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(71) Applicant: **ALCATEL LUCENT**
92100 Boulogne-Billancourt (FR)

(72) Inventors:
- **Venkateswaran, Vijay**
  Dublin, Dublin 15 (IE)
- **Dartois, Luc**
  91620 Nozay Route de Villejust (FR)
- **Boyon, Benoit**
  78141 Velizy Saulnier, BP 57 (FR)

(74) Representative: **Greenwood, Matthew David et al**
Bryers LLP
7 Gay Street
Bath BA1 2PH (GB)

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(54) Antenna feed and method of configuring an antenna feed

(57) An antenna feed for generating signals for an active antenna array including multiple antennas for transmitting transmission beams having selected tilt angles, the antenna feed comprising a digital beamformer operable to receive an input broadband signal and to generate a plurality N of output broadband signals, each having an associated phase and amplitude, a plurality N of transmission signal generators, each operable to receive one of the said plurality N of output broadband signals and to generate a corresponding plurality N of first RF signals, and a feed network comprising a power split network operable to receive the said plurality N of first RF signals and to generate a plurality P of second RF signals, each having an associated amplitude and an associated phase shift applied using a variable phase shifter of a phase shift network, the feed network operable to supply respective ones of the plurality P of second RF signals to respective ones of the multiple antennas, M, of the said active antenna array, where P≥M>N.

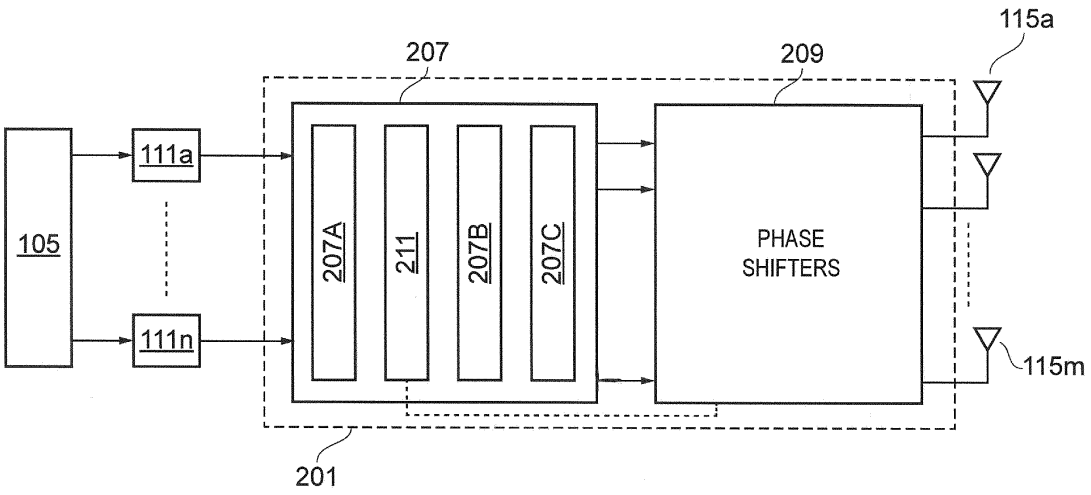


FIG. 2

EP 2 975 688 A1

Description

TECHNICAL FIELD

[0001] The present invention relates to an antenna feed and a method of configuring an antenna feed.

BACKGROUND

[0002] Antenna feeds are known. In order to support the spatial separation of signals from a static transmitter of, for example, a wireless telecommunications network, it is known to provide an array of antennas and utilise beamforming techniques. In particular, a signal may be provided which is subjected to varying phase and amplitude to generate multiple signals, each of which is provided to one of the antennas in the array in order to perform adaptive beamforming, virtual sectorisation and spatial multiplexing within a given cell. Such antenna arrays are typically referred to as active antenna arrays. These arrays significantly increase the coverage and capacity of a cellular network. However, the introduction of multiple transceivers at the transmitter typically increases the cost of the radio frequency (RF) front-end of a cellular base station.

SUMMARY

[0003] According to an example, there is provided an antenna feed for generating signals for an active antenna array including multiple antennas for transmitting transmission beams having selected tilt angles, the antenna feed comprising a digital beamformer operable to receive an input broadband signal and to generate a plurality N of output broadband signals, each having an associated phase and amplitude, a plurality N of transmission signal generators, each operable to receive one of the said plurality N of output broadband signals and to generate a corresponding plurality N of first RF signals, and a feed network comprising a power split network operable to receive the said plurality N of first RF signals and to generate a plurality P of second RF signals, each having an associated amplitude and an associated phase shift applied using a variable phase shifter of a phase shift network, the feed network operable to supply respective ones of the plurality P of second RF signals to respective ones of the multiple antennas, M, of the said active antenna array, where $P \geq M > N$. The feed network can be an RF beamformer in which the power split network comprises multiple power split stages to receive the plurality N of first RF signals, and generate the said plurality P of second RF signals using power splitters at respective ones of the stages of the network, wherein respective ones of the power slitters are operable to divide a received RF signal over at least two separate paths to provide respective split signals, wherein the power splitter are operable to provide the split signals to a subsequent stage of the RF beamformer to generate the plurality P of second RF signals. The variable phase shifter of the phase shift network can be logically positioned in between two of the multiple power split stages of the power split network. The phase shift network can further comprise a fixed phase shifter to apply a predetermined phase shift to respective ones of the plurality P of second RF signals. The fixed phase shifter can be logically positioned after the power split network. The plurality P of second RF signals can be supplied to respective antennas of the plurality M of antennas of the said active antenna array to transmit a transmission beam with a requested beam tilt angle. The feed network can comprise multiple feed instances each operable to provide selected ones of the P signals to the plurality M of antennas to transmit multiple different transmission beams simultaneously. Respective ones of the multiple feed instances can be operable to simultaneously transmit the multiple different transmission beams with respective different beam tilt angles. The digital beamformer can be an adaptive beamformer operable to weigh the plurality N of output broadband signals to provide a desired main lobe and side lobe distribution for a transmission beam.

[0004] According to an example, there is provided a method of configuring an antenna feed for generating signals for an active antenna array including multiple antennas for transmitting transmission beams having selected tilt angles, the method comprising receiving an input broadband signal at a digital beamformer and generating a plurality N of output broadband signals, each having an associated phase and amplitude, receiving one of the said plurality N of output broadband signals at each one of a plurality N of transmission signal generators and generating a corresponding plurality N of first RF signals, receiving the said plurality N of first RF signals at a feed network comprising a power split network and generating a plurality P of second RF signals, each having an associated amplitude, applying a phase shift to the said plurality N of first RF signals using a variable phase shifter of a phase shift network to apply, and supplying respective ones of the plurality P of second RF signals to respective ones of the multiple antennas, M, of the said active antenna array, where $P \geq M > N$. The method can further include receiving the plurality N of first RF signals in multiple power split stages of the power split network, generating the said plurality P of second RF signals using power splitters at respective ones of the stages of the network by dividing a received RF signal over at least two separate paths, and providing those to a subsequent stage to generate the plurality P of second RF signals. The method can further include logically positioning the variable phase shifter of the phase shift network in between two of the multiple power split stages of the power split

network. The method can further include applying a predetermined phase shift to respective ones of the plurality P of second RF signals using a fixed phase shifter of the phase shift network. The method can further include logically positioning the fixed phase shifter after the power split network. The method can further include supplying the plurality P of second RF signals to respective antennas of the plurality M of antennas of the said active antenna array, whereby to enable transmission of a transmission beam with a requested beam tilt angle.

BRIEF DESCRIPTION OF THE DRAWINGS

[0005] Embodiments will now be described, by way of example only, with reference to the accompanying drawings, in which:

Figure 1 is a schematic representation of a general architecture for an antenna feed according to an example;

Figure 2 is a schematic representation of the antenna feed network of figure 1 according to an example;

Figure 3 shows two graphs comparing the beam pattern performance for antenna feed networks designed with four transceivers with (figure 3a) and without (figure 3b) combiners

Figure 4 is a schematic representation of a configuration of an antenna feed network based active antenna array showing an arrangement with combiners (left hand side of figure 4) and a low-complexity approach without combiners according to an example (right hand side of figure 4);

Figure 5 is a schematic representation of a hybrid phase shifter arrangement according to such an example;

Figure 6 is a schematic representation of a multiple beam arrangement according to an example;

Figure 7 shows two graphs comparing the beam pattern performance for a multiple beam arrangement according to an example;

Figure A1 is a schematic representation of an antenna feed according to an example;

Figure A2 is a schematic representation of an antenna feed;

Figure A3 illustrates decomposition of a divider bank into three stages of power dividers according to an example; and

Figure A4 illustrates factorization into multiple stages of directional couplers.

DESCRIPTION

[0006] Example embodiments are described below in sufficient detail to enable those of ordinary skill in the art to embody and implement the systems and processes herein described. It is important to understand that embodiments can be provided in many alternate forms and should not be construed as limited to the examples set forth herein.

[0007] Accordingly, while embodiments can be modified in various ways and take on various alternative forms, specific embodiments thereof are shown in the drawings and described in detail below as examples. There is no intent to limit to the particular forms disclosed. On the contrary, all modifications, equivalents, and alternatives falling within the scope of the appended claims should be included. Elements of the example embodiments are consistently denoted by the same reference numerals throughout the drawings and detailed description where appropriate.

[0008] The terminology used herein to describe embodiments is not intended to limit the scope. The articles "a," "an," and "the" are singular in that they have a single referent, however the use of the singular form in the present document should not preclude the presence of more than one referent. In other words, elements referred to in the singular can number one or more, unless the context clearly indicates otherwise. It will be further understood that the terms "comprises," "comprising," "includes," and/or "including," when used herein, specify the presence of stated features, items, steps, operations, elements, and/or components, but do not preclude the presence or addition of one or more other features, items, steps, operations, elements, components, and/or groups thereof. Unless otherwise defined, all terms (including technical and scientific terms) used herein are to be interpreted as is customary in the art. It will be further understood that terms in common usage should also be interpreted as is customary in the relevant art and not in an idealized or overly formal sense unless expressly so defined herein.

[0009] Broadly speaking, for an active antenna array that can be used to beam form signals to be transmitted for a

telecommunications network, it is difficult to provide a simplified feed network which can efficiently generate signals through adaptive beamforming for a range of different tilt angles. In particular, in order to efficiently provide beamforming for a range of different tilt angles it is typically required to provide a single transceiver coupled with each antenna of an antenna array, but this is not always possible, creates additional weight and increases cost and power consumption, which is undesirable, particularly when the transceivers are co-located with the antennas on a broadcast mast of the network.

[0010] Accordingly, in an example, an arrangement is provided in which fewer transceivers are utilized, and signals generated by those fewer number of transceivers are provided to an antenna array via an antenna feed network. In a particular example, fewer transceivers than the number of antennas in the antenna array are provided. The transceivers can be driven by a digital signal processor or digital beamformer which receives an input broadband signal to be transmitted by the antenna array with a requested tilt angle. In an example, the tilt angle may be provided separately or as part of the broadband signal.

[0011] The digital beamformer receives the input broadband signal and generates a plurality of output broadband signals, each having an associated phase and amplitude. Multiple transmission signal generators, each operable to receive one of the plurality of output broadband signals can then be used to generate multiple corresponding first RF signals. A feed network comprising a power split network is operable to receive the first RF signals and generate multiple second RF signals, each having an associated amplitude and an associated selected phase shift that can also be applied applied using a variable phase shifter of a phase shift network. The feed network supplies respective ones of the second RF signals to respective antennas of a plurality of antennas of the active antenna array.

[0012] Figure 1 is a schematic representation of a general architecture for an antenna feed according to an example.

[0013] A digital input broadband signal 101 is provided to a digital beamformer (signal processor) 105. The digital signal 101 is a broadband signal provided by a telecommunications network 103. Also provided to the digital beamformer 105 is a desired tilt angle 107. It will be appreciated that the desired tilt angle may be encoded in the digital signal 101.

[0014] The digital beamformer 105 generates a plurality N of output broadband digital signals 109a-N, one for each of a plurality of transmission signal generators, such as transceivers, 111a-N. Each broadband signal 109a-N can have a differing amplitude and/or phase shift, depending on the tilt angle 107.

[0015] Each transceiver 111a-N generates an RF signal RF1 to RFN thereby forming a plurality N of first RF signals which are provided to an antenna feed network 113. The antenna feed network 113 generates a plurality P of second RF signals, RFO1-m, each of which is provided to an associated antenna 115a to 115M. In an example, the number M of antennas exceeds the number N of transceivers.

[0016] Accordingly, a radiofrequency antenna feed network is used to connect a reduced number of transceivers with an increased number of antennas. Different instantiations provide required beam patterns, sectorisation and sidelobe levels which are typically only seen with arrangements where a dedicated and separate transceiver chain is provided for each antenna within the antenna array.

[0017] Figure 2 is a schematic representation of the antenna feed network of figure 1 according to an example. The antenna feed network 201 feeds signals from each transceiver 111a-N to a set of antennas 115a-m. According to an example, the antenna feed network 201 can be broadly decomposed into multiple RF filter banks, depending on the primary function of each bank. More particularly, the antenna feed network 201 comprises a bank of power dividers 207 coupled with a bank of phase shifters 209.

[0018] Each of the banks may be characterized into one or more multiple stages. For example, the bank of power dividers 207 can be characterized into 3 stages 207A, 207B, 207C. Stage 207A receives signals from the transceivers 111a-N and generates an increased number of RF signals. This increased number of RF signals is provided to stage 207B, which in turn generates an increased number of RF signals and provides these to stage 207C. Generally, the bank of power dividers 207 generates P RF signals, where P is greater than or equal to N. That is, $P \geq M > N$ according to an example.

[0019] Each of the P signals is provided to a bank of phase shifters 209, which provides interconnecting wires to reorder the sequence of the signals received from the bank of power dividers 207 and applies a required phase shift to each of those signals. The bank of phase shifters 209 outputs P RF signals. The bank of phase shifters 209 are fixed phase shifters according to an example. That is, the phase shifters 209 shift an input signal by a fixed amount.

[0020] According to an example, a variable phase shifter can be provided. The variable phase shifter can shift the phase of an input signal by a variable amount that can be selected or tuned as desired. A variable phase shifter 211 can be considered as part of the phase shift network 209, but can be logically positioned in between two of the multiple power split stages of the power split network 207. For example, in figure 2, the variable phase shifter 211 is provided in a logical position between power split stages 207A and 207B. Other positions within the power dividers 207 are possible as will be appreciated. In an example, the output of stage 207C is input to the phase shifters 209, which are not variable. The output of the phase shifters 209 is input to the antennas 115a-m.

[0021] The antenna feed architecture as described with reference to figures 1 and 2 provides a simplified, reduced cost and reduced power consumption approach to provide adaptive beamforming of the transmission beam transmitted

by the antenna array. It will be appreciated that the phase shifts applied by the digital beamformer 105, the power division ratios applied by the bank of power dividers 207 and the interconnects and phase shifts applied by the bank of phase shifters 209 may be calculated in any number of different ways. An approach to generating these parameters according to an example is described below in more detail.

[0022] The amplitude and phase of the RF signal output by each transceiver 111a-N is different for different sectorisation tilt angles, and an antenna feed according to an example connects a reduced number of transceivers to an array with an increased number of antennas without the use of couplers. The transceivers typically contain adaptive beamformers, and in combination with the feed network and the antenna array, generate the desired beam to satisfy the coverage as well as capacity requirements of most sectors, for example, macro cell wireless networks. In particular, the feed network is primarily a fixed beam former with a variable phase shift component, and in combination with the transceivers and digital signal processor achieves adaptive beamforming.

