Sensing-Communication Coexistence vs. Integration

Meng Liu, Student Member, IEEE, Minglei Yang, Member, IEEE, Zhaoming Zhang, Huifang Li, Fan Liu, Member, IEEE, Arumugam Nallanathan, Fellow, IEEE, and Lajos Hanzo, Life Fellow, IEEE

Abstract—The design options of sensing-communication coexistence (SCC) vs. integrated sensing and communications (ISAC) relying on non-orthogonal downlink transmission (NO-DLT) and transmit antenna selection are investigated. To be practical, both realistic residual hardware impairments (RHI) and imperfect successive interference cancellation are considered. The performance of the SCC and ISAC NO-DLT frameworks is characterized by deriving both the exact and the asymptotic outage probabilities (OPs) as well as the probability of successful target detection (PoD). Our numerical results demonstrate the accuracy of the theoretical analysis and confirm that: 1) The ISAC NO-DLT framework is superior to the SCC NO-DLT framework both in terms of its OP and PoD; 2) The realistic RHI degrades the S&C performance of both the SCC and ISAC NO-DLT frameworks.

I. INTRODUCTION

Since the fifth-generation (5G) new radio (NR) standardization has reached maturity, the research commonly has turned to exploring new avenues. Sensing and communication (S&C) systems tend to become more and more similar in terms of their hardware, channel characteristics, and signal processing. Moreover, the combination of S&C can further improve the efficiency of wireless resources [1]. To this end, it is envisioned that radio sensing will develop into an essential function of the future communication systems in support of the vehicle-to-everything, smart homes, and environmental monitoring, etc [2]. Sensing-communication coexistence (SCC) as well as integrated sensing and communication (ISAC) are considered as feasible frameworks for the sixth-generation (6G) systems [3, 4].

Under the SCC framework, the S&C stations operate independently of each other, while in the ISAC framework, a unified hardware platform is used for accomplishing both S&C functionalities simultaneously [5–8]. In [5], a power allocation algorithm was designed for maximizing the communication throughput while satisfying the minimum receive signal-to-interference-plus-noise ratio (SINR) required for reliable sensing in the SCC system. The sum-rate and the probability of detection (PoD) for sensing were analyzed in the full-duplex SCC system scenario of [6]. In [7], the non-cooperative target was regarded as an illegal eavesdropper, and then both the receive signal-to-noise ratio (SNR) of the target and the secrecy rate of the system were investigated in the ISAC system, respectively. The authors of [8] derived both the sensing estimation rate and communications water-filling inner bound of the ISAC systems in which the sensing station was employed as a relay for communication.

The contributions [5–8] considered traditional orthogonal downlink transmission (O-DLT) for communication. However, O-DLT is unable to satisfy the massive access of requirements imposed by the rapids of Internet-of-Things (IoT) devices. As a remedy, non-orthogonal downlink transmission (NO-DLT) was proposed as a promising concept for 6G to ensure that the spectral efficiency and quality of service for the IoT devices can be further improved [9]. Specifically, NO-DLT provides services for multiple IoT devices by sending superimposed signals in the same time/frequency/code domain resources from the transmitter (TX). At the receiver (RX), typically successive interference cancelation (SIC) is adopted for detecting the signals [9]. In [10], the transmit beamformer was designed for maximizing the secrecy sum rate of ISAC NO-DLT systems, where the sensing target was considered to act as an eavesdropper degrading the confidentiality of the communication function. However, the authors of [5–8, 10] assumed having perfect radio frequency (RF) equipment, which is overly idealistic. In realistic practical scenarios, the equipment may be affected by phase noise, in-phase and quadrature-phase imbalance, and amplifier non-linearities. Although some advanced compensation algorithms have mitigated the influence of the hardware impairments (HIs), leaving residual HI (RHI) is inevitable [11].

