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KEYHOLES AND MIMO CHANNEL MODELLING

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ABSTRACT:
It has been shown that the concept of "keyholes" in MIMO channels could be responsible for a reduced transmit/receive diversity performance, and therefore a reduction of capacity with respect the maximum available radio link capacity. In the present work, we further analyse this effect in terms of coupling between an input wave impinging on the keyhole, and several output waves due to diffraction. As a consequence in terms of MIMO channels modelling within a small antenna size approximation, the channel transfer function matrix can be expressed in terms of steering matrices for the emitting and receiving arrays, and of a connecting matrix relating the complex wave amplitudes at the two antennas. The rectangular or non diagonal character of this matrix is due to junctions in the wave paths, which leads to a keyhole when the matrix is full. This approach also allows to take into account explicitly the double directional dependence of the radio scenario[1] , and lends itself easily to geometric based stochastic modelling of MIMO channels. The generalization to nonideal multiport antennas, incorporating coupling and enhanced correlations between sensors, is immediate.

I INTRODUCTION

Multiple input/multiple output (MIMO) radio communications based on transmit/receive diversity are attracting increasing interest, due to the enhanced capacity or radio link robustness they promise in comparison with conventional SISO or SIMO techniques[2] However the enhanced performance is crucially dependent on the existence of an appreciable diversity at both ends of the radio link. This is favored by a large angular spread of the multipath structure, translating into a small decorrelation length, and is best obtained in indoor environments due to the intricate complexity of multiple reflection-refraction-diffraction events experienced by the electromagnetic waves. However it has been shown on the basis of simple examples [3] , that large angular spreads were not sufficient to guarantee high capacity when "keyholes" were present between the emitter and the receiver. In essence the forced transmission of all waves through a single hole, or by a single electromagnetic mode, reduces the diversity to one, instead of a value equal to the number of waves or the number of sensors. In the present work, we further analyse this effect in terms of coupling between each incoming wave and all waves exiting the keyhole, because diffraction is responsible for radiation into a continuum of angles. This leads us to investigate the consequence of diffraction, or more generally of the presence of junctions and subpaths in radio propagation, on MIMO channel modelling.

II A SIMPLE EXAMPLE OF KEYHOLE

Let us consider a simple hypothetical example of keyhole, made of a slit in an otherwise perfectly absorbing plane of material, with an omnidirectional emitter (Tx) and an omnidirectional receiver (Rx) in the two half-spaces (fig. 1). Perfectly conducting top and bottom walls are also present, responsible for specular reflections. The Tx and Rx positions are chosen in such a way, and the width of the slit is sufficient, that three rays contribute to the radio link, i.e. the direct LOS one and two double reflected ones. It is trivial to compute the transfer functions $H(\omega)=\sum_{i=1}^{3} A_i \exp(-jkl_i)$ relating the voltage on the receiving antenna port to the voltage on the transmitting antenna port, where $k=\omega/c$ is the wave vector, l_i is the length of path *i* and A_i is the free space attenuation for this path. However when the slit is narrow enough, diffraction continuously spreads the exiting rays into a wide angle, with amplitudes decreasing away from the straight path. These amplitudes can be analytically approximated by physical optics (Kirchhoff approximation), neglecting the perturbation of the incoming wavefront by the screen. In this situation and also neglecting higher order reflections, Rx is illuminated by three sources (one real slit and two image slits) or equivalently three waves, each of them being the superposition of one main wave (LOS or specularly reflected) and of two diffracted waves (fig. 2).

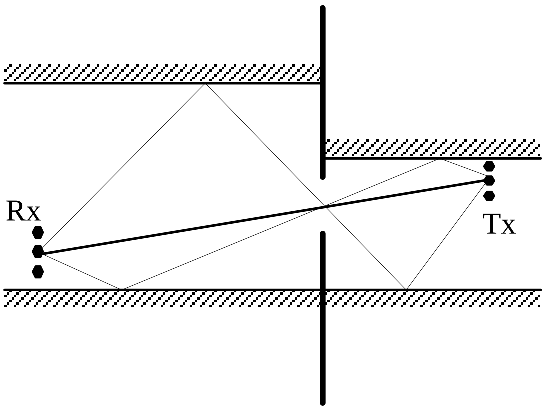

fig. 1: hypothetical scenario with 3 wave paths, one LOS and 2 double reflected ones.

