# WPAN Interference cancellation from OFDM based WLAN : Strategic comparisons of coexistence techniques

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***Abstract*— WLAN and WPAN operate in the same 2.4GHz ISM band. So whenever they operate nearby interfere with each other, results in performance degradation for both the system in terms of bit error rate or throughput. In this paper we give an algorithm for estimating and canceling the narrowband WPAN interference from OFDM based WLAN receiver in AWGN, multipath fading channels is given. First a generalization of the algorithm is given which is applied in the simulation with specific interference modeling of Bluetooth or Zigbee to cancel out the interference. The algorithm performs best when the interference power of the WPAN system is high as the estimation of frequency, power, and phase of the WPAN signal are more accurate. BER curves, for SIR = -10dB, are almost the same as the BER curves without interference, implies full interference cancellation. Next we compare the results of using existing techniques for interference mitigation or avoidance and the approach proposed in this paper. Simulation result show when WPAN system Bluetooth is using adaptive frequency hopping and WLAN is using proposed cancellation technique, BER is decreasing with increase in SIR values for both the system which guarantees coexistence.**

***Index Terms*— WLAN,WPAN, Zigbee, Bluetooth, interference, OFDM, GFSK, OQPSK**

## I. Introduction

THE 2.4GHz ISM band is congested with wireless system like WLAN, WPANs (Bluetooth, Zigbee).When multiple RF systems operating near each other and transmitting at the same time, at same frequency range interfere with each other. If the power of interfering signal is high, it may corrupt the other systems' signals significantly. As a consequence performance of the interfered systems deteriorates severely. So the use of the unlicensed ISM band for simultaneous use by multiple standards is a serious issue Therefore, for coexistence, each of these systems should use some interference mitigation or avoidance techniques.

WLAN system operate in 2.4GHz band is based on IEEE802.11g standard. It is for communication within 100m range. It is uses OFDM modulation for transmitting data at maximum speed of 54Mbps , having high immunity for inter carrier interference or intersymbol interference. Bluetooth is based on IEEE802.15.1 standard operates in the ISM band and has gained popularity for wireless connectivity for the short range (within 10m) and low power applications. Zigbee is another WPAN system is based on IEEE802.15.4 standard, used for sensor networking. Zigbee operates in 16 channels each with a 5 MHz band and a raw data rate of 250 kbps using OQPSK modulation.

Even though literature gives an analysis and the severity of the problem, there are no promising solutions to the coexistence problem. The performance in terms of throughput, BER or SER of the devices degrades severely in presence of an interfering device operating nearby. Papers [9]-[23] show how the devices operating in free unlicensed ISM band are affected in presence of other devices operating in the same frequency bands.

The OFDM based WLAN occupies 22 MHz band whereas Bluetooth or Zigbee occupies a band of 1 MHz and 2 MHz respectively. Hence for the WLAN system, Bluetooth or Zigbee is a narrowband interferer. Based on this property there are techniques present in literature to avoid or mitigate interference. One such technique is use of erasures [9] or deterministic nulling particularly valid for OFDM based system to avoid narrowband interference. In using erasures, OFDM based WLAN has to detect the interfering signals frequency or the sub-carriers which are affected significantly by the interference signal and then the interference affected sub-carriers have to be replaced by null sub-carriers, meaning nothing is transmitted using these interference affected sub-carriers. When the interference signal power is such that only a few sub-carriers are affected, this technique can reduce the BER for high SNR by dropping the overall throughput of the system. But in case the interfering power is high enough to affect larger numbers of the OFDM sub-carriers, the performance of WLAN in terms of throughput or BER will not improve using erasures. The reason behind this is, when narrowband interferer's power is high, not only the sub-carriers falling inside the narrowband signal's band will be affected but the adjacent sub-carriers of the narrowband system will also be affected. This is because leakage power in the adjacent sub-carriers due to narrowband signal is high

Manuscript received September 10, 2011

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enough to corrupt the adjacent sub-carriers. In fact it may so happen that for all the OFDM sub-carriers, the ratio of effective sub-carrier power to interference power decreases to such an extent that error free recovery for even a single sub-carrier is not possible.

Adaptive frequency hopping is another technique for interference avoidance which is presently used by Bluetooth. In the adaptive frequency hopping technique [23] Bluetooth scans the channel and mark the bands occupied by interferer as “BAD”. Bluetooth avoids these frequencies. Another technique used in OFDM based systems to cancel out narrowband interference is by predicting the interference signal using null sub-carriers [17]. Since interference is a signal for another system, if it is properly modeled, the information transmitted by the interference can be extracted. The null sub-carriers do not carry any information in OFDM system. These null carriers can be used to extract the information of interference signal. Once the information of the narrowband interference is extracted and interference component for each sub-carrier is estimated, then subtracting those interference components from received sub-carrier signal would result in interference free sub-carrier. This technique is good when narrow band signal’s frequency fall in null sub-carrier part of the OFDM signal. But when narrow band signal’s frequencies fall in the data sub-carriers part of the OFDM system this technique does not give good performance. If null sub-carriers are used to extract the interference signal while the narrowband interferer falls in data sub-carrier part, both noise and interference component increase making effective ratio of sub-carrier’s signal to noise power low.

In [17] a narrow band interference (NBI) suppression method is proposed where NBI is caused by a narrowband digital communication system. The “transmitted data” of NBI signal is estimated and its waveform reconstructed by measuring interference information on certain unmodulated sub-carriers. Next the estimated disturbance in frequency domain is subtracted from received signal. Clearly, if noise power is high, leading to erroneous computation of data vector**.** Another disadvantage with this technique is that it is inefficient when the interference signal hops on sub-carriers other than the null sub-carriers. Under such circumstances, extracting the transmitted data using the pseudo inverse enhances both the interference component and the noise component. This again leads to erroneous estimation. In the proposed algorithm null carriers are used to extract out interference using minimum distance principle. Both the limitation of the algorithm of ref[17] is not present in our algorithm.

The organization of the paper is given : In Section I Overview of OFDM based WLAN sysem, GFSK based Bluetooth system, OQPSK based Zigbee system is given. In section III Interference mitigation algorithm with modeling of WPAN interference signal and modeling of channel is given. In section IV simulation results in BER plots, tables is given . Section V concludes the paper by summarizing simulation results.

## II. OVERVIEW OF OFDM BASED WLAN SYSTEM, GFSK BASED BLUETOOTH SYSTEM, OQPSK BASED ZIGBEE SYSTEM

### A. *OFDM based WLAN system*

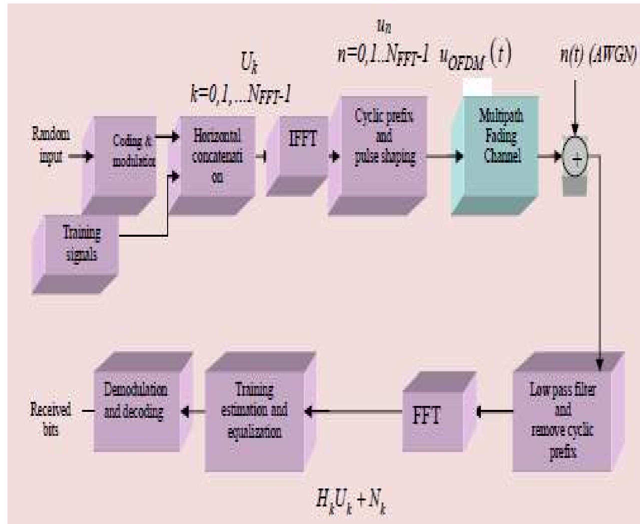



Fig 1 Block diagram of WLAN transreceiver

In this section an overview of OFDM followed by its features like immunity to ISI and ICI is described. In OFDM systems data are coded with a certain code rate that modulate carriers using the MQAM modulation technique. For the 54Mbps data rate, OFDM signal code rate is 3/4 and modulation scheme is 64QAM. In 64QAM modulator generates a complex valued symbol that is passed through IFFT Block. The output of the IFFT block is given by [4] :

$$u(t)=\sum_{k=0}^{N_{FFT}-1} U(k)exp\left(\frac{j2\pi kt}{T_{OFDM}}\right), \qquad 0\le t\le T_{OFDM} \qquad \text{(II-1)}$$

