Spectrum Sensing for Full-Duplex Cognitive Radio Systems

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Abstract. Full-Duplex (FD) transceiver has been proposed to be used in Cognitive Radio (CR) in order to enhance the Secondary User (SU) Data-Rate. In FD CR systems, in order to diagnose the Primary User activity, SU can perform the Spectrum Sensing while operating. Making an accurate decision about the PU state is related to the minimization of the Residual Self Interference (RSI). RSI represents the error of the Self Interference Cancellation (SIC) and the receiver impairments mitigation such as the Non-Linear Distortion (NLD) of the receiver Low-Noise Amplifier (LNA). In this manuscript, we deal with the RSI problem by deriving, at the first stage, the relation between the ROC curves under FD and Half-Duplex (HD) (when SU stops the transmission while sensing the channel). Such relation shows the RSI suppression to be achieved in FD in order to establish an efficient Spectrum Sensing relatively to HD. In the second stage, we deal with the receiver impairments by proposing a new technique to mitigate the NLD of LNA. Our results show the efficiency of this method that can help the Spectrum Sensing to achieve a closed performance under FD to that under HD.

Keywords: Full-Duplex · Self-Interference Cancellation · Non-Linearity Distortion · Spectrum Sensing · Cognitive Radio

1 Introduction

Recently, the Full Duplex (FD) transmission has been introduced in the context of Cognitive Radio (CR) to enhance the Data-Rate of the Secondary (unlicensed) User (SU). In FD systems, SU can transmit and sense the channel at the same time. Classically, Half Duplex (HD) was used, therefore the SU should stop transmitting to sense the status of the Primary (licensed) User (PU). Recent

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advancements in the Self-Interference Cancellation (SIC) make the application of FD in CR possible. Due to many imperfections, a perfect elimination of the self-interference cannot be reached in real world applications [1,2]. In CR, the SU makes a decision on the PU status using a Test Statistic (TS) [3]. This TS depends on the PU signal and the noise. Any residual interference from the SU signal can affect the TS norm and leads to a wrong decision about the presence of PU.

In wireless systems, the FD is considered as achieved if the Residual Self Interference (RSI) power becomes lower than the noise level. For that an important SIC gain is required (around 110 dB for a typical WiFi system [2]). This gain can be achieved using the passive suppression and the active cancellation. The passive suppression is related to many factors that reduce the Self Interference (SI) such as the transmission direction, the absorption of the metals and the distance between the transmitting antenna, T_x, and the receive antenna, R_x. The active cancellation reduce the Self-Interference (SI) by using a copy of the known transmitted signal. The estimation of channel coefficients becomes an essential factor in the active cancellation process. Any error in the channel estimation leads to decreasing the SIC gain.

Experimental results show that hardware imperfections such as the non-linearity of amplifiers and the oscillator noise are the main limiting performance factors [2,4–6]. Therefore, the SIC should also consider the receiver imperfections. The authors of [2] modify their method previously proposed in [7] to estimate the channel and the Non-Linearity Distortion (LND) of the receiver Low -Noise Amplifier (LNA). Their method requires two training symbol periods. During the first period, the channel coefficients are estimated in the presence of the NLD. The non-linearity of the amplifier is estimated in the second period using the already estimated channel coefficients. It is worth mentioned that the estimation of the NLD parameters in the second phase depends on the one of the channel coefficients done in the first phase. However the estimation of the channel coefficients in the first phase can be depending on unknown NLD parameters. To solve the previous dilemma, we propose hereinafter an estimation method of the NLD in such way that the estimation of the channel cannot be affected by the NLD.

[8–13] deal with the application of FD in CR. In [8–10,12], the RSI is modeled as a linear combination of the SU signal without considering hardware imperfections. In [10,13] the Energy Detection (ED) is studied in FD mode and the probability of detection, (P_d), and false alarm, P_{fa}, are found analytically. According to our best knowledge, there was no analytic relationship between the RSI, P_d and P_{fa} for both HD and FD mode.