The authors of [3–8] laid the solid foundations for both SCC and ISAC systems. However, the S&C performance analysis of the SCC and ISAC frameworks based on NO-DLT is still widely unexplored. In light of the above, we fill in these research gaps in the SCC and ISAC NO-DLT literature. As a further extension, a transmit antenna selection (TAS) strategy based on the IoT device fairness is proposed for reducing the number of activated RF chains and the power consumption as well as hardware cost, while maintaining the benefits of multiple antennas, while satisfying IoT devices having different priorities [12]. To be more practical, by taking into account the equipment aging and the environmental uncertainties, we elaborate on quantifying the residual hardware impairment (RHI) effects, which result in imperfect SIC (ipSIC) at the RXs. It should be emphasized that both the SCC and ISAC NO-DLT frameworks considered can be used in vehicle-to-vehicle and vehicle-to-everything communications [13, 14]. In order to evaluate the performance of the SCC and ISAC NO-DLT frameworks, the closed-form and approximate outage probabilities (OPs) of the IoT devices are obtained for communication. In addition, the sensing performance is analyzed by deriving the expression of the PoD. Our contributions are boldly and explicitly contrasted to the relevant state-of-the-art at a glance in Table I. Our numerical results show that due to the interference and other non-ideal factors, the OPs of the IoT devices exhibit error floors. Moreover, we demonstrate that the outage performance and the sensing capability of the ISAC outperform those of the SCC NO-DLT framework.

Notations: In this paper, \( \mathcal{CN}(a, b) \) denotes a complex Gaussian random variable with mean \( a \) and variance \( b \); \( \mathcal{C} \)
represent complex numbers; $\mathbb{E}[X]$ is the expectation of $X$; $\Gamma(a, b)$ represents the incomplete gamma function.

## III. SYSTEM MODEL

In Fig. 1(a) and Fig. 1(b), we consider the SCC and ISAC frameworks based on the NO-DLT supporting multiple IoT devices, respectively. For the SCC NO-DLT framework, the communication station $S$ and the sensing station $R$ operate independently of each other and are deployed at different locations. Specifically, $S$ aims for communicating with $N$ IoT devices ($D_1, D_2, \ldots, D_N$), while $R$ makes the decisions concerning the existence of a non-cooperative target $T$ within the sensing range. It is assumed that all the IoT devices and $R$ are equipped with a single-antenna, while $S$ is equipped with $N_S$ antennas with $N_S > N$. For notational convenience, we use $h_{S,D_m}$ to denote the channel between the $k$-th antenna of $S$ and the $m$-th IoT device. Furthermore, $h_{RT}$ and $h_{TR}$ denote the sensing channels of the $R \rightarrow T$ and $T \rightarrow R$ links, $h_{RD_m}$ represents the interference channel spanning from $R$ to the $m$-th IoT device, $h_{SR}$ is the interference channel impinging from the $k$-th antenna of $S$ upon $R$, and $h_{TD_m}$ is the interference channel reflected from $T$ to the $m$-th IoT device. By contrast, for the ISAC NO-DLT framework, $S$ performs both communication and target sensing functions simultaneously. We assume that $S$ is equipped with $N_S$ antennas. Additionally, $h_{ST}$ and $h_{SR}$ denote the sensing channels and $h_{SI}$ represents the self-interference (SI) channel associated with the transmitting and receiving signals simultaneously. It is assumed that the small-scale fading channels between all equipment pairs are modeled by the uncorrelated Rayleigh distribution. We use $h_i = h_i \sqrt{1 + d_i^2}$ and $\rho_i = |h_i|^2$, $i \in \{S,D_m, RT, TR, RD_m, TD_m, SR, ST, TS\}$ to denote the channel coefficients and gains, where $h \sim \mathcal{CN}(0, \Omega)$. Finally, $\alpha$ is the path-loss exponent and $d_i$ denotes the distance.

For the TAS strategy proposed, the best antenna is selected for maximizing the received SINR of $D_m$. More particularly, to further improve the fairness of the IoT devices for the NO-DLT system considered, we assume that the farthest IoT device $D_N$ has the highest priority, followed by $D_{N-1}, \ldots, D_1$ has the lowest priority. On this basis, the best antenna for $D_N$ is selected from $N_S$ antennas, and $D_1$ selects from the remaining $N_S - N + 1$ antennas. Hence, the best antenna for $D_m$ can be chosen as

$$k^* = \arg \max_{k \in \{1, 2, \ldots, N_S\}} \rho_{S_k, D_m}$$

(1)

### A. SCC NO-DLT framework

As seen from Fig. 1(a) for the SCC NO-DLT framework, $S$ and $R$ transmit their own signals so that the S&C can be carried out. As for the communication function, $S$ intends to send $x_S = \sum_{n=1}^{N} a_n P_S x_n$ to the IoT devices, while $x_r \in \mathbb{C}$ is the expected signal of $D_n$ with $\mathbb{E}[|x_n|^2] = 1$, $P_S$ denotes the transmit power of $S$, and $a_n$ represents the power allocation coefficient of $x_n$ with $\sum_{n=1}^{N} a_n = 1$ and $a_1 < a_2 < \ldots < a_N$. As for the sensing function, $R$ sends $x_r = \sqrt{P_R} x_r$ to sense the non-cooperative targets, where $x_r \in \mathbb{C}$ is the sensing signal with $\mathbb{E}[|x_r|^2] = 1$ and $P_R$ denotes the transmit power of $R$. From a global perspective, the signal received at the $m$-th IoT device is expressed as

$$y_{D_m}^C = h_{SD_m} (x_S + \eta_{SD_m}) + h_{TD_m} h_{RT} (\delta_{RD_m} x_r + \eta_{RD_m}) + \eta_{SR},$$