Where $T_{OFDM}$ is the symbol period of OFDM signal and *U (k)* is the signal before IFFT corresponding to the $k^{th}$ carrier. The carriers $exp\ (j2\pi kt/T_{OFDM})$ are orthogonal over 0 to $T_{OFDM}$. So the signal can be recovered at the receiver for the $l^{th}$ sub-carrier using the orthogonality property as given in [4]:

$$U_k=\sum_{k=0}^{N_{FFT}-1} U(k)\frac{1}{T_{OFDM}}\int_0^{T_{OFDM}} exp\left(\frac{j2\pi(k-l)t}{T_{OFDM}}\right)dt \qquad \text{(II-2)}$$

Let $\mathbf{U}=\left[U_0U_1......U_{N_{FFT}-1}\right]^T$, $\mathbf{u}=\left[u_0u_1.......u_{N_{FFT}-1}\right]^T$ be $N_{FFT}\times 1$ vector of transmitted symbols before and after IFFT respectively and $\boldsymbol{F}$, $\boldsymbol{F^H}$ be $N_{FFT}\times N_{FFT}$ FFT and IFFT matrix respectively[1].

$$\mathbf{u}=\boldsymbol{F^H}\bullet\boldsymbol{U} \qquad \text{(II-3)}$$

Cyclic prefix is inserted by pre-multiplying the matrix **u** with matrix $\boldsymbol{T_{cp}}$ of dimension $(N_{cp}+N_{FFT})\times N_{FFT}$, where $N_{cp}$ is length of cyclic prefix. This produces the prefix inserted vector $\boldsymbol{u_{cp}}=\boldsymbol{T_{cp}}\bullet\boldsymbol{u}$, which is next passed through a pulse shaping

filter of impulse response $h_{ps}(t)$ and then the signal is transmitted over the channel.
At the receiver, the received signal is first taken through a low pass filter and cyclic prefix is removed by pre-multiplying with matrix $\boldsymbol{R_{cp}}$ of dimension $N_{FFT} \times (N_{cp} + N_{FFT})$. The FFT operation is performed to recover the signal for the corresponding sub-carriers next. Let $\mathbf{R} = \left[ R_0 R_1 .......R_{N_{FFT}-1} \right]^T$ be the $N_{FFT} \times 1$ received vector after performing FFT operation. Then,

$$\boldsymbol{R} = \boldsymbol{F} \bullet \boldsymbol{R_{cp}} \bullet \boldsymbol{H_0} \bullet \boldsymbol{T_{cp}} \bullet \boldsymbol{F^H} \bullet \boldsymbol{U} + \boldsymbol{N} \qquad \text{(II-4)}$$

Where

$$\boldsymbol{H_0} = \begin{bmatrix} h_0 & 0 & 0 \ldots\ldots & \ldots\ldots & 0 \\ h_1 & h_0 & 0 \ldots & \ldots\ldots & 0 \\ . & . & & & . \\ . & & . & & . \\ h_{L-1} & \ldots\ldots & h_0 & . \; . & 0 \\ . & . & & . & \\ . & & . & & . \\ 0 & . \; . \; . \; . & h_{L-1} & \ldots\ldots & h_0 \end{bmatrix} \qquad \text{(II-5)}$$

$\boldsymbol{H_0}$ *is* the $(N_{FFT} + L) \times (N_{FFT} + L)$ channel matrix and $\boldsymbol{N}$ is vector of AWGN. $\boldsymbol{h} = (h_{L-1} .......h_1 \; h_0)^T$ denotes the tap values of the equivalent channel filter which is given by convolution of the impulse response of the transmitter pulse shaping filter $h_{ps}$, channel impulse response $h_{chann}$ and receiver low pass filter impulse response $h_{lpf}$

$$h(t) = h_{ps}(t) \otimes h_{chann}(t) \otimes h_{lpf}(t). \qquad \text{(II-6)}$$

Now the inner matrix $\tilde{\boldsymbol{H}} = R_{cp} \bullet H \bullet T_{cp}$ is a circulant matrix and from the properties of circulant matrices, equation (II-6) can be derived.

$$\boldsymbol{F} \bullet \tilde{\boldsymbol{H}} \bullet \boldsymbol{F^H} = diag\left\{ H\left(e^{j0}\right), H\left(e^{j2\pi/N_{FFT}}\right), \ldots, H\left(e^{j2\pi(N_{FFT}-1)/N_{FFT}}\right) \right\} \qquad \text{(II-7)}$$

Thus use of cyclic prefix maintains the orthogonality of the carrier. As a result, the information symbol is just weighted by a factor of channel transfer function and the output after FFT is given by:

$$R_k = H_k U_k + N_k \qquad \text{(II-8)}$$

Equation (II-**7)** can be rewritten as:

$$\boldsymbol{R} = \boldsymbol{H} \bullet \boldsymbol{U} + \boldsymbol{N} \qquad \text{(II-9)}$$

Where H = *diag [H(0), H(1),..., H($N_{FFT-1}$)]*, $\boldsymbol{N}$ = *[N(0), N(1),….N($N_{FFT}$-1)] '*
From the equation(II-8) it is seen that only single tap equalizer is needed to nullify the effect of channel for OFDM signal.

*B. GFSK based Bluetooth System*

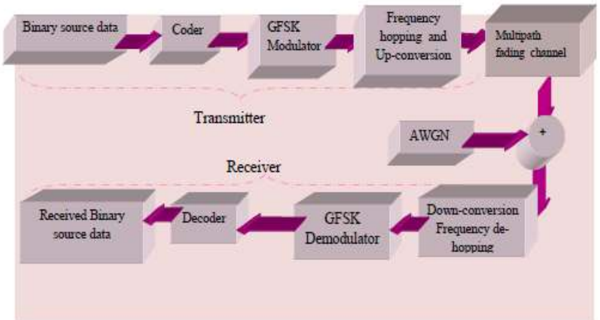



Fig 2: Block diagram of Bluetooth transceiver

The basic block diagram of a Bluetooth transceiver with GFSK modulation is given figure (2). The GFSK modulator is identical to FSK modulator except that before the base band pulse goes into the FSK modulator, it is passed through a Gaussian filter. This pulse shaping results in narrow bandwidth with sharp cut off frequencies which suppresses the high frequency components of the transmitted signal. The pulse shaped data is then passed to binary FSK modulator and then up-converted to a particular hopping frequency. In the ISM frequency band, Bluetooth uses 79 hopping frequencies which are 1 MHz wide each and a data rate of 1 Mbps. As minimum duration of one Bluetooth packet is 625µs and Bluetooth hop frequency remains same through the packet duration, its hop rate is 1600 hops/s. Bluetooth hopping frequencies are given by (2402 + $k$) *MHz*, where $k$ = 0, 1, 2, 3….. 78.
The frequency hopped GFSK modulated signal is written as follows [12][14]:

$$S_{BT}(t, \boldsymbol{D}) = \left( A\cos\left(2\pi\left(f_c + f_l\right)t + \Phi(t; \boldsymbol{D})\right)\right) \qquad \text{(II-10)}$$

and its low pass equivalent waveform is given by:

$$S_{BT}(t, \boldsymbol{D}) = \mathrm{Re}\left( A e^{\left(j\left(2\pi f_l t + \Phi(t; \boldsymbol{D})\right)\right)} \right) \qquad \text{(II-11)}$$

Where, $A = \sqrt{2E_b/T_B}$, $E_b$ is the energy per data bit, $T_B$ is the symbol period, $f_c, f_l$ is the carrier and hopping frequencies of the Bluetooth signal respectively and $l \in \{0, 1..78\}$. $\Phi(t, \boldsymbol{D})$ is the time varying phase of the Bluetooth signal which can be expressed in the interval $nT_B \leq t \leq (n+1)\, T_B$ as:

$$\Phi(t, \boldsymbol{D}) = 2\pi h D_n q\left(t - nT_B\right) + \theta_n \qquad \text{(II-12)}$$

Where, $q(t) = \int_{-\infty}^{t} g(\tau) d\tau$ (II-13)

