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Joint Design and Operation of Shared Spectrum Access for Radar and Communications

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***Abstract*—A new theoretical foundation for the joint design and operation (JDO) of shared spectrum access for radar and communications (SSPARC) is presented. The JDO SSPARC framework entails advanced radar-comms channel estimation, along with a real-time adaptive space-time transmit and receive optimization procedure to maximize forward channel signal-to-noise while simultaneously minimizing co-channel interference. Additionally, a new expression for radar capacity is introduced that when combined with traditional communication capacity provides a unified measure of the total capacity of the combined radar-comms network. High fidelity site-specific electromagnetic radiation propagation simulations are conducted to provide a sense of the real-world potential gains achievable with JDO SSPARC as compared with non-optimized approaches.**

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I. Introduction

The increasing global demand for useful radio frequency (RF) spectrum by wireless communications has led to a conflict with traditional primary users of legacy spectrum such as radar. This has prompted new research into new approaches to potentially sharing spectrum between radar and communications. To that end, the Defense Advanced Research Projects Agency (DARPA) has created a major new project titled Shared Spectrum Access for Radar and Communications (SSPARC) [1]. The overall scope of the SSPARC program includes techniques that may be incorporated by existing systems with minimal changes, to the most advanced spectrum sharing approaches that entail a "back to the drawing board" approach. The JDO SSPARC approach described herein is an example of the latter. An additional goal of the DARPA SSPARC program is to not just share spectrum, but use spectrum in a much more efficient way that results in improved performance for both radar and communications.

To that end, this paper introduces a new and fundamental approach to both the joint design *and* operation of a radar-comms (RC) SSPARC network. In Section II, a new expression for the radar "capacity" is derived that allows for a joint radar-comms capacity measure for the entire network. In Section III, a mathematical framework for the joint optimization of an *N*-node RC network is developed and is shown to result in generalized eigenvalue problem for the unconstrained case, or a constrained Rayleigh quotient optimization for the general constrained optimization case. The framework makes clear the need for, and dependency on, multidimensional channel information. To that end, in Section IV, a comprehensive framework for advanced JDO SSPARC channel estimation is outlined that builds on advances in both knowledge-aided (KA) processing, and the latest cognitive radar/comms technologies [2-5]. Lastly, in Section V, a high fidelity and site-specific electromagnetic (EM) propagation simulation and analysis is conducted to gauge the potential performance improvements achievable in a realistic setting as compared to conventional (non-optimized) approaches.

II. A New Measure of Radar Capacity

Radar is most assuredly a special case of a communication system. So it stands to reason that a measure of capacity directly analogous to that developed by Hartley and Shannon should not be too much of a stretch [6]. Though Woodward's landmark book examined radar from an information theoretic framework (and introduced the ambiguity function along the way!), it did not develop a measure of capacity directly analogous to that of comms [7]. Rather, a general measure of the information gain from a measurement was developed. In this section an intuitively appealing and relatively simple measure of radar capacity is introduced.

For an MTI radar, one can consider each independent resolution cell (e.g., range-angle-Doppler) as a binary information storage unit, i.e., "0" = target absent, "1" = target present. Viewed in this manner, the maximum capacity of an MTI radar performing a periodic search is given by the Hartley capacity measure [6]:

$$C_R = \log N \tag{1}$$

where N is the total number of independent MTI resolution cells with a minimum prescribed signal-to-noise-ratio (SNR) for an assumed target RCS, denoted as $\mathrm{SNR}_{\min}$, and is given by

$$N \propto \left(BR_{\max}\right)\left(\frac{2\pi}{\Delta\theta}\right)\left(\frac{\mathrm{PRF}}{\Delta f_D}\right) \tag{2}$$

where where $\Delta\theta$ and Δf_D denote the angle (e.g., azimuth) and Doppler resolution respectively, B is the operating bandwidth, and $R_{\max}$ is the maximum range with SNR equal to $\mathrm{SNR}_{\min}$, and can be derived from the so-called radar range equation [8]

