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Beamforming-type base station power control for personal communication networks

P.D. Karaminas and A. Manikas

Indexing terms: Personal communication networks, Beamforming arrays, Power control

A new beamforming-type power control approach, at the base station of a personal communication network, is proposed. This approach provides an increased system capacity by performing a versatile form of spatial filtering which enhances the spectrum utilisation efficiency.

Introduction: Recently, a new approach to power control was introduced in order to enhance the capacity, i.e. number of simultaneously served users, of personal communication networks. The new method uses superresolution beamformers and array signal processing at the base stations to suppress co-channel interference by synthesising an antenna pattern with high gain towards the direction of the desired mobile unit and deep nulls towards the directions of the interfering ones. So far however, all the attempts which have been made to apply this beamforming-type power control in mobile communications make unrealistic assumptions in trying to oversimplify the overall situation, or they use reference signals to estimate the direction of arrival of the desired mobile unit [1–3]. In this Letter, a new beamforming-type power control approach for mobile radio networks is introduced. This approach is robust to pointing errors and does not suffer from the power inversion problem. Furthermore, it can be classified as a signal subspace type approach with main objective to cancel completely (asymptotically) the unknown co-channel interferences and at the same time to estimate their directions. The estimation of the direction-of-arrivals (DOAs) of all the signals impinging on the array is based on the array pattern. Finally, the algorithm is based on the maximisation of the SIR in contrast to the Wiener-Hopf algorithm which uses SNIR as its performance criterion.

Complete co-channel interference cancellation: Consider a base station with an array of N antennas, receiving signals from M mobile units. It can be shown that the resolution capabilities of the used array structures are normally unable to resolve the multiple rays coming from the inter-cell interfering mobile units. This is mainly due to the large distance between the base station and these mobiles. In contrast, rays originating from the desired mobile unit are usually resolvable. Therefore the array input $\underline{x}(t)$ can be expressed as

$$\underline{x}(t) = [\underline{S}_D, \underline{S}_I] \begin{bmatrix} \underline{m}_D(t) \\ \underline{m}_I(t) \end{bmatrix} + \underline{n}(t) \quad (1)$$

where $\underline{S}_D = [\underline{S}(\theta_D + \Delta\theta_{D1}, \phi_D + \Delta\phi_{D1}), \underline{S}(\theta_D + \Delta\theta_{D2}, \phi_D + \Delta\phi_{D2}), \dots, \underline{S}(\theta_D + \Delta\theta_{DK}, \phi_D + \Delta\phi_{DK})]$ is the $N \times K$ source position matrix corresponding to the K rays coming from the desired mobile unit, $(\Delta\theta_{Dk}, \Delta\phi_{Dk})$ $k = 1, 2, \dots, K$ are the bearings of the k th reflector/scatterer in the vicinity of the desired mobile unit measured with respect to the bearings (θ_D, ϕ_D) , $\underline{S}_I = [\underline{S}_{I1}, \underline{S}_{I2}, \dots, \underline{S}_{IM}]$ is the source position matrix of the interfering mobile units, $\underline{m}_D(t)$ is the $K \times 1$ vector of the message signals corresponding to the K scattered rays of the desired mobile unit, $\underline{m}_I(t)$ represents the message signals of the interfering mobile units $\underline{m}_I(t) = [m_{I1}(t), m_{I2}(t), \dots, m_{IM}(t)]^T$, and $\underline{n}(t)$ is $N \times 1$ vector of the additive white Gaussian noise at the locations of the array elements.

The N -dimensional vectors $\underline{S}(\theta, \phi)$ are known as source position vectors (SPVs) or manifold vectors. For an array of omnidirectional antenna elements with unity gain and zero phase receiving signals from far-field sources, the SPVs are given by

$$\underline{S}(\theta, \phi) = \exp\{-j \underline{r}^T \underline{k}(\theta, \phi)\} \quad (2)$$

where (θ, ϕ) are the bearings of the source, $\underline{r}$ denotes the locations of the array elements, and $\underline{k}(\theta, \phi)$ is the wavenumber vector.