Here $g(\tau)$ is the shaping pulse which is obtained by filtering a rectangular pulse with Gaussian filer and $D_n \in \{+1, -1\}$ denotes random data input and $h$ is modulation index. In equation(II-12) the second part $\theta_n$ denotes the accumulation (memory) of all the symbols up to time $(n-1)\, T_B$ and first part of equation(II-12) is dependent on the data bit $D_n$ at time instant $nT_{B.}$. The time bandwidth product of Bluetooth is given by $WT_B = 0.5$. The response of the *g (τ) is* given by:

$$g(\tau)=0.5\left[erfc\left(\pi\sqrt{\frac{2}{\log 2}}WT_B\left(\frac{\tau}{T_B}-0.5\right)\right)-erfc\left(\pi\sqrt{\frac{2}{\log 2}}WT_B\left(\frac{\tau}{T_B}+0.5\right)\right)\right]$$

*C. QPSK based Zigbee system*

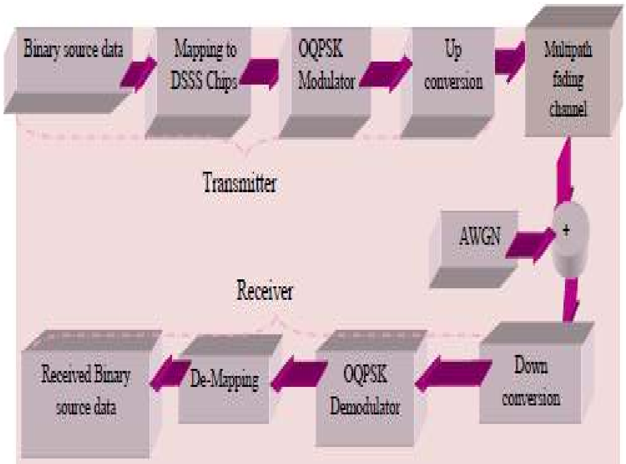



Fig 3: Block diagram of Zigbee Transceiver

Zigbee operates in 16 channels each with a 5 MHz band and a raw data rate of 250 kbps using OQPSK modulation. To make the signal robust and less susceptible to interference Zigbee uses Direct Sequence Spread Spectrum (DSSS) for its operation. A 4-bit data is mapped into 32–chip-length pseudo noise (PN) sequence. The 250 kbps bit rate results in 62.5 k symbol/s and each symbol represents a 32 chip sequence with a chip rate of 2 M chip/s. Though the available bandwidth for Zigbee system is 5 MHz per channel, effective band occupied by Zigbee is 2 MHz and the remaining 3 MHz band is used for removing inter symbol interference (ISI) OQPSK is a form of QPSK that transmits two bits at a time using in phase (I) and quadrate phase (Q) components of a signal. The only difference between QPSK and OQPSK is that the Q phase component is delayed one bit with respect to I phase component in OQPSK. This delay allows only one zero crossing to occur at a time which gives OQPSK phase transition limit to $90^0$ instead of $180^0$ in QPSK. For half sine wave pulse this results in one phase of the signal being at its peak while the other is at zero allowing a more reliable demodulation. The OQPSK signal is given by:

$$s(t) = d_I(t)\,p(t)\cos(2\pi ft) + d_Q(t)\,p(t-T_Z)\cos(2\pi ft) \qquad \text{(II-14)}$$

Where $$p(t)=\sin\left(\frac{\pi t}{2T_Z}\right) \quad 0\le t\le 2T_Z; \qquad =0 \quad elsewhere \qquad \text{(II-15)}$$

Then,

$$s(t) = d_I(t)\sin\left(\frac{\pi t}{2T_Z}\right)\cos(2\pi ft) + d_Q(t)\cos\left(\frac{\pi t}{2T_Z}\right)\cos(2\pi ft) \qquad \text{(II-16)}$$

Where $d_I(t)$, $d_Q(t)$ (+ 1 or -1) are input binary data corresponding to in phase and quadrature phase components of the transmitted signal

## III. INTERFERENCE CANCELLATION ALGORITHM

In this first a generalized version of the narrowband cancellation algorithm is given. Then it is shown how to model the WPAN interference (Bluetooth or Zigbee) signal. After that a method for estimating different parameters of interference using training signal is given. Frequency is estimated first to find out which part of the spectrum the interference signal is present. Then using the estimated frequency, interfering signal power, phase, and equivalent channel tap values are estimated.

*A. Generalized interference cancellation algorithm:*

In OFDM based WLAN receiver the received signal after FFT can be written as:

$$R_k = H_k U_k + I_k + N_k \qquad \text{(III-1)}$$

$I_k$ denotes the interference component for the $k^{th}$ sub-carrier. In an OFDM system, 12 sub-carriers are null carriers among the 64 sub-carriers. The received data corresponding to the null sub- carriers can be written as:

$$R_k = I_k + N_k \qquad \text{(III-2)}$$

For low noise the received signal corresponding to $k^{th}$ null sub-carrier represents the interference only. It is shown in the chapter 4 that interference $I_k$ can be modeled as :

$$I_k = \sum_{n=0}^{p-1} a_{n,k} D_n + C_k \qquad \text{(III-3)}$$

where $D_n$ is the unknown binary (belongs to + 1 or -1) transmitted data of WPAN interference signal. $a_{n,\ k}$ is the coefficient corresponding to $n^{th}$ binary input data, $k^{th}$ sub-carrier and $C_k$ is a constant corresponding to $k^{th}$ sub-carrier. The details of finding $a_{n,k}$, $C_k$ by estimating different narrowband properties like power, frequency, phase is given in a subsequent section with specific model of narrowband WPAN signal. Assuming that narrowband signal has been identified accurately with its spectrogram [18] the generalized interference cancellation algorithm is as given below:

**Step1**: Estimate the signal frequency, power, and initial phase of the narrow band WPAN signal using training signal of the OFDM system.

**Step2**: With specific model of narrowband interference in OFDM receiver and using the estimated parameters found in step 1 obtain the $a_{n,k}$, and constant $C_k$.

**Step3**: For each combination of data input vector $\boldsymbol{x_j} = \{D_0 D_1 D_2 \ldots D_{p-1}\}$, where $j \in \{0,1,..,N-1\}$, $N = 2^p$, find the interference component $I_{k,x_j}$ corresponding to $k^{th}$ null sub-carrier using the equation given below:

$$I_{k,x_j} = \sum_{n=0}^{p-1} a_{n,k} D_n + C_k \qquad \text{(III-4)}$$

**Step 4**: Find the difference between the received signal $R_k$ for $k^{th}$ null sub-carrier and estimated interference component of the $k^{th}$ null sub-carrier (obtained from step 3) which is given by:

$$E_{k,x_j} = R_k - I_{k,x_j} \qquad \text{(III-5)}$$

**Step 5**: Find the magnitude of $E_{k,x_j}$. Average $\left|E_{k,x_j}\right|$ over all the chosen set of null carriers corresponding to a particular set of input vector $x_j$ is given by:

$$E_{x_j} = \frac{1}{p} \sum_{k=0}^{p-1} \left|E_{k,x_j}\right| \qquad \text{(III-6)}$$

**Step 6:** Choose $\boldsymbol{x_j} = \{D_0 D_1 D_2 \ldots D_{p-1}\} \in \{\boldsymbol{x_0}, \boldsymbol{x_1}, \ldots, \boldsymbol{x_{N-1}}\}$ that will minimize the average error $E_{x_j}$ and obtain the interference component for all the sub-carriers of the OFDM signal for the above chosen $x_j$ and the estimated values of $a_{n,k}$, $C_k$.

**Step 7**: Subtract the above interference component for each sub-carrier from the received signal in OFDM system.

For the null subcarriers the received signal gives the Interference component only when AWGN is very low. The particular combination of binary data vector, $\boldsymbol{x_j} \in \{\boldsymbol{x_0}, \boldsymbol{x_1}, \ldots, \boldsymbol{x_{N-1}}\}$ that gives the interference for the $k^{th}$ null subcarrier nearer the value to the received signal component will be chosen as the transmitted binary data vector from the Bluetooth system for the particular symbol duration of OFDM system. This is the logic behind step 6.