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(consistent units assumed). Note that in (2) we have assumed the radar scans 360-deg azimuthal (2π radians). This can obviously be modified to suit the specific application. $R_{\max}$ can in turn be expressed in terms of $\text{SNR}_{\min}$ and the total noise σ_n^2, namely,

$$R_{\max} = \frac{\kappa}{\sqrt{\sigma_n} \cdot \sqrt[4]{\text{SNR}_{\min}}} \tag{3}$$

where κ is a fixed constant of proportionality that subsumes all of the radar range equation parameters (antenna gain, losses, etc.). Substituting (3) into (2) then into (1) yields the final basic MTI radar capacity expression

$$C_R = \frac{1}{T_s}\left(\frac{\kappa B}{\sqrt{\sigma_n} \cdot \sqrt[4]{\text{SNR}_{\min}}}\left(\frac{2\pi}{\Delta\theta}\right)\left(\frac{\text{PRF}}{\Delta f_D}\right)\right) \tag{4}$$

where T_s denotes the scan or update time so that the capacity is in the conventional "bit rate" form [6]. Note that for a given application, $\text{SNR}_{\min}$ is specified ahead of time, and does not usually vary during operation. Typical values for an MTI radarfor $\text{SNR}_{\min}$ would be between 10 and 15 dB depending on acceptable false alarm levels, for a given assumed target RCS.

Note that an analogous expression for radar capacity can be readily formulated for synthetic aperture radar (SAR) by simply recognizing that the equivalent of a "bit" is a SAR "pixel"—again the minimum independent resolution cell with a prescribed minimum SNR (sometimes referred to as terrain-to-noise ratio (TNR)). Additionally, the most general radar "bit" for either MTI or SAR is inherently an "M-ary" versus binary piece of information since the amplitude (and possibly polarization, etc.) contains additional information that can be very useful such as target ID in MTI, or contrast in SAR. Obviously for the M-ary case, there is an increase in capacity relative to the binary case considered in (4).

Finally for the case where both radar and communications are operating simultaneously the combined expression for the total capacity is given by

$$C_{\text{tot}} = \alpha \frac{1}{T_s}\left(\frac{\kappa B_r}{\sqrt{\sigma_n} \cdot \sqrt[4]{\text{SNR}_{\min}}}\left(\frac{2\pi}{\Delta\theta}\right)\left(\frac{\text{PRF}}{\Delta f_D}\right)\right) + \beta \cdot B_c \log\left(1 + \text{SNR}_c\right) \tag{5}$$

where B_r and B_c denote the radar and communication bandwidths respectively, and SNR_c denotes the additive Gaussian white noise (AGWN) communications channel SNR. For the assumed co-channel interference model, the coupling between radar and communications occurs in the σ_n term for radar, and the SNR_c term for communications. Thus, in a co-design environment where the cross coupling terms would depend on operating powers, bandwidths, and relative locations and terrain, one would seek to jointly maximize capacity while maintain a minimally acceptable level of radar and communications performance. If either or both the radar and communication systems are capable of performing *both* radar and communications functions then the joint capacity expression in (5) contains and even greater degree of coupling than just the co-channel noise.

In the next section we introduce a theoretical framework for optimizing the transmit-receive functions of an N-node RF SSPARC network. The resulting design equations provide a tight bound on theoretical performance (as opposed to generally approximate methods such as Cramer-Rao [9]).

III. Theory of JDO SSPARC

Consider a notional geographically distributed radar-comms (RC) scenario in which there is significant potential for co-channel interference (see Fig. 1). If the available and limited RF spectrum is to be shared by all parties in a dynamic fashion, knowledge and adaptive functionality will be required to avoid highly sub-optimal performance. For any RC node (radar or wireless communication device), at any given moment, there is a set of desired channels (target channel for radar, forward link for comms), and other channels that are not of interest which might include radar clutter, and other co-channel links.

