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Analysing Self Interference Cancellation in Full Duplex Radios

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***Abstract*—Full duplex communication promises a theoretical 100% throughput gain by doubling the number of simultaneous transmissions. Such compelling gains are conditioned on perfect cancellation of the self interference power resulting from simultaneous transmission and reception. Generally, self interference power is modelled as a noise-like constant level interference floor. However, experimental validations have shown that the self interference power is in practice a random variable depending on a number of factors such as the surrounding wireless environment and the degree of interference cancellation. In this study, we derive an analytical model for the residual self interference power, and demonstrate various applications of the derived model in analysing the performance of a Full Duplex radio. In general, full duplex communication is found to provide only modest throughput gains over half duplex communication in a dense network scenario with practical self interference cancellation models.**

***Index Terms*—Full duplex, self interference model, interference model, 5G.**

I. Introduction

Historically, full duplex communication had been considered impractical due to the overwhelming loopback interference from the transmission-end. Recent advances in self-interference cancellation (SIC) allow suppressing this loopback interference to within tolerable limits, thereby making full duplex (FD) communication appealing with viable cost. In that respect, FD has the potential of becoming a significant breakthrough in the design of a novel 5th Generation (5G) radio access technology [1]–[3].

Self-interference power in FD radios are generally in the order of a billion-fold stronger than the desired signal [2]. Recent progress in SIC techniques have enabled suppressing this interference power to within satisfactory levels through a combination of passive interference suppression, and active cancellation in both, the analog, and the digital domain [1], [2], [4]. This has resulted in research into FD communication receiving wide attention in recent years. For instance, the performance of wireless networks with FD capable radios has been investigated in [3]–[8] among others. Most studies investigating the performance of FD communication either assume perfect SIC [3], [5], [6], or consider the self interference power to be suppressed to a constant noise-floor like level [7], [8].

Assuming perfect self interference cancellation, the authors in [5] show that FD capabilities can provide close to 100% throughput (TP) gains in a cellular system with relatively isolated cells. However, much of these gains are lost when realistic inter-link interference and spatial reuse patterns are considered [6]. Similar findings were observed in our previous studies investigating the system level performance of dense small cell network with FD capable radios [3], [9]. In contrast, references [7], [8] investigate FD performance by considering the interference as a constant floor. In particular, reference [7] investigates the TP gain of FD communication in a large wireless network as a function of the SIC level using stochastic geometry tools; while the TP performance of a two-hop relay channel with FD enabled relay node for different SIC levels are analysed in [8].

Experimental evaluations have revealed that the residual self interference power after SIC is in practice a random variable and not a constant level interference floor [4], [10]. Inspired by the experimental findings presented in [10], this contribution derives an analytical model for the random residual self interference power of a FD radio. SIC generally involves two phases of cancellation, namely passive suppression and active cancellation. Passive suppression involves isolating the transmit and receive antennas electromagnetically, for example through physical separation of the transmit and receive antennas, or interference shielding via the placement of radio frequency (RF) absorbers. On the other hand, active suppression techniques exploit a node's knowledge of its own transmit signal to cancel the self-interference [4]. The proposed self interference model captures the impact of both of these SIC phases. An isolated FD transceiver pair is considered in order to focus the analysis on the impact of the self interference power by segregating it from out-of-cell interference sources.

Organization: Section II introduces the system model, followed by performance analysis of a Full Duplex Radio in the presence of residual self interference power in Section III. Numerical results demonstrating the validity of the analytical findings are presented in Section IV. Finally, Section V discusses closing remarks and future outlook.

II. Signal Model

We consider a full duplex transceiver pair. Each node is assumed to have a single transmit and a single receive antenna with antenna separation of d_{AS} centimeters (cm). The transceiver pairs are separated by a distance of r

meters, as depicted in Figure 1.

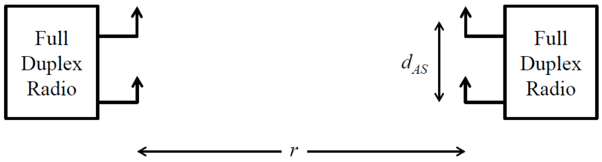

Fig. 1. System Model showing a standalone Full Duplex Transceiver pair.

A. The Residual Self Interference Power

Most existing contributions on FD performance analysis consider the self interference to be a constant-level noise-floor like signal. In practice, empirical studies [4], [10] have shown that the amount of SIC is in fact a random variable (r.v.) that depends on the random wireless fading channel, and the amount passive and active interference cancellation. A distribution for the post-SIC residual self interference power is characterized in this subsection.