[0023] Subsequently, the adaptive beamforming leads to sectorisation and enhanced coverage at a fraction of the complexity and cost of arrangements where a separate transceiver chain is provided for each antenna.

[0024] The power dividers 207 distribute the transceiver power amplifier outputs with the appropriate power ratios towards multiple antennas. According to an example, each bank of power dividers can be composed of multiple stages of Wilkinson power dividers and each stage of power dividers comprises at least N Wilkinson power dividers. The power dividers used in an example are 3-port networks, with 1 input and 2 outputs. Each of these dividers are designed to be either a balanced divider (providing a 3dB ratio at each output) or an unbalanced divider.

[0025] In an example, the number of stages of power dividers is limited to 3 in order to minimize the overall losses in the network. To achieve a specific beam pattern, each signal RF1 to RFN can be divided into 2 signals using a Wilkinson divider at stage 207A. This action is repeated subsequently at each stage such that the power divided signals output by the bank of power dividers and their power ratios enable the required beam patterns at different tilt angles.

[0026] The phase shifters 209 (including variable phase shifter 211 which is logically positioned in between stages of power dividers 207) shift the phase of the power divided signals to achieve a desired beam shape.

[0027] Phase shifters 209 can be, for example, transmission lines, micro-strip lines or other phase shifting devices. The length of these lines is dictated by the phase shifts required, which in turn is estimated to achieve specific beam patterns. The bank of phase shifters 209 can also contain an interconnecting matrix of wires. The function of the interconnecting matrix of wires is to ensure that the rest of the network has no requirement for any further crossovers or interconnects and to ensure that the overall number of crossovers and interconnects in the entire network is reduced to a minimum.

[0028] Figure 3 shows two graphs comparing the beam pattern performance for antenna feed networks designed with four transceivers with (figure 3a) and without (figure 3b) combiners. In an example, an antenna feed network satisfies the spatial mask demands in existing 3GPP and LTE standards for a beam tilt range $R_\theta \in \{1°, \cdots, 11°\}$. Note that the setup in figure 3b without combiners fully satisfies the demands. This is true for all beamtilts in R_θ. Accordingly, a low-complexity setup without combiners according to an example will always satisfy the required spatial mask in existing 3GPP and LTE standards for a beam tilt range $R_\theta \in \{1°, \cdots, 11°\}$.

[0029] Figure 4 is a schematic representation of a configuration of an antenna feed network based active antenna array showing an arrangement with combiners (left hand side of figure 4) and a low-complexity approach without combiners according to an example (right hand side of figure 4). The arrangement according to an example does not have cable crossovers and does not have combiners, thus significantly reducing the implementation complexity of the overall setup and minimizing substrate loss due to combiners. For simplicity, we assume that all PAs are uniform, however such a setup can be used with non-uniform PAs.

[0030] The arrangement depicted on the right hand side of figure 4 is an arrangement according to an example in which the number of transceivers is N_{pa}. This can provide a starting point for an N_{pa} - 1 transceiver setup that can be used in combination with a set of non-uniform power amplifiers (PAs) and phase shifters.

[0031] Figure 5 is a schematic representation of a hybrid phase shifter arrangement according to such an example. A set of passive phase shifters 501 that apply a fixed phase shift to an input signal are connected to the central PAs 503 to achieve enhanced beamforming while operating with further reduced number of transceivers (transceivers 1 to 3).

[0032] In the example of figure 5, output signal 506 from transceiver 2 is divided into 2 branches 505, 507 that are applied with variable phase shifts using variable phase shift components 509, 511. In an example, components 509, 511 can be provided by a single variable phase shift component. The output (513, 515) of components 509, 511 is used with two PAs 517, 519 respectively.

[0033] An alternative arrangement according to an example is to place the power dividers after the central PA, and subsequently use a high power PA (say 20 W). Note that such a setup improves the power efficiency of the overall system, since it is reasonably easier to design a high power PA with higher efficiency than a set of low-power PAs.

[0034] The introduction of non-uniform PAs provides an additional degree of freedom to optimize the overall antenna feed network- digital beamformer (AFN-DBF) arrangement; initially the amplitude and phase could be tapered during the AFN design phase as well as amplitude and phase tapering at the DBF. The introduction of non-uniform PAs provides

an additional degree of freedom with non-uniform PAs in the design phase. This approach can be generalized to using PAs with non-uniform output power, non-uniform antenna elements with non-uniform spacing between each other, and antenna elements with non-uniform gain response. All these techniques will further enhance the beam pattern performance while operating at reduced complexity.

[0035] A hybrid AFN with passive phase shifter and N_{pa} - 1 transceivers according to an example can perform close to that of an optimal AFN arrangement with N_{pa} transceivers. However, one must note that the AFN intended for LTE can be embedded with MIMO and multi-beam capability. Multiple beams are realized using multiple antenna array columns, where each column is connected to an AFN and subsequently to a set of transceivers.

[0036] Passive phase shifts using electrical tilts cannot be applied for individual beams in a multi-beam AFN, which is a limitation whenever variable phase shifts using electrical or electromechanical means are used due to physical and structural limitations. They can only be applied for all columns simultaneously. Such a limitation essentially means that the hybrid phase shifter AFN and DBF setup with N_{pa} - 1 transceivers leads to a sub-optimal performance when used multiple beams on multiple columns. However, according to an example this performance degradation can be accounted for by joint optimization of the passive phase shifter and the DBF weights to achieve the necessary phase shift.

[0037] In this connection, figure 6 is a schematic representation of a multiple beam arrangement according to an example. To get simultaneous beams, multiple AFN's W_1 and W_2 are jointly designed and provide signals to multiple antenna arrays 601, 603 to achieve desired beam tilts for the multiple antenna array. Corresponding digital beamformers 605, 607 are provided as depicted in figure 6. As will be appreciated, more than two AFNs and antenna arrays may be used.

[0038] In this case, the objective is to jointly design the phase shifter, specified by the variable phase shifter matrix Φ as well as the DBF weights $\vartheta(\theta_1)$ and $\vartheta(\theta_2)$. In this case, the DBFs are $N_{pa} \times 1$ vectors and the AFNs are $N_t \times N_{pa}$ matrices. The phase shifter Φ is a $N_{pa} \times N_{pa\text{-}1}$ matrix, where the passive phase shift values are embedded inside. Note that in an example there is only one phase shifter for a 2 x 2 MIMO setup with 2 AFNs. For such as multi-AFN case, the spectral masks can be represented:

$$[\ \Delta_1\ \Delta_2\ ] = A(\theta)\ [W_1, W_2]\ \Phi\ [\vartheta_1(\theta_d); \vartheta_2(\theta_d)]$$

[0039] The initial $N_{pa,} \times 1$ DBF values are estimated independently from the single beam case, and the individual phase shifts are extracted from them. If this value is designated as initial phase shift Φ_0, it can be estimated by a simple LS estimate:

$$\Phi_0 = [W_1, W_2]^{\dagger} A(\theta)^{\dagger} [\ \Delta_1\ \Delta_2\ ]\ [\vartheta_1(\theta_d); \vartheta_2(\theta_d)]^{\dagger}$$

[0040] Note that Φ_0 does not correspond to the phase shift constraints as required by the macro-array setup, such as a diagonal matrix with only phase shifts and the phase shift matrix is approximated or re-designed to account for these constraints and denoted as Φ.

[0041] Subsequently, for a given $W_{i,}$ $i \in \{1, 2\}$ and Φ, the DBF is re-estimated as in a single beam case for a combined

$$\vartheta_1(\theta_d) = (W_i\ \Phi)^{\dagger}\ A(\theta)^{\dagger} [\ \Delta_1\ \Delta_2\ ]$$

[0042] In practice, the LS estimate is replaced by interior point algorithms as the original AFN and DBF design.

[0043] More details on a methodology used for arriving at a particular antenna feed design can be found at Annex A, which considers a single dimensional feeder network. A two dimensional network can be composed of two-feeder networks that are connected by a phase shift matrix. This network can be designed to satisfy two spectral masks Δ_1 and Δ_2. Accordingly, two digital beamformers, ϑ_1 and ϑ_2, can be used to simultaneously obtain two different beams using the matrix and W_1, W_2.

[0044] Figure 7 is a pair of graphs illustrating the performance of an arrangement such as depicted in figure 6. More particularly, figure 7a illustrates beam pattern performance of a 3 x 12 hybrid AFN setup with combiners, and figure 7b illustrates a beam pattern performance of 3 x 12 hybrid AFN setup according to an example without combiners.

[0045] According to an example, the number of transceivers used in an antenna feed can be reduced, thereby reducing the overall cost in an active antenna array (AAA) setup while making sure that the performance of the AAA setup is

close to that expected in a modular AAA solution. This primarily requires designing an RF feeder network (also referred to as an RF beamformer) to achieve the desired performance while accounting for all the constraints and challenges observed in a cellular base station.

[0046] An RF feeder network according to an example operates in combination with a digital beamformer (DBF) to achieve joint RF-digital beamforming. A generic antenna feed connected to an AAA macro-cell base station in order to provide an arbitrary set of beamtilts and to enable vertical sectorisation, while operating with a reduced number of transceivers is therefore provided. An AAA setup containing N_t = {10, 12, 14} antennas can be connected to a DBF containing N_{pa} = {2, 3, 4} transceivers through the antenna feeder network (AFN).

[0047] The RF feeder network (RF beamformer) is typically a fixed beamformer, and is used in combination with an adaptive DBF to provide a joint digital and RF beamforming focusing signals towards various sectors within a given macro-cell. This represents a joint DBF-AFN design. In an example, the DBF can be implemented in an FPGA, where the DBF weights can be modified for each beamtilt.

[0048] Practical constraints for efficient operation in cellular base stations such as tapering the power amplifier inputs to specific power levels in order to satisfy PA linearity and enhance their efficiency can be considered, and, according to an example a joint DBF-AFN design maximizes the signal to noise ratio (SNR) of the signals from macro-cell base station towards the UE for the specified sector.

[0049] An optimal Joint DBF-AFN design provides optimal beam-pattern performance, where in an example optimality is specified by signal energy or effective isotropic radiated power (EIRP) from the antenna arrays along the specified sector. This can ensure that UEs present in a specific sector will be able to clearly receive signals with improved SNR values. Side lobe levels outside the specified sector can be suppressed, and this can ensure that signal energy from a base station is not wasted or radiated towards any other sector. This ensures that the interference levels observed at user equipment present in other sectors or neighboring sectors is minimised.

[0050] According to an example, the RF-AFN can be implemented using microwave components such as power dividers (Wilkinson dividers), micro-strip lines, suspended strip lines, and so on. The output of PAs (and transceivers) can be divided into an increasing number of branches using a bank of power dividers. Subsequently, these power divided branches can be shifted by a bank of phase shifters to achieve beamforming. As noted above, a variable phase shift device can be used to apply a selected phase shift before, after or between power divider stages. The power dividers can be either balanced or unbalanced power dividers, and the phase shifts can be implemented either using micro-strip lines or suspended di-electrics/suspended striplines and subsequently connected to the antennas.

[0051] An antenna feed according to an example minimises the number of cross-overs of the signal paths between PAs and antennas, thus reducing the implementation complexity, and the feed network can be optimized by balancing some or all of the dividers to minimize loss. A remaining set of unbalanced power dividers can be used to account for desired optimal beam pattern performance.

[0052] According to an example, a multi-beam multi-column AFN can be provided. Multiple columns of either similar or different AFNs can be used with multiple transceivers in each column to realize a 4x4 MIMO or 2x2 MIMO for example, as required in LTE, thereby improving capacity/coverage. Such an arrangement provides MIMO, spatial multiplexing and space-time coding with multiple columns of AFN, while providing spatial/vertical sectorization within a single AFN column. Such a 2-dimensional multi-input multi-beam (MIMB) system can improve interference levels as well as overall capacity.

[0053] An AFN according to an example will lead to optimal beam pattern performance whenever it is used independently to realize individual beams for a specific downtilt or is used jointly to realize multi-user or MIMO beamforming.

[0054] A hybrid phase shifted AFN can be provided according to an example. One of the active transceivers and the DBF can be replaced by one or more passive mechanical or electromechanical phase shifters. Such configurations can be decided in a deployment phase. An optimal AFN plus a variable passive phase shifter can be used for optimal beamforming with reduced AAA. Such an approach of passive phase shifters and reduced AAA is referred to as hybrid-AFN. The hybrid-AFN arrangement can also be designed for use with PAs having non-uniform power output levels. For instance, the AFN can have a PA radiating at a higher power compared to other PAs. In this case, signals from this PA can be divided with a Wilkinson divider and subsequently followed by a variable hybrid phase shifter. Subsequently, it can be fed into the previously designed AFN. The performance of a hybrid- AFN setup with N_h hybrid phase shifters and N_{pa} transceivers converges to an optimal solution with an AFN configuration having $N_{pa} + N_h$ DBF active transceivers. This results in cost savings up to N_h transceivers.

[0055] According to an example, the optimal hybrid phase AFN can be designed for arbitrary N_{pa} and N_h, with a performance as noted above. In an example, the number of power divider stages containing 3-port Wilkinson dividers is limited by $\lceil \log_2 N_t/N_{pa} \rceil$, where $\lceil . \rceil$ specifies the ceil function. Thus the substrate losses are also limited by this factor. Within a given bank of power dividers connecting a transceiver with a set of antennas the propose AFN design does not contain any cross-over wires, thus reducing the complexity.

[0056] The optimal AFN connections, balanced and unbalanced power divider weights, phase shifts, and DBF phase shifts as well as amplitude weights can be estimated using a constrained interior point algorithm. For details of the interior

point algorithm to estimate beamformer weights, please refer to the Appendix of Annex A. The optimisation technique includes a specific set of constraints to limit the EIRP within the main sectors, to limit the interference and power levels along the undesired sectors, and to taper the power levels of signals at PA inputs.

[0057] The optimisation constraints are represented either as linear equalities or inequalities and interior point convex optimization methods are used to reach the optimal solution. Alternatively, these values can also be designed using stochastic gradient descent algorithms. The algorithms designing RF-AFN and DBF always account for antenna spacing, type of antennas, frequency band of operation, grating lobe and side lobe requirements, beamtilt range required, desired number of transceiver, and so on to achieve optimal beamforming. The algorithm can be applied for any combination of number of transceivers and number of antennas; and the solution will always converge to an optimal beam pattern performance. The approach can be generalized to using PAs with non-uniform output power, antenna elements with non-uniform spacing between each other and antenna elements with non-uniform gain response. All these techniques will further enhance the beam pattern performance while operating at reduced complexity.

Annex A

[0058] *Notation:* Lower and upper case bold letters denote vectors and matrices. An over-tilde ($\tilde{(\cdot)}$) denotes RF signals, while time indexes (.) and [.] respectively denote analog and digital signals. Superscripts $(.)^T$, $(.)^H$, $(.)^\dagger$ and $\|.\|$ respectively denote transpose, Hermitian transpose, pseudo-inverse and Frobenius norm operations. The matrix $\mathbf{I}_K$ denotes an identity matrix while 0 and 1 respectively denote matrix/vectors of zeros and ones.