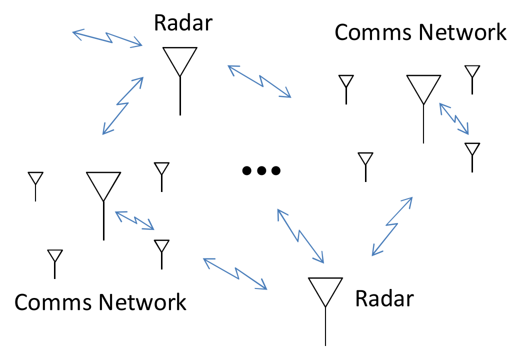

Fig. 1. Notional combined multi-radar and multi-comms network (RC) system

We begin with the following set of definitions:

$$\begin{aligned} H_{ij} &= \text{transfer function from the } j\text{-th to the } i\text{-th nodes.} \\ \mathbf{s}_j &= \text{signal transmitted by } j\text{-th node} \\ \mathbf{y}_i &= \text{recieved signal at } i\text{-th node} \\ \mathbf{n}_i &= \text{additive receive noise at } i\text{-th node} \end{aligned} \tag{6}$$

The dimensionality of the above variables depends on the choice of degrees-of-freedom (DoFs) to be included in the optimization. For example, if each node had say L spatial DoFs on both transmit and receive, the vectors in (6) would be L dimensional and complex valued, and the constituent transfer matrices would be $L \times L$ (and complex valued). We first consider the above model over a short period of time for which the channels can assumed to be non-varying, though they can still be stochastic. In this way we can arrive at a tight theoretical performance bound as discussed below.

Interpretation of the constituent block matrices $\{H_{ij}\}$ is based on whether the nodes are radar or comms. For example, it is convenient to define the submatrices with equal indices as follows:

$$H_{ii} = \varnothing, \text{ for comms nodes}$$
$$H_{ii} \neq \varnothing, \text{ for monostatic radar nodes} \tag{7}$$

where $\varnothing$ denotes the null (zero) matrix or vector. In other words, a comms node does not radiate to itself. However, for a monostatic MTI radar $\{H_{ii}\}$ represents the total signal-dependent radar channel that in general consists of targets and clutter. More specifically,

$$H_{ii} = \sum_{n=1}^{N_{\mathrm{TOT}}} H_{T_{n,i}} + H_{C_i} \tag{8}$$

where $\{H_{T_{n,i}}\}$ and $\{H_{C_i}\}$ denote the target and clutter channels for the i^{th} radar.
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With the above definitions, the entire composite RC channel matrix is given by

$$H = \begin{bmatrix} H_{11} & H_{12} & \cdots & H_{1N} \\ H_{21} & H_{22} & & \\ \vdots & & \ddots & \vdots \\ H_{N1} & \cdots & & H_{NN} \end{bmatrix} \tag{9}$$

If channel reciprocity is assumed, then $H_{ij} = H_{ji}$ (symmetric but not Hermitian). Moreover, some number of diagonal and off diagonal submatrices will be null. As mentioned previously, for comms nodes, the corresponding diagonal submatrices will be null. Also, for monostatic radar, the "reverse" channel is also null. As with co-channel comms nodes, it is assumed that radar nodes can interfere with each other and other comms nodes.

The total composite transmit signal defined in (6) has the structure

$$\mathbf{s} = \begin{bmatrix} \mathbf{s}_1 \\ \mathbf{s}_2 \\ \vdots \\ \mathbf{s}_N \end{bmatrix} \tag{10}$$

where $\mathbf{s}_i$ denotes the spatio-temporal transmit signal from the i^{th} RF node. Note some nodes may be receive only, or simply not transmitting at any given moment, in which case the corresponding entries are the null vector.

For both radar and comms there are generally both "forward," or desired channels, and "co-channels" or undesired propagation channels. A fundamental goal of JDO SSPARC is to maximize transmission through forward channels, while simultaneously minimizing co-channel interference. Thus for the i^{th} transmit node, there can be as many as N forward channels if the i^{th} node is simultaneously acting as a monostatic radar and/or communications transmitter and/or multistatic radar (for which the other nodes are passive receivers). In general, the number of forward channels for the i^{th} transmit node will be less than N, say N_{FC_i}. Additionally, for the i^{th} transmit channel, there will be some number of co-channels (at most N-1) for which it is desired to minimize interference. The situation is depicted in Fig. 2.