Let I denote the post-SIC residual self interference power. The r.v. I can be decoupled into the product $I = P_I \Omega_\phi(P_I)$, where P_I is the received self interference power at the antenna front-end following passive suppression of the transmitted signal, while $\Omega_\phi(P_I)$ denotes active interference cancellation. The active interference cancellation, which is a function of P_I, can be of two types: either pure analog cancellation (AC), or analog cancellation supplemented by digital cancellation (ACDC) [10]. The subscript $\phi \in \{\text{AC, ACDC}\}$ in $\Omega_\phi(P_I)$ indicates the active cancellation type.

1) Self Interference Power at Receiver Front End: The residual self interference power at the antenna front-end following passive interference suppression can be expressed as $P_I = P_T \kappa d_{AS}^{-\alpha} g$, where the constant variables κ and α are respectively the path loss constant at reference distance, and the path loss exponent. Note that, such a path loss model is valid for short dipole antennas at $d_{AS} > \lambda/2\pi$, where λ is the wavelength. This condition is satisfied for $d_{AS} > 10$ cm in the below 6 GHz 5G spectrum of interest [11].

The randomness in P_I is due to the random fading power of the wireless channel between the transmit and the receive antenna as given by $g = |h|^2$, where h is the channel fading gain. According to the empirical results reported in [10], h is found to follow a Rician distribution with a strong line of sight component (K factor). For such a large K value, it is possible to approximate the Rician distribution with a Gaussian PDF of the following form [12, pg. 117]

$$f_h(\rho) = \frac{1}{\sqrt{\pi(1-v^2)}} \exp\left(-\frac{(\rho - v)^2}{(1-v^2)}\right), \tag{1}$$

where $v^2 = \frac{K}{1+K}$. By a change of variable involving $g = |h|^2$, the distribution of P_I readily follows as

$$f_{P_I}(x) = \frac{1}{2\sqrt{\pi\tau x(1-v^2)}} \exp\left(-\frac{(\sqrt{x/\tau} - v)^2}{1-v^2}\right), \tag{2}$$

where the constant $\tau = P_T \kappa d_{AS}^{-\alpha}$.

2) Active Interference Cancellation: The amount of active interference cancellation is a function of the received self interference power (P_I) and is given *in dB values* as [10]

$$\Omega'_\phi(P'_I) = -\lambda_\phi P'_I - \beta'_\phi, \tag{3}$$

where P'_I is P_I in dBm values, and λ_ϕ and β'_ϕ (also in dB values) are active interference cancellation parameters. Let us define the constant $\beta_\phi \triangleq 10^{-\frac{\beta'_\phi + 30\lambda_\phi}{10}}$. Converting $\Omega'_\phi(P'_I)$ and P'_I to linear values, Eq. (3) becomes $\Omega_\phi(P_I) = \beta_\phi P_I^{-\lambda_\phi}$. The parameter values as reported in [10] are: $\lambda_{AC} = 0.21$ dB/dBm, $\lambda_{ACDC} = 0.12$ dB/dBm, $\beta'_{AC} = 37.42$ dB, and $\beta'_{ACDC} = 35.49$ dB. It should be highlighted here that, though the above parameter values have been derived through extensive empirical studies, they still depend on specific conditions and experimental setups.

B. Characterizing the Distribution of I

The self interference power (normalized by noise power, N_0) after both stages of self interference cancellation can be expressed as a function of the single r.v. P_I as

$$I = P_I \Omega_\phi(P_I) = \xi_\phi P_I^{1-\lambda_\phi}, \tag{4}$$

where the constant $\xi_\phi = \beta_\phi / N_0$. The subscript ϕ is henceforth omitted for the ease of presentation.

The distribution of P_I is given in Eq. (2). Deriving the distribution of I through conventional change of variable leads to an equation that is not easily tractable. However, numerical simulations show that the empirical distribution of the residual self interference power after SIC closely matches a gamma distribution. We therefore propose to model the random residual self interference power as a gamma r.v. having the following distribution [13]

$$f_I(x; m, \mu_I) = \frac{m^m x^{m-1}}{\mu_I^m \Gamma(m)} \exp\left(-\frac{mx}{\mu_I}\right). \tag{5}$$

The gamma distribution is characterized by parameters m and mean μ_I, where $m = \frac{\mu_I^2}{\mathbb{E}[I^2] - \mu_I^2}$ and mean $\mu_I = \mathbb{E}[I]$. $\Gamma(m) \triangleq \int_0^\infty t^{m-1} \exp(-t)\, \mathrm{d}t$ is the Gamma function.