(2)

where $\delta_{RD_m}$ is the scattering coefficient of the target by $R \rightarrow T \rightarrow D_m$, $\eta_{SD_m} \sim \mathcal{CN}(0, N_0)$ denotes the additive white Gaussian noise (AWGN) at $D_m$ and $N_0$ is the noise power. Furthermore, $\eta_{SR} \sim \mathcal{CN}(0, \kappa_{SR}^2 P_S)$ and $\eta_{RD_m} \sim \mathcal{CN}(0, \kappa_{RD_m}^2 \rho_{RD_m} P_S)$ represent the distortion of $S$ and $D_m$ by $S \rightarrow D_m$, while $\kappa_{SR}$ and $\kappa_{RD_m}$ characterize the levels of RHI at the TX and RX. To elaborate further, $\eta_{RD_m} \sim \mathcal{CN}(0, \kappa_{RD_m}^2 P_R)$ and $\eta_{RD_m} \sim \mathcal{CN}(0, \kappa_{RD_m}^2 \rho_{RD_m} P_R)$ represent the distortion of $R$ and $D_m$ by $R \rightarrow T \rightarrow D_m$ and $\rho_{RD_m} \sim \mathcal{CN}(0, \kappa_{RD_m}^2 \rho_{RD_m} P_R)$ represents the distortion of $D_m$ by $R \rightarrow D_m$ [11]. After some further mathematical manipulations, (2) can be rewritten as

$$y_{D_m}^C = h_{SD_m} (x_S + \eta_{SD_m}) + h_{RD_m} (x_R + \eta_{RD_m})$$

(3)

where $\eta_{SD_m} \sim \mathcal{CN}(0, \kappa_{SD_m}^2 P_S)$ and $\eta_{RD_m} \sim \mathcal{CN}(0, \kappa_{RD_m}^2 P_R)$. The aggregate levels of RHI with $\kappa_{SD_m} = \sqrt{\kappa_{SR}^2 + \kappa_{RD_m}^2}$ and $\kappa_{RD_m} = \sqrt{\kappa_{RD_m}^2 + \kappa_{RD_m}^2}$.

According to the NO-DLT protocol, SIC is adopted at the RXs for detecting the signals. Therefore, the received SINR of $x_j$ at $D_m$, $m \leq j \leq N$, can be expressed as

$$\gamma_{D_{m-j}} = \frac{\rho_{SD_m} a_j}{\rho_{SD_m} |s \phi_1 + \rho_{RD_m} a_0 \phi_2 + \rho_{PTD_m} a_0 \rho_{TD_m} a_0 \phi_3 + 1},$$

(4)

where $\rho_{SD_m} |s \phi_1$ represents the inter-user interference with $\phi_1 = \sum_{m=1}^{N_S} a_n + \sum_{m=1}^{N_S} a_n = \sum_{m=1}^{N_S} a_n + a_0 \phi_2 + \kappa_{SD_m}$, $\rho_{RD_m} a_0 \phi_2$ denotes the interference caused by $R$ in conjunction with $s \phi_1$.

$\phi_1$ is the azimuth angle [15].

$\phi_2$ is the azimuth angle [15].

$\phi_3$ is the azimuth angle [15].

$\phi_4$ is the azimuth angle [15].
\[ S \] is the signal transmitted by \( B \). ISAC NO-DLT framework

In the \(N\) to communicate with \( \phi \) echo, while the channels \( h_{RR} \) and \( h_{ST} \) are the scattering coefficient of the target by \( \mathcal{R} \) and \( \mathcal{S} \), respectively.

\[ \delta_{RR} = 1 + \kappa_{RD}^2, \quad \rho_{RT} \rho_{TD} \gamma_{R} \phi_{3} \] is the interference caused by the target scattering along with \( \phi_{3} = \delta_{RD}^2 + \kappa_{RD}^2 \). \( \gamma_{S} = P_{S}/N_{0}, \quad \gamma_{R} = P_{R}/N_{0}, \) and \( \varepsilon \in [0, 1] \) represents the level of SIC.