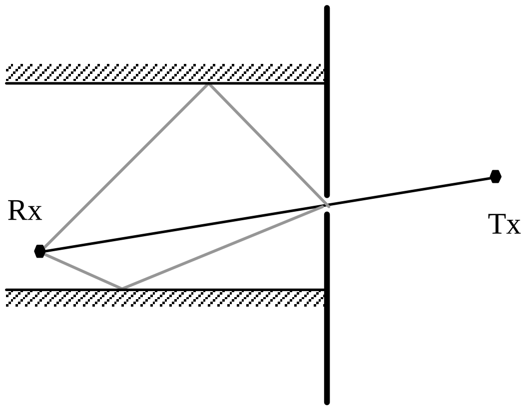

fig. 2: diffraction of one wave path into 3 sub-paths

The transfer function can now be written in this generalized case:
$H(\omega)=\sum_{i=1,\,j=1}^{3} R_i.T_j.K_{ij} \exp(-jkl_i)\exp(-jkl'_j)$, where R_i and T_j are now the free space attenuations for path *i* or *j* on the Rx or Tx sides respectively, and K_{ij} is the complex attenuation of a wave going through the slit computed in the Kirchhoff approximation.
Assuming now that the Rx and Tx radiators are only slightly displaced around a reference origin by the vectors $\vec{r}_r$ and $\vec{r}_t$, we have:

$$(1) \qquad H(\omega,\vec{r}_r,\vec{r}_t)=\sum_{i=1,\,j=1}^{3} R_i.T_j.K_{ij} \exp(-jkl_i)\exp(-jkl'_j)\exp(-j\vec{k}_i\cdot\vec{r}_r)\exp(-j\vec{k}_j\cdot\vec{r}_t)$$

with $\vec{k}_i$ and $\vec{k}_j$ the received and emitted wavevectors respectively.
The previous expression defines a MIMO channel transfer function matrix when $\vec{r}_r$ and $\vec{r}_t$ are vector positions of receiver and emitter uncoupled antenna radiators. It is straightforward to compute the capacity through the expression [2] $C=\log_2\left[\det\left(I_{n_r}+\frac{SNR}{n_t}H.H^+\right)\right]$, where n_r

and n_t are the number of receiving and transmitting radiators[1]. This expression being sensitive to small-scale fading need computation of a statistics of the capacity, e.g. by varying the multisensors antenna positions over a spatial area of a few wavelengths squared.

Let us take a simple numerically computed example for the sake of illustration: two identical triple linear arrays of sensors are used for the emitter and the receiver. When their inter-sensor separation is a full wavelength λ, corresponding to a good, and the slit width is 5λ, corresponding to a very small diffraction, the median capacity for 3 dB SNR is 3.54 b/s/Hz. On the other hand if the slit width is reduced to λ/4, the median capacity falls to 2.5 b/s/Hz. Finally further reducing the sensors separation to λ/10 yields an unchanged capacity with a median again of 2.5 b/s/Hz. As can be seen in table 1, giving the medians of the normalized channel transfer function singular values, the two latter cases are close to a single degree of freedom (SDF) channel. This is not surprising, as explained in [chizik]: when the slit is very narrow the channel transfer function can be factorised in the form

$$H = \begin{pmatrix} B_1 \\ B_2 \\ B_3 \end{pmatrix} K \begin{pmatrix} A_1 & A_2 & A_3 \end{pmatrix} = K \begin{pmatrix} A_1 B_1 & A_2 B_1 & A_3 B_1 \\ A_1 B_2 & A_2 B_2 & A_3 B_2 \\ A_1 B_3 & A_1 B_3 & A_3 B_3 \end{pmatrix}$$

where A_i and B_j are complex attenuations between the transmitting or receiving sensors and the slit, and K is the slit complex attenuation expressing the ratio of output to input electric fields[2]. Obviously, the two coefficients of lowest degree of the characteristic polynomial are nil, yielding two zero eigenvalues. This is irrespective of the correlation between sensors, although in practice a high correlation will further enhance the SDF character.

Assuming any two path attenuations are uncorrelated and taking any two entries of the channel matrix, we find for instance:
$E(A_1B_1A_3B_2) - E(A_1B_1)E(A_3B_2) = E(A_1)E(B_1)E(A_3)E(B_2) - E(A_1)E(B_1)E(A_3)E(B_2) = 0$,
which means that the matrix entries are uncorrelated [3] .