### B. *Interference model of Bluetooth System*

from equation (II-10) using complex valued data at the output of the GFSK modulator can be written as:

$$S_{BT}^{c}(t, \boldsymbol{D}) = A e^{\left(j(2\pi f_l t + \theta_n)\right)} \begin{pmatrix} \cos\left(2\pi h D_n q(t - nT_B)\right) \\ + j \sin\left(2\pi h D_n q(t - nT_B)\right) \end{pmatrix} \qquad \text{(III-7)}$$

As $D_n$ is + 1 or -1 and *cos (x) = cos (-x), sin (x) = -sin (-x)*; Equation(III-7) can be rewritten as:

$$S_{BT}^{c}(t, \boldsymbol{D}) = A e^{\left(j(2\pi f_l t + \theta_n)\right)} \begin{pmatrix} \cos\left(2\pi h q(t - nT_B)\right) \\ + j D_n \sin\left(2\pi h q(t - nT_B)\right) \end{pmatrix} \qquad \text{(III-8)}$$

Let $A e^{\left(j(2\pi f_l t)\right)} \cos\left(2\pi h q(t - nT_B)\right) = P_l\ (t)$ (III-9)

$$jA e^{\left(j(2\pi f_l t)\right)} \sin\left(2\pi h q(t - nT_B)\right) = Q_l\ (t) \qquad \text{(III-10)}$$

For $P_l\ (t)$ and $Q_l\ (t)$ the subscript "$l$ "denote $l^{th}$ hopping frequency $f_l$

Then equation (III-7) becomes

$$S_{BT}^{c}(t, \boldsymbol{D}) = \left(1 \bullet P_l\ (t) + D_n \bullet Q_l\ (t)\right) \bullet e^{j\theta_n} \qquad \text{(III-11)}$$

If there are $N$ samples for one Bluetooth symbol period, then in matrix form equation can be written as:

$$\boldsymbol{S^c_{BT_{n,l}}} = \begin{bmatrix} P_l(nT_B) & Q_l(nT_B) \\ P_l(nT_B + T_S) & Q_l(nT_B + T_S) \\ . & \\ . & \\ P_l(nT_B + (N-1)T_S) & Q_l(nT_B + (N-1)T_S) \end{bmatrix} \qquad \text{(III-12)}$$

$$= \boldsymbol{B^0_{PQ_{n,l}}} \bullet \boldsymbol{Y}_n \bullet e^{j\theta_n}$$

where subscript $n$, $l$ denotes the signal transmitted at $n^{th}$ time instant with $l^{th}$ hopping frequency by Bluetooth system.

and, $\boldsymbol{Y}_n = \begin{bmatrix} 1 & D_n \end{bmatrix}^T$

$$\boldsymbol{B^0_{PQ_{n,l}}} = \begin{bmatrix} P_l(nT_B) & Q_l(nT_B) \\ P_l(nT_B + T_S) & Q_l(nT_B + T_S) \\ . & \\ . & \\ P_l(nT_B + (N-1)T_S) & Q_l(nT_B + (N-1)T_S) \end{bmatrix} \qquad \text{(III-13)}$$

Similarly samples corresponding to next Bluetooth symbol (after GFSK modulation) can be written as:

$$\boldsymbol{S^c_{BT_{n+1,l}}} = \boldsymbol{B^0_{PQ_{n+1,l}}} \bullet \boldsymbol{Y}_{n+1} \bullet e^{j\theta_{n+1}} \qquad \text{(III-14)}$$

$$= \boldsymbol{B^0_{PQ_{n+1,l}}} \bullet \boldsymbol{Y}_{n+1} \bullet e^{j(\theta_n + \phi_{S_n})} \qquad \text{(III-15)}$$

$$= \boldsymbol{B^1_{PQ_{n+1,l}}} \bullet \boldsymbol{Y}_{n+1} \bullet e^{j\theta_n} \qquad \text{(III-16)}$$

Where $\boldsymbol{B^1_{PQ_{n+1,l}}} = \boldsymbol{B^0_{PQ_{n+1,l}}} \bullet e^{j\phi_{S_n}}$ (III-17)

Where $\phi_{S_n}$ is the phase difference of the $N^{th}$ sample of the vector $\boldsymbol{S^{c^4}_{BT_{n,l}}}$ and the initial phase $\theta_n$. Let $\boldsymbol{S^{c^p}_{BT_{n,l}}}$ denote the vector of GFSK modulated Bluetooth samples corresponding to one OFDM symbol period, where superscript $p$ denotes the number of Bluetooth signal in one OFDM symbol period. According to IEEE802.11g standard [7] one OFDM symbol period is $T_{OFDM} = 4\mu s$ and one Bluetooth symbol period is $T_B = 1\mu s$. Thus in one OFDM symbol period there are 4 Bluetooth symbols ($p = 4$).

$\boldsymbol{S^{c^4}_{BT_{n,l}}}$ has $N_1 = 4N$ data samples, $\boldsymbol{B^4_{BT_{n,l}}}$ is a $N_1 \times 8$ deterministic matrix, and $\boldsymbol{Y^4_n}$ is a $8 \times 1$ matrix of unknown random data and phase. If $S^{c^4}_{BT_{n,l}}$ is a vector of the samples of the 4 Bluetooth modulated symbol then,

$$S^{c4}_{BT_{n,l}} = \begin{bmatrix} S^c_{BT_{n,l}} \\ S^c_{BT_{n+1,l}} \\ . \\ . \\ S^c_{BT_{n+3,l}} \end{bmatrix} = \begin{bmatrix} B^0_{PQ_{n,l}} & & & \\ & B^1_{PQ_{n+1,l}} & & \\ . & & . & \\ . & & & \\ & & & B^3_{PQ_{n+3,l}} \end{bmatrix} \begin{bmatrix} Y_n \\ Y_{n+1} \\ . \\ . \\ Y_{n+3} \end{bmatrix} e^{j\theta_n} \quad \text{(III-18)}$$

$$S^{c^4}_{BT_{n,l}} = B^4_{BT_{n,l}} \bullet Y^4_n \bullet e^{j\theta_n} \quad \text{(III-19)}$$

Where vector $Y^4_n = \left[1 \;\; D_n \;\; 1 \;\; D_{n+1} \;\; 1 \;\; D_{n+2} \;\; 1 \;\; D_{n+3}\right]^T$.

Clearly $Y^4_n$ depends on data vector $x_j = \{D_n D_{n+1} D_{n+2} D_{n+3}\} \in \{-1-1-1-1, -1-1-11, ...., 1111\} = \{x_0, x_1, ...., x_{15}\}$

Where $j \in \{0,1,....,15\}$.

Equation (4-20) can be rewritten as:

$$S^{c^4}_{BT_{n,l,j}} = B^4_{BT_{n,l}} \bullet Y^4_{n,j} \bullet e^{j\theta_n} \quad \text{(III-21)}$$

Where subscript $j$ denotes the dependence on a particular data input vector $x_j$. The received Bluetooth data vector $S^{c^4}_{BT_{n,l,j}}$ in OFDM receiver is passed through a low pass filter of impulse response $h_{lpf}(t)$. Let $T^{OFDM}_s$ and $T_s$ be the sampling time of the OFDM and Bluetooth signal respectively. Then the filtered Bluetooth samples in OFDM system are:

$$i(n,l,j) = \sum_{m=0}^{N_l-1} h_{lpf}\left(nT^{OFDM}_s - mT_s\right) S^{c^4}_{BT_{n,l,j}}(m) \quad \text{(III-22)}$$

After removing cyclic prefix of length $L$ in OFDM receiver, which is done by pre-multiplying by cyclic prefix remove matrix $R_{cp} = \left[0_{N_{FFT}\times L} \; I_{N_{FFT}\times N_{FFT}}\right]$, where $0$ is a null matrix and $I$ is an identity matrix. The interference vector $i^{N_{FFT}}_{n,l,j}$ is given by:

$$i^{N_{FFT}}_{n,l,j} = R_{cp} \bullet h_{lpf} \bullet S^{c^4}_{BT_{n,l,j}} = h_l \bullet B^4_{BT_{n,l}} \bullet Y^4_{n,j} \bullet e^{j\theta_n}$$

Where, $i^{N_{FFT}}_{n,l,j} = \left[i_{n_0,l,j} \;\; i_{n_1,l,j} \;..\; i_{n_{N_{FFT}-1},l,j}\right]^T$, $h_l = R_{cp} \bullet h_{lpf}$

Matrix $h_l$ is of dimension $N_{FFT}\times N_l$. According to IEEE802.11g standard for OFDM system $N_{FFT} = 64$ and cyclic prefix is ¼th of the symbol duration. Therefore, $L = 16$ is used in our analysis. After performing FFT on $i^{N_{FFT}}_{n,l,j}$, interference on the $k^{th}$ sub-carrier is given by:

$$I^{BT}_{k,l,j} = \sum_{n=0}^{N_{FFT}-1} i_{n,l,j} e^{-j\frac{2\pi kn}{N_{FFT}}} \quad \text{(III-23)}$$