I. SYSTEM MODEL AND PROPOSED ARCHITECTURE

A. Data Model

[0059] Consider an $N_t \times 1$ vector $\tilde{\mathbf{x}}(t)$ denoting the RF signal radiated from the antenna array at time t. In the modular AAA setup, $\tilde{\mathbf{x}}(t)$ is obtained using an $N_t \times 1$ DBF vector $\mathbf{u}(\theta_d) = [u_1(\theta_d), \cdots, u_{N_t}(\theta_d)]^T$ operating on a data stream $s[k]$ at time $t = kT$, and followed by N_t 'RF chains'. The DBF vector $\mathbf{u}(\theta_d)$ is usually designed θ_d. Thus

$$\tilde{\mathbf{x}}(t) = \mathcal{RF}\{\mathbf{x}[k]\}$$

where

$$\mathbf{x}[k] = \mathbf{u}(\theta_d)s[k].$$

Typically, $\mathbf{u}(\theta_d)$ is designed to produce a mainlobe centered at θ_d.

[0060] Consider a setup with N_{trx} transceivers connected to N_t radiating elements through a passive AFN (For example, $N_t = 11$ and $N_{trx} = 5$). Details of the AFN will be explained in Sec. III. In this case an $N_{trx} \times 1$ DBF vector $\vartheta(\theta_d) = [\vartheta_1(\theta_d), \cdots, \vartheta_{N_{trx}}(\theta_d)]^T$ operates on the data stream s[k], followed by N_{trx} digital-RF transformation blocks and the AFN matrix W, and subsequently radiated as a $N_t \times 1$ vector

$$\tilde{\mathbf{x}}_r(t) = \mathbf{W}\tilde{\mathbf{y}}(t) \qquad (1)$$

where

$$\tilde{\mathbf{y}}(t) = \mathcal{RF}\{\mathbf{y}[k]\}$$

and

$$\mathbf{y}[k] = \boldsymbol{\vartheta}(\theta_{\mathrm{d}})s[k].$$

[0061] We refer to the setup as shown in Fig. A1(b) as a partially-adaptive beamformer, since the AFN is estimated for a specific architecture at the outset and kept fixed. Subsequently, the DBF $\mathcal{G}(\theta_{\mathrm{d}})$ is adaptively designed for each beamtilt. Assuming the antenna elements are equally spaced, the array response can be modeled as a function of angle θ_i as

$$\mathbf{a}(\theta_i) = g(\theta_i)\begin{bmatrix} 1 \\ e^{j\frac{2\pi}{\lambda}\delta\cos(\theta_i)} \\ \vdots \\ e^{j\frac{2\pi}{\lambda}\delta(N_{\mathrm{t}}-1)\cos(\theta_i)} \end{bmatrix}$$

where δ is the spacing between adjacent antennas, λ is the wavelength in meters and $g(\theta_i)$ is the antenna characteristic. Note that the 3GPP transmission standard allows antenna characteristic $g(\theta_i)$ with a 3-dB beamwidth of either 65° or 110°.

B. Modular AAA Architecture

[0062] As mentioned before, our objective is to reduce the number of transceivers and thereby reduce the cost and power consumed by the antenna array. The performance of the reduced DBF-AFN setup is compared with a modular AAA setup having $N_{\mathrm{trx}} = N_{\mathrm{t}}$ transceivers. Let $\mathbf{u}(\theta_{\mathrm{d}})$ be a $N_{\mathrm{t}} \times 1$ beamforming vector in the modular AAA setup operating on $s[k]$ to achieve a mainlobe at a user beamtilt θ_{d}.

[0063] The performance of the modular AAA setup depends on the channel capacity, as well as the adaptive sectorization of the beamformer $\mathbf{u}(\theta_{\mathrm{d}})$. This performance requirement is specified by the operational constraints and is referred to in this paper as a spectral mask $\Delta_{\theta_{\mathrm{d}}}$. The constraints that make up $\Delta_{\theta_{\mathrm{d}}}$ are explained in Sec. I-C. The spectral mask includes information regarding the gain and directivity along θ_{d} as well as the SLLs. In modular AAA architecture, the objective is to design the adaptive beamformer $\mathbf{u}(\theta_{\mathrm{d}})$ minimizing the overall mean-squared error:

$$\mathbf{u}_0 = \arg\min_{\mathbf{u}(\theta_{\mathrm{d}})} \|\boldsymbol{\Delta}_{\theta_{\mathrm{d}}} - \mathbf{A}(\theta)\mathbf{u}(\theta_{\mathrm{d}})\| \qquad (2)$$

where

$$\mathbf{A}(\theta) = \begin{bmatrix} \mathbf{a}^T(-\pi) \\ \vdots \\ \mathbf{a}^T(\pi)] \end{bmatrix}$$

is a $N_\theta \times N_{\mathrm{t}}$ matrix obtained by stacking the array response vectors. A well known approach to estimate $\mathbf{u}(\theta_{\mathrm{d}})$ in (2) is $\mathbf{u}_0 \approx \mathbf{A}_d^{\dagger}\boldsymbol{\Delta}_{\theta_{\mathrm{d}}}$ using the least-squares approach [10]. However, that approach would not lead to optimal solutions that consider microwave component design, would not account for the linear range of PA operation and might not always satisfy the required SLLs.

[0064] In this paper, we will design the modular AAA beamformer weights subject to operational constraints using iterative convex optimization techniques [11]. The operational constraints include desired power levels and 3-dB beamwidth along θ_{d}, SLLs and dynamic range of PA output. The optimal modular AAA solution using the convex optimization techniques has been shown in [12], [13], so the details of the beamformer design for modular AAA are omitted in this section.

[0065] The modular AAA architecture is not among the main contributions of this paper, however it will serve as our

reference design for performance comparisons. The modular AAA weights can also be obtained as a special case of joint DBF-AFN design approaches of Sec. II-B, by setting **W** = I and $N_t = N_{trx}$.

C. AFN Architecture: Problem Formulation

[0066] Our aim is to jointly design the optimal AFN matrix W and DBFvector $\vartheta(\theta_d)$ to satisfy the desired set of spectral masks ($\Delta_{\theta d}$). Unlike the modular AAA beamformer $\mathbf{u}(\theta_d)$, the AFN is fixed and can only satisfy a preset range of beamtilts $\mathcal{R}_\theta = \{\theta_1, \cdots, \theta_{N_\theta}\}$, where N_θ corresponds to the number of sectors and $\theta_d \in \mathcal{R}_\theta$. The fixed AFN and adaptive $\vartheta(\theta_d)$ must be designed to satisfy the spectral masks corresponding to all beamtilts i.e. $\Delta_{\theta d}$, $\forall \theta_d \in \mathcal{R}_\theta$.

[0067] If we do not consider the constraints regarding the dynamic range of the PA, this problem can be seen as a least squares (LS) fit jointly designing {W, $\vartheta(\theta_d)$} to minimize the overall mean-square error (MSE):

$$\{\mathbf{W}, \boldsymbol{\vartheta}(\theta_d)\} = \arg\min_{\boldsymbol{\vartheta}(\theta_d)} \|\boldsymbol{\Delta}_{\theta_d} - \mathbf{A}(\theta)\mathbf{W}\boldsymbol{\vartheta}(\theta_d)\| \quad \forall \theta_d \in \{\theta_1, \cdots, \theta_{N_\theta}\} \tag{3}$$

We would like to keep the overall number of transceivers N_{trx} to a minimum, in order to minimize the overall cost. The optimization of the cost function (3) includes the following constraints:

[C1] The number of transceivers N_{trx} is restricted to a minimum.
[C2] The SLLs are constrained to be at-least 15 dB below the mainlobe. This is to ensure that most of the power is directed towards the desired sector, as well as to limit the interference to neighboring cells/sectors. The 3-dB beamwidth is constrained to be less than 4° for a macro-cell and less than 15° for a small cell setup.
[C3] The AFN design and spectral mask Δ_d must satisfy the constraints [C1] and [C2] over the entire downtilt range $\theta_d \in \mathcal{R}_\theta$ where $\mathcal{R}_\theta = \{\theta_1, \cdots, \theta_{N_\theta}\}$.
[C4] The power amplifiers (PAs) must operate in a linear mode [14] and their output power must be limited to a range 0 dB $\leq |\vartheta_k(\theta_d)|^2 \leq$ 1 dB.
[C5] The number of AFN stages must be limited to 3. This is done to ensure low-complexity networks and minimize propagation of loss in the network.
[C6] The input signals at the last stage of the AFN must be matched to account for insertion loss.

[0068] The objectives are to (1) design the AFN and DBF weights to constrain the beampattern satisfying the spectral mask $\Delta_{\theta d}$, as well as to restrict the dynamic range of PA output and (2) translate the designed AFN weights to a microwave feeder circuit that provides the desired beam pattern while minimizing the insertion loss. We proceed with the AFN design in the following order:

P1-a We initially relax the loss in microwave circuits and PA efficiency. Given a specific architecture and performance requirements, what are the bounds on the minimum number of transceivers?
P1-b For an arbitrary N_{trx} and limited dynamic range of PA, can we design DBF weights with optimal beam pattern performance?
P2-a How do we redesign the AFN using microwave components such as power dividers and DCs? What is the underlying structure of such AFN factorizations?
P2-b For a macro-cell setup, is it possible to design an AFN to minimize the insertion loss while providing the desired beam pattern? How does the AFN design vary for a small cell setup, where the angular spread between different beam patterns is much greater than that of the macro AAA setup?
P3 Is it possible to provide an RF design example for macro-cell and small-cell AFN, minimizing insertion loss and Δ_d, and to compare the performance measurements of AFN the simulation results

[0069] The above two problems form the core of this paper and their solutions are covered in the next three sections. Problem [P2] is subdivided, depending on the objectives of the cellular architecture, and a detailed synthesis and analysis of such architectures as well as RF design examples are provided in Sec. III.

II. ALGORITHMS FOR JOINT OPTIMIZATION OF AFN AND DBF WEIGHTS

[0070] In this section, we consider problems [P1-a] and [P1-b], and estimate the AFN order as well as the AFN and

the DBF weights.

A. Bounds On The Number Of Transceivers

[0071] The introduction of an AFN reduces the order of the adaptive DBF to N_{trx}. Before we proceed to derive the AFN and DBF weights, it is important to derive theoretical bounds on the minimum number of transceivers N_{trx} achieving the desired SLL for a given $\mathcal{R}_\theta$. We start with the MSE cost function in (3) and assume that we have obtained the optimal beamformer weights $\mathbf{u}(\theta_d)$ for the modular AAA. We will later briefly explain the design procedure in Sec. II-C, the design procedure for modular AAA is also specified in [12], [13]. From (2), we get the approximation

$$\mathbf{\Delta}_{\theta_\mathrm{d}} \approx \mathbf{A}(\theta)\mathbf{u}(\theta_\mathrm{d}). \tag{4}$$

[0072] The joint DBF-AFN optimization (3) can be rewritten utilizing the LS approximation of (4) as specified in [15]:

$$\begin{aligned}
\{\mathbf{W},\, \boldsymbol{\vartheta}(\theta_\mathrm{d})\} &\approx \arg \min_{\mathbf{W},\boldsymbol{\vartheta}(\theta_\mathrm{d})} \|\mathbf{A}(\theta)\mathbf{u}(\theta_\mathrm{d}) - \mathbf{A}(\theta)\mathbf{W}\boldsymbol{\vartheta}(\theta_\mathrm{d})\|^2 \quad \forall\, \theta_\mathrm{d} \in \{\mathcal{R}_\theta\} \\
&\approx \arg \min_{\{\mathbf{W},\boldsymbol{\vartheta}(\theta_\mathrm{d})\}} \|\mathbf{A}(\theta)\|^2 \| \,\mathbf{u}(\theta_\mathrm{d}) - \mathbf{W}\boldsymbol{\vartheta}(\theta_\mathrm{d})\,\|^2 \quad \forall\, \theta_\mathrm{d} \in \{\mathcal{R}_\theta\} \\
&\approx \arg \min_{\{\mathbf{W},\boldsymbol{\vartheta}(\theta_\mathrm{d})\}} \| \,\mathbf{u}(\theta_\mathrm{d}) - \mathbf{W}\boldsymbol{\vartheta}(\theta_\mathrm{d})\,\|^2 \quad \forall\, \theta_\mathrm{d} \in \{\mathcal{R}_\theta\}
\end{aligned} \tag{5}$$

The following Lemma characterizes the necessary conditions for optimal AFN weights in the beamtilt range $\mathcal{R}_\theta$.

[0073] *Lemma 1:* Consider the scenario of [P1-a]. Assuming that the AFN is made of ideal and lossless components and the PAs have infinite range. Given $\mathcal{R}_\theta = \theta_1, \cdots, \theta_{N_\theta}$ the optimal weights of the AFN must lie in the space spanned by the dominant basis vectors of $\Theta = |\mathbf{u}(\theta_1), \cdots, \mathbf{u}(\theta_{N_\theta})]$:

$$\mathbf{W} \in \mathrm{col\ span}\{\mathbf{\Theta}\}.$$

[0074] *Proof:* Note that W has to reasonable performance for all values of $\theta_d \in \mathcal{R}_\theta$. Stacking the cost function in (5) for the entire beamtilt range $\theta_d \in \mathcal{R}_\theta$:

$$\begin{aligned}
\{\mathbf{W},\, \boldsymbol{\vartheta}(\theta_\mathrm{d})\} &= \arg\min \| \,[\mathbf{u}(\theta_1), \cdots, \mathbf{u}(\theta_{N_\theta})\,] - \mathbf{W}\,[\boldsymbol{\vartheta}(\theta_1), \cdots, \boldsymbol{\vartheta}(\theta_{N_\theta})\,]\,\|^2 \\
&\Leftrightarrow \arg\min \|\mathbf{\Theta} - \mathbf{W}\mathbf{\Upsilon}\|^2
\end{aligned} \tag{6}$$

where Θ and Υ are respectively N_t x N_θ and N_{trx} x N_θ matrices. Let us assume that $N_t \geq N_\theta$ and compute the singular value decomposition (SVD) of Θ:

$$\mathbf{\Theta} = [\mathbf{U}]\ [\mathbf{\Sigma}]\ [\mathbf{V}]^H = [\mathbf{u}_1, \cdots, \mathbf{u}_{N_{\mathrm{trx}}}, \cdots] \begin{bmatrix} \sigma_1 & & \\ & \sigma_2 & \\ & & \ddots \end{bmatrix} \mathbf{V}^H,$$

where U and V contain the left and right singular vectors and E correspond to their singular values, typically arranged

in descending order. For $N_t \geq N_\theta$, the optimal modular AAA weights $\mathbf{u}(\theta_d)$ can be approximated as a linear combination of the left singular vectors in U. If we have N_{trx} transceivers, then choosing $\mathbf{W} = [\mathbf{u}_1, \cdots, \mathbf{u}_{N_{trx}}]$ would provide the best N_{trx}-rank representation of Θ. From the reduced rank expression of W, the modular AA response $\mathbf{u}(\theta_d)$ can be approximated as $\mathbf{u}(\theta_d) \approx \mathbf{W}\vartheta(\theta_d)$, if $\sigma_{N_{trx}}+1 \approx 0$.

[0075] For this reason, the AFN design focussing on performance close to modular AAA is obtained by choosing N_{trx} leading to $\sigma_{N_{trx}}+1 \approx 0$. ■ A few remarks are in order:

- In the array signal processing literature, choosing dominant eigenvectors (usually in the digital domain) are referred to as reduced rank approaches [8],[16]. Such techniques are used to reduce complexity.
- Lemma 1 assumes that $N_t \geq N_\theta$. If $N_t < N_\theta$, choosing N_t extreme values in the range $\mathcal{R}_\theta$ and proceeding similarly will give an approximate W.
- Lemma 1 approach can also be seen as a more systematic and robust approach to come up with the weights of the Blass matrix, as in [7].
- Note that this bound on the optimal W does not consider feeder loss, dynamic range of the PAs, and the number of possible interconnects in the overall network. However, it does provide a starting point for modifications in the next sections that consider practical issues.