Iinterestingly the amplitudes of the channel coefficients are not all uncorrelated, e.g.:

$$E\left(|A_1B_1|.|A_1B_2|\right) - E\left(|A_1B_1|\right)E\left(|A_1B_2|\right) = E\left(|A_1|^2\right)E\left(|B_1|\right)E\left(|B_2|\right) - E\left(|A_1|\right)^2 E\left(|B_1|\right)E\left(|B_2|\right)$$
$$= E\left(|B_1|\right)E\left(|B_2|\right)\left(E\left(|A_1|^2\right) - E\left(|A_1|\right)^2\right) \neq 0$$

Nevertheless this amplitude correlation is not responsible for the SDF. This can be easily seen by some thought experiment: let us artificially randomise the phase of just one of the entries of the channel matrix. Instead of –47, -28 and +9.5 dB, we now find –40, -3.9 and +9.4 dB. In other words, the channel has turned into a double degree of freedom (DDF), although the amplitudes are still correlated. This is due to the fact that although the determinant is still zero, this is no longer true of the coefficient of degree one in the characteristic polynomial.

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[1] This expression is valid when no channel state information is available to the transmitter. *H* is here normalized by the rule $\sum_{i,j} \left|H_{ij}\right|^2 = n_r.n_t$, which incorporates both receiver and transmitter array gains in the capacity.

[2] when the slit is very narrow, K_{ij} mentioned above reduces to a unique value *K*.

singular values (dB)	lowest SV	medium SV	highest SV
case A: wide slit, little correlation	-7	+4.8	+7.6
case B: narrow slit, little correlation	-47	-28	+9.5
case C: narrow slit, strong correlation	-111	-41	+9.5
case D: narrow slit, little correlation (one phase randomised)	-40	-3.9	+9.4

Table 1: median of the three singular values of the channel matrix, for four situations differing by the slit width (importance of diffraction) or inter-sensor spacing (correlations)

Interestingly, the statistical distribution of capacities in this case is very wide, starting from the SDF value (2.5 b/s/Hz) up to 3.5 b/s/Hz, i.e. only slightly less than the full triple degree of freedom case (maximum of 3.7 b/s/Hz). This is reminiscent of an ordinary fading phenomenon, creating dips in radio transmission for certain receiver positions or at some time instants. For certain values of the randomised phase, the coefficient of degree one of the characteristic polynomial of the channel matrix gets nearly zero, therefore the channel basically degenerates to a SBF, with reduced capacity. On the other hand this phase randomisation sometimes allow high values of the medium singular value, resulting in high capacity. This helps understand the peculiarity of keyholes: the rank deficiency they are responsible of is not "accidental", as is the case of a fading event. On the contrary it is a permanent deficiency of the channel, due to the geometric/electromagnetic characteristic of the environment. In the example chosen, this deficiency is due to diffraction, responsible for a "mixing" of waves exiting the emitter into each of the waves reaching the receiver, thereby factorising the channel matrix into a product of two vectors. Quantitatively speaking, we see on fig. 4 that most of the capacity collapse arises when the slit tightens from about two to about one wavelength.

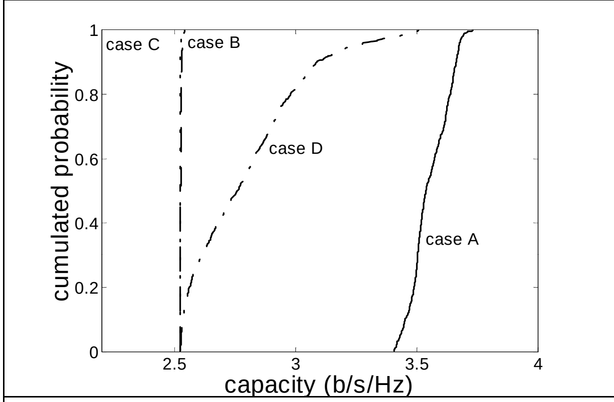

fig. 3: CDF of the capacity in the various cases

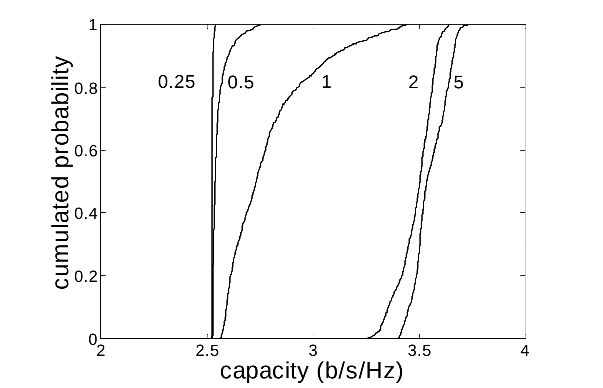

fig. 4: CDF of the capacity as a function of the slit width, expressed in wavelengths.

