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New Algorithm to Improve the Coexistence between IMT-Advanced Mobile Users and Fixed Satellite Service

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Abstract. Novel technique to modulate the nulls in the radiation pattern of IMT-Advanced base station (BS) towards the fixed satellite service is (FSS) affirmed in this paper. Designing a new algorithm to extract the nulls in the forbidding area and other base on MUSIC algorithm to estimate the direction of mobile user and control the handover technique is our major concern. A scenario of two mobile users (MS) moving around one FSS had been exclusively studied and simulated. We are looking through the mechanical of interference and our ultimate goals is to develop mechanize for interference mitigation by deploying a special division multiple access technique SDMA with MU-MIMO base on OFDM. Calculating the shortest separation distance after identifying the critical point and compare the result with the recent recommendation had shown how magnificent coexistence and spectrum sharing we can get. Analysis, discussion and conclusion had been made to explain.

Keywords: MUSIC Algorithm, Mobile users, SDMA, MU-MIMO, OFDM.

1. Introduction and Background

Wireless interference is the one of the main factors which degrades the performance of all the wireless communications devices like cell phones, laptops and Bluetooth's. So, there have been always a certain challenges and few possible solutions. According to the world radio communication 2007 (WRC-07) for ITU with resolution 228 the frequency band 3400-4200MHz (C-Band) is the most expected candidate for IMT-Advanced systems, on the other hand we have around 160 geostationary satellites operating in the band 3400 - 4200 MHz. However, C-band used over 40 years for FSS because of its low atmospheric absorption, highly reliable space-to-earth communication, wide service coverage and geographical areas with severe rain fade conditions. While, IMT-Advanced it's a combination of different access technologies; complement each technology in an optimum way and provide a common and flexible service platform for different services and applications. In contrast, C-band is also important for IMT-Advanced because it allows use of smaller antenna for terminals and base stations, favorable to implement multiple antenna techniques and enabling high spectrum efficiency. Some techniques had been used before to solve such interference like: from 2003 up to 2005 mitigation by shielding and filtering had been used to prevent the harmful interference from FWA towards FSS. From 2005 up to now different techniques like: tilting angle, beamforming and opportunistic MU-MIMO (multi user- multi input multi output) had been used.

Due to the continuing demand for high data rates and spectral efficiency, MU-MIMO-OFDM (Orthogonal frequency division multiplexing) is a promising access schemes to support IMT-Advanced systems. Because it is splitting high data rate sequence into a number low rate sequence, transmitted over a number of subcarriers, symbol duration increase for low rate parallel subcarriers, multipath delay spread is decreased, ISI (inter symbol interference) is eliminated be guard interval at the start of each OFDM symbol, OFDM is cyclically extended to avoid ICI, highly frequency selective channel transformed into large set of flat fading, Non-frequency selective narrowbwnd channels, and OFDM eliminate the need to the complex rake receiver. Therefore, Utilization of OFDM technique is expected to offer improved performance in combating adverse frequency selective fading encountered in wideband MIMO wireless systems [1].

Major impediment in MIMO based systems is the cost of hardware, because multiple antenna elements and RF chains are deployed at each terminal. Therefore, a promising technique named antenna selection has been proposed to reduce the hardware complexity (save on RF chains) while retaining many diversity benefits [2]. It employs a number of RF chains, each of which is switched to serve multiple antennas [3]. In recent years, there has been increasing interest in applying antenna subset selection to MIMO systems over frequency selective channels [4], [5]. Several algorithms have been developed for selecting the optimal antenna subset. MU-MIMO aims at improving the overall system throughput by supporting multiple users in a cell over the same time-frequency resource. If the users are sufficiently spatially separated, MU-MIMO significantly outperforms Single-User (SU) MIMO.

The need to separate co-channel signals by an antenna array arises in many situations and in cellular communications in particular. Separating signals from different users by their spatial signatures can increase significantly system capacity beyond what can be achieved by temporal processing alone. The multiple exploitation of the same resource can be done either explicitly by forming directional beams to several users (in this case often denoted as Space Division Multiple Access (SDMA), or implicitly by utilizing the spatial diversity of the multipath radio channel by appropriate precoding / beamforming techniques at the transmitter (often denoted as spatial multiplexing). Combinations of both approaches are possible. In general, beamforming-based solutions are advantageous in outdoor wide-area cellular environments with usually low angular spread, whereas spatial multiplexing performs better in pico-cellular or indoor scenarios with rich multipath propagation. Extensive work has been done on techniques for separating co-channel signals as described in [6] and [7].