In matrix form the above equation is written as:

$$\begin{aligned} I^{BT}_{l,j} &= F \bullet i^{N_{FFT}}_{n,l,j} = F \bullet h_l \bullet S^{c4}_{BT_{n,l,j}} \\ &= F \bullet h_l \bullet B^4_{BT_{n,l}} \bullet Y^4_{n,j} \bullet e^{j\theta_n} \end{aligned} \quad \text{(III-24)}$$

Where $I^{BT}_{l,j} = \left(I^{BT}_{0,l,j}, I^{BT}_{1,l,j}, ... I^{BT}_{N_{FFT}-1,l,j}\right)^T$

$F$ is the FFT transform matrix. Let $V_{k,l} = F_k \bullet h_l \bullet B^4_{BT_{n,l}}$, where $F_k$ corresponds to the $k^{th}$ row of FFT matrix $F$. From , interference on $k^{th}$ sub-carrier of an OFDM symbol is given by:

$$I^{BT}_{k,l,j} = V_{k,l} \bullet Y^4_{n,j} \bullet e^{j\theta_n} \quad \text{(III-25)}$$

The magnitude, of the interference $I^{BT}_{k,l,j}$ (4-20) for the $k^{th}$ sub-carrier is independent of the initial phase $\theta_n$. It depends only on the product of the matrix $V_{k,l}$ and matrix $Y^4_{n,j}$. For a particular hopping frequency $f_l$ magnitude or power of the Bluetooth interferer on $k^{th}$ WLAN sub-carrier will depend upon the data input vector $x_j$.

Section 4.2.1 is on finding the interference vector present in OFDM subcarriers. The derivation of equation(III-24) gives the interference vector present in all the subcarriers of OFDM signal. The derivation shows that interference vector present in a particular OFDM symbol duration is a function of binary data vector $\left[D_n \;\; D_{n+1} \;\; D_{n+2} \;\; D_{n+3}\right]^T$ transmitted by Bluetooth during that time interval. The vector $Y^4_n = \left[1 \;\; D_n \;\; 1 \;\; D_{n+1} \;\; 1 \;\; D_{n+2} \;\; 1 \;\; D_{n+3}\right]^T$ depends on $j^{th}$ combination binary data vector transmitted by Bluetooth. Matrix $V_{k,l}$ can be obtained from different parameters of Bluetooth like power, frequency, modulation index and parameters of WLAN receiver low pass filter, FFT matrix, and cyclic prefix removal matrix.

As mentioned earlier in chapter 4, the vector of Bluetooth interference is given by:

$$I^{BT}_{l,j} = F \bullet h_l \bullet B^4_{BT_{n,l}} \bullet Y^4_{n,j} \bullet e^{j\theta_n} \quad \text{(III-26)}$$

where $I^{BT}_{l,j} = \left(I^{BT}_{0,l,j}, I^{BT}_{1,l,j}, ... I^{BT}_{N_{FFT}-1,l,j}\right)^T$

and interference component corresponding to $k^{th}$ sub-carrier, $l^{th}$ hopping frequency and $j^{th}$ combination of binary data vector is given by:

$$I^{BT}_{k,l,j} = F_k \bullet h_l \bullet B^4_{BT_{n,l}} \bullet Y^4_{n,j} \bullet e^{j\theta_n} \quad \text{(III-27)}$$

where $F_k = \left[1 \;\; e^{-j\frac{2\pi k}{N_{FFT}}} \;\; e^{-j\frac{2\pi k.2}{N_{FFT}}} ... e^{-j\frac{2\pi k(N_{FFT}-1)}{N_{FFT}}}\right]$ is the subset of $F$ and the vector $Y^4_{n,j}$ is of dimension 8×1, consisting of binary data vector of the Bluetooth system and superscript "4" denotes the number of binary Bluetooth data transmitted within the symbol duration of OFDM signal.

$h_l = R_{cp} \bullet h_{lpf}$ where $R_{cp}$ is a cyclic prefix removal matrix and $\boldsymbol{h_{lpf}}$ is matrix generated from tap values of low pass filter. The matrix $\boldsymbol{h_1}$ is of dimension $N_{FFT}\times N_1$. The matrix $\boldsymbol{B^4_{BT_{n,l}}}$ is of dimension $N_1\times 8$. The matrix $\boldsymbol{B^4_{BT_{n,l}}}$ is generated depending upon modulation properties of the Bluetooth system as given in equation (III-13) of the chapter 4 which is given below:

Now $$\boldsymbol{B^4_{BT_{n,l}}} = \begin{bmatrix} \boldsymbol{B^0_{PQ_{n,l}}} & & & \\ & \boldsymbol{B^1_{PQ_{n+1,l}}} & & \\ . & . & .\boldsymbol{B^2_{PQ_{n+2,l}}} & \\ . & . & . & \boldsymbol{B^3_{PQ_{n+3,l}}} \end{bmatrix} \qquad \text{(III-28)}$$

Where $\boldsymbol{B^0_{PQ_{n,l}}}$, $\boldsymbol{B^1_{PQ_{n+1,l}}}$, $\boldsymbol{B^2_{PQ_{n+2,l}}}$, $\boldsymbol{B^3_{PQ_{n+3,l}}}$ are $N\times 2$ matrices each, whose 1st and 2nd column consist of $N$ number of samples of $P_l\ (t)$ and $Q_l\ (t)$ as given in equation (III-9) ) and (III-10) Let $\boldsymbol{B^4_{BT_{n,l}}} = [\boldsymbol{C_1\ C_2\ C_3\ C_4\ C_5\ C_6\ \ C_7\ C_8}]$,

Where $$\begin{bmatrix} \boldsymbol{C_1} & \boldsymbol{C_2} \end{bmatrix} = \begin{bmatrix} \boldsymbol{B^0_{PQ_{n,l}}} \\ \\ 0_{3N\times 2} \end{bmatrix} \qquad \text{(III-29)}$$

and each of the column vector $\boldsymbol{C_1}$, $\boldsymbol{C_2}$ …. $\boldsymbol{C_8}$ is of dimension of $4N\times 1$ and given by:

Where $$\boldsymbol{C_1} = \begin{bmatrix} P_l(nT_B) \\ P_l(nT_B+T_S) \\ . \\ . \\ P_l(nT_B+(N-1)T_S) \\ 0 \\ . \\ . \\ 0 \end{bmatrix} \qquad \text{( III-30)}$$

, $$\boldsymbol{C_2} = \begin{bmatrix} Q_l(nT_B) \\ Q_l(nT_B+T_S) \\ . \\ . \\ Q_l(nT_B+(N-1)T_S) \\ 0 \\ . \\ . \\ 0 \end{bmatrix} \qquad \text{(III-31)}$$

Similarly, $$\begin{bmatrix} \boldsymbol{C_3} & \boldsymbol{C_4} \end{bmatrix} = \begin{bmatrix} \boldsymbol{0}_{N\times 2} \\ \boldsymbol{B^1_{PQ_{n+1,l}}} \\ \boldsymbol{0}_{2N\times 2} \end{bmatrix} \qquad \text{(III-32)}$$

$$\begin{bmatrix} \boldsymbol{C_5} & \boldsymbol{C_6} \end{bmatrix} = \begin{bmatrix} \boldsymbol{0}_{2N\times 2} \\ \boldsymbol{B^2_{PQ_{n+2,l}}} \\ \boldsymbol{0}_{N\times 2} \end{bmatrix} \qquad \text{(III-33)}$$

and $$\begin{bmatrix} \boldsymbol{C_7} & \boldsymbol{C_8} \end{bmatrix} = \begin{bmatrix} \boldsymbol{0_{3N\times 2}} \\ \boldsymbol{B^3_{PQ_{n+3,l}}} \end{bmatrix} \qquad \text{(III-34)}$$