III MIMO CHANNEL MODELLING WITH KEYHOLES AND WAVE JUNCTIONS

1/ The "wave connecting matrix"

It now appears very clearly that due to diffraction keyholes, there is not a one to one correspondance between emitted and received waves. In the picture of fig. 2, a single emitted wave is responsible for three received ones due to the "junction" represented by the slit. In the general case and in the small antenna approximation[3], we will have N received waves and M emitted ones. The MIMO channel matrix can be written:

(2) $\mathbf{H}(\omega) = \mathbf{a_r}(\omega)\mathbf{W}(\omega)\mathbf{a_t^T}(\omega)$,

where $\mathbf{a_r}$ and $\mathbf{a_t}$ are the steering matrices for the receiving and transmitting arrays, of size $n_r \times N$ and $M \times n_t$ respectively: $a_{rij} = \exp(-j\vec{k}_i \cdot \vec{r}_{rj})$ and $a_{tij} = \exp(-j\vec{k}_i \cdot \vec{r}_{tj})$,

and $\mathbf{W}$ is an $N \times M$ matrix connecting each received wave to each emitted one. W_{ij} is thus the ratio of the complex amplitude of received wave $\vec{k}_i$ to the amplitude of the emitted wave $\vec{k}_j$ for Tx and Rx radiators located at the origins of their respective referentials. In the above 3×3 waves example (eq. (1)), we have thus $W_{ij} = R_i.T_j.K_{ij}\exp(-jkl_i)\exp(-jkl_j)$.

Usual geometrical descriptions of channel models including MIMO [4] [5] connect each emitted wave to a single received wave, and *vice-versa*. This implies a restriction to the events experienced by any given wave, which can only scatter into a single other wave. In such a case, the wave connecting matrix has a single element per column/line, which guarantees that it is square and is full rank, although of course some singular values may still be very small.

However diffraction, but also partial transmission/refraction is in practice at the origin of multiple exiting waves for a single initial one. Such phenomena will fill the wave connecting matrix, eventually resulting in a low-rank of this matrix.

By Fourier transformation, we rewrite eq. (2) in the delay domain at baseband:

(3) $\mathbf{H}(\tau) = \mathbf{a_r}(\tau) \otimes \mathbf{W}(\tau) \otimes \mathbf{a_t^T}(\tau)$

where now the steering matrices have for their entries Dirac operators of the form $\delta(\tau - \tau_{ij})$ delaying the response of sensor j for the wave i by the inter-antenna delay $\tau_{ij} = \frac{\vec{k}_i \cdot \vec{r}_j}{c\left\|\vec{k}_i\right\|}$, and the wave connecting matrix is explicitly expressed in the delay domain. Its entries are also of the form $\delta(\tau - \tau'_{ij})$ times a complex constant, where τ'_{ij} is the path delay, i.e. $\tau'_{ij} = (l_i + l_j)/c$ in the example. Since the convolution product has replaced ordinary product, it is obvious that the entries of matrix $\mathbf{H}(\tau)$ are linear combinations of Dirac operators $\delta(\tau - \tau_{ij} - \tau'_{kl})$. Expression (3)for the channel matrix is thus the basis of MIMO channel modelling. If the system has an infinite bandwidth all paths are resolved, and for each value of the various delays the matrix has rank one. However in the case of a band-limited system the rank of the matrix for a given delay value, and therefore the transmit-receive diversity, will tend to increase when the bandwidth is reduced. It is also possible to extract a MIMO tapped delay line model from the delay dependent channel matrix in (3).