The overwhelming majority of this work assumes that the signals are point sources having zero angular spread, and that the propagation from transmitter to receiver occurs along one path or a few distinct paths (multi path). A more practical model used in the cellular communications environment represents the propagation as a continuum of different paths, where the signal arriving at the receiver antenna can be characterized by a continuous spatial distribution of energy (angular spread) around the nominal direction of arrival. Relatively little work has been done on co-channel signal separation for the case of distributed signal sources. Previously [8] developed a performance bound for specifically the isotropic scattering case. More recently, [4] proposed a structured estimate of the array response, which performed better in a distributed fading environment to a system that used an unstructured estimate. [9] looked at the effect that temporal and spatial correlation has on the DOA estimation, and [10] investigated the performance of the general class of semi-blind subspace-based estimators, like those that we are using here.

Therefore, in this paper we study the effect of angular spread on the ability of an array to separate co-channel signals. While, SDMA –UPC (unitary Pre-coded matrix) –MU-MIMO can be adjusted with limited feedback and null steering technique modeling to mitigate the interference from IMT-Advanced to FSS receiver. In particular we use a framework where an isotropic scattering environment was assumed (essentially an angular spread of 180 degrees) which is characteristics of a pico cell environment. Here the angular spread can take on any value between zero (point source) and 180 degrees. Calculating the shortest separation distance after identifying the critical point for fixed services and compare the result with the recent recommendation had shown how magnificent coexistence and spectrum sharing we can get. Finally we developed a mechanism for hand over technique to avoid the interference. Deployment of the interference scenarios in the Matlab and ICS telecom software to get the minimum separation distance base on our assumption had shown a magnificent result which is compatible with our expectations. Analysis, discussion and conclusion had been made to explain the results and achievements. The modification process is composed of 3 steps: (1) estimate the used direction, (2) computation of null points of original precoding vectors, (3) steering null into the direction of the FSS ES, (4) selection of modified precoding vectors, (5) smooth hand over to avoid service failure caused by nulls.

2. Algorithm Design

The basic concept of the algorithm is to form nulls in the spatial spectrum that correspond to the direction angles of the victim FSS ES. In this paper, for convenience the term 'DOE' denotes the direction angles of the victim FSS ES. First, the IMT-Advance BS has to obtain DOE information in order to perform nullsteering. DOE information can be obtained by adopting a popular spatial spectrum estimation direction finding method which is MUSIC Algorithem in our case. The main steps of the algorithm are outlined as follows:

1- Estimating Beamforming Weights: The beamforming weights that will cancel the direction of FSS-ES must be estimated from noisy snapshot data. Below, samples at the i[th] antenna are specified by s_i, where

$$s_i = u_i + v_i; \qquad i = 1; 2; 3$$

Pairing the samples from antenna i and m, where again $i \neq m$ we define a 2×1 vector

$$s_{i,m} = [s_i, s_m]^T .$$

Next, those weights are used to search for a target with threshold detection in each bin of interest that is obscured by the jammer (FSS-ES).

2- Weight Estimation Method: Having constructed our system model such that the jammer powers (other mobile users) for any particular snapshot are equal in magnitude, the only quantities that need to be estimated are the phases $W_{1,2}$, $W_{1,3}$, and $W_{2,3}$.

The least complicated way to estimate these is to measure the phase differences between the signals at the individual antennas for a collection of N snapshots, take an average, and then add π fo $W_{i,m}$

$$\hat{W}_{i,m} = \pi + \frac{1}{N}\sum_{k=1}^{N} \angle(S_i^k) - \angle(S_m^k)$$

Which results in the vector estimate

$$\hat{W}_{i,m} = \frac{1}{\sqrt{2}}\left[1, e^{j\hat{w}_{i,m}}\right]^T$$

In the weighted sum, the jammer's power is negligible, but the target's directional phase information is distorted. For this reason the phase information has to be recovered in a novel manor. So, if a target is detected, its threshold is recorded and we proceed to step 3, if no target is detected, we start over with the next coherent processing interval (CPI=P(pulses)/PRF(pulse repetition frequency)).