This paper deals with the Spectrum Sensing in real world applications. At first we analytically address the impact of the RSI power on the detection process. For that objective, we derive a relation between the RSI power, the probabilities of detection and false alarm under HD and FD modes. Secondly, we analyze the NLD impact on the channel estimation and the Spectrum Sensing Performance. Hereinafter, a novel method is proposed to suppress the NLD of LNA without affecting the channel estimation process. Further, our proposed method outperforms significantly the method proposed in [2]. In addition, using our method, the receiver requires only one training symbol period to perform the estimation of the channel and the NLD estimation.

2 System Model

In the spectrum sensing context, we usually assume two hypothesis: H_0 (PU signal is absent) and H_1 (otherwise). In our works, we assume that PU signal and SU signals are wideband signals such as OFDM. Throughout this paper, uppercase letters represent frequency-domain signals and lower-case letters represent signals in time-domain. By focusing only on the additive receiver distortion which is dominated by the NLD of the LNA [2], the received signal can be modeled as follows:

$$Y_a(m) = HS(m) + W(m) + D(m) + \eta X(m) \tag{1}$$

H is the channel between the SU transmitter antenna T_x and the SU receive antenna R_x, $S(m)$ is the SU signal, $W(m)$ is an Additive White Gaussian Noise (AWGN), $D(m)$ represents the NLD of the LNA, $X(m)$ is image of the PU signal on R_x and $\eta \in \{0, 1\}$ is the PU indicator ($\eta = 1$ under H_1 and $\eta = 0$ under H_0).

After the SIC and the circuit imperfections mitigation, the obtained signal, $\hat{Y}(m)$, can be presented as follows;

$$\hat{Y}(m) = \xi(m) + W(m) + \eta X(m) \tag{2}$$

Where $\xi(m)$ is the RSI and is defined as $\xi(m) = (H - \hat{H})S(m) + D(m) - \hat{D}(m)$. $\hat{H}$ and $\hat{D}(m)$ are the estimated channel and the NLD respectively.

Ideally $\hat{H} = H$ and $\hat{D}(m) = D(m)$, therefore Eq. (2) becomes: $\hat{Y}(m) = W(m) + \eta X(m)$, which corresponds to an HD mode. Any mistake in the cancellation process may lead to a wrong decision about the PU presence.

3 The RSI Effect on the Spectrum Sensing

In order to decide the existence of the PU, the Energy Detector (ED) compares the received signal energy, T, to a predefined threshold, λ.

$$T = \frac{1}{N}\sum_{m=1}^{N} |\hat{Y}(m)|^2 \underset{H_0}{\overset{H_1}{\gtrless}} \lambda \tag{3}$$

By assuming the $i.i.d$ property of $\epsilon(n)$, $w(n)$ and $x(n)$, then $\xi(m)$, $W(m)$ and $X(m)$ become $i.i.d.$ (See Appendix (A.1)). In this case, the distribution of T should asymptotically follow a normal distribution for a large number of samples, N, according to the central limit theorem. Consequently, the probabilities of False Alarm, P_{fa}^F, and the Detection, P_d^F, under the FD mode can be obtained as follows (See Appendix (A.2)):

$$P_{fa}^F = Q(\frac{\lambda - \mu_0}{\sqrt{V_0}}) = Q\left(\frac{\lambda - (\sigma_w^2 + \sigma_d^2)}{\frac{1}{\sqrt{N}}(\sigma_w^2 + \sigma_d^2)}\right) \tag{4}$$

$$p_d^F = Q(\frac{\lambda - \mu_1}{\sqrt{V_1}}) = Q\left(\frac{\lambda - (\sigma_w^2 + \sigma_d^2 + \sigma_x^2)}{\frac{1}{\sqrt{N}}(\sigma_w^2 + \sigma_d^2 + \sigma_x^2)}\right) \tag{5}$$