<u>2/ Computation of the channel correlation matrix</u>

In the narrow band limit, the MIMO channel correlation matrix of dimension $NM \times NM$ writes:

$\mathbf{R} = E\{\mathbf{a_r}(\omega)\mathbf{W}(\omega)\mathbf{a_t^T}(\omega) \times \mathbf{a_r^*}(\omega)\mathbf{W}^*(\omega)\mathbf{a_t^H}(\omega)\}$, where * stands for complex conjugation, H for complex conjugation and transposition, and x is the tensor product.

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[3] This stems from the use of a steering matrix, whereby only the phase changes from one sensor to another

Let us temporarily assume that **W** is a purely diagonal matrix with entries W_i (*i=1, ... NM*). The correlation between any pair of entries of the channel matrix can be computed in the following way:

$$\mathbf{R}_{mnm'n'} = E\left\{\left(\sum_i W_i \exp\left[\mathrm{j}\cdot\vec{\mathrm{k}}_{\mathrm{ri}}\cdot\vec{\mathrm{r}}_{\mathrm{rm}} + \mathrm{j}\cdot\vec{\mathrm{k}}_{\mathrm{ti}}\cdot\vec{\mathrm{r}}_{\mathrm{tn}}\right]\right)\cdot\left(\sum_i W_j^* \exp\left[-\mathrm{j}\cdot\vec{\mathrm{k}}_{\mathrm{rj}}\cdot\vec{\mathrm{r}}_{\mathrm{rm'}} - \mathrm{j}\cdot\vec{\mathrm{k}}_{\mathrm{tj}}\cdot\vec{\mathrm{r}}_{\mathrm{tn'}}\right]\right)\right\}$$

$$= \sum_{i,j} E\left\{W_i\cdot W_j^*\cdot\exp\left[\mathrm{j}\cdot\vec{\mathrm{k}}_{\mathrm{ri}}\cdot\vec{\mathrm{r}}_{\mathrm{rm}} - \mathrm{j}\cdot\vec{\mathrm{k}}_{\mathrm{rj}}\cdot\vec{\mathrm{r}}_{\mathrm{rm'}} + \mathrm{j}\cdot\vec{\mathrm{k}}_{\mathrm{ti}}\cdot\vec{\mathrm{r}}_{\mathrm{tn}} - \mathrm{j}\cdot\vec{\mathrm{k}}_{\mathrm{tj}}\cdot\vec{\mathrm{r}}_{\mathrm{tn'}}\right]\right\}$$

where the r subscript applies to the receiver array and the t subscript to the transmitter array. Assuming the expectation is computed by averaging over a certain small-scale local zone for which the waves have a given direction and amplitude:

$$\mathbf{R}_{mnm'n'} = \sum_{i,j} W_i\cdot W_j^*\cdot E\left\{\exp\left[\mathrm{j}\cdot\vec{\mathrm{k}}_{\mathrm{ri}}\cdot\vec{\mathrm{r}}_{\mathrm{rm}} - \mathrm{j}\cdot\vec{\mathrm{k}}_{\mathrm{rj}}\cdot\vec{\mathrm{r}}_{\mathrm{rm'}}\right]\right\}\cdot E\left\{\exp\left[\mathrm{j}\cdot\vec{\mathrm{k}}_{\mathrm{ti}}\cdot\vec{\mathrm{r}}_{\mathrm{tn}} - \mathrm{j}\cdot\vec{\mathrm{k}}_{\mathrm{tj}}\cdot\vec{\mathrm{r}}_{\mathrm{tn'}}\right]\right\}$$

we have for instance[4]:

$$E\left\{\exp\left[\mathrm{j}\cdot\vec{\mathrm{k}}_{\mathrm{ri}}\cdot\vec{\mathrm{r}}_{\mathrm{rm}} - \mathrm{j}\cdot\vec{\mathrm{k}}_{\mathrm{rj}}\cdot\vec{\mathrm{r}}_{\mathrm{rm'}}\right]\right\} = \exp\left[-\mathrm{j}\cdot\vec{\mathrm{k}}_{\mathrm{rj}}\cdot\left(\vec{\mathrm{r}}_{\mathrm{rm'}} - \vec{\mathrm{r}}_{\mathrm{rm}}\right)\right]\cdot E\left\{\exp\left[\mathrm{j}\cdot\left(\vec{\mathrm{k}}_{\mathrm{ri}} - \vec{\mathrm{k}}_{\mathrm{rj}}\right)\cdot\vec{\mathrm{r}}_{\mathrm{rm}}\right]\right\}$$

$$= \exp\left[-\mathrm{j}\cdot\vec{\mathrm{k}}_{\mathrm{rj}}\cdot\left(\vec{\mathrm{r}}_{\mathrm{rm'}} - \vec{\mathrm{r}}_{\mathrm{rm}}\right)\right]\cdot\delta_{ij}$$

which expresses uncorrelated scattering for two waves differing by their wavevector.