Derivation of the Gamma Distribution Parameters

The Mean of I: The mean residual self interference power $\mu_I = \mathbb{E}[I]$ is derived by taking the expectation of I over the distribution of P_I as given by Eq. (2), which yields

$$\mu_I = \int_0^\infty \frac{\xi x^{1-\lambda}}{2\sqrt{\pi\tau x(1-v^2)}} \exp\left(-\frac{\left(\sqrt{x/\tau} - v\right)^2}{1-v^2}\right) \mathrm{d}x$$
$$= \frac{\xi \tau^{1-\lambda}}{\sqrt{\pi}} \int_{-\sqrt{K}}^{\infty} \left(v + t\sqrt{1-v^2}\right)^{2-2\lambda} \exp\left(-t^2\right) \mathrm{d}t, \tag{6}$$

where the second step follows from a change of variable involving $t = \left(\sqrt{x/\tau} - v\right)/\sqrt{1-v^2}$. Since the above integral cannot be reduced to a closed-form solution, we resort to evaluating it numerically using the Gauss Hermite quadrature integral [14, Eq. (25.4.46)]. The resulting mean residual interference power is

$$\mu_I \cong \frac{\xi \tau^{1-\lambda}}{\sqrt{\pi}} \sum_{x \in \mathcal{X}} w(x) \left(v + x\sqrt{1-v^2}\right)^{2-2\lambda}, \quad (7)$$

where the set $\mathcal{X} = \{x_i | x_i \geq -\sqrt{K}\}$, with x_i being the i^{th} root of the N order Hermite polynomial $H_N(x)$, and $w(x_i) = 2^{N-1}N!\sqrt{\pi}/(N[H_{N-1}(x_i)])^2$ being the respective weight. The accuracy of the numerical integration is determined by the Gauss Hermite approximation order N. In this contribution, we let $N \geq 20$ to ensure sufficient numerical accuracy.

The Second Moment of I: The second moment of I, $\mathbb{E}[I^2]$ is defined as

$$\mathbb{E}[I^2] = \int_0^\infty \frac{\left(\xi x^{1-\lambda}\right)^2}{2\sqrt{\pi \tau x(1-v^2)}} \exp\left(-\frac{(\sqrt{x/\tau} - v)^2}{1-v^2}\right) \mathrm{d}x. \quad (8)$$

Following similar procedures as in (6), $\mathbb{E}[I^2]$ numerically evaluates to

$$\mathbb{E}[I^2] \cong \frac{\left(\xi \tau^{1-\lambda}\right)^2}{\sqrt{\pi}} \sum_{x \in \mathcal{X}} w(x) \left(v + x\sqrt{1-v^2}\right)^{4-4\lambda}. \quad (9)$$

Knowing the first two moments of I allows us to fully characterize the approximated gamma distribution of the post-SIC residual self interference power in a full duplex radio. Empirically derived distributions of the residual self interference power, and their gamma approximations for different parameter values are shown in Fig. 2; wherein a close match is observed between both the distributions[1].

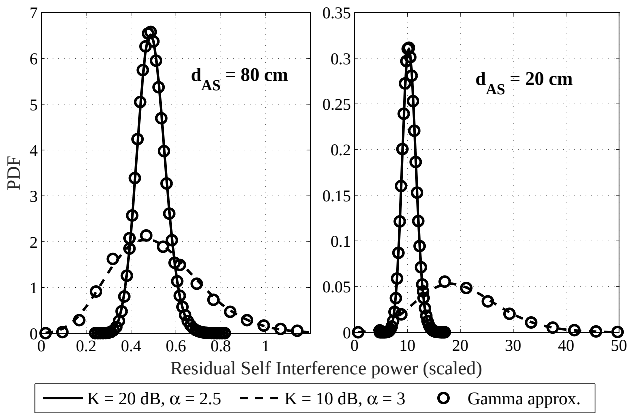

Fig. 2. Empirically derived distributions of the residual self interference, and their gamma approximations for different parameter values.

III. Performance Analysis of a FD Radio

The TP performance of a FD radio in the presence of residual self interference power is analysed in this section. More specifically, we present the bandwidth normalized ergodic and instantaneous TP in terms of bits per channel use (bpcu). The performance is compared against that of a half duplex (HD) radio node.

[1] P_I has been equally scaled in both sub-plots for the ease of presentation.