Taking into account that there is usually a line of sight between the reflectors/scatterers around the mobile unit and the base station antenna, the following propagation model can be used for the k th element of the vector $\underline{m}_D(t)$:

$$m_{Dk}(t) = \frac{cR_k}{4\pi F_c d_{Dk}} e^{-j(\Delta\varphi_{refk} + 2\pi \frac{\Delta d_{Dk}}{c} F_c)} u_D \left(t + \frac{\Delta d_{Dk}}{c} \right) \quad (3)$$

where $u_D(t)$ is the message signal transmitted by the desired mobile unit, F_c is the propagation frequency, c is the speed of light in the propagation medium, d_{Dk} is the distance between the base station and the k th scatterer, R_k and $\Delta\varphi_{refk}$ are the amplitude and phase part of the reflector coefficient respectively, and $\Delta d_{Dk} = |(\text{direct path}) - (k\text{th path})|$ is the path difference between the direct and the k th paths. The inter-cell interfering signal $m_{Ik}(t)$ can be expressed as

$$m_{Ik}(t) = \sqrt{L_{phIk} L_{sIk} L_{fIk}} e^{j\varphi_{fIk}} u_{Ik}(t) \quad (4)$$

where $u_{Ik}(t)$ is the message signal transmitted by the k th interfering mobile unit, $L_{phIk} \approx 1/d_{Ik}^4$ is the propagation path loss factor, L_{sIk} is the fast fading factor which is log-normal distributed with mean 0 dBs and variance 6dBs, L_{fIk} is the amplitude component of the fast fading which is Rayleigh distributed, and φ_{fIk} is the phase component of the fast fading which is uniformly distributed between 0 and 2π . Assuming that the signals propagate through distinct paths, the factors L_{phIk} , L_{sIk} , L_{fIk} , φ_{fIk} , R_k , $\Delta\varphi_{refk}$, $\Delta\varphi_{pathk}$, Δt_k , can be considered constant during an observation interval of L snapshots.

As mentioned above, the objective of this study is the complete cancellation of co-channel interferences and the estimation of their directions. Therefore the beamformer should estimate the appropriate weight vector which synthesises an antenna pattern with deep nulls toward the unknown directions of the interfering mobile units. To do so, the beamformer requires a pilot signal coming from an *a priori* known direction. The pilot signal must be uncorrelated with any other signal impinging on the array and can be transmitted by a neighbouring base station. The weights to receive the signal from the desired mobile unit, and cancel all the others, can be estimated by using the following algorithm:

- (i) A base station transmits a pilot signal continuously at the uplink carrier frequency(ies).
- (ii) Over an observation interval of L snapshots, the data covariance matrix $\mathbf{R}_{xx}$ is formed.
- (iii) The received power P_p of the pilot signal is estimated by minimising the following cost function:

$$\xi(a) = \sum_{\substack{i=1 \\ e^{j\varphi_{i2}} > 0}}^N (1 + e^{j\varphi_{i2}} (\mathbf{R}_{xx} - e^{j\varphi_{min}}(\mathbf{R}_{xx}) \mathbf{I}_N - a \underline{S}_P \underline{S}_P^H)) + 10 \log_{10} \left(\sum_{\substack{i=1 \\ e^{j\varphi_{i2}} < 0}}^N |e^{j\varphi_{i2}} (\mathbf{R}_{xx} - e^{j\varphi_{min}}(\mathbf{R}_{xx}) \mathbf{I}_N - a \underline{S}_P \underline{S}_P^H)| \right) \quad (5)$$

where $e^{j\varphi_{min}}(\mathbf{R}_{xx})$ is the minimum eigenvalue of the $\mathbf{R}_{xx}$.