Therefore: $\mathbf{R}_{mnm'n'} = \sum_i |W_i|^2\cdot\exp\left[-\mathrm{j}\cdot\vec{\mathrm{k}}_{\mathrm{ri}}\cdot\left(\vec{\mathrm{r}}_{\mathrm{rm'}} - \vec{\mathrm{r}}_{\mathrm{rm}}\right) + \mathrm{j}\cdot\vec{\mathrm{k}}_{\mathrm{ti}}\cdot\left(\vec{\mathrm{r}}_{\mathrm{tn'}} - \vec{\mathrm{r}}_{\mathrm{tn}}\right)\right]$, to be divided by $\sum_i |W_i|^2$ to obtain a normalized correlation.

In the general case of a full wave connecting matrix, the sums carry on two indexes instead of one, yielding the expression:

$$\mathbf{R}_{mnm'n'} = E\left\{\left(\sum_i W_{i,i'} \exp\left[\mathrm{j}\cdot\vec{\mathrm{k}}_{\mathrm{ri}}\cdot\vec{\mathrm{r}}_{\mathrm{rm}} + \mathrm{j}\cdot\vec{\mathrm{k}}_{\mathrm{ti'}}\cdot\vec{\mathrm{r}}_{\mathrm{tn}}\right]\right)\cdot\left(\sum_i W_{j,j'}^* \exp\left[-\mathrm{j}\cdot\vec{\mathrm{k}}_{\mathrm{rj}}\cdot\vec{\mathrm{r}}_{\mathrm{rm'}} - \mathrm{j}\cdot\vec{\mathrm{k}}_{\mathrm{tj'}}\cdot\vec{\mathrm{r}}_{\mathrm{tn'}}\right]\right)\right\}$$

and finally $\mathbf{R}_{mnm'n'} = \sum_{i,i'} |W_{i,i'}|^2\cdot\exp\left[-\mathrm{j}\cdot\vec{\mathrm{k}}_{\mathrm{ri}}\cdot\left(\vec{\mathrm{r}}_{\mathrm{rm'}} - \vec{\mathrm{r}}_{\mathrm{rm}}\right) + \mathrm{j}\cdot\vec{\mathrm{k}}_{\mathrm{ti'}}\cdot\left(\vec{\mathrm{r}}_{\mathrm{tn'}} - \vec{\mathrm{r}}_{\mathrm{tn}}\right)\right]$, to be normalized by $\sum_{i,i'} |W_{i,i'}|^2$.

From the knowledge of the wave connecting matrix at a given location and frequency, we thus immediately deduce the narrowband channel correlation matrix.

3/ Case of an array of coupled sensors

Before proceeding, let us note that the assumption used above of uncoupled array sensors is not compulsory and can be relaxed. If we generalize the antenna description to a linear passive multiport with n_t sensor inputs, its complete characterization in transmission is achieved by the knowledge of n_r complex radiation patterns. Each of the latter specifies the complex amplitude of a far field wave emitted in any direction, for an input voltage of 1V on one sensor and 0V on the others. Therefore in the case of *M* emitted waves, the antenna contribution to the channel transmission is given by an $M \times n_t$ complex gain matrix, simply

[4] The Kronecker delta implies uncorrelated scattering for two received waves of differing directions.

taking the place of the transmission steering matrix in (2). By reciprocity, the same conclusion applies to the receiving antenna array. Let us further stress that the complex gain matrix inherently contains whatever coupling between sensors, which takes part to the coherent construction of the radiation pattern within the antenna, as well of course as the dephasing between sensors due to their differing spatial location (fig. 5).