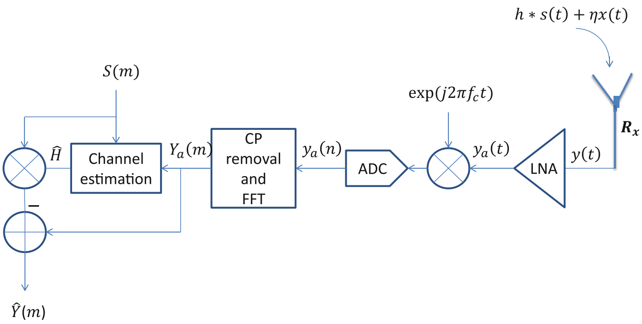

Fig. 1. Classical SIC circuit for OFDM receiver

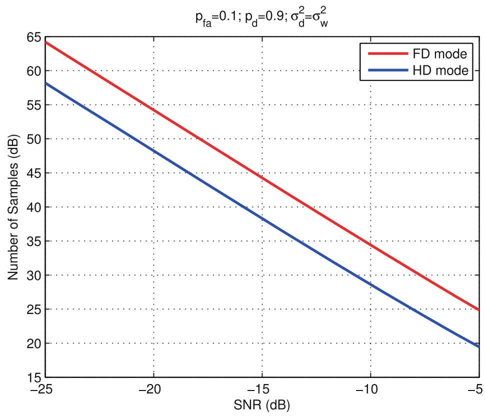

Fig. 2. The number of samples required to reach $P_d = 0.9$ and $P_{fa} = 0.1$ (Color figure online)

Where μ_i and V_i are the mean and the variance of T under H_i respectively, $i \in \{0; 1\}$, $\sigma_d^2 = E[|\xi(m)|^2]$ represents the RSI power, $\sigma_w^2 = E[|W(m)|^2]$ and $\sigma_x^2 = E[|X(m)|^2]$. The SNR, γ_x, is defined as: $\gamma_x = \frac{\sigma_x^2}{\sigma_w^2}$. If the SIC is perfectly achieved, *i.e.* $\sigma_d^2 = 0$, P_{fa}^F and P_d^F take their expressions under the HD mode.

Figure (2) shows the required number of samples to reach $P_d = 0.9$ and $P_{fa} = 0.1$ under the HD and FD modes for different values of SNR. In FD mode, we set $\sigma_d^2 = \sigma_w^2$ as the target values of σ_d^2 in digital communication. Figure (2) shows that the number of required samples slightly increases under the FD modes. For example if $\gamma_x = -5$ dB, then 85 samples are enough to reach the target $(P_d; P_{fa})$ under the HD mode while under FD mode, around 300 samples are needed.

Let us define the Probability of Detection Ratio (PDR), δ, for the same probability of false alarm under FD and HD modes, as follows:

$$\delta = \frac{P_d^F}{P_d^H} \; with \; P_{fa}^F = P_{fa}^H = \alpha \tag{6}$$

Where P_d^H and P_{fa}^H are the probabilities of detection and false alarm under HD respectively, $0 \leq \alpha \leq 1$ and $0 \leq \delta \leq 1$. As with an excellent SIC, the ROC can mostly reach in FD the same performance of HD. In order to show the effect of RSI on δ, let us define the RSI to noise ratio γ_d as follows:

$$\gamma_d = \frac{\sigma_d^2}{\sigma_w^2} \tag{7}$$

Using (4) and (6), the threshold, λ, can be expressed as follows:

$$\lambda = \left(\frac{1}{\sqrt{N}} Q^{-1}(\alpha) + 1\right)(\sigma_w^2 + \sigma_d^2) \tag{8}$$

By replacing (8) in (5), γ_d can be expressed as follows:

$$\gamma_d = \frac{(1+\gamma_x)Q^{-1}(\delta P_d^H) - Q^{-1}(\alpha) + \sqrt{N}\gamma_x}{Q^{-1}(\alpha) - Q^{-1}(\delta P_d^H)} \tag{9}$$

If $\delta = 1$, then we can prove that γ_d becomes zero, which means that the SIC is perfectly achieved. Figure (3) shows the curves of γ_d for various values of PDR, δ, with respect to the SNR, γ_x, for $P_d^H = 0.9$ and $\alpha = 0.1$. This figure shows that as δ increases γ_d decreases. To enhance the PDR, the selected SIC technique should mitigate at most the SI. In fact, for $\gamma_x = -5$ and a permitted loss of 1 % (*i.e.* $\delta = 0.99$), γ_d is about −15 dB.