B. Algorithms To Optimize AFN And DBF

[0076] Estimating N_{trx} for the required range of $\mathcal{R}_\theta$ via Lemma 1 is the first step in the AFN design. Once we have established the minimum N_{trx} for the desired SLL, the next step is the design of AFN and DBF weights accounting for the constraints and satisfying the spatial mask Δ_{θ_d}. The focus of this sub-section is the AFN design while accounting for main lobe energy, PA range, SLL, etc. We start with the cost function supplied in (3) and propose an interior-point algorithm to jointly estimate W and $\vartheta(\theta_d)$.

[0077] *Constrained AFN optimization:* For simplicity of notation, we denote $\mathbf{u}(\theta_d) \approx \mathbf{W}\vartheta(\theta_d)$; and proceed to estimate $\mathbf{u}(\theta_d)$. Subsequently, the set of $\mathbf{u}(\theta_d)$ is fixed $\forall_{\theta_d}$ to estimate W using Lemma 1.

[0078] A well known technique for the beamformer design is the Capon minimum-variance distortionless response (MVDR) approach [10]. In this case, the objective is to design the weights of $\mathbf{u}(\theta_d)$ such that the convolution of $\mathbf{u}(\theta_d)$ with the antenna array response $\mathbf{A}(\theta)$ provides a mainlobe steered towards the desired sector, while minimizing the overall variance of signal radiated from the antenna towards other sectors. Mathematically, the above two conditions can be combined and written as

$$\mathbf{u}(\theta_d) = \arg\min_{\mathbf{u}(\theta_d)} \|\mathbf{u}^H(\theta_d)\mathbf{A}^H(\theta)\mathbf{A}(\theta)\mathbf{u}(\theta_d)\| \quad \text{subject to} \quad \|\mathbf{u}^H(\theta_d)\mathbf{a}(\theta_d)\|^2 = 1. \tag{7}$$

[0079] The coverage of the signals from the antenna array towards the desired sector can be enhanced by specifying the 3-dB or half power, beamwidth ($\theta_{3\,dB}$) constraint along the main lobe in the expression (7) i.e.,

$$\|\mathbf{u}^H(\theta_d)\mathbf{a}(\theta_d)\|^2 = 1 \quad \text{and} \quad \|\mathbf{u}^H(\theta_d)\mathbf{a}(\theta_{3\,dB})\|^2 = 1/2.$$

Typically, $\theta_d - \theta_{3,\,dB} \leq 5°$ in a macro-cell setup and $\theta_d - \theta_{3,\,dB} \leq 15°$ in a small cell setup.

[0080] In order to design the beamformer to direct signals only towards the specific sector, we include SLLs in (3). To achieve a specific SLL (say ε_{dB} = 20 dB below the mainlobe) over a range of angles accounting for side lobes θ_{SLL}, the beamformer

$$\|\mathbf{u}^H(\theta_d)\mathcal{A}(\boldsymbol{\theta}_{SLL})\|^2 \leq \epsilon \tag{8}$$

where $\mathcal{A}(\theta_{SLL})$ denotes the array response for θ_{SLL} and $\varepsilon = 10^{(-\varepsilon_{dB}/10)}$ i.e. $\varepsilon = 0.01$ for ε_{dB} = 20 dB.

[0081] Combining all the above constraints, the central optimization problem becomes

$$\mathbf{u}_0(\theta_\mathrm{d}) \quad = \quad \arg\min_{\mathbf{u}(\theta_\mathrm{d})} \quad \mathbf{u}^H(\theta_\mathrm{d})\mathbf{A}^H(\theta)\mathbf{A}(\theta)\mathbf{u}(\theta_\mathrm{d}) \tag{9}$$

subject to

$$|\mathbf{u}^H(\theta_\mathrm{d})\mathbf{a}(\theta_\mathrm{d})| = 1,\ |\mathbf{u}^H(\theta_\mathrm{d})\mathbf{a}(\theta_{3,\,\mathrm{dB}})| = 1/\sqrt{2} \tag{10}$$

$$|\mathbf{u}^H(\theta_\mathrm{d})\mathcal{A}(\boldsymbol{\theta}_\mathrm{SLL})| \leq [\epsilon, \cdots, \epsilon] \tag{11}$$

with (10) specifies the main-beam constriants and (11) specifies the SLL constraints. The above cost function can be recast as a convex optimization problem [11] and solved numerically to obtain the optimal solution. The solution is obtained using the well known interior point algorithm [11]; note that similar techniques to estimate modular AAA weights $\mathbf{u}(\theta_\mathrm{d})$ have been proposed in [12], [13]. For details, see Appendix.

[0082] A few remarks are in order:

- For simplicity, we estimate $\mathbf{u}(\theta_\mathrm{d})$. Once the modular AAA weights $\mathbf{u}(\theta_\mathrm{d})\ \forall\, \theta_\mathrm{d} \in \mathcal{R}_{\theta,}$ the AFN estimated using Lemma 1.
- A more comprehensive approach is to jointly optimize **W** and $\mathcal{I}(\theta_\mathrm{d})$, by representing $\mathcal{I}(\theta_\mathrm{d})$ as a function of of W similar to [9]

$$\boldsymbol{\vartheta}(\theta_\mathrm{d}) \approx \boldsymbol{\Delta}(\theta_\mathrm{d})\ [\mathbf{W}\mathbf{A}(\theta_\mathrm{d})]^\dagger$$

  and optimize for W.
- Note that the interior point optimization is not the main contribution of the paper and for a detailed performance analysis of these approaches, refer to [12], [13].

C. *DBF Design for given AFN*

[0083] Once the optimal AFN is designed as in II-B for $\forall\, \theta_\mathrm{d} \in \mathcal{R}_{\theta,}$ the adaptive DBF weights are estimated for each beamtilt θ_d. Note that the DBF $\mathcal{I}(\theta_\mathrm{d})$ is a function of the AFN W, which in turn is a function of the array response $\mathbf{a}(\theta_\mathrm{d})$. Our current objective of designing DBF weights and minimizing the overall cost in (5) is transformed into

$$\boldsymbol{\vartheta}_0(\theta_\mathrm{d}) = \arg\min_{\boldsymbol{\vartheta}(\theta_\mathrm{d})} \|\boldsymbol{\Delta}(\theta_\mathrm{d}) - \mathbf{H}(\theta)\boldsymbol{\vartheta}(\theta_\mathrm{d})\|^2 \quad \text{where} \quad \mathbf{H}(\theta) = \mathbf{A}(\theta)\mathbf{W} \tag{12}$$

can be seen as the beamspace [8] array response for a given W and the mask of spatial requirements $\Delta(\theta_\mathrm{d})$.

[0084] If the DBF $\mathcal{I}(\theta_\mathrm{d})$ is not limited by any constraints, the straightforward solution of the above cost is the LS estimate:

$$\boldsymbol{\vartheta}_0(\theta_\mathrm{d}) = [\mathbf{H}(\theta)]^\dagger\, \boldsymbol{\Delta}(\theta_\mathrm{d}).$$

[0085] *DBF design with PA constraints:* The PAs are usually limited by their ability to operate in a linear mode for only a limited range of input gain and amplitudes (typically with amplitude variation between 0 dB to 2 dB), and the DBF weights should comply with these output levels. We explicitly include these constraints on output power from each transceiver or DBF weights in the overall expression similar to the cost function in (7) and limit the PA output

$$|\vartheta_k(\theta_\mathrm{d})|^2 \approx \frac{1}{N_\mathrm{trx}} \quad \forall\, k \in \{1, \cdots, N_\mathrm{trx}\}.$$

[0086] Some comments are in order regarding the DBF design:

- In addition, we can also introduce the mainlobe and SLL constraints similar to (9 - 11) in the above formulation.
- The optimization in Appendix can be expressed including the per PA power constraint. In order to find the optimal $\mathcal{I}(\theta_\mathrm{d})$ satisfying the constraints, we proceed in the similar way as specified in Sec. II-B and as explained in [13]. Note that for such algorithms to yield optimal solution, we need to explicitly show that the problem is convex.
- Please note that per antenna power constraint is not convex (unlike the inequality and linear constraints as proposed in [13]). As a special case, the expression can be represented using magnitude and phase expressions and this magnitude constraint can be represented as a convex problem by exploiting the freedom to choose the phase. For further details on the applicability of such algorithms refer to [11].

III. AFN ARCHITECTURES

[0087] Note that Section II provides some important directions on the design of this AFN, however, they it does not consider the RF limitations and loss due to microwave components used for the implementation of such networks. These constraints as well as the design objectives vary for different scenarios and it is not possible to directly apply the results of Sec. II to design the RF feeder network This section proposes design changes for specific architectures and factorizes the AFN using a combination of microwave components.

A. Two Stage Beamforming: Macro and Small cell AFN

[0088] The DBF-AFN arrangement can be seen as two-stage beamforming towards a specific sector. The first stage i.e. DBF is an adaptive transformation for each beamtilt with a straightforward implementation (say using FPGA as in our experimental setup). The second stage i.e. AFN is made up of microwave components, and its implementation is not trivial, especially when the objectives are to minimize the overall loss and provide distinct beampatterns towards different sectors.

[0089] The AFN is comprised of a combination of commonly used microwave elements such as power dividers (Wilkinson dividers or WDs), phase shifters (micro-strip lines) and hybrid DCs [17, Ch. 7]. Typical implementations of coupler/divider elements result in loss of 0.1-0.2 dB. However most critical in the AFN is the insertion/return loss. The insertion loss occurs due to the amplitude and phase mismatch of the incoming signals at each combiner and depending on the beamtilt range of the AFN, these loss can increase to 3 dB or more.

[0090] In addition, the AFN has to be designed to account for a specific cellular arrangement. For example in a macro-cell AAA setup, the range of beamtilt between adjacent sectors is small ($\mathcal{R}_\theta$ < 20°) and the distance between the mobile user and base station is typically large. For such scenarios, the emphasis is to design *macro-AFN* to achieve a pointed beam, focusing on minimizing the overall loss. Alternatively, in a small-cell AAA setup, the beamtilt range is large (60° - 90°) and the emphasis is on increasing the angular coverage of the AAA setup. In such a setup, the focus is more towards fixed 3-4 beams covering wide angular region. At this stage the joint design problem {W, $\mathcal{I}(\theta_\mathrm{d})$} can be reclassified depending on the type of cellular architecture as

D1 AFN designed to minimize insertion loss over a relatively small downtilt range, subsequently optimizing $\mathcal{I}(\theta_\mathrm{d})$. - This approach is typically suited for *macro-cell* cases.
D2 AFN designed to form widely spaced beams at {+30°, 0°, -30°}.

- This approach is typically suited for *small-cell* cases.

Intutively [D1] and [D2] would lead to distinct redesigns of the AFN algorithms proposed in Sec. II. We focus the rest of this section for the design of W. The subsequent redesign of $\mathcal{I}(\theta_\mathrm{d})$ is straightforward from Sec. II and hence omitted.

B. Macro AAA

[0091]

1) Redesign of **W** *to minimize insertion loss:* A generic $N_\mathrm{t} \times N_\mathrm{trx}$ AFN matrix where N_t, >> N_trx AFN can be represented

using a bank of 3-port networks containing a maximum of

- $(N_t - 1)$ power dividers connected to each PA or $N_{trx}(N_t - 1)$ dividers in total.
- $(N_{trx} - 1)$ combiners connected to each antenna or $N_t(N_{trx} - 1)$ combiners in total.
- $N_{trx}N_t$ phase shifts to achieve the desired beam pattern.

[0092] Such an arrangement would result in a significant increase in AFN size and interconnect complexity. In addition, it is highly unlikely that the incoming signals at each DC will be matched in both amplitude as well as phase. This mismatch would significantly increase the insertion loss.

[0093] *Claim 1*: Consider scenario [P2], where the AFN has been factorized into a bank of DCs ($\mathbf{R}_{fb}$) as shown in Fig. A2. Each bank is further divided into several stages of DCs $\mathbf{R}_{c,i.}$ Irrespective of the AFN setup and beamtilt range, a $N_{trx} \times 1$ vector $\mathcal{I}(\theta_d)$ can minimize the insertion loss in $N_{trx} - 1$ combiners.

[0094] *Proof:* From linear estimation theory, it is straightforward that $\mathcal{I}(\theta_d)$ has N_{trx} degrees of freedom. From Lemma 1 and (7), it is possible to design $\mathcal{I}(\theta_d)$ to achieve a mainlobe pointed towards θ_d using *one* of the N_{trx} degree. Subsequently, the amplitude and phase of the remaining $(N_{trx} - 1)$ elements in DBF can be used to align the amplitude and phase at $(N_{trx} - 1)$ DCs. Note that this digital alteration of the amplitude and phase weights at the DBF minimizes the insertion loss in the specified $(N_{trx} - 1)$ DCs. However, this does not necessarily reduce the overall insertion loss in the rest of the network. ■

[0095] Claim 1 acts as a starting point and provides a lower bound on the possible number of combiners for any downtilt range minimizing loss in $N_{trx} - 1$ ports. Note that, the SLL performance will be poor if we use only *one* dimension to optimize for beam pattern and the rest to optimize for the insertion loss. Intuitively, it makes sense to use all the available N_{trx} degrees of freedom available in $\mathcal{I}(\theta_d)$ to optimize for SLL. This would mean a fundamental redesign of W to minimize insertion loss.

2) Multistage AFN decomposition: The connections within the AFN designed in Sec. II do not take Claim 1 into account. The AFN decomposition satisfying Claim 1 is modeled using a $N_t \times N_{trx}$ spatial interconnect matrix S. The objective is to redesign the AFN weights satisfying S as well as Claim 1. One such example of the spatial interconnect map is, in a 11×5 case

$$\mathbf{S} = \begin{bmatrix} \mathbf{1}_3 & \mathbf{0}_2 & \mathbf{0}_4 & \mathbf{0}_6 & \mathbf{0}_4 \\ \mathbf{0}_4 & \mathbf{1}_3 & \mathbf{1}_3 & \mathbf{1}_3 & \mathbf{0}_4 \\ \mathbf{0}_4 & \mathbf{0}_6 & \mathbf{0}_4 & \mathbf{0}_2 & \mathbf{1}_3 \end{bmatrix}$$

where 1_l and 0_l respectively correspond to $l \times 1$ vector of ones and zeros. The AFN satisfying the interconnects S is redesigned using a modified version of successive orthogonal projection (also referred to as multi-stage Wiener decomposition) [16] as follows:

- Note that $[\mathbf{u}_1, \cdots, \mathbf{u}_{N_{trx}}, \cdots] = \text{Basis}\{\Theta\}$.
  - From the result of Lemma 1.
- for $k \in \{1, \cdots, N_{trx}\}$
  - Partition SVD(Θ) = $[\mathbf{u}_1, \mathbf{U}_N]$.
  - Extract the AFN weights satisfying the spatial interconnects: $\mathbf{w}_k = \mathbf{s}_k \odot \mathbf{u}_1$.
  - Normalize each column of w_k.
  - $\mathbf{W} = [w_1, \cdots w_k]$.
  - Compute the orthogonal projection: $u = (\mathbf{I} - \mathbf{W}\mathbf{W}^H)\Theta$.
- end k
- Final AFN: $\mathbf{W} = [\mathbf{w}_1, \cdots \mathbf{w}_{N_{trx}}]$.

[0096] The multistage AFN decomposition is chosen, since it provides a low-complexity implementation (compared to brute search techniques) and, as shown in [16], converges to optimal solution for increasing N_{trx}. Note that while the

spatial interconnect map changes for additional claims the methodology is generic and can be applied for any S.

