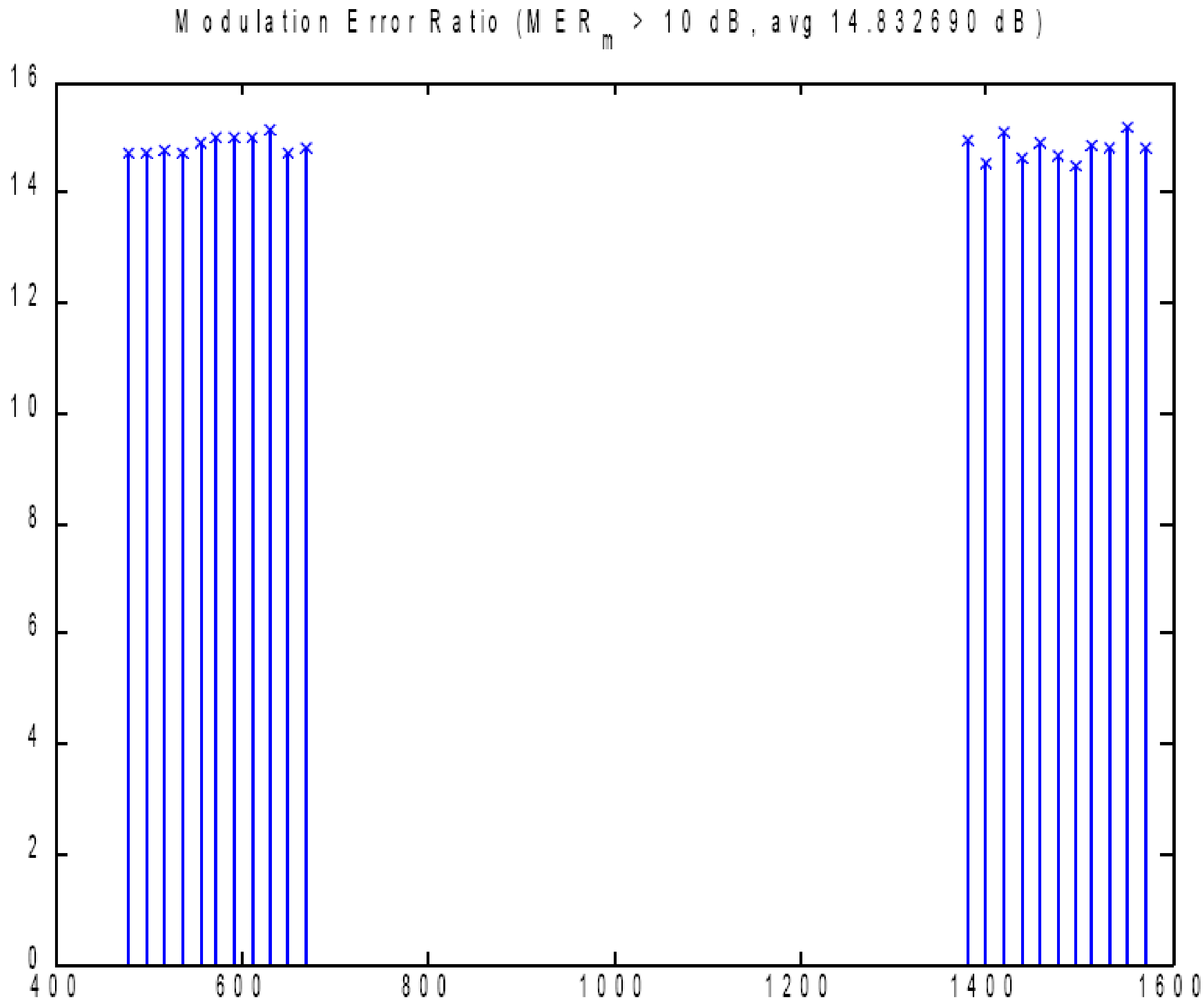


Figure B-18: Reference MER with PAPR Reduction Disabled – Cd/No = 68 dB-Hz