Now Equation(III-27) can be rewritten as:

$$I^{BT}_{k,l,j} = \boldsymbol{F_k}\bullet\boldsymbol{h_1}\bullet\left[\boldsymbol{C_1 C_2\ C_3 C_4\ C_5\ C_6\ C_7 C_8}\right]\bullet\boldsymbol{Y^4_{nj}}\bullet e^{j\theta_n} \qquad \text{(III-35)}$$

$$\boldsymbol{Y^4_n} = \begin{bmatrix} 1 & D_n & 1 & D_{n+1} & 1 & D_{n+2} & 1 & D_{n+3} \end{bmatrix}^T,$$

Therefore, the interference $I^{BT}_{k,l,j}$ can be expressed as:

$$\begin{aligned} I^{BT}_{k,l,j} &= e^{j\theta_n}\bullet\boldsymbol{F_k}\bullet\boldsymbol{h_1}\bullet\left[\boldsymbol{C_1}+\boldsymbol{C_3}+\boldsymbol{C_5}+\boldsymbol{C_7}\right] \\ &+ e^{j\theta_n}\bullet\boldsymbol{F_k}\bullet\boldsymbol{h_1}\bullet\boldsymbol{C_2}\bullet D_n + e^{j\theta_n}\bullet\boldsymbol{F_k}\bullet\boldsymbol{h_1}\bullet\boldsymbol{C_4}\bullet D_{n+1} \\ &+ e^{j\theta_n}\bullet\boldsymbol{F_k}\bullet\boldsymbol{h_1}\bullet\boldsymbol{C_6}\bullet D_{n+2} + e^{j\theta_n}\boldsymbol{F_k}\bullet\boldsymbol{h_1}\bullet\boldsymbol{C_8}\bullet D_{n+2} \end{aligned} \qquad \text{(III-36)}$$

$$I^{BT}_{k,l,j} = \sum_{n=0}^{3} a_n D_n + C_k \qquad \text{(III-37)}$$

Where $$a_0 = e^{j\theta_n}\bullet\boldsymbol{F_k}\bullet\boldsymbol{h_1}\bullet\boldsymbol{C_2} \qquad \text{(III-38)}$$

$$a_1 = e^{j\theta_n}\bullet\boldsymbol{F_k}\bullet\ \boldsymbol{h_1}\bullet\boldsymbol{C_4} \qquad \text{(III-39)}$$

$$a_2 = e^{j\theta_n}\bullet\boldsymbol{F_k}\bullet\boldsymbol{h_1}\bullet\ \boldsymbol{C_6} \qquad \text{(III-40)}$$

$$a_3 = e^{j\theta_n}\bullet\boldsymbol{F_k}\bullet\boldsymbol{h_1}\bullet\ \boldsymbol{C_8} \qquad \text{(III-41)}$$

And

$$C_k = e^{j\theta_n}\bullet\boldsymbol{F_k}\bullet\ \boldsymbol{h_1}\bullet\left[\boldsymbol{C_1}+\boldsymbol{C_3}+\boldsymbol{C_5}+\boldsymbol{C_7}\right] \qquad \text{(III-42)}$$

*C. Interference Model of Zigbee System*

Similarly, using equation(II-16) can be rewritten as[24]:

$$I^{Z}_{k,l,j} = \sum_{n=0}^{7} a_n D_n \qquad \text{(III-43)}$$

Where

$$a_0 = \boldsymbol{F_k}\bullet\boldsymbol{h_1}\bullet\boldsymbol{C_1};\ a_1 = \boldsymbol{F_k}\bullet\boldsymbol{h_1}\bullet\boldsymbol{C_2};\ a_2 = \boldsymbol{F_k}\bullet\boldsymbol{h_1}\bullet\boldsymbol{C_3};\ a_3 = \boldsymbol{F_k}\bullet\boldsymbol{h_1}\bullet\boldsymbol{C_4};$$
$$a_4 = \boldsymbol{F_k}\bullet\boldsymbol{h_1}\bullet\boldsymbol{C_5};\ a_5 = \boldsymbol{F_k}\bullet\boldsymbol{h_1}\bullet\boldsymbol{C_6};\ a_6 = \boldsymbol{F_k}\bullet\boldsymbol{h_1}\bullet\boldsymbol{C_7};\ a_7 = \boldsymbol{F_k}\bullet\boldsymbol{h_1}\bullet\boldsymbol{C_8}$$

*D. Estimation of WPAN signal's frequency*

Let $B$ be the total frequency band and $\Delta f$ the frequency spacing of each of the FFT bin (sub-carrier spacing). If the estimation frequency band ranges from $f_{start}$ to $f_{finish}$ and $k^{th}$ FFT bin or sub-carrier has maximum energy, then frequency of narrowband signal is given by:

$$f_h = round\left(f_{start} + \left(\Delta f\cdot k\right)\right) \qquad \text{(III-44)}$$

where, $f_{finish} - f_{start} = B$ MHz and $f_h$ is the estimated frequency of narrowband signal.

*E. Estimation of WPAN signal amplitude and phase*

Let the narrowband transmitted signal be:

$$S = B \bullet Y \bullet e^{j\theta_n} = B \bullet Y_1 \qquad \text{(III-45)}$$

$$\text{where } Y_1 = Y \bullet e^{j\theta_n}$$

The received narrowband data vector $S$ in OFDM receiver is passed through a low pass filter of impulse response $h_{lpf}(t)$. Let $T_s^{OFDM}$ and $T_s$ be the sampling time of the OFDM and narrowband signal respectively. Then the filtered narrowband samples in OFDM system is[11]:

$$i(n) = \sum_{m=0}^{N_1-1} \left( h_{lpf}\left(nT_s^{OFDM} - mT_s\right) S(m) \right) \qquad \text{(III-46)}$$

The filtered samples pass through cyclic prefix removal block. After passing through the cyclic prefix removal block interference is given by:

$$\mathbf{i}^{N_{FFT}} = R_{cp} \bullet h_{lpf} \bullet S \qquad \text{(III-47)}$$

$$= h_2 \bullet Y_1 \qquad \text{(III-48)}$$

Where,

$$i^{N_{FFT}} = \left[ i_0 \; i_1 \; .. \; i_{N_{FFT}-1} \right]^T, \quad h_2 = R_{cp} \bullet h_{lpf} \bullet B$$

The matrix $h_2$ is of dimension $N_{FFT} \times 2p$.

Let $u_{Train}$ be the set of training signal after IFFT in the transmitter of an OFDM system.

$$u_{Train} = F^H \bullet U_{Train} = \left[ u_0 \, u_1 \ldots u_{N_{FFT}-1} \right]^T \qquad \text{(III-49)}$$

If $u1_{Train}$ is the received vector at the receiver before FFT operation then it is given by:

$$u1_{Train} = R_{cp} \bullet H_0 \bullet T_{cp} \bullet u_{Train} \qquad \text{(III-50)}$$

$$= \tilde{H} \bullet u_{Train} \qquad \text{(III-51)}$$

where $\tilde{H}$ is $N_{FFT} \times N_{FFT}$ circulant matrix which is given below:

$$\tilde{H} = \begin{bmatrix} h_0 & 0 . & . & . & . & . & . & . & .0 & h_{L-1} & . & . & h_2 & h_1 \\ h_1 & h_0 & 0 . & . & . & . & . & . & . & .0 & h_{L-1} . & & h_3 & h_2 \\ . & . & . & . & . & & . & & . & . & . & . \\ h_{L-2} & h_{L-3} & .h_1 & h_0 & 0 . & . & . & . & . & . & 0 & h_{L-1} \\ h_{L-1} & . & . & . & .h_1 & h_0 & 0 . & . & . & . & .0 . & . & . & . & . & .0 \\ 0 & h_{L-1} & . & . & h_1 & h_0 & 0 . & . & . & . & . & . & . & .0 \\ . & & . & & . \\ . & & . & & . \\ 0 & . & . & & . & . & 0 & h_{L-1} . & . & . & . & .h_1 & h_0 \end{bmatrix} \qquad \text{(III-52)}$$