[0097] Consider multi-stage AFN decomposition such as Fig. A2, where components with similar functions are combined respectively into a filter-bank of power dividers D_{fb}, a bank of phase shifters P_1 and a filter-bank of DCs R_{fb} :

$$\begin{aligned}\mathbf{W} &= \mathbf{D}_{\mathrm{fb}} \times \mathbf{P}_1 \times \mathbf{R}_{\mathrm{fb}} \\ &= (\mathbf{D}_{w1} \times \mathbf{D}_{w2} \times \mathbf{D}_{w3}) \times \mathbf{P}_1 \times (\mathbf{R}_{c1} \times \mathbf{R}_{c2}).\end{aligned}$$

In the above expression the subscripts, D_{wi} denotes a filter-bank of power dividers for stage *i* and R_{ci} denotes a filter-bank of hybrid couplers/ combiners for stage *i*.

3) Factorization of power dividers D_{fb} *and phase shifters* P_1: Since $N_{trx} << N_t$, the first stage of the AFN consists primarily of power dividers to increase the number of input signals. The magnitude and phase values of elements in the AFN W, i.e. w_i, correspond to power ratios and phase shifts of the signals from the *i*th PA. Let N_{si} be the number of non-zero elements in the column w_i, or s_i. When $N_{si} > 3$, the implementation of a $(N_{si} + 1)$ port WD, might lead to impractical realizations. For this reason, each PA output is successively factorized into multiple stages made up of 3-port WDs as shown in Fig. A3. For reduced complexity, the design algorithms implement balanced WDs in the first two stages of $\mathbf{D_{fb}}$ followed by unbalanced WDs in the third stage. The total number of stages in $\mathbf{D_{fb}}$ with 2-port WDs is limited by $\log_2 [\max(N_{si}) + 1]$.

[0098] The output signals from $\mathbf{D_{fb}}$ undergo a phase shift denoted by diagonal matrix $\mathbf{P_1}$. The elements in $\mathbf{P_1}$ correspond to lengths of strip line, which inturn correspond to the phase shifts of **W.** Note that the signals undergoing these phase shifts in $\mathbf{P_1}$ are already modified by the corresponding power ratios. At this point, the transmit signal *s*[*k*] in (1) is modified by the DBF $\vartheta(\theta_d)$ and followed by the AFN $\mathbf{D_{fb}}$ and $\mathbf{P_1}$ to obtain the achieve the desired beamtilt and pattern. At this stage, the total number of signals is $N_s = \sum_{i=1}^{N_{trx}} N_{s_i}$. Note that it is necessary to have $N_s >> N_t$ to achieve the desired beam pattern for various downtilts.

4) AFN as a linear phase FIR filter: The N_s outputs from $\mathbf{P_1}$ are modified by $\mathbf{R_{fb}}$ and fed to N_t antennas. For efficient and lossless operation of $\mathbf{R}_{fb}$, it is necessary that the input signal at each DC has to be matched in terms of amplitude and phase and any mismatch will result in insertion loss. Note that the AFN is fixed, but the phase and amplitude of $\vartheta(\theta_d)$ is varied for each beamtilt. Such an arrangement when used with the standard 3-port DCs will always result in insertion loss. For this reason, we use hybrid elements, such as rat-race couplers or branch hybrids (four port networks) instead of a standard three port combiner. In a rat-race coupler, Ports 2 and 3 are input ports and the sum of the inputs is coupled to standard output (Port 1). A fourth port (also referred to as the reflection port or isolation port) extracts the signals that would have otherwise led to insertion loss whenever there is a mismatch between the input signals [17, Pg. 480]. Thus, at a given stage $R_{c,\,i}$, any phase or amplitude mismatch can be captured using the isolation port of the hybrid coupler. We exploit the signal processing properties of the antenna array and DBF with the redundant fourth port [17, Ch. 7] to extract/reroute the insertion loss seen in Stage *i* and minimize the overall loss in subsequent stages.

[0099] *Claim 2:* Consider scenario P2, where the AFN is subdivided into 4-port DCs.

- The elements of the combined response of $\mathbf{W}\vartheta(\theta_d)$, $\forall\theta_d$ will have a linear phase progression.
- The number of DCs in each stage $R_{c,\,i}$ must not exceed $N_{trx} - 1$ to minimize the insertion loss.

[0100] *Proof:* Consider the modular AAA setup with N_t antennas and array response $\mathbf{A}(\theta)$ as specified in (2) and the modular AAA beamformer $\mathbf{u}(\theta_d) = [u_0(\theta_d), \ldots, u_{N_t-1}(\theta_d)]^T$. Note that $a(\theta_i)$ has linear phase progression since the antenna elements are uniformly spaced. The transfer function of $\mathbf{u}(\theta_d)$ operating on the antenna array can be written as

$$U(\theta_i) = \sum_{k=0}^{N_t-1} u_k(\theta_d)\, e^{-j2\frac{\pi}{\lambda}k\cos(\theta_i)} \quad \forall\theta_i \in [-\pi, \pi]. \tag{13}$$

[0101] The beamformer $\mathbf{u}(\theta_d)$ can be seen as a spatial finite impulse response (FIR) filter operating on the antenna

array and the above expression can be seen as the spatial equivalent of a filter transfer function. From filter design theory, we know that matched filters for such uniformly spaced antennas will have a *symmetric* magnitude response along the central tap $N_t/2$ or the central antenna element in this case where $|u_k(\theta_d)| = |u_{N_t-k}(\theta_d)|$. In addition, these filters will also have a linear phase progression, eventually facilitating implementation using low-complexity FFT's.

[0102] We extend this *linear phase* and *symmetric magnitude* argument for our AFN-DBF setup. The DBF $\mathfrak{I}(\theta_d)$ operates on reduced dimension antenna array $\mathbf{H}(\theta) = \mathbf{WA}(\theta)$ and its transfer function is similar to $U(\theta_i)$ in (13). The symmetric linear phase property ensures that the elements of $\mathfrak{I}(\theta_d)$ as well as the columns of $\mathbf{H}(\theta)$, will have linear phase progression and the joint DBF-AFN operation can be seen as a convolution operation in spatial domain. Exploiting the shift invariance property of the convolution operation, the combined DBF-AFN $\mathbf{W}\mathfrak{I}(\theta_d)$ operating on antenna array $\mathbf{A}(\theta)$ will also have a linear phase progression. Due to the AFN shift invariance property, a linear phase progression in $\mathfrak{I}(\theta_d)$ will lead to linear phase signals at the isolation ports as well as the output ports of the corresponding DCs.

[0103] The combination of symmetric magnitude response, shift invariance and linear phase progression of the DBF-AFN implies that the signal from the isolation port of the rat-race Coupler k is in phase with respect to the signal at the output port of the rat-race Coupler $N_{trx} - k \; \forall, k \in \}1, \cdots, N_{trx} - 1\}$. This property means that the above two signals from Stage i can be combined to mitigate the insertion loss at the subsequent stage $\mathbf{R}_{c,i+1}$ as shown in Fig. A4. ■ Considering that $\mathbf{R}_{c,i}$ typically has dimensions close to $N_t >> N_{trx}$, this result implies that $\mathbf{R}_{c,i}$ minimizing insertion loss has to be a sparse matrix.

5) Factorization of hybrid couplers $\mathbf{R}_{fb}$: From Claim 2, the Port 4 output at Stage i can be subsequently recirculated as input to the next stage $R_{c,i+1}$ to account for the insertion loss. These hybrid elements can either be branch hybrids (commonly used in Butler matrix implementations [4]) or rat-race hybrids. In our setup, we use rat-race hybrid elements, the motivation being

- Four-port rat-race couplers can also be seen as radix-2 DFT or FFT implementations. Higher order DFTs can then be obtained using a bank of such couplers.
- The FFT analysis provides a generic approach to construct $\mathbf{R}_{fb}$ using a careful arrangement of rat-race couplers at different positions to mitigate the overall loss. For example, the arrangement of rat-race couplers in $\mathbf{R}_{fb}$ as in Fig. A4 provides a linear phase progression and symmetric magnitude response.

C. Small-cell AFN

[0104] In a small-cell setup, the number of RF chains are limited to 2 or 3 and the number of antennas are limited to 4-6 (typically as an horizontal arrangement). The objective is to provide coverage along $N_\theta = 3$ - 4 fixed beams spaced 30° apart from each other (say $\theta_d \in \{-30°, 0°, +30°\}$). Each beam has a wider 3-dB beamwidth (nearly 15° and the focus is more on improving angular coverage (unlike D1 where the focus is on minimizing loss). The requirements and overall setup make this design fundamentally different from that of D1.

1) Existing architectures: One well known RF AFN to provide distinct (and orthogonal) beams is through the use of a Butler matrix [4]. This matrix has N_t inputs and N_t outputs, connecting N_t PAs and N_t antennas and can be implemented using either branch or rat-race hybrid couplers resulting in very low loss. Typical implementations have $N_t = 4$ antennas spaced half a wavelength ($\lambda/2$) apart. Note that in the majority of Butler-matrix-type lossless implementations, only one PA is turned *on* at a given time to generate the desired beam patterns. Such a design sacrifices the radiated power from the array for lossless implementation.

[0105] A systhematic approach to estimate the AFN weights for arbitrary beamtilts $R_{N_\theta} = \{\theta_1, \theta_2, \theta_3\}$ is the AFN algorithm proposed in Lemma 1 taking the SLL and PA requirements into consideration:

$$\mathbf{W} \in \text{col span}\{\mathbf{\Theta}\} \quad \text{where} \quad \mathbf{\Theta} = [\mathbf{u}(\theta_1), \mathbf{u}(\theta_2), \mathbf{u}(\theta_3)]. \tag{14}$$

Note that W can be obtained from the basis vectors of Θ using a QR decomposition or SVD [18]. This QR decomposition approach is similar to the approach in [7], where the authors propose an RF matrix (also referred to as Blass matrix) from the Gram-Schmidt orthogonalization [18] of different beampatterns. However they do not account for operational constraints (such as SLL suppression). In order to minimize the insertion loss, they transform the Blass matrix into a modified Butler matrix, with only one PA operating at a time.

[0106] Thus the existing family of AFNs to generate distinct/orthogonal beams come in two categories

- Lossless Butler matrix implementations, where there is only one PA operating at a time, but reduces the overall radiated power by $\frac{1}{N_{\text{trx}}}$.
- Lossy Blass matrix with N_{trx} transceivers operating simultaneously, but incurs insertion loss in the combiners of nearly 5 dB.

The underlying question is to check whether we can design an AFN with significant reduction in insertion loss while having all PAs operate at a time in order to provide the desired beam patterns with maximum radiated power and reduced loss.

2) Generalized Butler matrix using Givens rotations: Note that the $\mathbf{R}_{\text{fb}}$ in a lossy Blass matrix AFN will have more than N_{trx} - 1 combiners and the Claims 1-2 will not be satisfied. Additionally, the DBF-AFN arrangement will not always have a linear phase progression (since the degrees of freedom/redundancy is limited).

[0107] In general, our objective is to come up with generic factorizations of the AFN satisfying downtilts R_θ with a reduced number of combiners such that the insertion loss is reduced. Given a N_{t} x N_{trx} AFN, the number of combiners connected to an antenna i, $i \in \{1, \cdots, N_{\text{t}}\}$ is the number of non-zero elements of the the row $\mathbf{W}(i,:)$ (commonly referred to as a *row weight)*. AFN factorizations with reduced row weights would simplify matching the combiner inputs and subsequently minimize the insertion loss while achieving the desired beam pattern. From the AFN designed in (14), we search for the unmatched/out-of-phase combiners and:

- Selectively remove the connections corresponding to these unmatched combiners.
  $\rightarrow$ This operation can be done by zeroing the specific entries in the AFN matrix that exceed a specific mismatch threshold. To zero a specific entry in the AFN matrix, we use the Givens rotation method [18].
- Subsequently, redesign the AFN W as specified in Sec. II, as well as the DBF $\vartheta(\theta_{\text{d}})$ such that the input signals at the remaining combiners are matched.

[0108] Once this is performed, we can factorize the AFN, as specified in Sec. III-B.1 into $\mathbf{D}_{\text{fb}}$, $\mathbf{P}_1$, and $\mathbf{R}_{\text{fb}}$.

[0109] We specify the Givens rotation elimination for $\vartheta(\theta_1) = [1, 1, 1]^T$ at downtilt θ_1. The combined DBF-AFN response can be represented as

$$\mathbf{W}\,\vartheta(\theta_1) = \begin{bmatrix} \begin{bmatrix} * & * & * \end{bmatrix} \\ * \leftarrow \boxed{*} \rightarrow * \\ \begin{bmatrix} * & * & * \end{bmatrix} \end{bmatrix} \begin{bmatrix} 1 \\ 1 \\ 1 \end{bmatrix}. \tag{15}$$

[0110] Consider above expression, where $w_{3,2}$ (enclosed in the box) is out of phase with the two other signals $w_{3,1}$ and $w_{3,3}$. Givens rotation allows us to remove a specific element (in this case $w_{3,2}$) using the N_{t} x N_{t} matrix

$$\mathbf{G}(2,3,\phi) = \begin{bmatrix} 1 & & & \\ & \cos(\phi) & \sin(\phi) & \\ & -\sin(\phi) & -\cos(\phi) & \\ & & & 1 \end{bmatrix} \quad \text{where} \quad \phi \forall \{-pi,\ pi\} \text{is estimated.}$$

[0111] A generalization of the above expression $G\ (i, j, \phi)$ is iteratively applied to zero the element in the ith row and the jth column, whenever the amplitude and phase mismatch exceeds a particular threshold, while making sure that the overall transfer function of the matrix does not change considerably. Thus we have

$$\mathbf{W} \rightarrow \mathbf{G}(3,2,\phi_1)\mathbf{W} \rightarrow \begin{bmatrix} * & * & \boxed{*} \\ * & * & * \\ * & 0 & * \\ * & * & * \end{bmatrix} \mathbf{G}(3,1,\phi_2)\mathbf{W} \rightarrow \begin{bmatrix} * & * & 0 \\ * & * & * \\ * & 0 & * \\ * & * & * \end{bmatrix} \tag{16}$$

[0112] In practice, we iteratively update W for all values of $\vartheta(\theta_d)$ i.e., in (15), we replace $\vartheta(\theta_d)$ with $\Upsilon = [\vartheta(\theta_1), \vartheta(\theta_2), \vartheta(\theta_3)]$. Some comments are in order:

- Note that after every rotation, the DBF $\vartheta(\theta_d)$ as well as the combined response $\mathbf{W}\vartheta(\theta_d)$ has to be redesigned to account for SLL requirements, PA limitations, etc.
- The number of Givens rotation iterations is a tradeoff between the quality of beam patterns (desired SLL/beamtilts) and the insertion loss levels. Considering that $N_{trx} = 3$ and $N_t = 6$, increasing the number of iterations might degrade beam-patterns and vice versa.
- Zeroing a specific entry essentially removes the connection. Similarly, the entries of T can be removed, and such an operation will lead us to the Butler matrix when Υ is diagonal.

IV. CONCLUSION

[0113] Existing RF feeder and beam forming networks are constructed in a mainly empirical way, which, depending on the experience of the designer and the complexity of the required network often gives reasonable results, but this process is very time consuming and does not guarantee the optimal solution in terms of beamforming performance, network complexity and minimized loss. A thorough understanding of the theoretical bounds as well as the microwave limitations will lead to the optimal solution enhancing the capacity and coverage of the communications system while operating at a reduced cost.