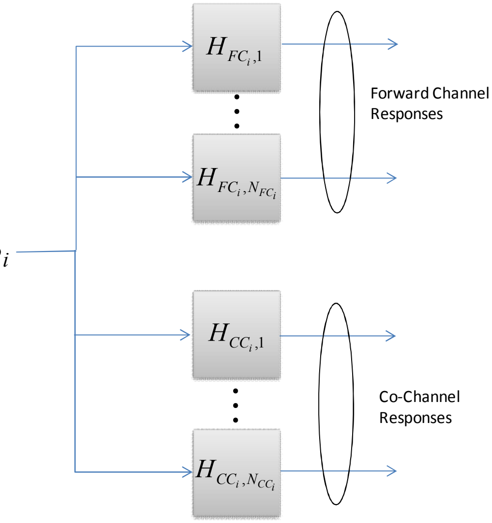

Fig. 2. Illustration of forward and co-channel signal paths.

Assuming for the moment that all channels have equal "importance" (relative weighting will be introduced later), one common design goal would be to maximize the signal power through the forward channels while simultaneously minimizing the response in the co-channels. For the unconstrained optimization case (other than finite norm) this leads to Rayleigh quotient maximization whose solution is a generalized eigenvector problem, i.e.,

$$\max_{\mathbf{s}_i} \frac{\mathrm{E}\left(\left\| H_{FC_i}\mathbf{s}_i \right\|^2\right)}{\mathrm{E}\left(\left\| H_{CC_i}\mathbf{s}_i \right\|^2\right)} = \max_{\mathbf{s}_i} \frac{\mathbf{s}_i' \,\mathrm{E}\left(H_{FC_i}' H_{FC_i}\right)\mathbf{s}_i}{\mathbf{s}_i' \,\mathrm{E}\left(H_{CC_i}' H_{CC_i}\right)\mathbf{s}_i} \tag{11}$$

where H_{FC_i}, H_{CC_i} denote the composite forward and co-channel transfer matrices respectively, $\|\ \|$ denotes the L_2 -norm, and $\mathrm{E}(\ )$ denotes the probabilistic expectation operator. For the N-node case, the composite forward and co-channel transfer functions H_{FC_i}, H_{CC_i} are given by (see Fig. 2)

$$H_{FC_i} = \sum_{n=1}^{N} w_{FC_i,n} H_{FC_i,n}$$
$$H_{CC_i} = \sum_{n=1}^{N} w_{CC_i,n} H_{CC_i,n} \tag{12}$$

where $H_{FC_i,n}, H_{CC_i,n}$ denote the forward and co-channel transfer matrices from the i^{th} transmit node to the n^{th} receive node, and $w_{FC_i,n}, w_{CC_i,n} \in [0,1]$ are relative weighting coefficients.

The solution to (11) is based on solving the following generalized eigenvalue problem

$$\mathrm{E}\left(H'_{FC_i} H_{FC_i}\right)\mathbf{s} = \lambda \mathrm{E}\left(H'_{CC_i} H_{CC_i}\right)\mathbf{s} \tag{13}$$

The eigenfunction solution with associated maximum eigenvalue yields the optimum space-time transmit signal. Note that if either (or both) of the matrices in (13) are nonsingular, the solution is that of an ordinary eigenvalue problem. It should also be noted that the kernel matrices in (13) are positive semi-definite or definite, and thus all eigenvalues are non-negative. 所有特征值是非负的

In general, additional constraints (beyond finite norm and length) on $\mathbf{s}$ may be imposed such as constant modulus, etc., that result in generally nonlinear programming problem of the form

$$\max_{\mathbf{s}_i} \frac{\mathrm{E}\left(\left\| H_{FC_i}\mathbf{s}_i \right\|^2_{W_{FC_i}}\right)}{\mathrm{E}\left(\left\| H_{CC_i}\mathbf{s}_i \right\|^2_{W_{CC_i}}\right)}$$

subject to: (14)

$$\mathbf{f}(\mathbf{s}) \geq \varnothing$$

One important special case is when the constraints are in the form of angular "keep-out" zones. This can arise for example if the locations of co-channel RF nodes are known or where there is a desire to achieve a low probability of intercept (LPI) capability. If we let $\mathbf{v}_1,\ldots,\mathbf{v}_J$, where $J < N$, denote the spatial steering vectors (narrowband) associated with the keep-out directions, (14) takes on the form

$$\max_{\mathbf{s}_i} \frac{\mathrm{E}\left(\left\| H_{FC_i}\mathbf{s}_i \right\|^2_{W_{FC_i}}\right)}{\mathrm{E}\left(\left\| H_{CC_i}\mathbf{s}_i \right\|^2_{W_{CC_i}}\right)}$$

subject to: (15)

$$V'\mathbf{s} = \varnothing$$

where the columns of V are the keep-out steering vectors. Fig. 3 shows an example for the case of one forward channel and two keep-out directions. The antenna consisted of a 16 element uniform linear array (ULA), with half-wavelength element spacing, operating from +/- 90-deg (+/- 0.5 normalized angle)) [5]. In practice, the beam-patterns would be far more complex due to non-line-of-sight (NLOS) propagation. 非视距传播