Signal to Interference Plus Noise Ratio: The considered TP performance measures are generally a function of the received signal-to-interference-plus-noise ratio (SINR) at the receiver node. The SINR with FD communication is $\gamma_{FD} = \zeta/(I+1)$, where ζ is the signal-to-noise ratio (SNR) of the desired signal power. In the case of HD communication, $\gamma_{HD} = \zeta$ since the residual self interference power $I = 0$.

The desired signal amplitude is assumed to follow a Nakagami-m fading distribution, which is a general fading distribution that includes a wide range of other distributions as special case via its parameters [13]. The SNR is correspondingly distributed according to a gamma distribution. Let n and μ_ζ respectively denote the gamma fading parameter and the mean of the SNR ζ. Furthermore, ζ is assumed to be independent from the residual self interference power I.

Assuming the Shannon rate can be achieved at every resource slot, the instantaneous TP conditioned on the received SINR is given by the famous Shannons formula for AWGN channels $R(\gamma) = \log(1+\gamma)$, where the logarithm is base 2.

A. Ergodic Throughput Analysis

The bandwidth normalized ergodic throughput can be obtained by taking the expectation of the instantaneous TP over the distributions of ζ and I. With FD communication, the ergodic TP is formulated as

$$\bar{R}_{FD} = 2\,\mathbb{E}\left[\log\left(1 + \frac{\zeta}{I+1}\right)\right], \quad (10)$$

where the factor 2 accounts for simultaneous transmission and reception across both uplink and downlink directions.

The direct method for calculating the above expectation requires multiple integrations over the distributions of ζ and I, which cannot be evaluated readily. Instead, we simplify the above expression by applying Hamdi's lemma to evaluate the capacity involving multiple integrals [15]. By virtue of the independence assumption between ζ and I, the ergodic throughput with FD communication can be expressed in the form of a single integral as

$$\bar{R}_{FD} = \frac{2}{\ln(2)} \int_0^\infty \frac{\mathcal{M}_I(s)\left(1 - \mathcal{M}_S(s)\right)}{s} \exp(-s)\,\mathrm{d}s, \quad (11)$$

where $\mathcal{M}_X(s) \triangleq \mathbb{E}_X\left[\exp(-sX)\right]$ is the Laplace transform of X and $\ln(x)$ is the natural logarithm. For gamma distributed r.v.s with parameter m and mean μ, $\mathcal{M}(s) = \left(1 + \frac{s\mu}{m}\right)^{-m}$ [13]. The integrand in (11) is a continuous and bounded non-negative quantity in the range of integration. Therefore, the ergodic throughput with FD communication can be easily computed using any suitable numerical integration techniques or available mathematical software by plugging in the appropriate parameter values of ζ and I.

On the other hand, the ergodic TP with HD communication is given by $\bar{R}_{HD} = \mathbb{E}\left[\log\left(1+\zeta\right)\right] = \int_0^\infty \log\left(1+x\right) \frac{n^n x^{n-1}}{\mu_\zeta^n \Gamma(n)} \exp\left(-\frac{nx}{\mu_\zeta}\right) \mathrm{d}x$. On utilizing the identities $\ln(1+x) = \mathrm{G}_{2,2}^{1,2}\left[x \middle| {}^{1,1}_{1,0}\right]$ and $e^{-x} = \mathrm{G}_{0,1}^{1,0}\left[x \middle| {}^{-}_{0}\right]$ [16, Eq. (11)], the ergodic TP with HD communication can be reformulated as

$$\bar{R}_{HD} = \frac{(n/\mu_\zeta)^n}{\ln(2)\,\Gamma(n)} \int_0^\infty x^{n-1}\, \mathrm{G}_{2,2}^{1,2}\left[x \middle| \begin{matrix}1,1\\1,0\end{matrix}\right] \mathrm{G}_{0,1}^{1,0}\left[\frac{n}{\mu_\zeta}x \middle| \begin{matrix}-\\0\end{matrix}\right] \mathrm{d}x, \quad (12)$$

where G[.] is the Meijer's G function as defined in [16, Eq. (5)]. Subsequently, the integral in Eq. (12) is solved via employing [16, Eq. (21)] as

$$\bar{R}_{HD} = \frac{(n/\mu_\zeta)^n}{\ln(2)\,\Gamma(n)}\, \mathrm{G}_{2,3}^{3,1}\left[\frac{n}{\mu_\zeta} \middle| \begin{matrix}-n, 1-n\\0, -n, -n\end{matrix}\right], \quad (13)$$

to obtain the ergodic TP with HD communication in closed-form with arbitrary fading parameter n and mean SNR μ_ζ.