Now (III-52) is rewritten as:

$$u1_{Train} = \tilde{u}_{Train} \bullet h \qquad \text{(III-53)}$$

$$\text{Where } \tilde{u}_{Train} = \begin{bmatrix} u_0 & u_{N_{FFT}-1} \ldots u_{N_{FFT}-(L-1)} \\ u_1 & u_0 \ldots\ldots\ldots u_{N_{FFT}-(L-2)} \\ . & . \quad\quad . \\ . & . \quad\quad . \\ u_{N_{FFT}-1} & u_{N_{FFT}-2} \; . \; . \; u_{N_{FFT}-L} \end{bmatrix} \qquad \text{(III-54)}$$

and $h = (h_0 \, h_1 \ldots\ldots h_{L-1})^T$, $\tilde{u}_{Train}$ is $N_{FFT} \times L$ consisting training of signals. Now in presence of AWGN noise and Bluetooth interference the received signal before FFT operation is written as:

$$\hat{u}1_{Train} = u1_{Train} + i^{N_{FFT}} + n1 \qquad \text{(III-55)}$$

where **n1** is AWGN noise

$$\hat{u}1_{Train} = \left(\tilde{u}_{Train} \bullet h\right) + \left(h_2 \bullet Y_1\right) + n1 \qquad \text{(III-56)}$$

$$= \left[ \tilde{u}_{Train} \quad h_2 \right] \begin{bmatrix} h \\ Y_1 \end{bmatrix} + n1 \qquad \text{(III-57)}$$

$$= Ay + n1 \qquad \text{(III-58)}$$

where $A = \left[ \tilde{\mathbf{u}}_{\mathbf{Train}} \quad h_2 \right]$ is of dimension $N_{FFT} \times (L + 2p)$.

and $y = \begin{bmatrix} h \\ Y_1 \end{bmatrix}$ is matrix of dimension $(L + 2p) \times 1$ consisting of unknown channel vector and unknown Bluetooth signal vector. Unknown parameters of matrix $y$ can be estimated accurately by (III-59) if noise is negligible:

$$y = \mathbf{A}^{\dagger} \bullet \left( \hat{\mathbf{u}}\mathbf{1}_{\mathbf{Train}} - \mathbf{n1} \right) \qquad \text{(III-59)}$$

Where $\mathbf{A}^{\dagger}$ is the pseudo inverse of **A.** Let $\widehat{Y}_1$ be the vector of estimated vector of Bluetooth signal from equation (III-59), then estimated Bluetooth samples corresponding to the p symbol will be:

$$\widehat{S} = B \bullet \widehat{Y}_1 \qquad \text{(III-60)}$$

The square of each of the samples of matrix $\widehat{S}$ gives the power and averaging it over all the samples give the average power. The square root of the average power is taken the signal magnitude. The phase of the first sample of matrix $\widehat{S}$ gives the value of initial phase $\theta_n$ .

## IV. Simulation Results

In this section simulation results comparing performance of WLAN, WPAN with different interference avoidance or mitigation techniques is given. It is assumed that the system whose performance is to be found has scanned the channel before transmitting data and has detected the particular WPAN interference[18] by estimating the bandwidth. For WLAN system, performance comparison with respect to different position of the interference in the WLAN spectrum is also simulated.

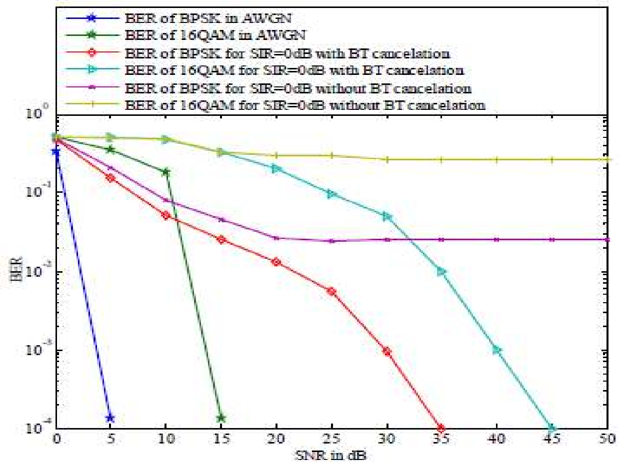



Fig 4: BER of uncoded16QAM and BPSK-OFDM in presence of Bluetooth in AWGN for SIR = 0dB.

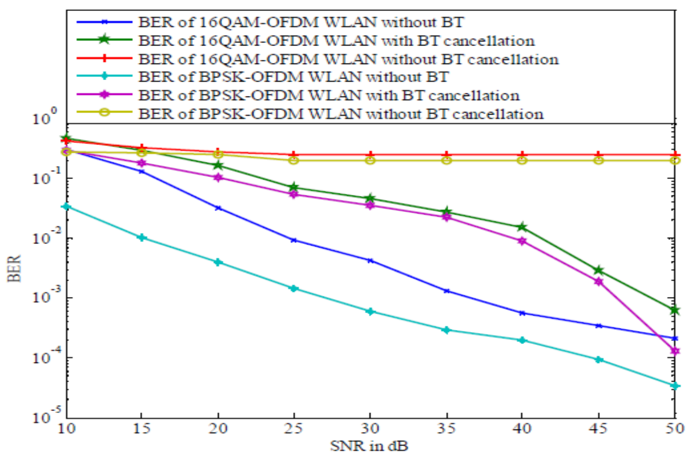
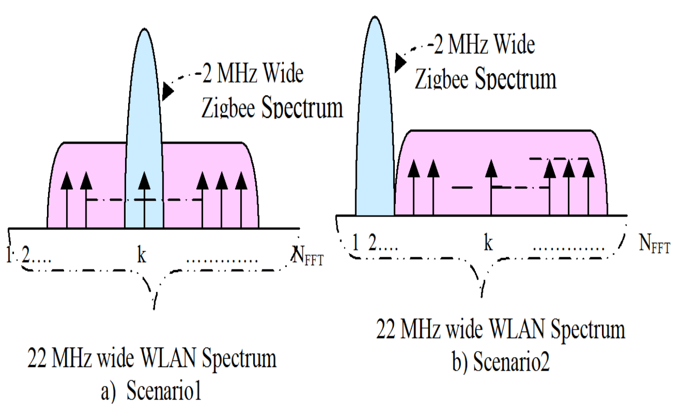



Fig5: Coded BER for 16QAM, BPSK OFDM based WLAN with Bluetooth cancellation in flat fading channel



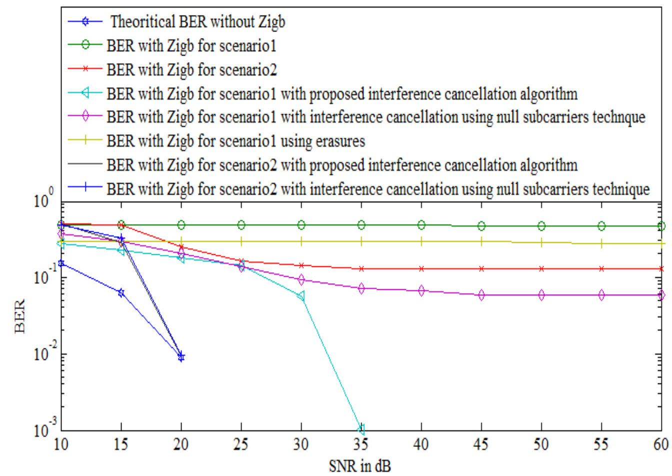




Fig6: Comparison of coded BER for OFDM based WLAN with Zigbee interference

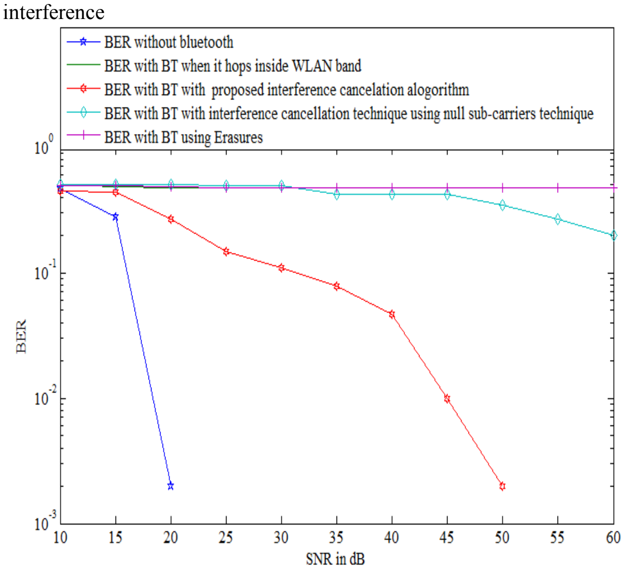



Fig 7: BER using different interference mitigation techniques of WLAN in presence of Bluetooth interference for BER fh = 39 and SIR = 0db in AWGN