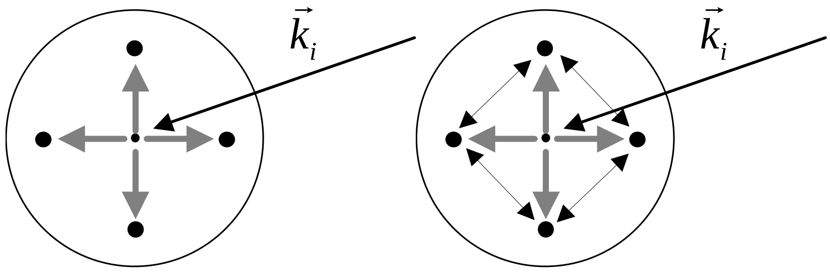

fig. 5: schematic view of a "virtual" and a "real" array of sensors. In the first case the sensor signals are dephased by $\exp(-j\vec{k}_i \cdot \vec{r}_{rj})$ w.r.t. the array center, symbolised by the grey arrows. In the second, multiple scattering of the wave due to inter-sensor coupling modifies this dephasing (symbolised by dashed lines).

<u>4/ Towards a stochastic MIMO channel model</u>

Going back to channel modelling, we state that the above approach lends itself naturally to stochastic modelling using a geometric based approach. The inputs of the model will be the following:

1. specification of the multiport Tx and Rx antennas, either through steering matrices (uncoupled sensors) or through complex gain matrices.
2. double directional model of emitted and received waves, specifying the statistical laws of angular distributions on both sides of the radio link.
3. statistical model for the wave connecting matrix $\mathbf{W}(\tau)$, specifying the distribution of complex entries of the matrix, especially the number of non zero entries for the various columns or lines and their relative amplitudes.
4. statistical model for the distribution of delays involved in the non zero entries of $\mathbf{W}(\tau)$.

Obviously these various inputs are not independent. Therefore the stochastic model will have to determine the relationships between the parameters involved in the various distributions, as a function of the targeted radio environment.

<u>5/ Model simplification</u>

As compared to SISO or SIMO modelling, the new source of complexity here lies in the size of the wave connecting matrix. In all generality, if there are say 10 significant DOAs at the receiver and 12 directions of departure (DOD) at the transmitter, 120 paths are possible with 120 concomitant delays. Obviously this is not very realistic, we would rather expect something like e.g. 8 junctionless paths, and 2 paths incorporating a Y junction. Therefore we have to decide of the number of these few junction paths, of their respective DOAs and DODs, their delays and the relative amplitudes of the sub-paths. We may also decide that the full complexity of a real propagation channel is exaggerated, and that only the most important rays will be retained. In this case we may limit the angular resolution of our model, group some rays into a single one, and create artificial junctions (fig. 6). The decision to do this will depend on the required angular precision of the model, in turn depending through Heisenberg relation on the maximum size considered for the array antennas.

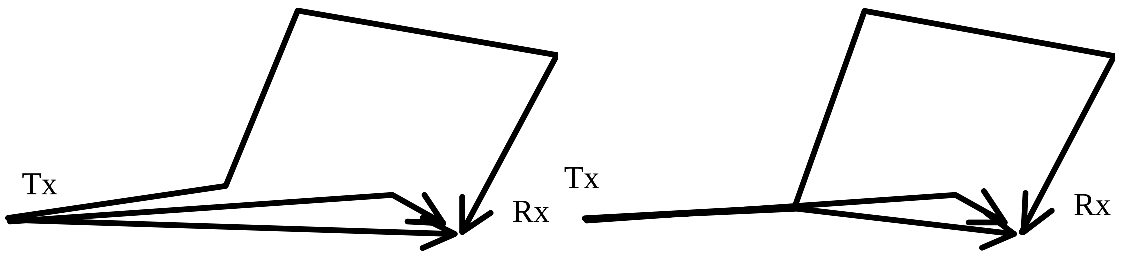

fig. 6: a real situation of waves (left), and its approximation using an artificial junction of waves, reducing the size of the wave connecting matrix.

We see that often enough the wave connecting matrix will have only one non-zero entry in most of its columns/lines. Only in the case of a "genuine" keyhole as in the example of section I, will it be a full matrix. Nevertheless these few extra non zero coefficients are important with respect to MIMO performance, and even though a superposition of keyholes is not in general a keyhole, these extra coefficients will often tend to reduce the wave connecting matrix rank. Knowing that the rank of a matrix product cannot be greater than the smallest rank of the multiplied matrices, this will impact the channel matrix rank and therefore the radio link capacity. This is the reason why realistic channel modelling is required. In any case the number of degrees of freedom of the channel is at most equal to the smallest of the two numbers of emitted and received waves, making rich scattering environments favourable to MIMO radio transmission.