B. Instantaneous Throughput Analysis

The bandwidth normalized ergodic throughput analysis in the previous subsection provides an overview of the average TP behaviour with FD and HD communication. In order to analyse the performance of FD communication in greater details, we investigate the respective instantaneous TP behaviour in this section.

The probability of the instantaneous FD TP exceeding the HD TP for a given SNR value is formulated as $\Pr[R_{FD} > R_{HD}|\zeta] = \Pr[2\log(1+\gamma_{FD}) > \log(1+\gamma_{HD})]$. Following some algebraic manipulations, the probability is reduced to $\Pr[R_{FD} > R_{HD}|\zeta] = F_I\left(\sqrt{\zeta+1}\right)$, where $F_X(x)$ is the cumulative density function (CDF) of the r.v. X evaluated at $X = x$. The probability that instantaneous FD TP exceeds the HD TP conditioned on a specific SNR value is then straightforwardly given by the CDF of a gamma r.v. as [14, Eq. 6.5.3]

$$\Pr[R_{FD} > R_{HD}|\zeta] = \frac{\gamma\left(m, \mu_I\sqrt{\zeta+1}/m\right)}{\Gamma(m)}, \quad (14)$$

where $\gamma(m,x) \triangleq \int_0^x t^{m-1}\exp(-t)\,\mathrm{d}t$ is the lower incomplete Gamma function.

Eq. (14) states the probability of observing an instantaneous TP gain with FD communication conditioned on the SNR ζ. The general probability of such an event can be obtained by averaging Eq. (14) over the r.v. ζ; as expressed in the following

$$\Pr[R_{FD} > R_{HD}] = \int_0^\infty F_I(\sqrt{t+1}) f_\zeta(t)\,\mathrm{d}t. \quad (15)$$

Using the relation between the CDF and Laplace transform of a r.v., $F_I(\sqrt{\zeta+1})$ can be found from the inverse Laplace transform of $\mathcal{M}(s)/s$ evaluated at $I = \sqrt{\zeta+1}$ [13, Eq. 1.6]. Eq. (15) can thus be rewritten as

$$\Pr[R_{FD} > R_{HD}] = \frac{1}{2\pi i}\int_{\nu-i\infty}^{\nu+i\infty} \frac{\mathcal{M}_I(s)}{s} \times \left[\int_0^\infty \exp\left(s\sqrt{t+1}\right) f_\zeta(t)\,\mathrm{d}t\right] \mathrm{d}s, \quad (16)$$

where $i = \sqrt{-1}$ is the imaginary number, and ν is chosen in the region of convergence of the integral in the complex s plane. Assuming a high SNR scenario $(\zeta \gg 1)$,[2] the inner integral in Eq. (16) can be approximated as $\int_0^\infty \exp\left(s\sqrt{t}\right) f_\zeta(t)\,\mathrm{d}t$. By virtue of the gamma distribution assumption on ζ and the definition of a moment generating function (MGF), $\int_0^\infty \exp\left(s\sqrt{t}\right) f_\zeta(t)\,\mathrm{d}t$ is in fact the MGF of a Nakagami-m distributed r.v.; and is given by $\mathcal{M}_{\sqrt{\zeta}}(-s) = \int_0^\infty \exp(sx)\frac{2(n/\Omega)^n}{\Gamma(n)}x^{2n-1}\exp\left(-\frac{nx^2}{\Omega}\right)\mathrm{d}x$, where Ω is the mean amplitude. Following similar steps as in Section III-A, the above integral can be solved in closed-form in terms of the Meijer's G function as

$$\mathcal{M}_{\sqrt{\zeta}}(-s) = \frac{2^{2n}\,(n/\Omega)^n}{\sqrt{\pi}\,s^{2n}\,\Gamma(n)}\, \mathrm{G}_{2,1}^{1,2}\left[\frac{4n}{s^2\,\Omega} \middle| \begin{matrix}\frac{1-2n}{2}, 1-n\\0\end{matrix}\right]. \quad (17)$$

Finally, the general probability of observing an instantaneous TP gain with FD communication can be succintly expressed in terms of $\mathcal{M}_I(s)$ and $\mathcal{M}_{\sqrt{\zeta}}(-s)$ as

$$\Pr[R_{FD} > R_{HD}] = \frac{1}{2\pi i}\int_{\nu-i\infty}^{\nu+i\infty} \frac{\mathcal{M}_I(s)\mathcal{M}_{\sqrt{\zeta}}(-s)}{s}\,\mathrm{d}s. \quad (18)$$

A summary of numerical methods to evaluate Eq. (18) can be found in [17].