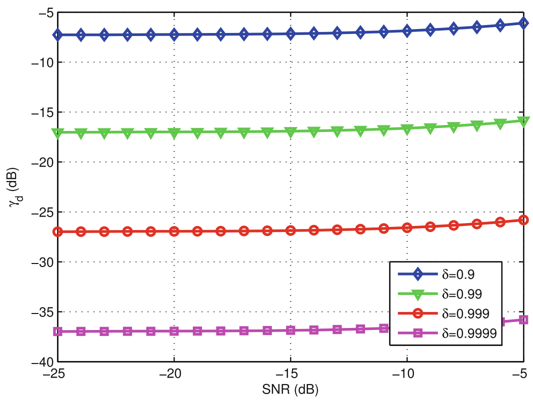

Fig. 3. Evolution of γ_d with respect to γ_x for various values of δ, $(P_{fa}^H\ ;\ P_d^H) = (0.1\ ;\ 0.9)$

4 The Effect of the Amplifier Distortion on Spectrum Sensing

In real world applications, the full duplex transceiver seems hard to be achieved due to hardware imperfections: the non-linearity of the amplifiers, the quantization noise of the Analog to Digital Converter (ADC), the phase noise of the oscillator, *etc*. The NLD of LNA is an important performance limiting factor [2,4–7]. According to NI 5791 datasheet [14], the NLD power is of 45 dB below the power of the linear amplified component. A new efficient algorithm is proposed in this section, it shows more reliable performance than that proposed in [2] and make the channel estimation performed without the influence of NLD.

4.1 Estimation of the Non-linearity Distortion of LNA

The LNA output can be written as an odd degrees polynomial of the input signal [15]. The NLD stands for the degrees greater than one. By limiting to the third degree and neglecting the higher degrees power [16], the NLD component can be written as follows:

$$d(t) = \beta y^3(t) \tag{10}$$

Where β is the NLD coefficient. The estimation of β can be helpful to suppress the LNA output. In this case, the channel estimation is no longer affected by the NLD. The overall output signal of the LNA, $y_a(t)$, can be expressed as follows:

$$y_a(t) = \theta y(t) + \beta y^3(t) \tag{11}$$

θ and $y(t)$ are the power gain and the input signal of the LNA respectively. To estimate θ and β, by a and b respectively, one can minimize the following cost function:

$$\mathcal{J} = E\left[\left(y_a(t) - (ay(t) + by^3(t))\right)^2\right] \tag{12}$$

By deriving $\mathcal{J}$ with respect to a and b we obtain:

$$\frac{\partial \mathcal{J}}{\partial a} = 0 \Rightarrow aE[y^2(t)] + bE[y^4(t)] = E[y_a(t)y(t)] \tag{13}$$

$$\frac{\partial \mathcal{J}}{\partial b} = 0 \Rightarrow aE[y^4(t)] + bE[y^6(t)] = E[y_a(t)y^3(t)] \tag{14}$$

Using Eq. (13) and (14), a linear system of equations can be obtained:

$$\begin{bmatrix} a \\ b \end{bmatrix} = A^{-1}B \tag{15}$$

Where:

$$A = \begin{bmatrix} E[y^2(t)] & E[y^4(t)] \\ E[y^4(t)] & E[y^6(t)] \end{bmatrix} ; \; B = \begin{bmatrix} E[y_a(t)y(t)] \\ E[y_a(t)y^3(t)] \end{bmatrix} \tag{16}$$

Once the non-linearity coefficient, β, is estimated, the non-linearity component can be subtracted from the output signal of the amplifier.