[0114] For this reason, we adopted a holistic approach and proposed a comprehensive approach to design an RF feeder network and reduce the number of transceivers while accounting the desirable features in next generation cellular base stations. Effectively, we showed how to determine the minimum number of transceiver elements required to achieve a given set of access requirements and a presented a unified view on the design of an AFN for a given set performance requirements. In order to prove the validity of the approach, we have designed two feeder networks based the proposed method and architecture. Note that these requirements and the designs differ a lot in terms of their system requirements:

- A feeder network for a small-cell base station, generating three static beams with a rather broad beam width in the horizontal direction with a very large beam tilt angle of (-30°, 0°, +30°). In this design the focus was on achieving a set of orthogonal beams while maintaining a side lobe suppression of at least 10dB.
- A beamformer for a macro-cell base station antenna with a sharp and narrow vertical beam and (unlike the small cell), a continuous beam tilt range of 15degrees, while maintaining a tight set of spectral mask and insertion loss requirements. Subsequently, we have measured the performance of the proposed feeder network

[0115] Despite these two very different sets of requirements in both applications, the two derived AFNs prove benefits in designing through a generic approach. Especially in the case of a macro-cell base station antenna we could show that the loss in the combiner stages of the AFN caused by the digital phase shift at the AFN input ports to achieve the beam tilt are kept to a minimum, which is essential for such applications where the amount of radiated power easily reaches 100W and more, and where combiner loss in the AFN not only result in reduced radiated power, but also an increasing challenge for thermal management, if the power that needs to be dissipated within the AFN.

[0116] Future directions of research in this line includes

- Note that in case of the small cell application these combiner loss are not critical in terms of thermal management with radiated power level ¡ 5W, where an insertion loss of 1dB results in approximately 0.6W of dissipated power, however they affect the range and receiver sensitivity. For this reason, future small cell networks must consider insertion loss minimization in addition to wide beamtilt ranges.
- Another aspect of our future work is the expansion of the method towards two-dimensional arrays [21], which would enable the cost effective construction of large beam steering arrays and application of advanced 2-dimensional beam steering methods (e.g. per-user-beamsteering) in order to reduce radiated power levels, to reduced intra- and inter-cell interference and to incease per-user data rates within a given cell.

- All the above methods must address the practical issues experienced in the Sec. ?? leading to the phase mismatch between different transceivers. Note that joint beamforming and calibration in a AAA system is a fundamental requirement for enhanced capacity and coverage.

APPENDIX

Interior point algorithm.

[0117] Our goal is to formulate the inequality constraint present in (10) as a set of equality constraints. These equality constraints can be subsequently used in combination with well known approaches such as Newton method or a variation of Lagrange multiplier optimization [11] to iteratively solve for $\mathbf{u}(\theta_d)$. Note that $\mathbf{u}(\theta_d)$ must satisfy $|\mathbf{u}^H(\theta_d)\mathbf{A}(\theta)| = 1$. We have

$$\frac{\mathbf{u}^H(\theta_\mathrm{d})\mathcal{A}(\boldsymbol{\theta}_\mathrm{SLL})}{\mathbf{u}^H(\theta_\mathrm{d})\mathbf{a}(\theta_\mathrm{d})} \leq [\epsilon, \cdots, \epsilon]$$

leads to

$$\mathbf{u}^H(\theta_\mathrm{d})\mathcal{I}(\theta_\mathrm{d}) \geq \mathbf{0} \quad \text{where} \quad \mathcal{I}(\theta_\mathrm{d}) = [\mathbf{a}(\theta_\mathrm{d})[\epsilon, \cdots, \epsilon] - \mathcal{A}(\boldsymbol{\theta}_\mathrm{SLL})]\,.$$

[0118] Similarly, we can rewrite the equality constraints as follows:

$$\mathbf{u}^H(\theta_\mathrm{d})\mathcal{A}_{eq} = \mathbf{e}^T \quad \text{where} \quad \mathcal{A}_{eq} = [\mathbf{A}(\theta), \mathbf{a}(\theta_{3\,dB})] \quad \text{and} \quad \mathbf{e}^T = \left[1, 1/\sqrt{2}\right].$$

[0119] The basic idea of the interior point approach [11] is to introduce an indicator function $L(\theta_d)$ and make the inequality constraints implicit in the cost function in (7). Thus the original cost in (7) is rewritten as

$$\begin{aligned} \mathbf{u}_0(\theta_\mathrm{d}) \quad = \quad & \arg\min\, \mathbf{u}^H(\theta_\mathrm{d})\mathbf{u}(\theta_\mathrm{d}) - \frac{1}{t}\log[\mathcal{I}(\theta_\mathrm{d})] \\ & \text{subject to} \quad \mathbf{u}^H(\theta_\mathrm{d})\mathcal{A}_{eq} = \mathbf{e}^T \end{aligned} \tag{17}$$

where t corresponds to a convergence parameter. This problem can be subsequently optimized in a way similar to Lagrange optimization or a Gradient based approach.

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Claims

1. An antenna feed for generating signals for an active antenna array including multiple antennas for transmitting transmission beams having selected tilt angles, the antenna feed comprising:

   a digital beamformer operable to receive an input broadband signal and to generate a plurality N of output broadband signals, each having an associated phase and amplitude;
   a plurality N of transmission signal generators, each operable to receive one of the said plurality N of output broadband signals and to generate a corresponding plurality N of first RF signals; and
   a feed network comprising a power split network operable to receive the said plurality N of first RF signals and to generate a plurality P of second RF signals, each having an associated amplitude and an associated phase shift applied using a variable phase shifter of a phase shift network, the feed network operable to supply respective ones of the plurality P of second RF signals to respective ones of the multiple antennas, M, of the said active antenna array, where $P \geq M > N$.

2. An antenna feed as claimed in claim 1, wherein the feed network is an RF beamformer in which the power split network comprises multiple power split stages to:

   receive the plurality N of first RF signals; and
   generate the said plurality P of second RF signals using power splitters at respective ones of the stages of the network, wherein respective ones of the power slitters are operable to divide a received RF signal over at least two separate paths to provide respective split signals, wherein the power splitter are operable to provide the split signals to a subsequent stage of the RF beamformer to generate the plurality P of second RF signals.

3. An antenna feed as claimed in claim 2, wherein the variable phase shifter of the phase shift network is logically positioned in between two of the multiple power split stages of the power split network.

4. An antenna feed as claimed in any preceding claim, wherein the phase shift network further comprises a fixed phase shifter to apply a predetermined phase shift to respective ones of the plurality P of second RF signals.

5. An antenna feed as claimed in claim 4, wherein the fixed phase shifter is logically positioned after the power split

network.

6. An antenna feed as claimed in any preceding claim, wherein the plurality P of second RF signals are supplied to respective antennas of the plurality M of antennas of the said active antenna array to transmit a transmission beam with a requested beam tilt angle.

7. An antenna feed as claimed in any preceding claim, wherein the feed network comprises multiple feed instances each operable to provide selected ones of the P signals to the plurality M of antennas to transmit multiple different transmission beams simultaneously.

8. An antenna feed as claimed in claim 7, wherein respective ones of the multiple feed instances are operable to simultaneously transmit the multiple different transmission beams with respective different beam tilt angles.

9. An antenna feed as claimed in any preceding claim, wherein the digital beamformer is an adaptive beamformer operable to weigh the plurality N of output broadband signals to provide a desired main lobe and side lobe distribution for a transmission beam.

10. A method of configuring an antenna feed for generating signals for an active antenna array including multiple antennas for transmitting transmission beams having selected tilt angles, the method comprising:

receiving an input broadband signal at a digital beamformer and generating a plurality N of output broadband signals, each having an associated phase and amplitude;
receiving one of the said plurality N of output broadband signals at each one of a plurality N of transmission signal generators and generating a corresponding plurality N of first RF signals;
receiving the said plurality N of first RF signals at a feed network comprising a power split network and generating a plurality P of second RF signals, each having an associated amplitude;
applying a selected phase shift to the said plurality N of first RF signals using a variable phase shifter of a phase shift network to apply; and
supplying respective ones of the plurality P of second RF signals to respective ones of the multiple antennas, M, of the said active antenna array, where $P \geq M > N$.

11. A method as claimed in claim 10, further including
receiving the plurality N of first RF signals in multiple power split stages of the power split network;
generating the said plurality P of second RF signals using power splitters at respective ones of the stages of the network by dividing a received RF signal over at least two separate paths; and
providing those to a subsequent stage to generate the plurality P of second RF signals.

12. A method as claimed in claim 11, further including logically positioning the variable phase shifter of the phase shift network in between two of the multiple power split stages of the power split network.

13. A method as claimed in any of claims 10 to 12, further including applying a predetermined phase shift to respective ones of the plurality P of second RF signals using a fixed phase shifter of the phase shift network.

14. A method as claimed in claim 13, further including logically positioning the fixed phase shifter after the power split network.

15. A method as claimed in any of claims 10 to 14, further including supplying the plurality P of second RF signals to respective antennas of the plurality M of antennas of the said active antenna array, whereby to enable transmission of a transmission beam with a requested beam tilt angle.

Amended claims in accordance with Rule 137(2) EPC.

1. An antenna feed for generating signals for an active antenna array including multiple antennas for transmitting transmission beams having selected tilt angles, the antenna feed comprising:

a digital beamformer operable to receive an input broadband signal and to generate a plurality N of output broadband signals, each having an associated phase and amplitude;

a plurality N of transmission signal generators, each operable to receive one of the said plurality N of output broadband signals and to generate a corresponding plurality N of first RF signals; and
a feed network comprising

a power split network operable to receive the said plurality N of first RF signals and to generate a plurality P of second RF signals, each having an associated amplitude and an associated phase shift, and
a phase shift network, the phase shift network or the feed network including a variable phase shifter for applying said associated phase shift, the variable phase shifter being adapted to shift the phase of an input signal by a variable amount that can be selected or tuned;
the feed network being operable to supply respective ones of the plurality P of second RF signals to respective ones of the M multiple output ports thereof, each output port corresponding to an antenna of M antenna of the said active antenna array, where $P \geq M > N$.

2. An antenna feed as claimed in claim 1, wherein the feed network is an RF beamformer in which the power split network comprises multiple power split stages to:

receive the plurality N of first RF signals; and
generate the said plurality P of second RF signals using power splitters at respective ones of the stages of the network, wherein respective ones of the power slitters are operable to divide a received RF signal over at least two separate paths to provide respective split signals, wherein the power splitter are operable to provide the split signals to a subsequent stage of the RF beamformer to generate the plurality P of second RF signals.

3. An antenna feed as claimed in claim 2, wherein the variable phase shifter of the phase shift network is logically positioned in between two of the multiple power split stages of the power split network.

4. An antenna feed as claimed in any preceding claim, wherein the phase shift network further comprises a fixed phase shifter to apply a predetermined phase shift to respective ones of the plurality P of second RF signals.

5. An antenna feed as claimed in claim 4, wherein the fixed phase shifter is logically positioned after the power split network.

6. An antenna feed as claimed in any preceding claim, wherein feed network is operable to supply via the output ports thereof the plurality P of second RF signals to respective antennas of the plurality M of antennas of the said active antenna array to transmit a transmission beam with a requested beam tilt angle.

7. An antenna feed as claimed in any preceding claim, wherein the feed network comprises multiple feed instances each operable to provide selected ones of the P signals to the plurality M of antennas to transmit multiple different transmission beams simultaneously.

8. An antenna feed as claimed in claim 7, wherein respective ones of the multiple feed instances are operable to simultaneously transmit the multiple different transmission beams with respective different beam tilt angles.

9. An antenna feed as claimed in any preceding claim, wherein the digital beamformer is an adaptive beamformer operable to weigh the plurality N of output broadband signals to provide a desired main lobe and side lobe distribution for a transmission beam.

10. A method of configuring an antenna feed for generating signals for an active antenna array including multiple antennas for transmitting transmission beams having selected tilt angles, the method comprising:

receiving an input broadband signal at a digital beamformer and generating a plurality N of output broadband signals, each having an associated phase and amplitude;
receiving one of the said plurality N of output broadband signals at each one of a plurality N of transmission signal generators and generating a corresponding plurality N of first RF signals;
receiving the said plurality N of first RF signals at a feed network comprising a power split network and generating a plurality P of second RF signals, each having an associated amplitude, the feed network further comprising a phase shift network, the phase shift network or the feed network including a variable phase shifter;
applying a selected phase shift to the said plurality N of first RF signals using the variable phase shifter, the variable phase shifter being adapted to shift the phase of an input signal by a variable amount that can be

selected or tuned; and
supplying respective ones of the plurality P of second RF signals to respective ones of the multiple antennas, M, of the said active antenna array, where P≥M>N.

11. A method as claimed in claim 10, further including

receiving the plurality N of first RF signals in multiple power split stages of the power split network;
generating the said plurality P of second RF signals using power splitters at respective ones of the stages of the network by dividing a received RF signal over at least two separate paths; and
providing those to a subsequent stage to generate the plurality P of second RF signals.

12. A method as claimed in claim 11, further including logically positioning the variable phase shifter in between two of the multiple power split stages of the power split network.

13. A method as claimed in any of claims 10 to 12, further including applying a predetermined phase shift to respective ones of the plurality P of second RF signals using a fixed phase shifter of the phase shift network.

14. A method as claimed in claim 13, further including logically positioning the fixed phase shifter after the power split network.

15. A method as claimed in any of claims 10 to 14, further including supplying the plurality P of second RF signals to respective antennas of the plurality M of antennas of the said active antenna array, whereby to enable transmission of a transmission beam with a requested beam tilt angle.