IV. Numerical Results

The throughput performance metrics of full duplex and half duplex communication are numerically validated through Matlab® based Monte Carlo simulation in this Section. At least $100,000$ independent snapshots of each scenario are simulated to ensure statistical reliability. Unless stated otherwise, the following general simulation parameters are assumed to reflect a typical propagation scenario: path loss exponent[3] $\alpha = 4$, reference path loss $\kappa = 40$ dB, gamma parameter for the desired signal channel $n = 2$, and Rician K component of the antenna separation channel $K = 20$ dB.

A. Ergodic Throughput Analysis

The Ergodic Throughput of FD communication for different antenna separation distances (d_{AS}) are presented in Fig. 3, along with the HD throughput. The two different methods of SIC discussed in [10] are considered, namely analog cancellation only (AC) and analog-plus-digital cancellation (ACDC). The ergodic TP curves are plotted against the mean SNR of the desired channel μ_ζ given in dB values.

The TP curve with HD communication displays the well known linear increase with increasing SNR (in dB values). In contrast, the ergodic TP curves of FD communication

[2] The amount of SIC is in the range of $70 - 80$ dB [10], meaning the the residual self interference is usually quite strong. Therefore, FD will only makes sense in the high SNR regime; which justifies the above assumption.

[3] Note that, the amplitude of transmitted signal decays with the square of the distance at such short distances.

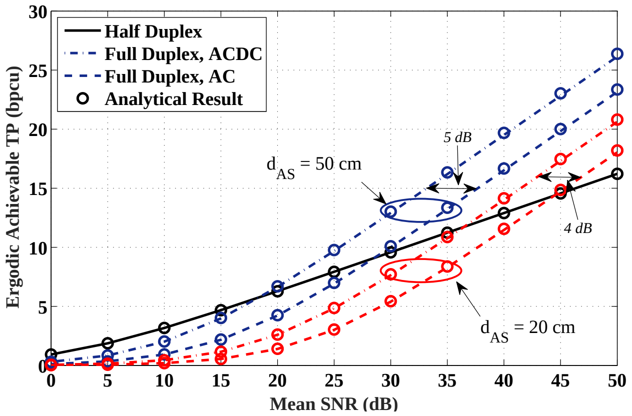

Fig. 3. The ergodic throughput of full duplex and half duplex communication for different mean SNR (μ_S) values.

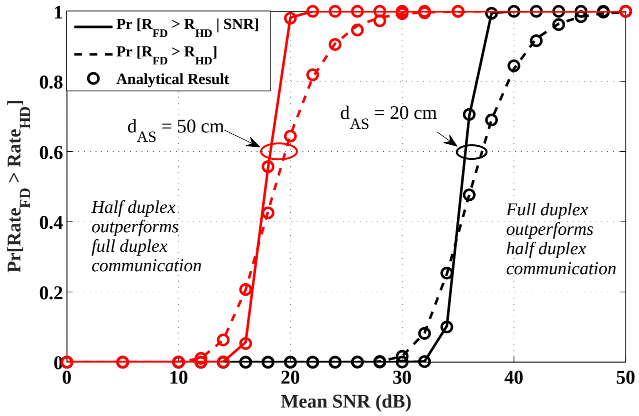

Fig. 4. Probabilities of full duplex communication outperforming half duplex communication for different antenna separation distances.

exhibit two distinct regions: a poor TP performance with slow growth in the low SNR regime, followed by a stronger linear growth region in higher SNR regime. The former behaviour results from the strong residual self interference power, which drives down the SINR to low values. As the SNR is increased, the linear growth in the TP due to simultaneous communication in both directions outperforms the loss in the SINR; thereby resulting in a throughput gain over HD communication. In fact, the slope of the ergodic TP curve with FD communication in the high SNR region is twice the slope with HD communication; which confirms that FD provides a two-fold multiplexing gain over HD communication.

In general, the TP gain of FD communication depends on the amount of SIC. Increasing the antenna separation distance or having multiple SIC stages (i.e., AC followed by digital cancellation) improves the TP performance of FD communication. In particular, it was reported in [10] that when AC gets worse and consequently remaining self-interference becomes larger, digital cancellation provides more gain. A similar observation is reflected in Fig. 3, where higher antenna separation results in greater passive cancellation, which corresponds to a lower P_I and consequently a worse AC. As a result, a higher TP gain due to the digital cancellation phase is observed for a higher antenna separation distance d_{AS}.