4.2 Numerical Results

Figure (4) shows the residual power of the NLD cancellation. The NLD power is fixed to 45 dB under the linear component [14]. This power is reduced to less than −300 dB after the application of our method. The method of [2] reduces the NLD

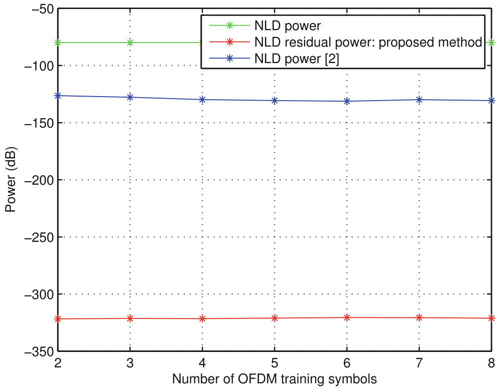

Fig. 4. The effect of the number of training symbols on the NLD residual power (Color figure online)

power by around 50 dB. β is estimated using various number of training symbols, N_e. In this simulation, OFDM modulations are used with 64 sub-carriers and a CP length equal to 16. The received power is fixed to -5 dBm and the noise power to -72 dBm [14]. As shown in Fig. (4), the residual power of NLD decreases with an increasing of N_e when the method of [2] is applied. However our method keeps a constant value of this power. Our technique outperforms significantly the method proposed in [2]. To show the impact of NLD on the channel estimation and the RSI power, Fig. (5) shows the power of $\hat{Y}(m)$ obtained in FD under H_0. The channel is estimated according to the method previously proposed by [17] as follows:

$$\hat{h} = IDFT\left\{\frac{1}{N_e}\sum_{k=0}^{N_e}\frac{Y_a^k(m)}{Y^k(m)}\right\} \; and \; \hat{H} = DFT\{\hat{h}(1, .., n_{tap})\} \tag{17}$$

Where $IDFT$ stands for the inverse discrete Fourier transform and n_{tap} is the channel order. The number of training symbols, N_e, is fixed to 4 symbols. To deal with a practical scenario, the number of sub-carrier is 64, the transmitted signal is of -10 dB (*i.e.* 20 dBm), and the noise floor is -102 dB (*i.e.* -72 dBm) [14]. The transceiver antenna is assumed to be omni-directional with 35 cm separation between T_x and R_x, so that a passive suppression of 25 dB is achieved [1]. According to the experimental results of [1], in a low reflection environment, 2 channel taps are enough to perform the SIC when the passive suppression is bellow 45 dB. Furthermore, the line of sight channel is modeled as Rician channel with K-factor about 20 dB. The non-line of sight component is modeled by a Rayleigh fading channel.

Figure (5) shows that our method leads to mitigate almost all the self interference, so that the power of $\hat{Y}(m)$ becomes very closed to the noise power. However, with the method of [2], the RSI power increases with the NLD power

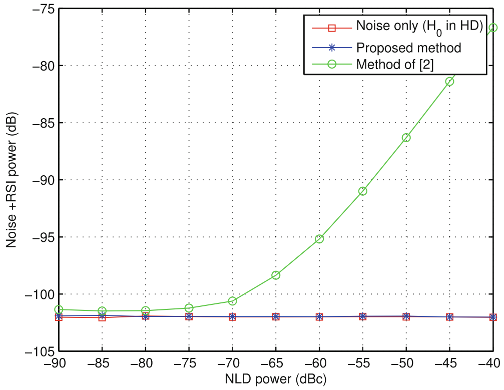

Fig. 5. The power of $\hat{Y}(m)$ under H_0 obtained after applying: (1) our proposed method, (2) the method of [2] is applied and (3) under HD mode