3- Signal Detection: The first step in searching for a target that is masked by an interfering signal is to estimate the beamforming weights. For this estimation to be accurate, the snapshots should not contain the target signal. If these snapshots are obtained improperly, the algorithm might be trained to cancel the target in the adapted beams. For this reason, a system's designer should consider the ranges and/or velocities at which a target of interest might be. If the interference source is a jammer, jammer-plus-noise snapshots can be reliably obtained from the furthest unambiguous range cells. After the training snapshots are collected, and the weights have been estimated, they will be used to search for a target signal within the jammer dominated received signal. Multiplying the weighting vectors with their respective received signal vectors results in the residue power detection variable

$$E\{p\} = \frac{1}{4}\sum_{m=2}^{3} E\{v_1^2 + v_m^2\} = \frac{1}{4}\sum_{m=2}^{3} 2\sigma^2 = \sigma^2$$

Therefore, if a target signal is not present in the received signals and the weight vectors are normalized, then the average residue power will be equal to the average noise power in the channels. If a target signal is present, then p will be larger and can be used for detection purposes. How much larger p becomes depends on the original target signal power and on the spatial angle separation between the target and jammer signals. Weighted sums in which the jammer has been canceled can be made from groups or from individual cells over any map region of interest; map can be created from values of p.

4- Relative phase information is estimated at three antennas. We need to find θ and δ by assuming the jammer signal is canceled and the noise negligible, we get a set of three equations

$$r_{1,2} = \frac{\alpha}{\sqrt{2}} e^{j\phi}(1 + e^{j(\theta+\omega_{1,2})})$$

$$r_{1,3} = \frac{\alpha}{\sqrt{2}} e^{j\phi}(1 + e^{j(\theta+\omega_{1,3})})$$

$$r_{2,3} = \frac{\alpha}{\sqrt{2}} e^{j\phi}(1 + e^{j(\theta+\omega_{2,3})})$$

Because the components of the weight vector are of equal magnitude, $r_{1,2}$ bisects the angle subtended by the target phasors shown in Figure below The bisection allows us to write an equation that only involves the unknown phases.

$$2\angle r_{1,2} = \phi + (\phi + \theta + \omega_{1,2})$$
$$\phi_1 = 2\angle r_{1,2} - \omega_{1,2} = \theta + 2\phi$$

Note that the quantity ϕ_1 is known, whereas the quantities on the right most side of the above equation are unknowns. The final two weighted sum equations for $r_{1,3}$ and $r_{2,3}$ can also be reduced to relationships involving only the phase angles. These operations respectively yield:

$$\phi_2 = \delta + 2\phi; \phi_3 = \theta + \delta + 2\phi;$$

Finally, estimating θ and δ is accomplished from

$$\theta = \phi_3 + \phi_2; \delta = \phi_3 - \phi_1$$

Once values for θ and δ are found, it is straight forward to obtain estimates for ϕ and α if desired, DOAs are calculated from the phase information

5- Compute the nulls θ_0 generated by n_T precoding vector $e_{g,m}(m = 0,1,...,n_T - 1)$.
6- Calculate n_T precoding vector $W_{g,m}$ (m=0,1,…,n_T -1) depending on DOE $\hat{\theta}$ and the null θ_0.
7- Select the n_T-1 precoding vectors, $W_{g,n}$ (n=0,1,…,n_T-2), forming nulls at DOE $\hat{\theta}$ from n_T precoding vectors $W_{g,m}$.

At the unitary precoding matrix U={E_0,…,E_{G-1}}, where E_g=[$e_{g,0}$,…,$e_{g,nT-1}$] is the g^{th} precoding matrix. $E_{g,m}$ is the m^{th} precoding vector in the matrix E_g.

$$e_{g,m} = \frac{1}{\sqrt{n_T}}[1e^{j\frac{2\Pi}{n_T}(\frac{g}{G}+m)} \ldots e^{j\frac{2\Pi}{n_T}(n_T-1)(\frac{g}{G}+m)}]^T$$

If the plane wave attacked the array at angle θ with a respect to the array normal the array propagation vector for a uniformly spaced linear array is defined by

$$v = [1e^{j2\Pi\frac{d}{\lambda}\sin\theta} \ldots e^{j2\Pi(n_T-1)\frac{d}{\lambda}\sin\theta}]^T$$

Where λ is a wavelength and d is is the space between the antennas array and we consider d= 0.5λ.

So, array factor in term of vector inner product:

$$F_m(\theta) = e_{g,m}^T v = \frac{1}{\sqrt{n_T}} \sum_{k=0}^{n_T-1} e^{j2\Pi k\{\frac{1}{n_T}(\frac{g}{G}+m)+\frac{d}{\lambda}\sin\theta\}}$$

θ_0 Satisfying $F_m(\theta)$=0 and means null generated by the precoding matrices and we need the nulls in the DOE $\hat{\theta}$ and null steering should be perform for the case of $\hat{\theta} \neq \theta_0$.