Finally, the presented ergodic TP curves showcase the inability to deliver the promised 100% TP gain of FD communication. In fact, such a gain is only achievable in the asymptotic high SNR regime as $\mu_\zeta \to \infty$.

B. Instantaneous Throughput Analysis

The instantaneous TP performance with ACDC SIC is evaluated in this section. The instantaneous TPs discussed in Section III-B are shown. Fig. 4 presents the probabilities of the events that the instantaneous TP with FD communication outperforms that with HD communication. The said probability conditioned on the SNR as given by Eq. (14) is indicated by $Pr\left[R_{FD} > R_{HD}|SNR\right]$; while the legend $Pr\left[R_{FD} > R_{HD}\right]$ refers to the probability averaged over the SNR distribution as expressed by Eq. (18).

For each of the presented scenarios, the area to left of the curve indicates the region where HD outperforms FD, and vice versa. Expectedly, the transition from one region to the other is smoother when averaged over the SNR distribution compared to the fixed SNR case.

It is interesting to observe the role of the antenna separation distance in naturally mitigating parts of the self interference power. Increasing d_{AS} from 20 to 50 cm (a factor of 4 dB) reduces the required SNR for meaningful FD communication by almost 20 dB (from 35 dB to around 15 dB). Thus, passive interference suppression through separating the transmit and the receive antenna, or by placing an RF insulator between them can be a significant contributing factor to the overall SIC in a FD node. Such a finding have also been independently reported in [4], where the authors have presented a detailed measurement-based study analysing the capabilities and limitations of passive SIC.

C. Impact of the Channel Fading Parameters

Finally, the impact of the different self interference channel fading parameters on the instantaneous TP performance is investigated in this section. The instantaneous TP performance averaged over the desired SNR distribution for different SIC types and different parameters of the self interference channel are presented in Fig. 5. The antenna separation distance, $d_{AS} = 30$ cm.

Increased variability in the self interference channel results in a greater variation in I, which leads to an increased fluctuation in the resulting SINR. Accordingly, the curves in Fig. 5 corresponding to $K = 10$ dB, and $\alpha = 3$ exhibit a lower slope compared to the curves of $K = 20$ dB, and $\alpha = 4$. Although not shown here, a similar effect is observed by increasing the fluctuation in the desired signal channel through a reduction in the n parameter value.

In contrast to the ergodic TP, we observe a higher gain of digital cancellation on top of AC with the instantaneous TP. Having a second stage of active digital cancellation results

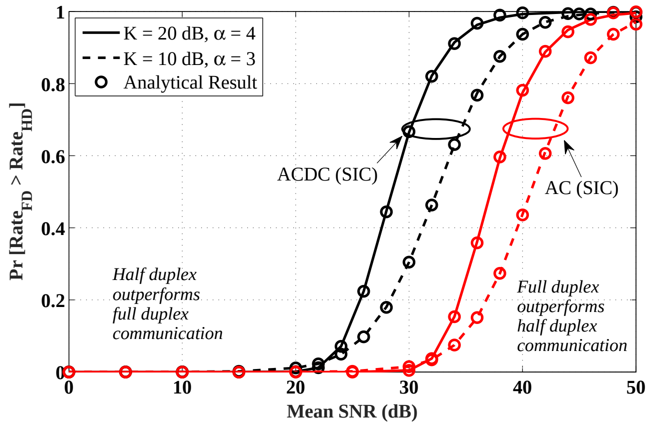

Fig. 5. Probabilities of full duplex communication outperforming half duplex communication for different parameters of the self interference fading channel; and different cancellation types.

in about 8 dB SNR gain compared to the ~ 4 dB gain for the case of the ergodic TP as observed earlier in Fig. 3. This highlights the importance of considering different performance metrics in order to obtain a diversified picture of the performance.

V. Conclusions and Outlook

The promise of doubling the network throughput makes full duplex communication an attractive feature for a 5G radio access technology. Suppressing the loopback interference from the transmission-end to a tolerable level is critical in making FD communication favourable. The residual self interference power in FD radios is analytically modelled, and cross validated through extensive Monte Carlo simulations in this contribution. Applications of the derived residual self interference model in evaluating the potential of FD communication are then demonstrated through ergodic and instantaneous TP analysis. The analytical findings are found to closely match the simulation results in all scenarios, thereby validating the accuracy of the analysis and the presented results.