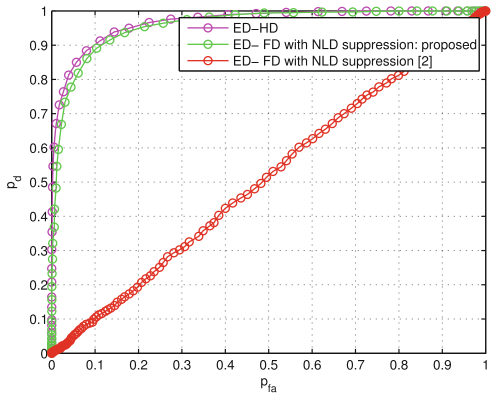

Fig. 6. The ROC curve after applying the proposed technique of NLD suppression (Color figure online)

because the NLD power is a limiting factor of the channel estimation which leads to a bad estimation of the channel.

To show the impact of the NLD on the Spectrum Sensing, Fig. (6) shows the ROC in various situations under $\gamma_x = -10$ dB. The simulations parameters in this figure are similar to those of Fig. (5), only the NLD power is 45 dB under the linear component according NI 5791 indications [14]. The method of [2] leads to a linear ROC, which means that no meaningful information about the PU status can be obtained. By referring to Fig. (5), the RSI power is of -82 dB for a NLD power of -45 dBc, which means that γ_d in this case is about 20 dB. This high RSI power leads to a harmful loss of performance (see Fig. (3)). From the other hand, our method makes the ROC in FD mode almost colinear with that of the ROC of HD mode, which means that all SI and receiver impairments is mitigated.

Figure (7) shows the PDR for a target $\alpha = 0.1$ and $P_d^H = 0.9$. The ratio δ increases with the SNR. At a low SNR of -10 dB, δ becomes closed to 1, so that a negligible performance loss is happen. As the SNR decreases the detection process in FD mode becomes more sensitive to the RSI power.

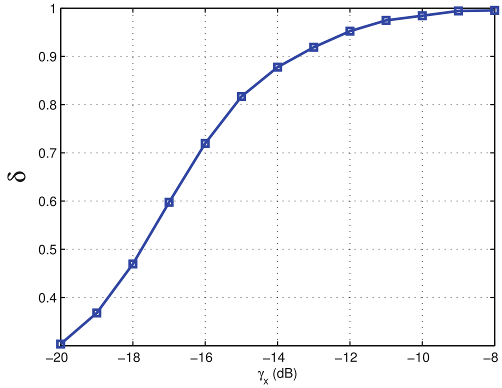

Fig. 7. The gain ratio: $\delta = \frac{P_d^F}{P_d^H}$ for different values of SNR

5 Conclusion

In this paper, we address the impact of the Residual Self Interference on the Spectrum Sensing for Full-Duplex Cognitive Radio. An analytic relation is derived relating the residual self interference with the probabilities of detection and false alarm under Full-Duplex and Half-Duplex modes. Furthermore, a new method is proposed to mitigate an important receiver impairment, which is the Non-Linear Distortion of the Low Noise Amplifier. This method shows its efficiency, leading the Spectrum Sensing performance in Full-Duplex mode to be closed to that under Half-Duplex mode.

A Appendix

A.1 *i.i.d.* property in Frequency Domain

Let $r(n)$ be an $i.i.d.$ time-domain signal. The DFT, $R(m)$, of the $r(n)$ is defined as follows:

$$R(m) = \sum_{n=1}^{L} r(n) e^{-j2\pi m \frac{n}{L}} \tag{18}$$

Where L is the number of samples of $r(n)$. According to the Central Limit Theorem (CLT), $R(m)$ follows asymptotically a Gaussian distribution for a large L. Based on [18], two normal variables are independent iff they are uncorrelated.