In general the constraints in (14) will be nonlinear and not amenable to analytical solution. One such example is a constant modulus constraint on the fast-time waveforms that allows for Class C (saturation) operation of the transmit amplifiers, and thus the most effective radiated power (ERP) [10]. Besides strictly numerical methods, one approximate solution to this that is asymptotically optimal is based on the method of stationary phase and nonlinear FM (NLFM) [11].

A. Optimal Receiver Structure

Once all RC *transmit* space-time waveforms have been designed, the optimal receiver functions can be designed. The reason the transmit waveforms are designed first is because they establish the colored noise component that must be accounted for in the generally colored noise matched filters [5, 9].

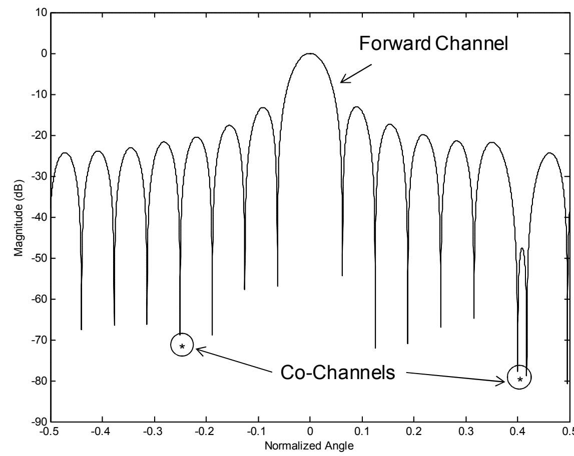

Fig. 3. Illustration of constrained optimum beamforming with co-channel "keep out" directions.

Although the basic theory for optimal reception is well established, JDO SSPARC will leverage the recent advances in knowledge-aided (KA), physics-based/model-based, and cognitive signal processing [4, 5, 12] to provide enhanced knowledge of the colored noise channel, represented by the space-time covariance matrix R_i for the i^{th} receive node. The associated optimal receiver weight vector $\mathbf{w}_i$ is given by (see [5, 9] for further details)

$$\mathbf{w}_i = R^{-1}\mathbf{s} \tag{16}$$

where $\mathbf{s}$ is a desired steering vector of interest.

IV. JDO SSPARC CHANNEL ESTIMATION

The previous section makes it clear that "*as goes channel knowledge, so goes performance*." Unlike the succinct governing design equations derived in the previous section, high-fidelity channel estimation has many facets and potential solutions, most of which are not mutually exclusive and can thus be combined into an overall effective approach. Thus we will first provide an overview summary of the different approaches, before delving into the mathematical and implementation details later in this section.

A. Overview of JDO SSPARC Channel Estimation

The following is a fairly comprehensive listing and brief description of adaptive channel estimation techniques that can be combined into the JDO SSPARC framework:

- Training Sequences: This is the classical approach used in wireless RF communication systems. A known signal is transmitted from one node to another. At the receive node and adaptive estimation problem is solved to derive the channel transfer function. Performance of this approach depends on a number of factors including the complexity of the training sequence, available computational receiver resources, and the degree of channel stationarity. The price paid is a reduction in overall capacity as compared to the case where the channel was known. In JDO SSPARC, this concept can easily be extended to allow for channel estimation between both comms, radar and comms-radar. This can also be combined with MIMO techniques and channel reciprocity (see below).

- MIMO: Spatial MIMO techniques can enhance the amount of information available by a channel estimator for a given training period. For example, instead of a single scalar transmit training waveform, several simultaneous training sequences can be transmitted from different antennas. A receiver would then use a bank of matched filters to reconstruct each transmit degree-of-freedom. An adaptive channel estimate that includes both spatial and temporal information can then be derived. This in turn can provide significant performance improvement over the aforementioned SISO approach.