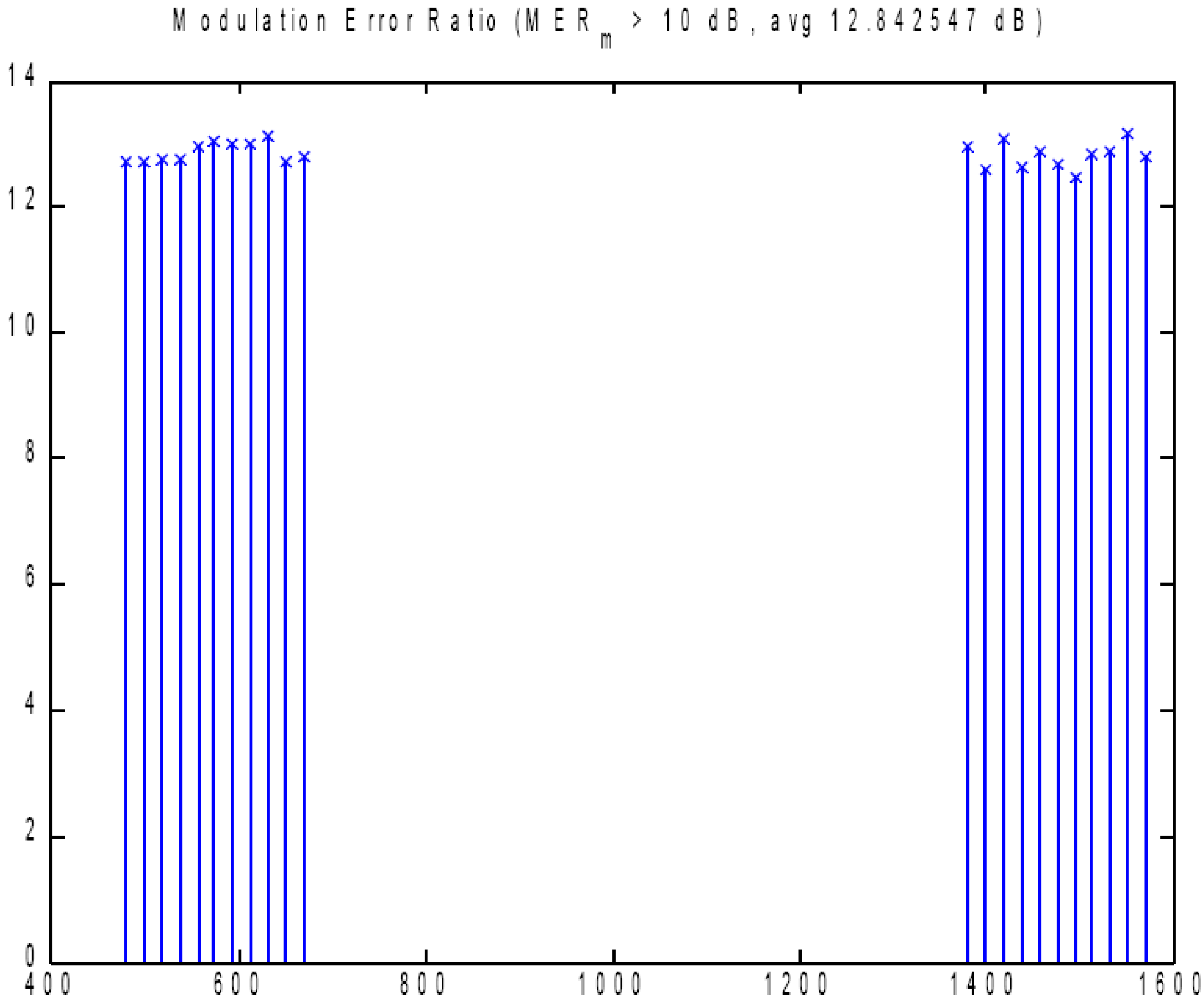


Figure B-19: Reference MER with PAPR Reduction Disabled – Cd/No = 64 dB-Hz