- (iv) The effects of the pilot signal are removed from the data covariance matrix, i.e. $\mathbf{R}_{D,I} = \mathbf{R}_{xx} - P_p \underline{S}_P \underline{S}_P^H$.
- (v) Based on the eigenvector decomposition of the matrix $\mathbf{R}_{D,I}$, the projection operator $\mathbf{P}_{E_N} = \mathbf{E}_N (\mathbf{E}_N^H \mathbf{E}_N)^{-1} \mathbf{E}_N^H$ is formed, where the matrix $\mathbf{E}_N$ has as columns the 'noise' eigenvectors of the $\mathbf{R}_{D,I}$. Then by weighting the array with the vector $\underline{w}_p = \mathbf{P}_{E_N} \underline{S}_P / \sqrt{[\underline{S}_P^H \mathbf{P}_{E_N} \underline{S}_P]}$, the DOAs are estimated from the array pattern.
- (vi) Based on the DOAs of (v), the weight vector $\underline{w} = (\mathbf{I}_N - \underline{S}_{I,P} (\underline{S}_{I,P}^H \underline{S}_{I,P})^{-1} \underline{S}_{I,P}^H) \underline{S}_D$ is formed, where $\underline{S}_{I,P} = [\underline{S}_{I1}, \underline{S}_{I2}, \dots, \underline{S}_{IM}]$. In this case, the signal $y(t) = \underline{w}^H \underline{x}(t)$ at the output of the beamformer will be asymptotically, (i.e. $L \rightarrow \infty$), free of interferences.

Results and conclusions: The performance of the proposed algorithm, as compared to conventional techniques, is tested by evaluating the outage probability for a personal communication network consisting of 19 clusters of hexagonal cells. A full load situation with no intra-cell frequency reuse is considered. Only the co-channel interference originated from the first tier is taken into account. The base stations, which are located in the centre of the cells, use uniform circular arrays of $N = 9$ elements. The mobile units use omnidirectional antennas for transmitting and receiving and they are uniformly distributed over the cell area.

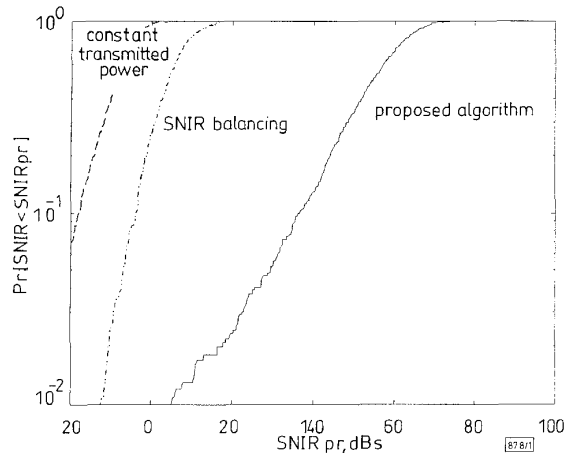


Fig. 1 Outage probabilities for 1000 trials

$C = 1$, $N = 9$, $R = 1500\text{m}$, $L = 100$ snapshots
 $P_t = 5\text{W}$, $P_r = 10^{-16}\text{W}$
 $d_{\text{microelement}} = \lambda/2 = 7.5\text{cm}$

Fig. 1 depicts the performance of the proposed algorithm as compared to the performance of the constant transmitted power and the SNIR-balancing algorithm for cluster size $C = 1$. The results show that the proposed algorithm outperforms the other two techniques by at least 35dBs for an outage probability of 5%. This implies an enormous capacity increase since the whole bandwidth can be reused in every cell of the system.

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24 September 1996

Electronics Letters Online No: 19961445

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Extinction ratio induced crosstalk system penalty in WDM networks

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Indexing terms: Wavelength division multiplexing, Crosstalk

The authors derive an analytical solution for crosstalk-induced penalty in wavelength routers with consideration of the extinction ratio. Both theoretical and experimental results suggest that the extinction ratio can induce a system penalty and should be kept to at least $\sim 15\text{dB}$ to maintain the penalty below 1dB for a total crosstalk level of $\sim 23\text{dB}$.