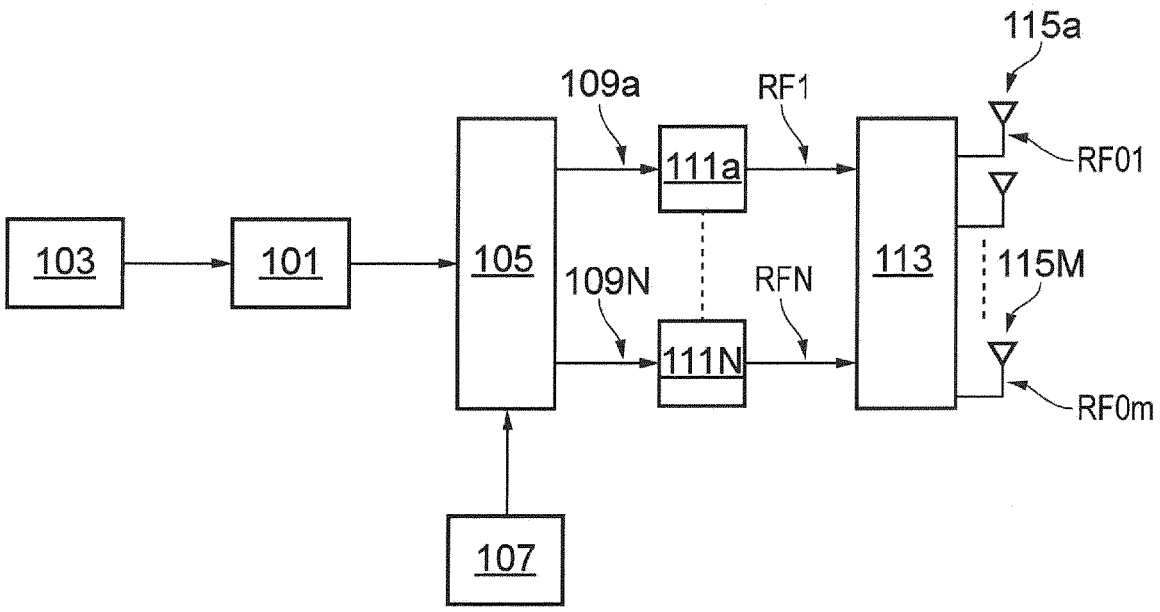

115a
109a
RF1
111a
RF01
103
101
105
113
115M
109N
RFN
111N
RF0m
107

FIG. 1

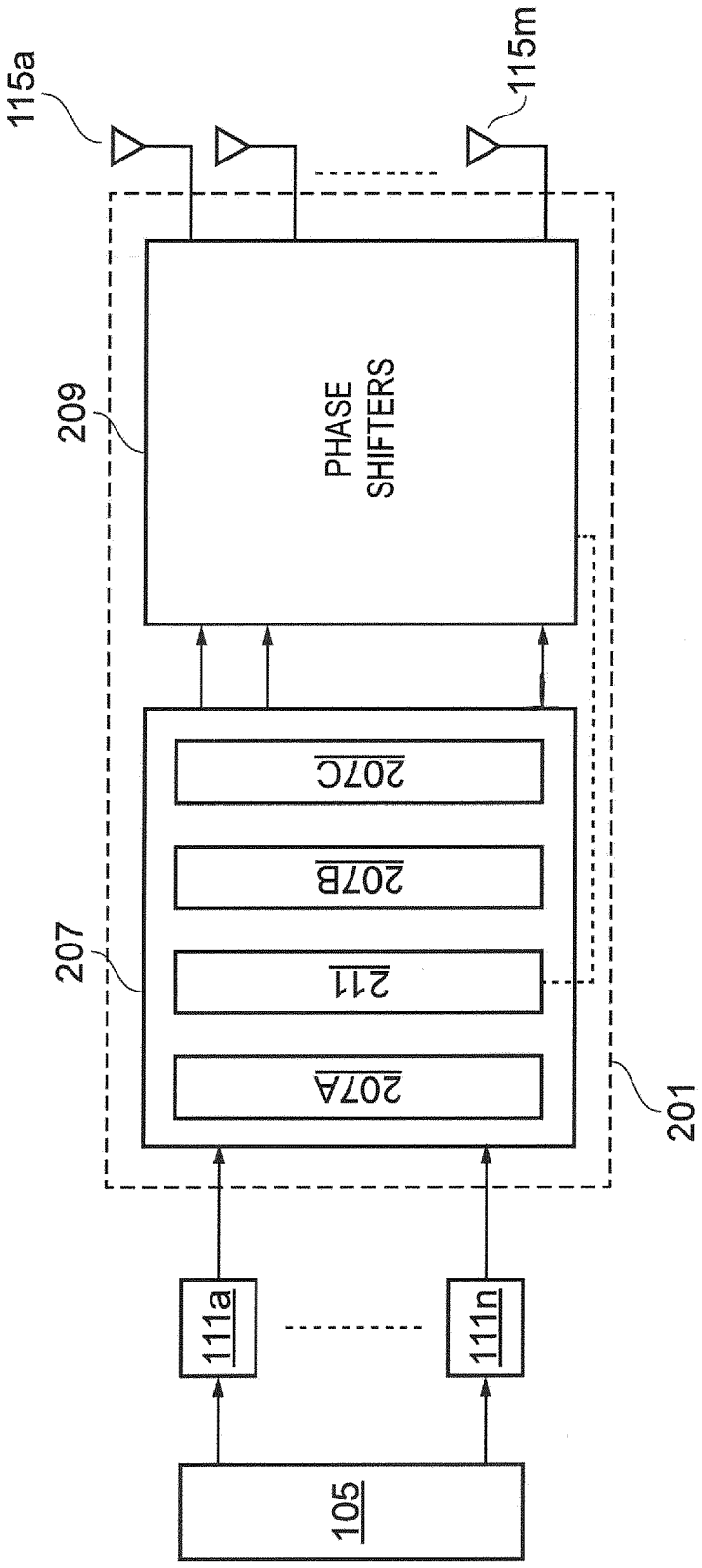
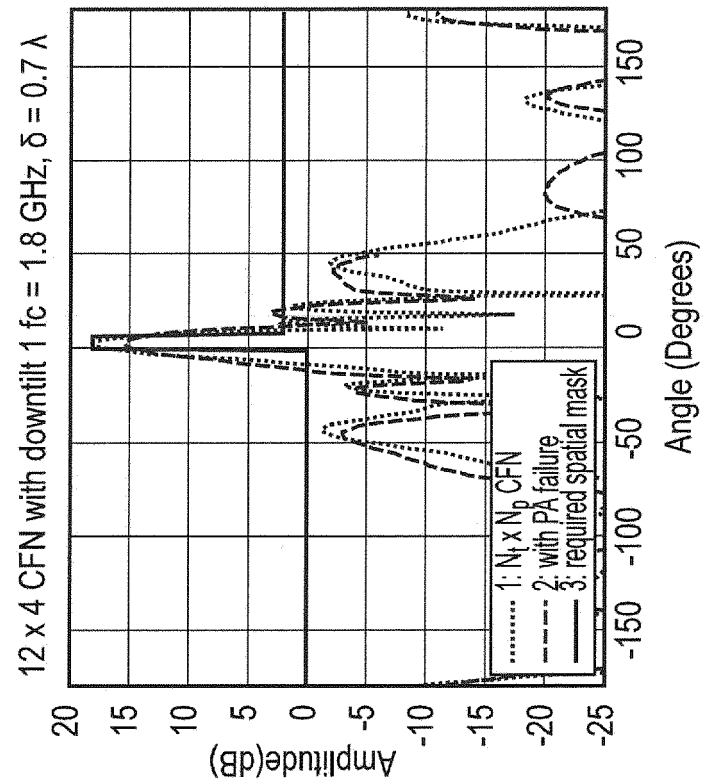

115a
115m
209
PHASE SHIFTERS
207
207C
207B
211
207A
201
111a
111n
105

FIG. 2


12 x 4 CFN with downtilt 1 fc = 1.8 GHz, δ = 0.7 λ
Amplitude(dB)
Angle (Degrees)
1: N_t x N_p CFN
2: with PA failure
3: required spatial mask
(b)

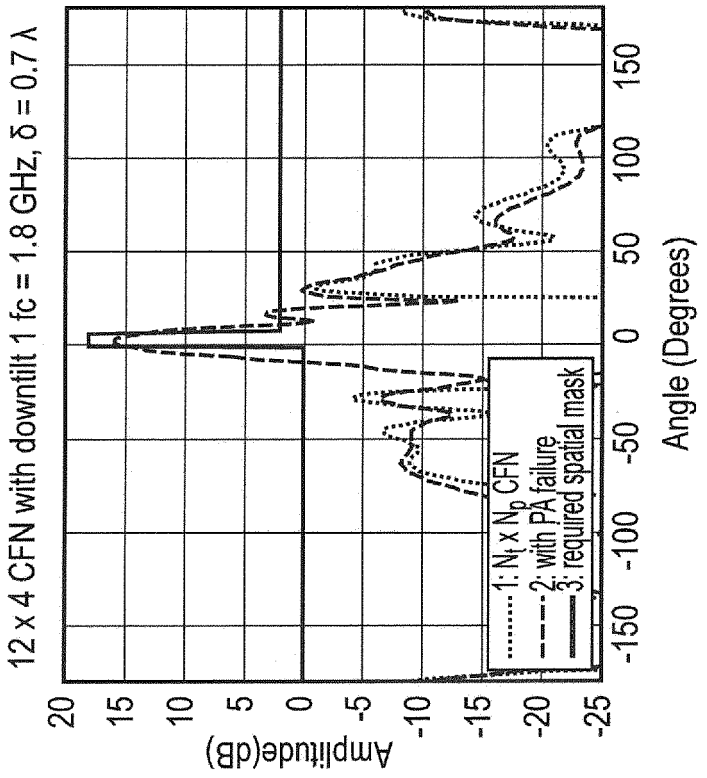

12 x 4 CFN with downtilt 1 fc = 1.8 GHz, δ = 0.7 λ
Amplitude(dB)
Angle (Degrees)
1: N_t x N_p CFN
2: with PA failure
3: required spatial mask
(a)

FIG. 3

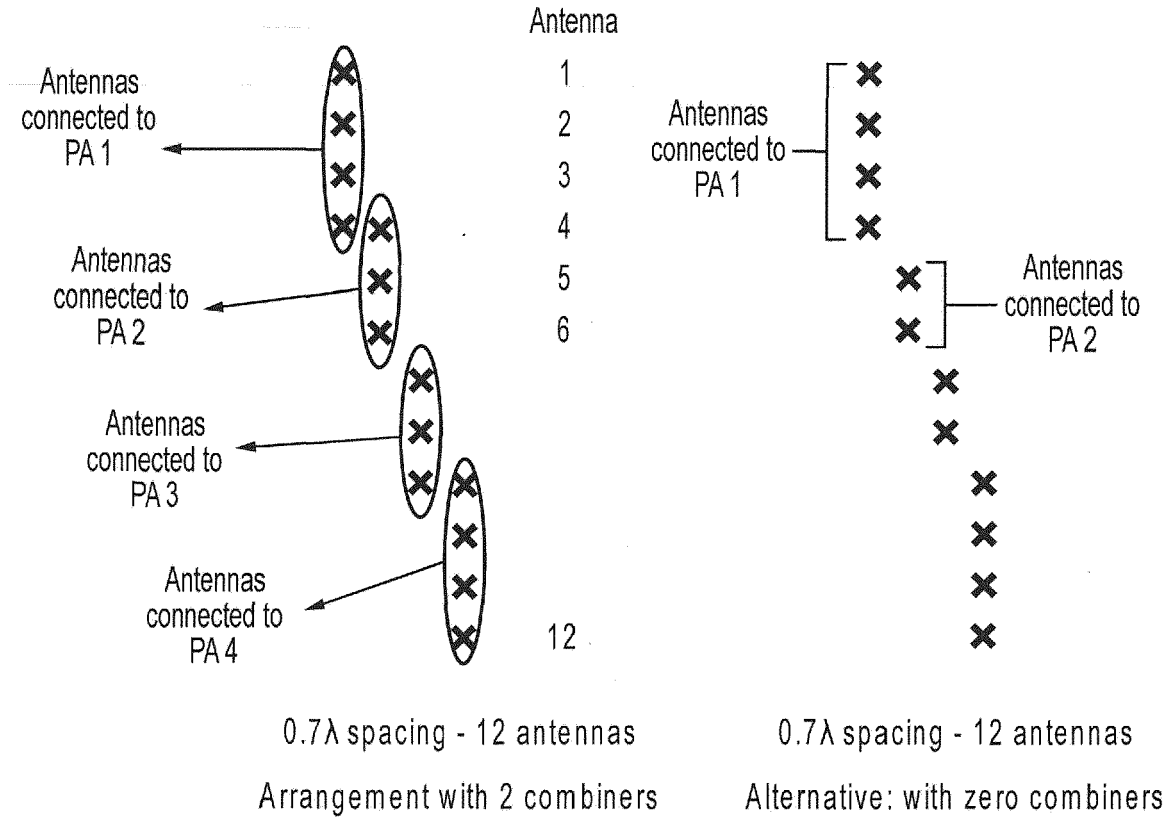

Antenna
1
2
3
4
5
6
12
Antennas connected to PA 1
Antennas connected to PA 2
Antennas connected to PA 3
Antennas connected to PA 4
0.7λ spacing - 12 antennas
Arrangement with 2 combiners
Antennas connected to PA 1
Antennas connected to PA 2
0.7λ spacing - 12 antennas
Alternative: with zero combiners

FIG. 4

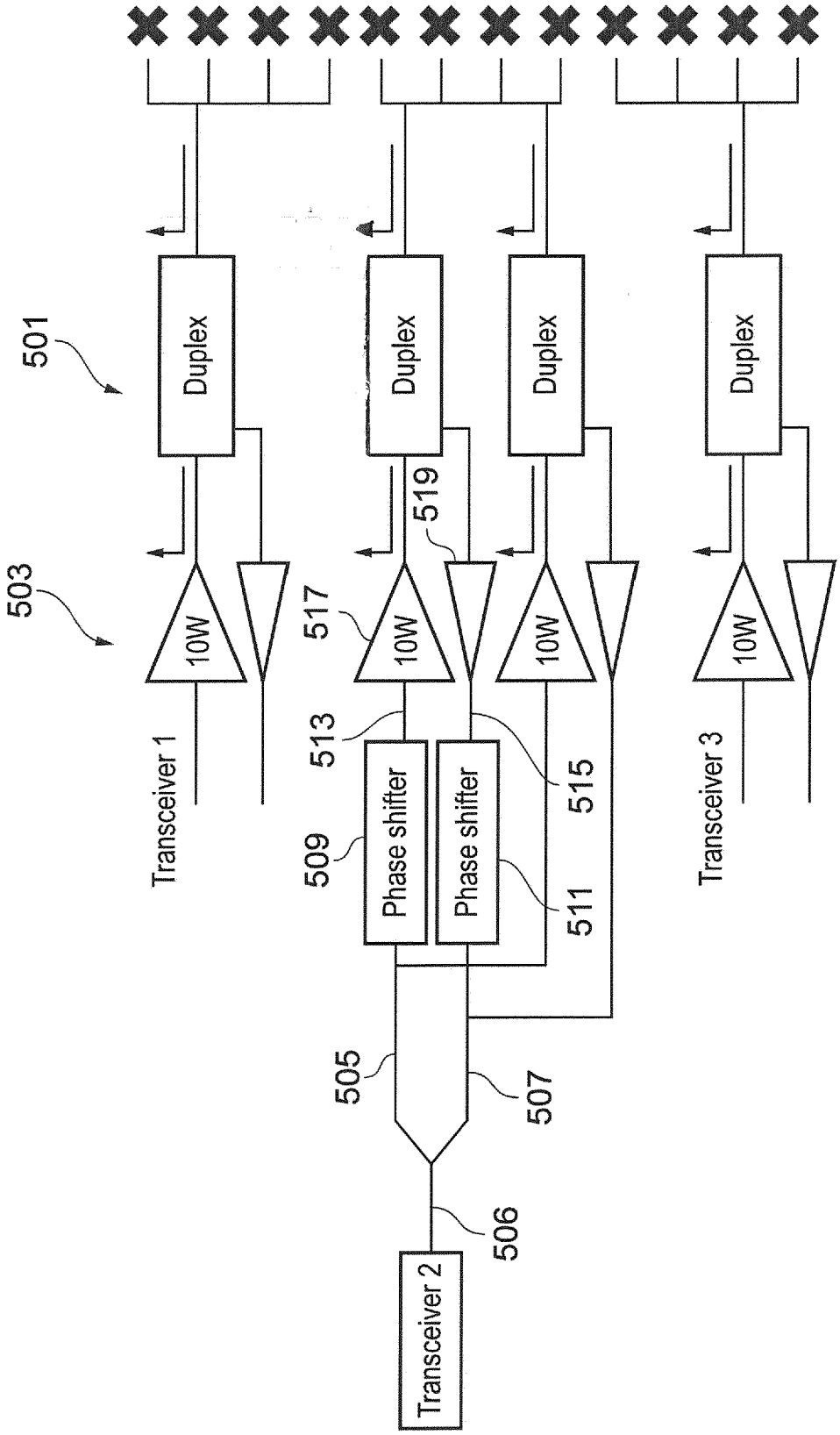

501
503
Duplex
Duplex
Duplex
Duplex
10W
10W
10W
10W
519
517
513
515
509
511
Phase shifter
Phase shifter
505
507
506
Transceiver 1
Transceiver 2
Transceiver 3