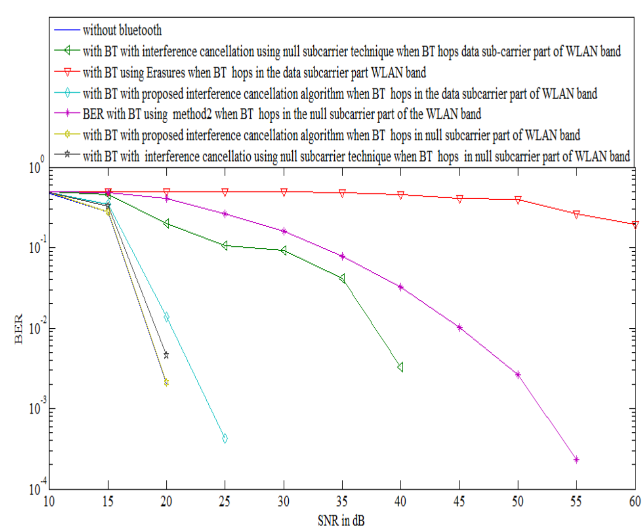



Figure 8: BER WLAN with Bluetooth interference with different technique and different Bluetooth hop frequencies

Table 1: BER of WLAN in AWGN with ADFH and proposed mitigation technique

| SIR in dB | BER with Bluetooth interference When Bluetooth hops randomly in any of the 79 frequency | BER with Bluetooth interference using proposed Cancellation Algorithm | BER with Bluetooth interference when ADFH is applied in Bluetooth systems | BER with Bluetooth interference when ADFH is applied in Bluetooth and cancellation algorithm is used in WLAN |
|---|---|---|---|---|
| 20 | 0. 0095 | 0. 0080 | 0 | 0 |
| 15 | 0. 0283 | 0. 0069 | 0 | 0 |
| 10 | 0. 0528 | 0. 0070 | 0. 0002 | 0 |
| 5 | 0. 0974 | 0. 0074 | 0. 0019 | 0 |
| 0 | 0. 1294 | 0. 0073 | 0. 0386 | 0. 0013 |
| -5 | 0. 1563 | 0. 0050 | 0. 1564 | 0. 0050 |
| -10 | 0. 1823 | 0. 0008 | 0. 1779 | 0. 0008 |
| -15 | 0. 2059 | 0. 0007 | 0. 2051 | 0. 0007 |
| -20 | 0. 2448 | 0. 0005 | 0. 2364 | 0. 0005 |

## V. Summary

From the simulation results it seen that performance of the WLAN or WPAN system depends on the interference power. The results also show that if the victim system uses some mitigation technique the performance improves significantly. It is seen that when WLAN uses the proposed narrowband cancellation algorithm, the BER of WLAN decreases or almost become zero for high SNR values.

The column 4 of Table 1 gives the results for the case when ADFH is used in BT and no interference mitigation technique is applied in WLAN. It shows the BER of WLAN is zero when SIR value is more than 15dB and below that BER increases. The reason behind this is the implementation of ADFH in Bluetooth system. Bluetooth decides a frequency to be bad when estimated SIR for Bluetooth system is less than some threshold value. So when WLAN power is high, compared to Bluetooth power, Bluetooth system easily detects the frequencies occupied by WLAN and hop to the frequencies other than those of WLAN. As WLAN power decreases, Bluetooth does not consider some or all the frequencies occupied by WLAN as bad. However at the same time WLAN experiences high interference power from the Bluetooth, resulting in an increase in BER for WLAN.

The column 5 of Table 1 gives the results for the case when ADFH is used in BT and proposed interference cancelation algorithm of chapter5 is applied in WLAN. The result shows a significant improvement in BER is seen particularly for the SIR values in 0 to 20 dB range. The reason is that when Bluetooth avoids all the bands occupied by WLAN interference, WLAN experiences very little or no threat from Bluetooth. In case when Bluetooth interferes with some of the WLAN frequencies (which happens for SIR of WLAN below 0 dB), the interference mitigation algorithm helps WLAN to nullify the Bluetooth interference. Thus the combined application of the proposed interference cancellation algorithm in WLAN and ADFH in Bluetooth system improves the overall performance of WLAN.

Table1 gives the results for the case when ADFH is used in BT and proposed interference cancelation algorithm of chapter5 is applied in WLAN. The result shows a significant improvement in BER is seen particularly for the SIR values in 0 to 20 dB range. The reason is that when Bluetooth avoids all the bands occupied by WLAN interference, WLAN experiences very little or no threat from Bluetooth. In case when Bluetooth interferes with some of the WLAN frequencies (which happens for SIR of WLAN below 0 dB), the interference mitigation algorithm helps WLAN to nullify the Bluetooth interference. Thus the combined application of the proposed interference cancellation algorithm in WLAN and ADFH in Bluetooth system improves the overall performance of WLAN.

A conclusion section is not required. Although a conclusion may review the main points of the paper, do not replicate the abstract as the conclusion. A conclusion might elaborate on the importance of the work or suggest applications and extensions.

## Appendix

In this section we present the details of the notations used in the paper.

$\boldsymbol{F}$ is a FFT matrix of dimension $N_{FFT} \times N_{FFT}$

$\boldsymbol{F^H}$ is a IFFT matrix of dimension $N_{FFT} \times N_{FFT}$

$\boldsymbol{F_k}$ is the subset of FFT matrix corresponding to the $k^{th}$ row of the FFT matrix is of dimension $1 \times N_{FFT}$

$I_k$ is a interference present in $k^{th}$ OFDM sub-carriers.

$\boldsymbol{x_j} = \left\{D_0 D_1 D_2 ....D_{p-1}\right\}$ is a binary data vector transmitted by WPAN system, $p$ is the number of binary data transmitted within symbol duration of the OFDM system and subscript $j$ denotes the $j^{th}$ combination of binary data vector where $j \in \left\{0,1,..,N-1\right\}$, $N = 2^p$. $D_n \in \left\{+1,-1\right\}$

$\boldsymbol{I_{l,j}^{BT}}$ is Bluetooth interference column vector having $N_{FFT}$ elements and each element corresponding to the interference of the $k^{th}$ OFDM sub-carriers. Superscript "BT" represents Bluetooth signal and subscript $l, j$ denotes the dependency on $l^{th}$ hopping frequency and $j^{th}$ combination of binary data vector of Bluetooth.

$\boldsymbol{I_{l,j}^{Z}}$ is Zigbee interference column vector having $N_{FFT}$ elements and each element corresponding to the interference for the $k^{th}$ OFDM sub-carriers. Superscript "*z*" represents Zigbee and subscript $l, j$ denotes the dependency on $l^{th}$ channel frequency and $j^{th}$ combination of binary data vector of Zigbee.

$\boldsymbol{S}$ is the column vector of WPAN transmitted samples having $N_l$ elements

$N_l$ is the number of WPAN samples which is $N \cdot p$

$N$ is the number of samples for one symbol period of WPAN

$p$ is the number of WPAN binary data transmitted within the symbol duration of WLAN system.

$\boldsymbol{Y_n^p}$ is the column vector consisting of binary data signal of WPAN system where superscript $p$ denotes the number of binary data transmitted by WPAN within the symbol duration of OFDM signal.

$\boldsymbol{U_{Train}}$ is a column vector representing WLAN training signal before IFFT operation with a length of $N_{FFT}$. $\boldsymbol{u_{Train}}$ is a column vector representing WLAN training signal after IFFT operation of length $N_{FFT}$.

$\boldsymbol{T_{cp}}$ is the cyclic prefix matrix of dimension $(N_{cp} + N_{FFT}) \times N_{FFT}$, where $N_{cp}$ is the length of cyclic prefix added to the OFDM symbol. $\boldsymbol{R_{cp}}$ is cyclic prefix removal matrix of dimension $N_{FFT} \times (N_{cp} + N_{FFT})$, $Ncp$ is the length of cyclic prefix.

$\boldsymbol{0_{m\times r}}$ denotes null matrix with $m$ rows and $r$ columns.

$\boldsymbol{H_0}$ is the equivalent channel matrix of dimension $(N_{FFT} + L) \times (N_{FFT} + L)$ where $L$ is the number of channel tap value


## ACKNOWLEDGMENT

The author would like to thank Prof H.S Jamadagni for his valuable suggestions.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