Let $\hat{\theta} = \theta_0 + \alpha$.

In the order of steering the null to $\hat{\theta}$, the array factor $F_m(\theta)$for the precoding vector $e_{g,m}$ have to be shifted to α, that is:

$$F_m(\theta - \alpha) = \frac{1}{\sqrt{n_T}} \sum_{k=0}^{n_T-1} e^{j2\Pi k\{\frac{1}{n_T}(\frac{g}{G}+m)+\frac{d}{\lambda}\sin(\theta-\alpha)\}}$$

$$F_m(\theta-\alpha)=\frac{1}{\sqrt{n_T}}\sum_{k=0}^{n_T-1}[e^{j2\Pi k\{\frac{1}{n_T}(\frac{g}{G}+m)+\frac{d}{\lambda}\sin\theta\}}\times e^{-j2\Pi k\frac{d}{\lambda}\cos\theta\sin\alpha}(e^{j2\Pi k\frac{d}{\lambda}\sin\theta})^{\cos\alpha-1}]$$

For $\theta=\hat{\theta}$ can be like:

$$F_m(\hat{\theta}-\alpha)=\frac{1}{\sqrt{n_T}}\sum_{k=0}^{n_T-1}[e^{j2\Pi k\{\frac{1}{n_T}(\frac{g}{G}+m)+\frac{d}{\lambda}\sin\hat{\theta}\}}\times e^{-j2\Pi k\frac{d}{\lambda}\cos\hat{\theta}\sin\alpha}(e^{j2\Pi k\frac{d}{\lambda}\sin\hat{\theta}})^{\cos\alpha-1}]=F_m(\theta_0)=0$$

So, We can say

$$F_m(\hat{\theta}-\alpha)=(e_{g,m}\odot s)^T v$$

Where $\odot$ denotes Hadamard (pointwise) product and

$$s=\left[1e^{-j2\Pi\frac{d}{\lambda}\cos\hat{\theta}\sin\alpha}(e^{j2\Pi\frac{d}{\lambda}\sin\hat{\theta}})^{\cos\alpha-1}\ldots e^{-j2\Pi(n_T-1)\frac{d}{\lambda}\cos\hat{\theta}\sin\alpha}(e^{j2\Pi(n_T-1)\frac{d}{\lambda}\sin\hat{\theta}})^{\cos\alpha-1}\right]$$

Therefore, adapted precoding vectors $W_{g,m}$ for forming the nulls at $\hat{\theta}$ can be calculated as:

$$W_{g,m}=e_{g,m}\odot s$$

Because the beams produced by n^T precoding vectors Wg,m are mutually orthogonal, only one of n^T beams does not construct null at DOE $\hat{\theta}$. Therefore, the nT-1 precoding vectors Wg,n (n = 0, 1, …, n^{T-2}), which form null at $\hat{\theta}$, are selected from Wg,m(m = 0, 1, …, n^{T-1}).

In conclusion, n^{T-1} precoding vectors Wg,n are used for data transmission of IMT-Advanced service.

3. Results and discussion

The proposed MUSIC algorithm first estimates a basis for the noise subspace and then determines the peaks the associated angles provide the DOA estimates. The MATLAB code for the MUSIC algorithm is sampled by creating an array of steering vectors corresponding to the angles in the vector angles. For any particular array shape, the weighted sums in the above analysis can be interpreted as a set of antenna radiation patterns with the set of magnitudes being analogous to adaptable antenna gains (if scaled by $1/\alpha$). Figure1 shows an example set of beams generated by an L-shaped array for the case when the algorithm cancels a jamming signal from -10± AZ and 25± EL. Before the adaptations, all of the magnitudes, or gains, would have been equal to 1.

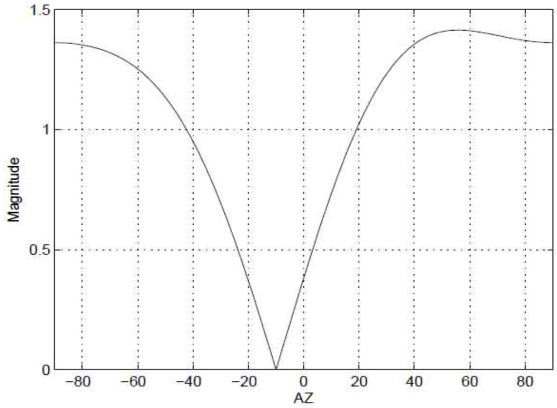


Figure 1. Beamformer output $|r_{1,2}|/\alpha$ as a function of target AZ when antenna 1 and 2 are separated by a half wavelength, and the jamming signal is at -10^0 AZ.