FIG. 5

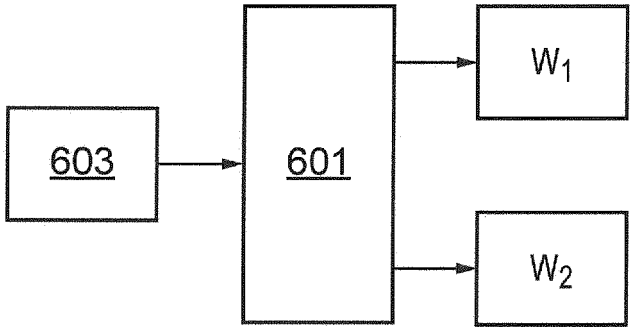


FIG. 6

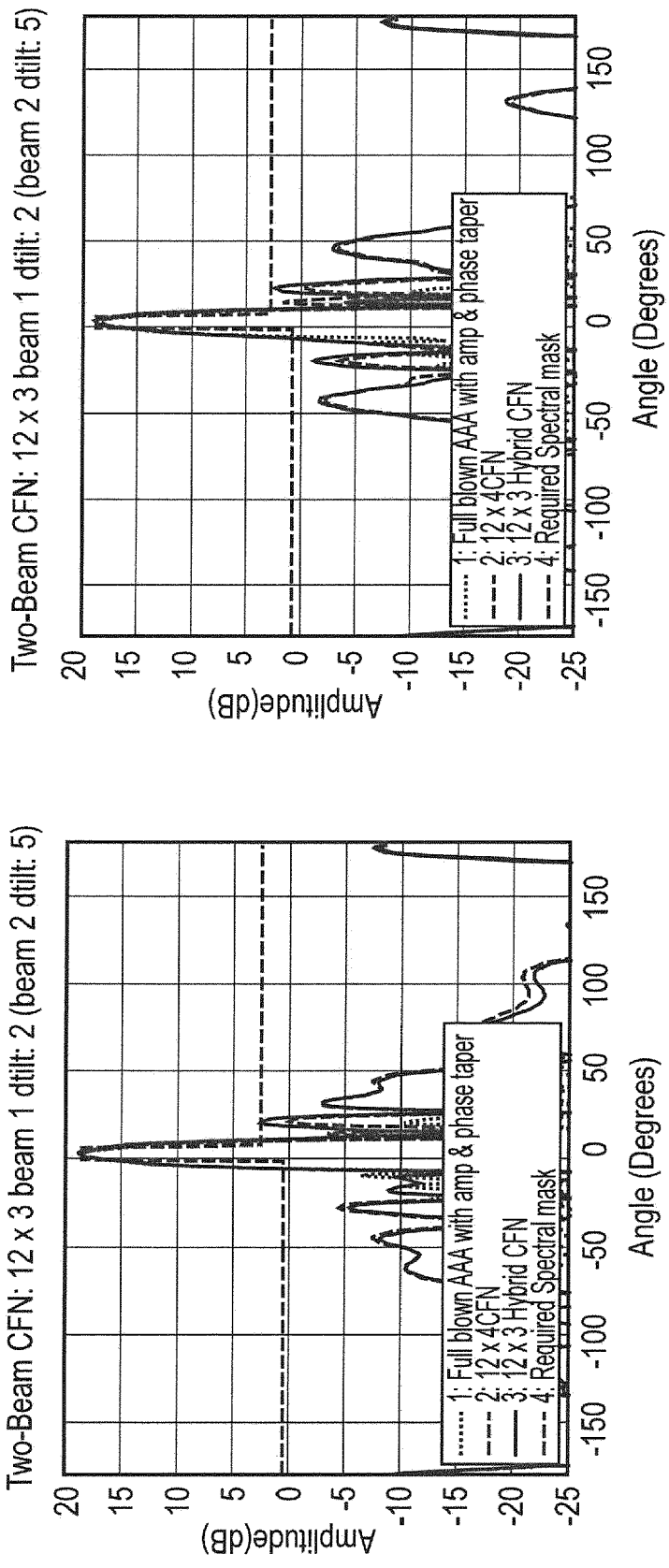
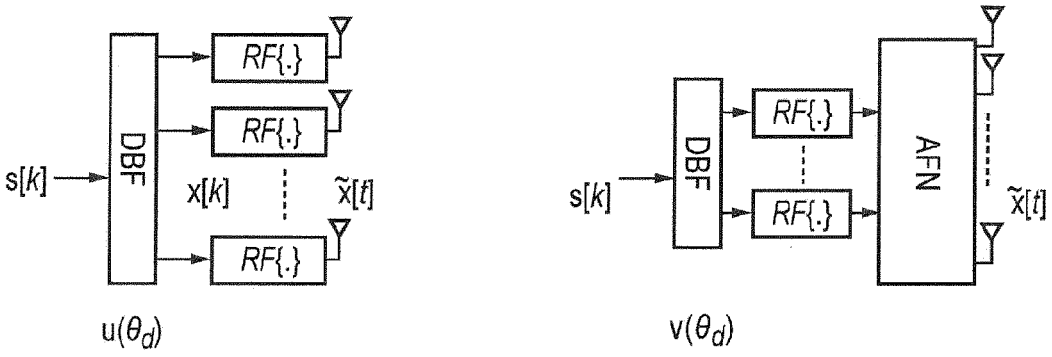

Two-Beam CFN: 12 x 3 beam 1 dtilt: 2 (beam 2 dtilt: 5)
Amplitude(dB)
Angle (Degrees)
1: Full blown AAA with amp & phase taper
2: 12 x 4CFN
3: 12 x 3 Hybrid CFN
4: Required Spectral mask
(a)
(b)

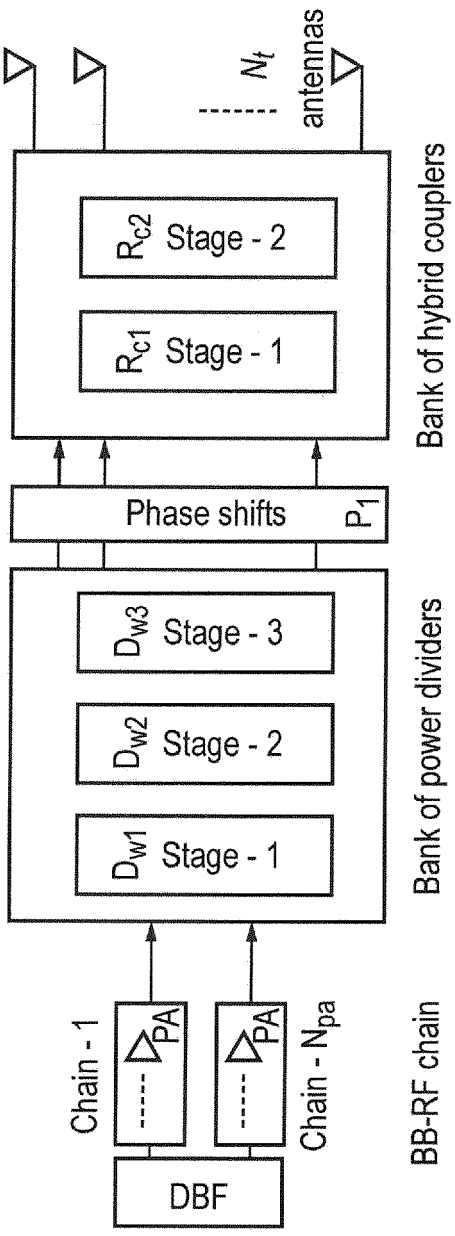
FIG. 7



$\tilde{x}[t] = RF\{u(\theta_d)s[k]\}$ $\tilde{x}_r[t] = WRF\{v(\theta_d)s[k]\}$

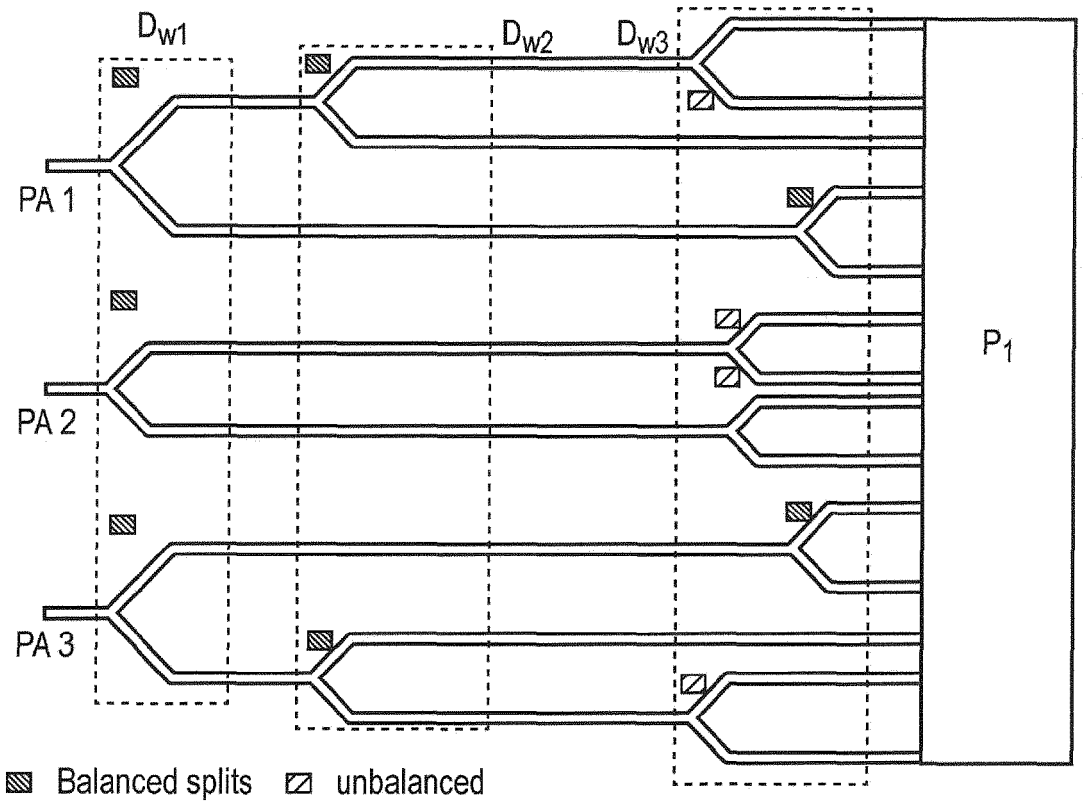
Adaptive antenna array architecture: (a) Modular AAA architecture with $N_{trx} = N_t$ tranceivers (denoted by $RF\{.\}$) digital beamformer $u(\theta_d)$ for all beamtilts $\theta_d \in R_\theta$ (b) Architecture with adaptive $u(\theta_d)$ beamforming signals connected to N_t x N_{trx} AFN and finally to N_t antennas.

FIG. A1



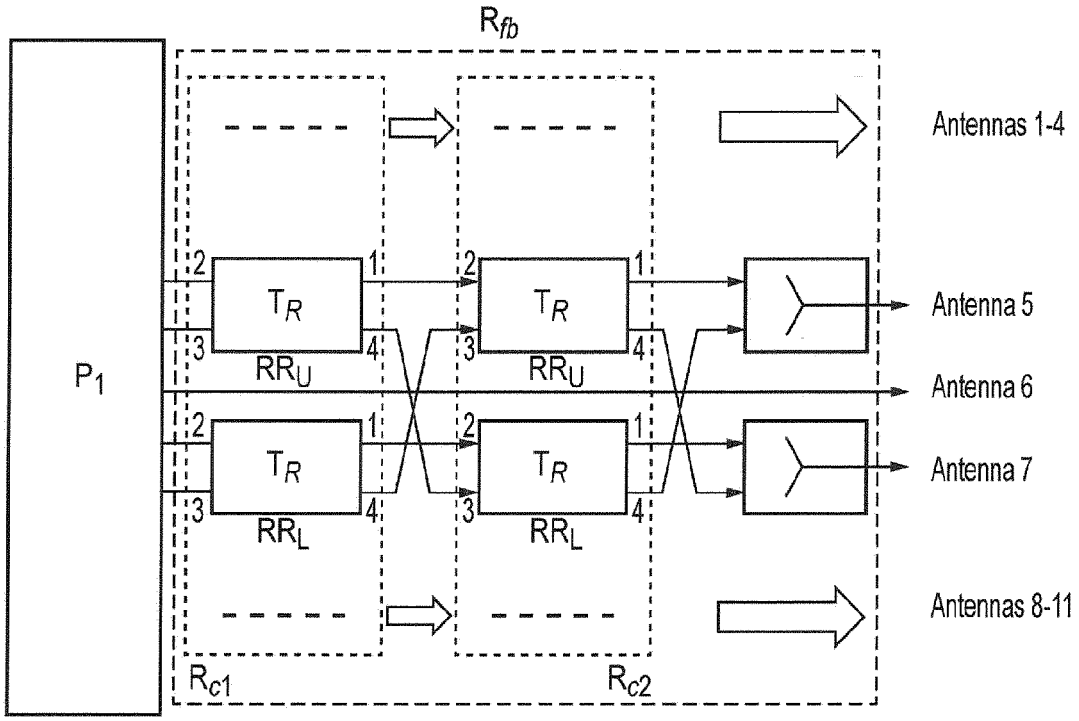
Factorization of an N_t x N_{trx} feeder network into a bank of dividers, phase-shifters, and DCs connected to N_t antennas. These stages are subsequently factorized into banks of 3-port or 4-port networks. Specific decompositions depend on the cellular architecture (such as macro-cell or small cell) and design objectives (such as insertion loss or beam pattern).

FIG. A2



Macro-cell AFN D_{fb} factorization: Decomposition of the divider bank into three stages of power dividers, which are subsequently decomposed into balanced ▧ and unbalanced Wilkinson power dividers ▨ . The weights, phase shifts, and power ratios are estimated by the algorithms proposed in Sec. III and Sec. IV

FIG. A3



Macro AFN R_{fb} factorization into multiple stages of directional couplers. Each stage is made-up of a 4-port rat-race couple (denoted by 2 x 2 matrix T_R) to compensate for insertion losses. For simplicity, only the combiners connecting Antennas 5 and 7 for a 11 x 5 setup is detailed.

FIG. A4

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EUROPEAN SEARCH REPORT

Application Number
EP 14 30 6148

DOCUMENTS CONSIDERED TO BE RELEVANT

Category	Citation of document with indication, where appropriate, of relevant passages	Relevant to claim	CLASSIFICATION OF THE APPLICATION (IPC)
X Y	EP 2 698 870 A1 (ALCATEL LUCENT) 19 February 2014 (2014-02-19) * paragraph [0066] - paragraph [0069]; figures 2,11 * * page 7, line 33 - line 34 * -----	1-3, 6-12,15 4,5,13, 14	INV. H01Q1/24 H01Q3/26 H01Q25/00 H01Q3/30 H01Q3/40
Y	AMIN M ABBOSH: "Ultra-Wideband Phase Shifters", IEEE TRANSACTIONS ON MICROWAVE THEORY AND TECHNIQUES, IEEE SERVICE CENTER, PISCATAWAY, NJ, US, vol. 55, no. 9, 1 September 2007 (2007-09-01), pages 1935-1941, XP011191455, ISSN: 0018-9480, DOI: 10.1109/TMTT.2007.904051 * figures 1-6 * -----	4,5,13, 14	H01Q21/00
A	EP 2 685 557 A2 (HUAWEI TECH CO LTD [CN]) 15 January 2014 (2014-01-15) * the whole document * -----	1-15	

TECHNICAL FIELDS SEARCHED (IPC)
H01Q

The present search report has been drawn up for all claims

Place of search	Date of completion of the search	Examiner
The Hague	13 January 2015	Sípal, Vít

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