Similar to (3), the signal received at \( R \) can be expressed as

\[ y_{R}^{C} = h_{TR}^{R} \delta_{RR} x_{R} + \eta_{RR} + \sum_{S \neq R} h_{SR} x_{S} + \eta_{SR} + n_{R}^{C} + \frac{P_{S}}{N_{0}} \]

(5)

where \( \eta_{RR} \sim \mathcal{CN}(0, \kappa_{RR}^2 P_{R}) \) and \( \eta_{SR} \sim \mathcal{CN}(0, \kappa_{SR}^2 P_{S}) \) are the aggregated RHI caused by \( R \rightarrow T \rightarrow R \) and \( S \rightarrow R \). \( \delta_{RR} \) is the scattering coefficient of the target by \( R \rightarrow T \rightarrow R \), and \( n_{R}^{C} \sim \mathcal{CN}(0, N_{0}) \) denotes the AWGN at \( R \).

B. ISAC NO-DLT framework

As seen from Fig. 1(b) for the ISAC NO-DLT framework, the signal transmitted by \( S \) is capable of simultaneously sup-\nporting C&S. Explicitly, \( S \) transmits \( x_{S} = \sum_{n=1}^{N} a_{n} P_{S} x_{n} \) to communicate with \( N \) IoT devices and concomitantly sense the non-cooperative targets. Therefore, the signal received at the \( m \)-th IoT device can be expressed as

\[ y_{Dm}^{1} = \sum_{t=1}^{N} h_{TDm} h_{ST}^{t} (\delta_{SDm} x_{S} + \eta_{SDm}) + h_{SDm} (x_{S} + \eta_{SDm}) + n_{Dm}^{1} \]

(6)

where \( \delta_{SDm} \) denotes the scattering coefficient of the target in \( S \rightarrow T \rightarrow D_{m} \) links, and \( n_{Dm}^{1} \sim \mathcal{CN}(0, N_{0}) \) is the AWGN at \( D_{m} \).

According to the NO-DLT protocol, the SINR at \( D_{m} \), \( m \leq j \leq N \), can be expressed as

\[ \gamma_{Dm,j} = \frac{\rho_{SDm} \gamma_{S} \phi_{1} + \rho_{TDm} \rho_{ST} \gamma_{S} \phi_{4} + 1}{\rho_{SDm} \gamma_{S} \phi_{1} + \rho_{TDm} \rho_{ST} \gamma_{S} \phi_{4} + 1}, \]

(7)

where we have \( \phi_{4} = N (\delta_{SDm}^2 + \kappa_{SDm}^2) \).

The signal received at \( S \) can be formulated as

\[ y_{S} = \sum_{t=1}^{N} h_{TS} h_{ST} (\delta_{RR} x_{S} + \eta_{SS}) + h_{SS} \sqrt{P_{S} x_{S}} + n_{S}^{1} \]

(8)

where \( \delta_{SS} \) represents the aggregated RHI and scattering coefficient of the target caused by \( S \rightarrow T \rightarrow S \), respectively.

\[ \eta_{SS} \sim \mathcal{CN}(0, \kappa_{SS}^2 P_{S}) \]

III. COMMUNICATION PERFORMANCE ANALYSIS

In this section, both the exact and asymptotic OPs are analyzed to evaluate the communication performance.

A. Exact OP

1) SCC NO-DLT framework: The outage event occurs at the \( m \)-th IoT device, when the expected signal of \( D_{j} \) cannot be decoded by \( D_{m} \), \( m \leq j \leq N \). Thus, the OP of \( D_{m} \) for the SCC framework can be expressed as

\[ P_{Dm}^{C} = 1 - Pr \left( \gamma_{Dm,m}^{1} > \gamma_{thm}, \ldots, \gamma_{Dm,m}^{1} > \gamma_{thm} \right), \]

(9)

where \( \gamma_{thj} \) denotes the outage threshold SINR of \( D_{j} \). Based on the above, we derive the exact OP expression \( P_{Dm}^{C} \) for the SCC NO-DLT framework in the following theorem.