The beamformers from 1 to 2 and from 3 to 1 will be equivalent to that shown in Figure 1. The beamformer from antenna 2 to 3 however differs significantly from the L-shaped case. It is plotted in Figure 2.

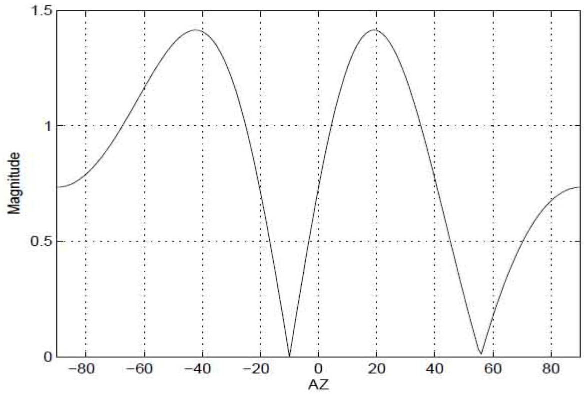


Figure 2. Shape of adapted beam obtained from sensors 2 and 3 in a uniform linear array configuration. Jamming signal at -10^0 AZ.

The over all MUSIC spectrum for two mobile user one of them on 200 and the second on 800 is represented in figure 3

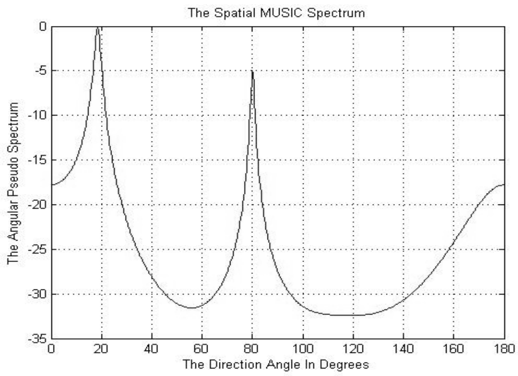


Figure 3: MUSIC spectrum for two mobile users

Based on the results of Simulation, we conclude that, while being a powerful algorithm and useful for complex multi-signal scenarios, MUSIC's computational complexity is unwarranted in the problem on which we are focused.

It is clear that Wg,m builds up nulls at $\hat{\theta}$ = -4.50, which is consistent with DOE. Figure 4 indicates that the proposed interference mitigation techniques adopt only three beams selected from the four beams. Finally, Figure 5 depicts the IMT-Advanced BS radiation pattern regardless of whether the proposed algorithm applied. The results confirm that, with the help of the proposed method, very little IMT-Advanced BS power is radiated to the FSS ES.

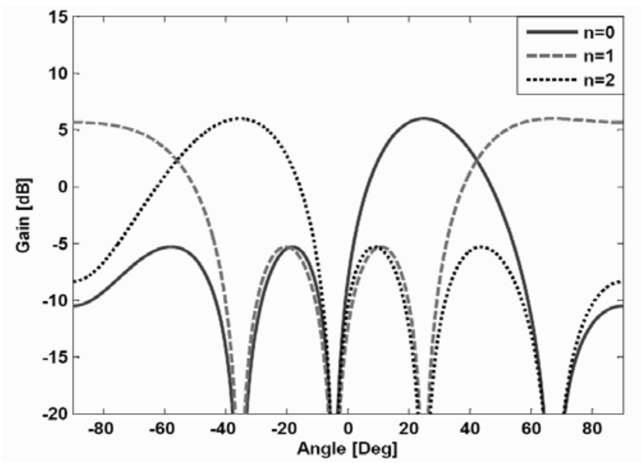


Figure 4 Three mutually orthogonal overlapped beams produced by the precoding vectors W1;n (n = 0,1,2)

Figure 5 below shows mutually orthogonal overlapped beams produced by the precoding vectors $W_{g,m}$ for $\hat{\theta}$ = -4.5^0 and α = 10^0.