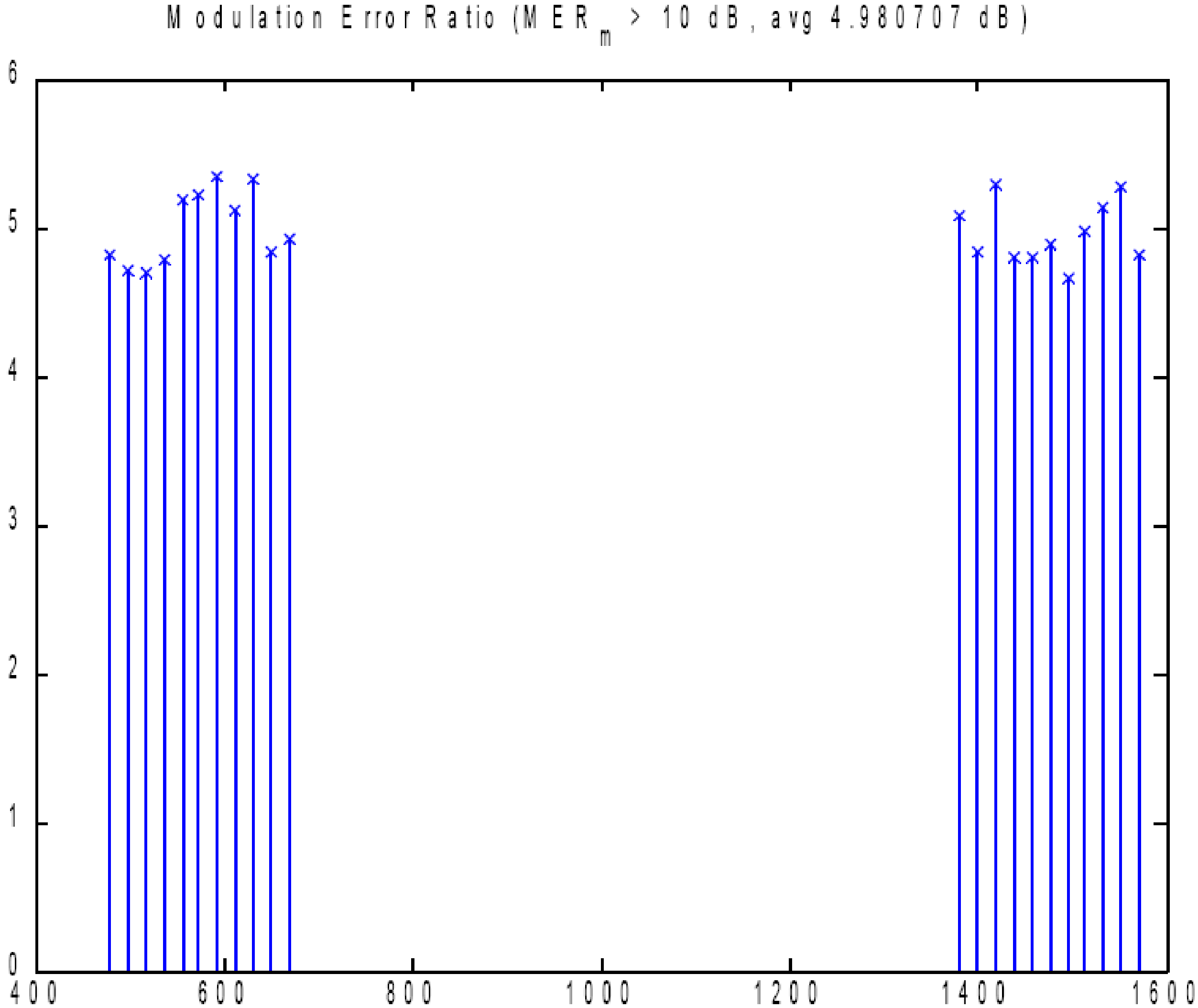


Figure B-20: Reference MER with PAPR Reduction Disabled – Cd/No = 56 dB-Hz