Theorem 1. For the SCC framework, \( P_{Dm}^{C} \) is given by

\[ P_{Dm}^{C} = 1 - \frac{N_{S} - N + m}{N_{S} - N + m} \left( 1 + \frac{1}{q} \right) e^{-\frac{\beta_{SDm}}{q} \gamma_{S} \phi_{4} \xi^{*} \phi_{4} + 1} \]

(10)

where \( \xi = \gamma_{thj} / \gamma_{S} (a_{j} - \gamma_{thj} \phi_{1}) \) with \( a_{j} > \gamma_{thj} \phi_{1}, \) otherwise \( P_{Dm}^{C} = 1 \).

Proof. See Appendix.

2) ISAC NO-DLT framework: Similar to (9), the OP of \( D_{m} \) for the ISAC NO-DLT framework is described as

\[ P_{Dm}^{I} = 1 - Pr \left( \gamma_{Dm,m}^{1} > \gamma_{thm}, \ldots, \gamma_{Dm,m}^{1} > \gamma_{thm} \right). \]

(11)

Based on the above, we derive the exact OP expression \( P_{Dm}^{I} \) for the ISAC NO-DLT framework in the following theorem.

Theorem 2. For the ISAC framework, \( P_{Dm}^{I} \) is given by

\[ P_{Dm}^{I} = 1 - \sum_{q=1}^{N_{S} - N + m} \left( 1 + \frac{1}{q} \right) e^{-\frac{\beta_{SDm}}{q} \gamma_{S} \phi_{4} \xi^{*} \phi_{4} + 1} \]

(12)

where \( \phi_{4}^{*} = \max_{m \leq j \leq N} \phi_{j} \), \( \xi = \gamma_{S} \phi_{4}^{*} \), and \( \phi_{j} = \gamma_{thj} / (a_{j} - \gamma_{thj} \phi_{1}) \) with \( a_{j} > \gamma_{thj} \phi_{1}, \) otherwise \( P_{Dm}^{I} = 1 \).
Proof. Similar to Theorem 1, substituting (7) into (11), (12) can be obtained after some mathematical manipulations. □

B. Asymptotic OP

1) SCC NO-DLT framework: When \(\gamma_S, \gamma_R \to \infty\), the SINR of \(x_j\) at \(D_m\) can be rewritten according to (4) as

\[
\gamma_{D_m-j} = \frac{\rho_{SD_m}\phi_1 + \rho_{RD_m}\phi_2 + \rho_{RT\phi_TD_m}\phi_3}{\rho_{SD_m}\phi_1 + \rho_{ST\phi_TD_m}\phi_4}.
\]

(13)

The asymptotic OP expression of \(D_m\) for the SCC NO-DLT framework is given in the following corollary.

Corollary 1. The asymptotic analytical OP expression of \(D_m\) for the SCC NO-DLT framework is given by

\[
P_{D_m,\infty}^{C} = 1 - \sum_{t=1}^{N_S-N+m} \left( N_S-N+m \right) \left( -1 \right)^{t+1} e^{\frac{\delta_{SD_m}}{\rho_{ST\phi_TD_m}}} (\beta_{SD_m} + \beta_{ST\phi_TD_m})
\]

\[
\times \left\{ \frac{\beta_{RT\phi_TD_m}}{\rho_{SD_m},\phi_1 + \rho_{ST\phi_TD_m},\phi_4} \right\}.
\]

(14)

where we have \(\varphi_{m,\infty} = \max_{m \leq j < N} \varphi_j,\infty\) with \(\varphi_j,\infty = \varphi_j\gamma_S\).

Proof. Substituting (13) into (9), similar to Theorem 1, (14) can be obtained after some mathematical manipulations. □

2) ISAC NO-DLT framework: When \(\gamma_S, \gamma_R \to \infty\), the SINR of \(x_j\) at \(D_m\) can be rewritten according to (7) as

\[
\gamma_{D_m-j} = \frac{\rho_{SD_m}\phi_1 + \rho_{ST\phi_TD_m}\phi_4}{\rho_{SD_m}\phi_1 + \rho_{ST\phi_TD_m}\phi_4}.
\]

(15)

The asymptotic expression OP of \(D_m\) for the ISAC NO-DLT framework is given in the following corollary.