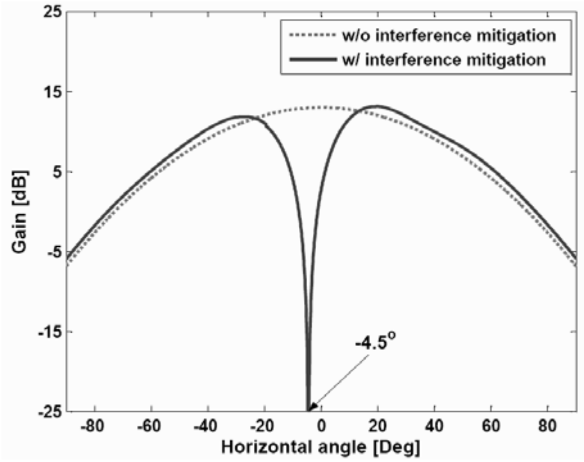
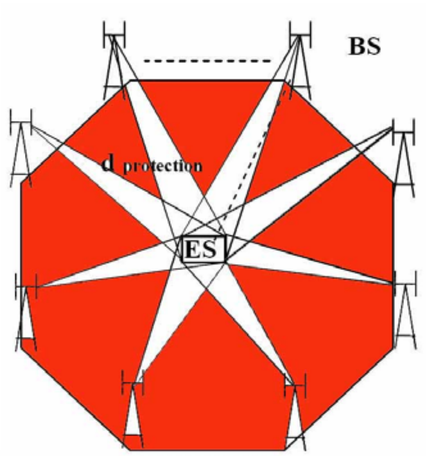


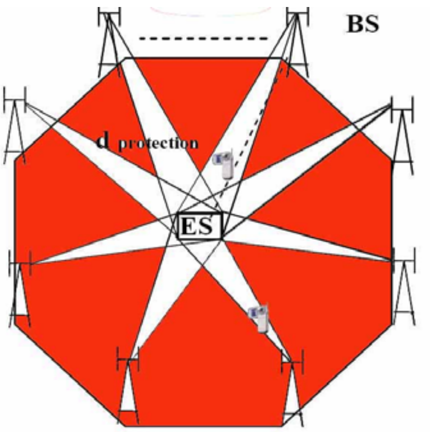
Figure 5 IMT-Advanced BS radiation patterns

4. Conclusion

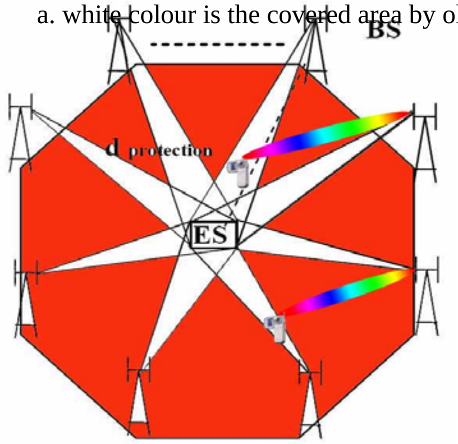
We generated nulls after we found the direction of arrived signal using the developed MUSIC algorithm. Then we generated the nulls towards the FSS-ES. One of the benefits which accrue from the use of smart beams is that users residing in different beams but in the same cell are able to reuse intra-cell frequency. This spatially separate of the signals, allow different users to share the same spectral resources, provided that they are spatially-separate at the base station. Therefore, Space Multiple Access (SDMA) allows multiple users to separate in the same cell, on the same frequency/time slot provided, using the adaptive antenna to separate the signals. This Algorithm will be deployed in the smart antenna BS scenario to improve the functionality of IMT-Advanced when it coexists with the current FSS frequency. According to the current preliminary calculations we improved that we can reduce the separation distance by 75% as illustrated in figure 6 (a, b, c and d).

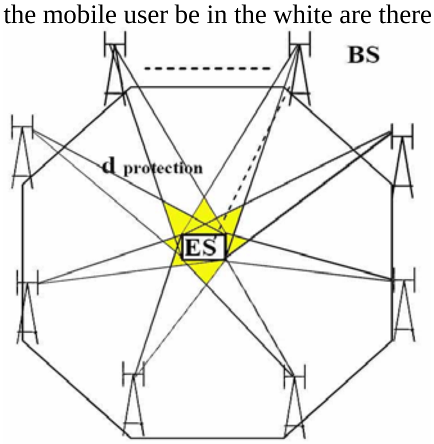


a. white colour is the covered area by old technique



b. if the mobile user be in the white are there will be





mobile user be in the white are there will be no coverage

c. New technique allow us to cover most of the white areas by handover to the nearest BS according to the received threshold

d. Yellow area represent the uncovered area (only 0.003km for suburban environment)

Figure 6: improvement of current Algorithm functionality

5. Acknowledgements

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