Introduction: A wavelength router is the key element in scalable wavelength division multiple access (WDMA) networks using wavelength reuse [1]. Both static and dynamic wavelength routers suffer a small but nonzero crosstalk from neighbouring inputs causing severe degradation in system performance. This type of crosstalk, which originates from other input ports carrying identical or similar wavelength with that of the signal, is difficult to eliminate by filtering. Planar SiO_2 waveguide devices demonstrate a crosstalk level of $\sim 25\text{dB}$ [2]. These unwanted signals will beat with the selected optical signal and introduce a new kind of noise in the receiver [3–7]. Previous work on system penalty evaluation either ignored the extinction ratio in their calculations [3, 6, 7], or did not isolate the effect of extinction ratio in their simulations [4, 5]. Here, we have derived a closed form analytical expression for the system penalty, incorporated with extinction ratio r (the power ratio of signal 'mark' to 'space') as a key factor. Both theoretical and experimental results suggest that the extinction ratio should be $> \sim 15\text{dB}$ when several routers are in cascade, assuming a typical crosstalk level of $\sim 23\text{dB}$ in SiO_2 -based routers. We further deduce that if the maximum number of nodes is limited to K for an extinction ratio of 15dB, reducing the extinction ratio to 7dB decreases this number to $K/2$.

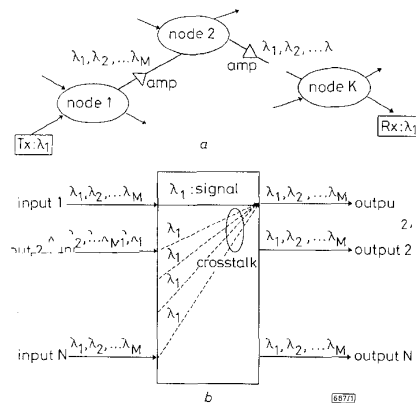


Fig. 1 Analytical model of our WDMA network and homo-wavelength crosstalk at each wavelength router

a Analytical model

b Homo-wavelength crosstalk at each wavelength router

System configuration and analysis: Fig. 1a shows a partial all-optical, wavelength reuse WDMA network. For transmitting λ_i from node 1 to node K , the wavelength router in each node will relate λ_i from node 1 towards node K . Each wavelength router has N input and N output ports and each port carries M wavelengths, $\lambda_1 - \lambda_M$ (Fig. 1b). All M wavelengths at N input ports will be demultiplexed and routed to different outputs according to either the pre-determined (static) or re-configurable (dynamic) pathways. The small but nonzero leakage at the same wavelength, with that of the signal from the other $N-1$ neighbouring input ports, will interfere with the signal.

We adopt a worst-case scenario where the interfering channels have the same polarisation with that of the signal. Without loss of generality, we assume an identical leakage among all ports at the router, that is, the crosstalk to signal ratio $C_i = C_j = C$ for all i and j ports. This ratio can be expressed as $C = P_i/P_s$, where P_i is the power leaking from any other port to the signal port and P_s is the transmitted signal power. We only consider the beat noise originating from the signal and crosstalk because the crosstalk-crosstalk beatings are insignificant [6]. With randomised phases arriving at the detector, the total crosstalk power C_{tot} after K nodes can be simplified [4–6] as $C_{\text{tot}} = CK(N-1)$. We assume that in our network, the signal attenuation in the fibre and wavelength router is fully compensated for by the in-line gain-flattened optical amplifiers, and the noise from the amplifiers is neglected.

The Q -factor for the optimised decision threshold [6] can be written as

$$Q = \frac{2i_p(r-1)/(r+1)}{\sqrt{I_{s-\text{talk},1}^2 + I_{th}^2} + \sqrt{I_{s-\text{talk},0}^2 + I_{th}^2}} \quad (1)$$

where i_p is the average photocurrent in the presence of crosstalk channels, I_{th} is the circuit thermal noise; $P_{s-\text{talk},1}$ and $P_{s-\text{talk},0}$ are