Corollary 2. The asymptotic analytical OP expression of \(D_m\) for the ISAC NO-DLT framework is expressed as

\[
P_{D_m,\infty}^{I} = 1 - \sum_{q=1}^{N_S-N+m} \left( N_S-N+m \right) \left( -1 \right)^{q+1} e^{\frac{\delta_{SD_m}}{\rho_{ST\phi_TD_m}}} (\beta_{SD_m} + \beta_{ST\phi_TD_m})
\]

\[
\times \left\{ \frac{\beta_{SD_m}}{\rho_{SD_m},\phi_1 + \rho_{ST\phi_TD_m},\phi_4} \right\}.
\]

(16)

Remark 1. It should be pointed out that the outage performance of the IoT devices is related to the target attributes. If the target is absent, both the SCC and ISAC NO-DLT frameworks degenerate into the traditional NO-DLT frameworks. Moreover, if \(\kappa_{SD_m} = \kappa_{ST\phi_TD_m} = \kappa_{RR} = \kappa_{SS} = \kappa_{SD} = \kappa_{ST\phi_TD_m} = \varepsilon = 0\) is established, the frameworks reduce to the ideal condition.

IV. SENSING PERFORMANCE ANALYSIS

For sensing, the fundamental tasks are target detection and parameter estimation. Specifically, target detection implies that the sensing RXs determine whether the target is present or not based on the power of the received signals. By contrast, parameter estimation is used to judge the target attributes, such as velocity and distance, based on the delay or Doppler frequency of the received signals. Hence, target detection is the premise of parameters estimation. Therefore, in this paper, we focus our attention on the problem of target detection. During the signal processing, due to the noise and the interference from \(S\), the sensing RXs may make wrong decisions, which is reflected in the sensing performance of the system. Indeed, typically the PoD and probability of false alarm (PoFA) are used to jointly describe the sensing performance [15]. The PoD \(P_d\) and the PoFA \(P_{fa}\) represent the probability of the sensing equipment making a correct or incorrect decisions when the target is within the sensing range. Specifically, \(P_d\) and \(P_{fa}\) are respectively given as \(P_d = \Pr (H_1 | H_1)\) and \(P_{fa} = \Pr (H_1 | H_0)\), where the alternative hypothesis \(H_1\) refers to the receive signals containing echoes, noise, and interference. By contrast, the null hypothesis \(H_0\) indicates that the receive signals do not contain echoes [15]. Based on this, we now study the sensing behavior by deriving the PoDs for the SCC and ISAC NO-DLT frameworks.

A. SCC NO-DLT framework

For the SCC NO-DLT framework, we assume that the null hypothesis \(H_0^C\) only includes the AWGN and the interference from \(S\), while the alternative hypothesis \(H_1^C\) additionally includes the target echoes. According to (5), the binary hypothesis sensing problem may be described as

\[
H_0^C: y_R^C = \sum_{i=1}^{N_S} h_{SR}(x_S + \eta_{SR}) + n_R^C,
\]

(17)

\[
H_1^C: y_R^C = h_{TR} h_{RT}(\delta_{RR} x_R + \eta_{RR}) + \sum_{i=1}^{N_S} h_{SR}(x_S + \eta_{SR}) + n_R^C.
\]

(18)

From (17), we can observe that the power received at \(R\) is

\[
\mathbb{E} \left[ |y_R^C|^2 \right] = \rho_{RT\phi_TR}(\delta_{RR}^2 + \kappa_{RR}^2) + N_{\rho_{SR}}(1 + \kappa_{SR}^2) P_S + N_0.
\]

(19)

Then, the information related to the target for the SCC framework may be inferred as

\[
\rho_{RT\phi_TR}(\delta_{RR}^2 + \kappa_{RR}^2) P_R = \mathbb{E} \left[ |y_R^C|^2 \right] - N_{\rho_{SR}}(1 + \kappa_{SR}^2) P_S - N_0.
\]

(20)

It should be noted that the noise and channels in (19) obey the Gaussian and Rayleigh distributions under statistical theory. However, due to the randomness of the instantaneous noise and channels, it is possible to make the wrong decision by the sensing RX. Therefore, according to the Neyman-Pearson criterion [15], the PoD may be maximized at a fixed PoFA. Based on (17), we can confirm that \(\mathbb{E} \left[ |y_R^C|^2 \right]\) is the non-central Chi-square random variable with a degree-of-freedom (DoF) which is three under \(H_0^C\) and five under \(H_1^C\). Hence, the false-alarm and sensing probabilities of \(R\) are respectively expressed as

\[
P_{fa}^C = Q_\frac{\sqrt{2N_{\rho_{SR}}P_S(1 + \kappa_{SR}^2) / N_0}}{\sqrt{\Delta}},
\]

(20)

\[
P_d^C = Q_\frac{\sqrt{2(\rho_{RT\phi_TR} \delta_{RR}^2 P_R + N_{\rho_{SR}} P_S) / P_n}}{\sqrt{\Delta}},
\]