Numerical results reveal that the promised 100% TP gain with FD communication is not achievable in practice. In a realistic setting, the TP gains of FD are affected by level of effective self interference cancellation achieved during the passive interference suppression and the active cancellation stage. Wireless channel variables such as the path loss exponent and fading parameters are also found to impact the SIC performance. The derived analytical results provide with a rather simple model to evaluate the potential performance of FD communication. As part of the future work, we plan to extend our study by analysing the impact of network level interferences on FD performance; and investigate algorithms that can enhance the potential of FD communication.

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References

[1] A. Sabharwal *et al.*, "In-band full-duplex wireless: Challenges and opportunities," *IEEE Journal on Selected Areas in Communications*, vol. 32, no. 9, pp. 1637–1652, Sep. 2014.

[2] S. Hong *et al.*, "Applications of self-interference cancellation in 5G and beyond," *IEEE Communications Magazine*, pp. 114–121, Feb. 2014.

[3] N. H. Mahmood, G. Berardinelli, F. Tavares, and P. Mogensen, "On the potential of full duplex communication in 5G small cell networks," in *Proc. IEEE 81st Vehicular Technology Conference: VTC-Spring*, Glasgow, Scotland, May 2015, pp. 1 – 5.

[4] E. Everett, A. Sahai, and A. Sabharwal, "Passive self-interference suppression for full-duplex infrastructure nodes," *IEEE Transactions on Wireless Communications*, vol. 13, no. 2, pp. 680–694, Feb. 2014.

[5] S. Goyal *et al.*, "Full duplex operation for small cells," *submitted to IEEE Transactions on Vehicular Technology*, Apr. 2015. [Online]. Available: http://arxiv.org/pdf/1412.8708.pdf

[6] X. Xie and X. Zhang, "Does full-duplex double the capacity of wireless networks?" in *Proc. The 33rd Annual IEEE International Conference on Computer Communications (INFOCOM'14)*, Toronto, Canada, Apr. 2014, pp. 253 – 261.

[7] Z. Tong and M. Haenggi, "Throughput analysis for wireless networks with full-duplex radios," in *Proc. Wireless Communications and Networking Conference (WCNC)*, New Orleans, USA, Mar. 2015, pp. 717 – 722.

[8] T. Riihonen, S. Werner, and R. Wichman, "Hybrid full-duplex/half-duplex relaying with transmit power adaptation," *IEEE Transactions on Wireless Communications*, vol. 10, no. 9, pp. 3074–3085, Sep. 2011.

[9] N. H. Mahmood, G. Berardinelli, P. Mogensen, and F. Frederiksen, "Throughput analysis of full duplex communication with asymmetric traffic in small cell systems," in *inProc. The Eleventh International Conference on Wireless and Mobile Communications (ICWMC)*, St. Julians, Malta, Oct. 2015, pp. 57–60.

[10] C. D. M. Duarte and A. Sabharwal, "Experiment-driven characterization of full-duplex wireless systems," *IEEE Transactions on Wireless Communications*, vol. 11, no. 12, pp. 4296–4307, Dec. 2012.

[11] P. Mogensen *et al.*, "Centimeter-wave concept for 5G ultra-dense small cells," in *Proc. VTC-Spring Workshop on 5G Mobile and Wireless Communication System for 2020 and Beyond (MWC2020)*, Seoul, South Korea, May 2014, pp. 1–6.

[12] G. Durgin, *Space-time Wireless Channels*, 1st ed. Upper Saddle River, NJ, USA: Prentice Hall Press, 2002.

[13] M. K. Simon and M.-S. Alouini, *Digital Communication over Fading Channels*, 2nd ed. New Jersey, USA: John Wiley & Sons, Dec. 2005.

[14] M. Abramowitz and I. Stegun, Eds., *Handbook of Mathematical Functions with Formulas, Graphs, and Mathematical Tables*, 2nd ed. New York, USA: Dover Publications, 1972.

[15] K. A. Hamdi, "A useful lemma for capacity analysis of fading interference channels," *IEEE Transaction on Communications*, vol. 58, no. 2, pp. 411–416, Feb. 2010.

[16] V. S. Adamchik and O. I. Marichev, "The algorithm for calculating integrals of hypergeometric type functions and its realization in reduce system," in *Proceedings of International Symposium on Symbolic and Algebraic Computation (ISSAC' 90)*, New York, USA, 1990, pp. 212–224.

[17] J. Abate and W. Whitt, "Numerical inversion of laplace transforms of probability distributions," *ORSA Journal on computing*, vol. 7, no. 1, pp. 36–43, 1995.