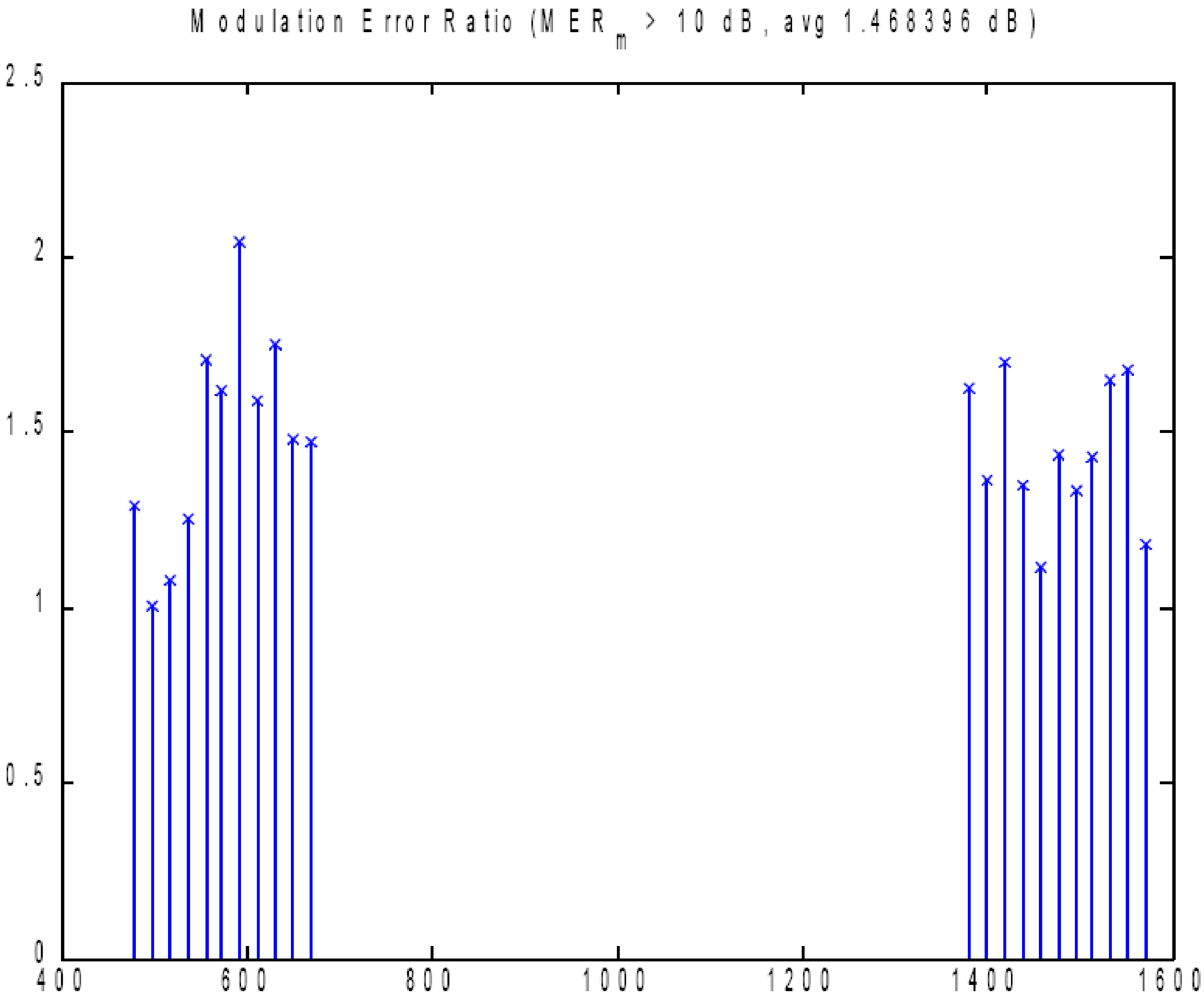


Figure B-21: Reference MER with PAPR Reduction Disabled – Cd/No = 52 dB-Hz