- Reciprocity: Except for very pathological cases (e.g., "ducting"), Maxwell's equations dictate the law of reciprocity for electromagnetic propagation. That is the forward channel is equal to the time reversed reverse channel for the same transmit waveform. In theory this means that one forward channel training sequence can be used to estimate the reverse channel. In practice the main impediment is the generally non-reciprocal nature of the active electronics in the transmit-receiver chains. However, with the advent of "smart RF" electronics (see for example [13]), it is now possible to have well calibrated electronics that can support the exploitation of reciprocity.

- Knowledge-Aided (KA) and Cognitive RF (CRF): KA and CRF channel estimation methods are far and away the most sophisticated and difficult to implement, but can in turn yield the biggest benefits that are well documented in the literature [2-5, 10, 12, 14]. Though there are a number of KA/CRF variants, in JDO SSPARC we will assume the most general instantiation that includes:

  - High-Fidelity Environmental Dynamic Database (EDDB): Environmental databases are increasingly becoming the norm in normal operations. The information they contain can include digital terrain maps, land-cover land-use (LCLU), meteorological, background traffic tracks, and RF "terrain" databases such as the DARPA Radiomap [15]. It is further assumed that mechanisms exist for the continual updating of the database. This is a special case of the assumed supervised learning function of the KA/CRF system.

  - KA/CRF Real-Time HPEC Architecture: We assume the existence of a real-time high-performance embedded computing (HPEC) architecture that is capable of exploiting the aforementioned database in real-time and perform all necessary online calculations. The DARPA KASSPER project successfully developed and demonstrated several such HPEC architectures. A key enabler is the existence of a "look ahead" scheduler and operating system that uses basic kinematic assumptions and RF scheduler information to overcome the inevitable memory access latencies inherent in KA methods. KA/CRF HPEC architectures are well documented in the literature [2-5, 10, 12, 14].

  - Fully Adaptive RF: Lastly, it assumed that adaptivity exists both in the receiver and the transmitter. The full power of KA/CRF in general, and JDO SSPARC in particular, is only realized when both the Tx/Rx functions are adaptive [4, 5].

Although all of the above could be incorporated in a JDO SSPARC system, the specific instantiations will vary significantly based on scenario specifics, and available resources.

V. High-Fidelity JDO SSPARC Design Example

In this section, high-fidelity ray tracing software was utilized to better gauge the theoretical performance bounds developed in Section III. While there is little opportunity for significant performance gains in benign direct line-of-sight (DLOS) propagation environments, non LOS (NLOS) scenarios provide a rich opportunity for space-time adaptive waveforms and receiver processing.

A five (5) node JDO SSPARC scenario was modeled using Wireless InSite™ [16], a high-fidelity RF ray tracing software package. The nodes were distributed in a 3 km x 2.5 km set of terrain and building structures from an existing model of the city of Philadelphia, PA, USA. The laydown is shown in Fig. 4, along with a sample of the propagation paths from Node 1 to 4.

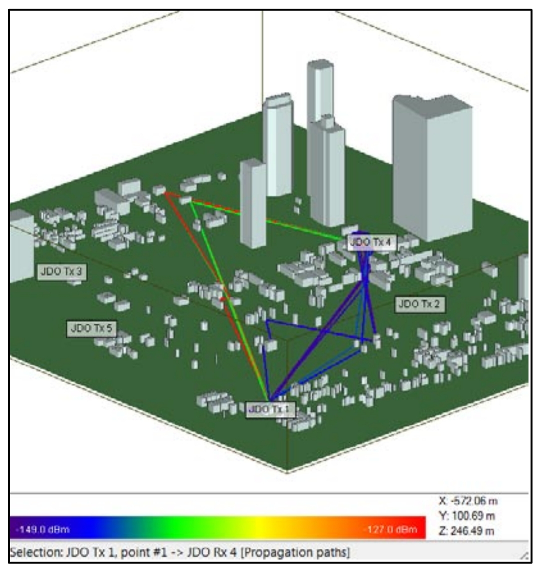

Fig. 4. View of the simulated urban scenario showing the laydown of the 5 RF nodes and propagation paths when Node 1 transmits and Node 4 receives.