Let $C(m_1, m_2)$ the correlation of $R(m_1)$ and $R(m_2)$ $\forall$ $m_1 \neq m_2$.

$$\begin{aligned}
C(m_1, m_2) &= E[R(m_1)R^*(m_2) \\
&= E\left[\sum_{n_1,n_2=1}^{L} r(n_1)r^*(n_2)e^{-j2\pi\frac{n_1m_1-m_2n_2}{L}}\right] \\
&= \sum_{n_1=n_2=1}^{L} E\left[|r(n_1)|^2\right]e^{-j2\pi\frac{(m_1-m_2)n_1}{L}} \\
&\quad + \sum_{n_1\neq n_2=1}^{L} \underbrace{E\left[r(n_1)r^*(n_2)\right]}_{=0,\ since\ r(n)\ is\ i.i.d.} e^{-j2\pi\frac{(n_1m_1-n_2m_2)}{L}} \\
&= E\left[|r(n_1)|^2\right] \underbrace{\sum_{n_1=1}^{L} e^{-j2\pi(m_1-m_2)\frac{n_1}{L}}}_{=0} = 0
\end{aligned} \tag{19}$$

$C(m_1, m_2) = 0$ $\forall$ $m_1 \neq m_2$, therefore $R(m_1)$ and $R(m_2)$ are uncorrelated and they become independent because of their Gaussianity distribution.

A.2 Probility of Detection and Probability of False Alaram

As by our assumption $\xi(m)$, $W(m)$ and $X(m)$, are asymptotically Gaussian *i.i.d.*, then $\hat{Y}(m)$ is also Gaussian and *i.i.d.*. Therefore the TS, T, of Eq. (3) follows a normal distribution according to CLT for a large N. Under H_0 (*i.e.* $X(m)$ does not exist), the mean, μ_0, and the variance, V_0 of T can be obtained as follows:

$$\mu_0 = E[T] = E\left[\frac{1}{N}\sum_{m=1}^{N} |\xi(m) + W(m)|^2\right] = \sigma_w^2 + \sigma_d^2 \tag{20}$$

$$\begin{aligned}
V_0 = E[T^2] - E^2[T] &= \frac{1}{N^2}E\left[\left(\sum_{m=1}^{N} |\hat{Y}(m)|^2\right)^2\right] - (\sigma_w^2 + \sigma_d^2)^2 \\
&= \frac{1}{N^2}E\left[\sum_{m_1=m_2=1}^{N} |\hat{Y}(m_1)|^4\right] \\
&\quad + \frac{1}{N^2}E\left[\sum_{m_1\neq m_2=1}^{N} |\hat{Y}(m_1)|^2\hat{Y}(m_2)|^2\right] - (\sigma_w^2 + \sigma_d^2)^2 \\
&= \frac{1}{N^2}\sum_{m_1=m_2=1}^{N} E\left[|\hat{Y}(m_1)|^4\right] - \frac{1}{N}(\sigma_w^2 + \sigma_d^2)^2
\end{aligned} \tag{21}$$

Since $\hat{Y}(m)$ is Gaussian, then its kurtosis $kurt(\hat{Y}(m))$ is zero.

$$kurt(\hat{Y}(m)) = E[|\hat{Y}(m)|^4] - E[\hat{Y}^2(m)] - 2E^2[|\hat{Y}(m)|^2] = 0 \tag{22}$$

Assuming that the real and the imaginary parts of $\hat{Y}(m)$ are independent and of the same variance then $E[\hat{Y}^2(m)]$ becomes zero. Therefore: $E[|\hat{Y}(m)|^4] = 2E^2[|\hat{Y}(m)|]^2 = 2(\sigma_w^2 + \sigma_d^2)^2$. Back to Eq. (21), the variance, V_0 becomes:

$$V_0 = \frac{1}{N}(\sigma_w^2 + \sigma_d^2)^2 \tag{23}$$

By following the same procedure, μ_1 and V_1 can be obtained as follows under H_1 ($X(m)$ exists):

$$\mu_1 = \sigma_w^2 + \sigma_d^2 + \sigma_x^2 \tag{24}$$

$$V_1 = \frac{1}{N}(\sigma_w^2 + \sigma_d^2 + \sigma_x^2)^2 \tag{25}$$

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