(21)

where we have \(P_n^C = \rho_{RT\phi_TR} \delta_{RR}^2 P_R + N_{\rho_{SR}} \kappa_{SR}^2 P_S + N_0\), \(\zeta\) denotes the sensing threshold and \(Q(\cdot, \cdot)\) is the Marcum Q-function.
B. ISAC NO-DLT framework

For the ISAC NO-DLT framework, assuming that the null hypothesis $H_0^1$ only includes the noise and SI from $S$, while the alternative hypothesis $H_1^1$ is additionally contaminated by the target echoes. According to (8), we arrive at

\[ H_0^1: y_S^1 = \sum h_{ST} \sqrt{P_{ST}S} + n_1 \]

\[ H_1^1: y_S^1 = \sum h_{ST} \sqrt{P_{ST}S} + \eta + n_1 \]

Similarly to (19), the information related to the target in the ISAC framework may be expressed as

\[ N_{PST} \rho S \sum (\delta^2_S + \kappa^2_S) P_S = \mathbb{E} \left[ \| y_S^1 \|^2 \right] - N_{PL} P_S - N_0. \]  

(23)

From (22), we can observe that the receive power of $S$ is a non-central Chi-square random variable with a DoF which is two under $H_0^1$ and four under $H_1^1$. Thus, the false-alarm and the correct PoD of $S$ are respectively expressed as

\[ P_{fa}^1 = Q_1 \left( \sqrt{2 \frac{N_{PS} P_{ST} S}{N_0}} \sqrt{2 \zeta / N_0} \right), \]  

(24)

\[ P_{du}^1 = Q_2 \left( \sqrt{2 \frac{N_{PS} (\delta^2_S + \rho S)}{P_{ST}^2}} \| y_S^1 \|^2 \right), \]  

(25)

where we have $P_{du}^1 = N \kappa^2_S \rho P_{ST} S + N_0$.

Remark 2. From (20), (21), (24), and (25), we can observe that for a given false-alarm probability, there exists a unique $\zeta$ corresponding to it, and then the PoD can be obtained. Additionally, it can be concluded that the PoFA is caused by the combination of the AWGN and the interference imposed by the communication TX on the sensing RX.

V. NUMERICAL RESULTS

In this section, some numerical results are provided for characterizing the accuracy of the theoretical analysis by Monte-Carlo computer simulations. According to the third generation partnership (3GPP) LTE-advanced [16] and practical communication scenarios, e.g., vehicle-to-vehicle communication in urban environment, and low-altitude unmanned aerial vehicle communication are used, which have the simulation parameters of Table II.\(^6\)

**TABLE II: Simulation parameters**

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
<th>Parameter</th>
<th>Value</th>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>$\kappa$</td>
<td>0.05</td>
<td>$\varepsilon$</td>
<td>0.01</td>
<td>$N_0$</td>
<td>1</td>
</tr>
<tr>
<td>$\gamma_{\alpha_1}$</td>
<td>2</td>
<td>$\alpha_1$</td>
<td>0.3</td>
<td>$\alpha_2$</td>
<td>0.7</td>
</tr>
<tr>
<td>$\delta_{\alpha R_{D}}$, $\delta_{\alpha D_n}$</td>
<td>2</td>
<td>$\Omega$</td>
<td>1</td>
<td></td>
<td></td>
</tr>
<tr>
<td>$d_{SR}$, $d_{D_1}$</td>
<td>10 m</td>
<td>$d_{SR_2}$, $d_{R_1}$</td>
<td>15 m</td>
<td>$d_{RT}$, $d_{TR}$</td>
<td>15 m</td>
</tr>
<tr>
<td>$d_{ST}$, $d_{R_2}$</td>
<td>15 m</td>
<td>$d_{R_1}$, $d_{D_2}$</td>
<td>20 m</td>
<td>$d_{TR_2}$</td>
<td>25 m</td>
</tr>
</tbody>
</table>

Fig. 2 illustrates the OPs of the IoT devices versus the transmit power of the communication station both for the ISAC and SCC frameworks at the same power consumption. The perfect match between our Monte Carlo simulation results and the theoretical values confirms the accuracy of the theoretical derivation in Section III. Observe from Fig. 2 that the proposed NO-DLT frameworks deliver exceptional outage performance compared to the traditional O-DLT. It can be further observed that the ISAC NO-DLT framework outperforms the SCC NO-DLT framework. This is because for the SCC framework, the direct irradiation of the sensing station may impose strong interference on the IoT devices, while the ISAC framework avoids this phenomenon. Additionally, we can also observe that the OPs of the IoT devices are bounded by the non-ideal factors of RHI, ipSIC as well as the interference among the equipment.