Table I below contains a description of the settings used to predict the interference CIRs in the modeled urban scenario. These are the standard model parameters for the full 3D solver that we previously verified by reviewing the results against our expectations of RF propagation. Our team has previously validated WI's received power predictions against calibrated measurements for a suburban propagation environment in New Jersey.

Table I. Specific simulation parameters

Wireless InSite™ Solver Settings	Tx/Rx Node Configuration
• Solver = 3D Shooting and Bouncing Rays • Ray Angle Spacing = 0.18° • 6 reflections/path (max) • 1 diffraction/path (max) • 0 Transmissions (buildings absorb rays)	• Isotropic antenna (Max Gain = 0 dBi) • Antenna Height = 2 m above ground level • Waveform = Continuous Wave (CW) • Carrier Frequency = 3000 MHz

Fig. 5 contains the full set of unity power normalized pair-wise CIRs between the five modeled nodes. This result shows that within this relatively small scale urban scenario that most of the multipath components occur within 2 micro-seconds of the arrival of the first ray. It should be noted that the first arriving component is not necessarily the direct line-of-sight path because we have disabled the "transmission" through buildings to account for the relatively high absorption by building materials at 3 GHz.

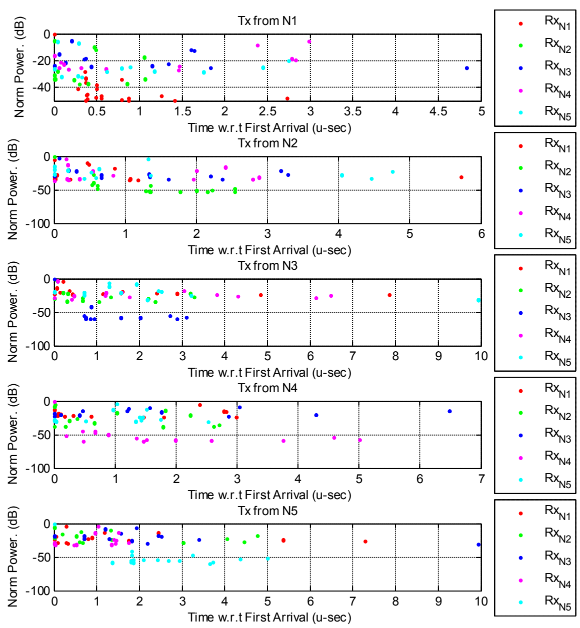

Fig. 5. Unity power normalized pair wise impulse responses between the five nodes.

To gauge the potential performance improvements possible using adaptive JDO SSPARC space-time waveforms, equation (13) was solved for a variety of transmit-receive scenarios. In each case significant eigenvalue spread was observed, often greater than 10 to 15 dB. This in turn implies there is significant opportunity for performance gains if space-time transmit-receive adaptation is performed via the JDO SSPARC design equations (or constrained variants thereof).

Fig. 6 shows a typical performance result for the case when Node 1 is "broadcasting" to Nodes 2 and 3, while simultaneously attempting to minimize its co-channel interference to Nodes 4 and 5. The 17 dB eigenvalue spread again indicates there is significant opportunity for space-time optimization.

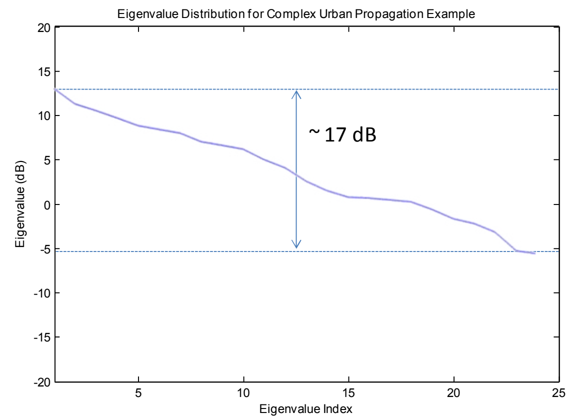

Fig. 6. Eigenvalue distribution for a particular wireless network topology. The significant spread indicates there is significant potential for performance enhancements using adaptive waveforms.

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