Another issue deals with the level of precision required in modelling. Obviously very low amplitude echoes will negligibly contribute to the radio link capacity. Discarding them will reduce the wave connecting matrix size. However since MIMO techniques are generally suited to narrowband transmission, all waves interfere and the sum of numerous low energy waves may not be negligible. It might be appropriate to replace them by one or a few Rayleigh fading waves of random DOA/DOD.

Let us consider an example scenario (fig. 7), e.g. an urban environment with an above rooftop base station and a below rooftop mobile terminal. Input 1 is specified by the antenna characteristics. According to the geometry considered, we expect the DOD from the mobile to exhibit a large angular spread, according to some statistical distribution [6] , and the direction of arrival to the base station to be little spread around the mean direction. This will define input 2.

We now have to list typical physical phenomena responsible for junctions in wave paths, e.g. roof edge or wall edge diffraction, street canyons, successive reflexion/transmission events etc … From physical laws, we can model in a statistical way the relative amplitudes of waves in the various sub-paths. This will define input 3.

Finally we can choose statistical laws for the total path delays, like Poisson or modified Poisson distributions, also taking into account the physical origin of the junction sub-paths with their associated relative delays. This will finally define input 4

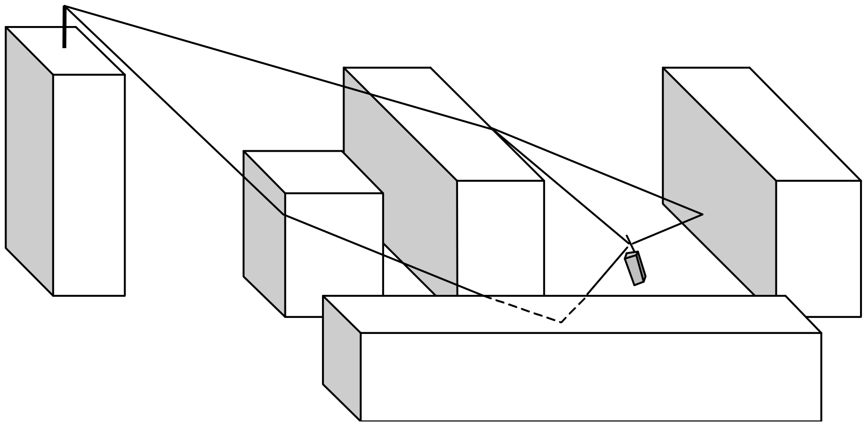

fig. 7: a suburban scenario including a junction between a main wave path and two diffracted subpaths

IV CONCLUSION

The consideration of the existence of keyholes has led us to a propose an improvement of geometric based stochastic modelling, particularly adapted to MIMO radio communications. The specificity of the model lies in the existence of a wave connecting matrix, essentially composed of a single non zero element per column/line, but where keyhole like physical situations are expressed by additional non zero coefficients, tending to make the channel matrix rank-deficient.
It is clear that much among the four inputs of section III and particularly a realistic modelling of the wave connecting matrix will benefit from experiments, but also from physical modelling based simulations like ray-tracing. These techniques have the major advantage to provide a very detailed information on wave paths and sub-paths for a given environment, taking into account reflection, refraction, diffraction and transmission events in a realistic way, and are therefore very powerful for MIMO modelling. However the major difficulty will be to define and parameterise in a realistic way the various statistical distributions involved.

To conclude, let us say that we expect such a geometric based MIMO channel modelling with keyholes (GMCMK) to be quite accurate and capable to cover a variety of environmental conditions, antennas and scenarios. This goes beyond presently used models based on the generation of a correlation matrix between sensors[7] , as currently considered in e.g. UTRA standardization bodies [8] .In effect, more sophisticated as discussed here might supply such models with consolidated values of the correlation matrices as a function of classified radio environments, allowing simple and reliable tests of technological solutions. In addition, GMCMK allows to investigate the role of its various constituents in the overall MIMO link performance, e.g. of the antenna internal coupling correlations, and discriminate it from the nature of the environment considered. This is important both from the point of view of the fundamental comprehension of MIMO performance, but also for its practical interest in terms of improvement of this performance.

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