Fig. 3 presents the OPs of the IoT devices versus the RHI and ipSIC at $P_S = P_R = 30$ dBm and $P_S = 60$ dBm for the SCC and ISAC frameworks, respectively. It can be seen that both the ipSIC and RHI degrade the outage performance of the IoT devices, and that the OPs are more sensitive to the performance of SiC than to the RHI. Furthermore, since the signal detection at the farthest IoT device is independent of the SIC performance, the variations of the ipSIC parameters have no effect on the OP of $D_N$.

Fig. 4(a) shows the PoD versus the transmit power of the sensing station at $P_{fa} = 10^{-6}$. We plot the PoD curves for both the SCC and ISAC frameworks of three scenarios: 1) Fixed $P_S = 20$ dBm; 2) Fixed $P_S = 0$; 3) $P_S = P_R$. Due to the absence of the interference from the communication station, the sensing performance of the ISAC framework is superior to the SCC framework. Moreover, when $P_S = 0$, the SCC framework transforms into a pure sensing framework, and the sensing capabilities of the SCC and ISAC frameworks are close to each other. Fig. 4(b) depicts the PoD versus the
false-alarm probability of the sensing station. In other words, the sensing behavior is represented by the receiver’s operating characteristic (ROC) curves. For the SCC framework, the interference arriving from the communication station is absent when $P_S^c = 0$. In this case, the area under the curve (AUC) reaches its maximum and the framework has excellent sensing performance. When $P_R = P_S$, the sensing performance is greatly eroded by the communication station, resulting in an almost linear relationship between $P_d$ and $P_{fa}$. Observe for the ISAC framework that the AUCs remain almost constant vs. the RHI and vs. the transmit power of the communication station, which indicates that the ISAC framework exhibits excellent robustness. Finally, as expected, the RHI degrades the sensing performance of the frameworks.

VI. CONCLUSIONS

In this paper, the performance of the SCC and ISAC NOMA DLT TAS frameworks was investigated from S&C. The exact and the asymptotic OPs and the PoD were derived in the context of RHI and ipSIC. The simulation results showed that the ISAC framework is superior to the SCC framework in terms of its outage and sensing performance at the same power consumption. Naturally, the RHI degrades both the SCC and ISAC performance.

APPENDIX: PROOF OF THEOREM 1

In the following, we first derive the cumulative distribution function (CDF) of the channel gains between the selected antenna and $D_m$.

By using the order statistics method, the CDF of $\rho_{SD_m}$ is given by

$$F_{\rho_{SD_m}}(x) = \sum_{l=0}^{N_S-m} \binom{N_S - m}{l} (-1)^l e^{-\frac{lx}{\sigma_{SD_m}}}. \quad (A.1)$$

Substituting (4) into (9), after some mathematical manipulations, the OP of $D_m$ may be formulated as

$$P_{D_m}^1 = 1 - \Pr(\rho_{SD_m} > \rho_{SD_m}^* | \rho_{RD_m} \gamma_R \phi_2 + \rho_{RT} \rho_{TD_m} \gamma_R \phi_3 + 1)$$

$$= 1 - \int_0^\infty f_{\rho_{SD_m}}(w) \int_0^{\rho_{SD_m}^*(w)} f_{\rho_{RD_m}}(x)$$

$$\times \int_0^{x} f_{\rho_{RT}}(y) \int_0^{\rho_{TD_m}^*(y)} f_{\rho_{TD_m}}(z) dw dx dy dz. \quad (A.2)$$

By using the probability density function of the Rayleigh fading channels, (A.2) may be further expressed as

$$P_{D_m}^1 = 1 - \int_0^\infty \int_0^{\rho_{SD_m}^*(w)} \int_0^{x} \int_0^{\rho_{TD_m}^*(y)} e^{-\frac{\beta_{RD_m} \beta_{RT} \gamma_{TD_m}}{\rho_{RT} \rho_{TD_m}}}$$

$$\times F_{\rho_{SD_m}}(\rho_{SD_m}^* \gamma_R \phi_2 + y \gamma_R \phi_3 + 1)) dx dy dz. \quad (A.3)$$

With the aid of (A.1) and [18, Eq. (3.462.15)], (10) can be obtained after some further mathematical manipulations.

REFERENCES