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Optimizing RF Efficiency of a Vector-Modulator-Driven Antenna Array

Veli-Pekka Kutinlahti , Anu Lehtovuori , and Ville Viikari

Abstract—The feed vector of an antenna array determines the array pattern and the active input impedance of each element. On the other hand, the output impedance of a variable-gain power amplifier, such as that used in a vector-modulator, depends on the amplification level. The resulting impedance mismatch is typically neglected when designing and controlling antenna arrays at the cost of reduced gain, bandwidth, and beam-steering range. This letter presents a simple method to find the array feed vector that maximizes the realized gain in a given direction in a case when the power amplifier output impedance depends on its amplification level. We verify the method with simulations for a test case consisting of a 4×1 patch array fed with ideal phase-shifters and realistic amplifiers. The simulation results of the case study show that the method improves transmission efficiency by 3.5 dB on average compared to the traditional phase shift.

Index Terms—Beam steering, efficiency, eigenanalysis, weighting.

I. INTRODUCTION

ENERGY efficiency of wireless systems is becoming increasingly important due to exponentially growing wireless data traffic. Even 65% of the total energy consumption of a base station is used for radio-wave generation, mostly in power amplifiers (PAs) [1]. The need for energy can be decreased by increasing the realized gain of the transmitting antenna. The realized gain defines the antenna's ability to convert available power to propagating fields in the desired direction, and it is the product of matching efficiency and antenna gain.

The array feed vector, which defines the directivity of the antenna array, also affects the active reflection coefficient. In case the elements are mutually coupled, as they always are in practice, an incoming voltage wave in one port excites outgoing voltage waves in other ports. The active port impedance finally depends on how all the outgoing voltage waves sum up in each port. The feed vector providing the highest directivity may sacrifice the matching efficiency, and vice versa.

To mitigate the active impedance problem, many approaches have been introduced to lower the mutual coupling in an array. Most common method to lower coupling in an antenna array

is the decoupling network [2]–[4]. Other methods include neutralization lines [5], band-gap structures between the elements [6], wave traps [7], high-impedance structures [8], etc. Some of the approaches are dissipative, i.e., they absorb the energy that would have otherwise leaked to another element. Some are reflective so that they isolate the energy within the one radiating element. Reflective approaches are generally narrowband, whereas resistive methods decrease the efficiency due to resistive losses. In either way, active impedance effects are lowered at the cost of otherwise reduced performance.

Large coupling between antenna elements has been recently demonstrated to be beneficial in certain circumstances. Examples include the connected dipole [9], the antenna cluster [10], and combining power on an antenna [11]–[14]. The connected dipole provides a wideband behavior and high directivity. The antenna cluster method utilizes many coupled radiators collaboratively as one wideband frequency tunable antenna. Combining power on an antenna can avoid the need for lossy power-combining on chip networks occupying a large chip area.

On the other hand, the impedance of a PA depends on the output power. Usually modern antenna feeding weights require amplitude modulation, in which case the amplifiers should exhibit a fairly constant output impedance across a certain output-power range. Typically, stable output impedance comes at the expense of a narrow band, lower efficiency, or more complex circuitry [15]–[20]. When the antenna and the feeding PA are highly integrated, this varying impedance should be taken into account when selecting feeding weights, in order to not cause increasing mismatch.

In this letter, we introduce a method to maximize the realized gain of an antenna array driven with PAs with gain-dependent output impedance. The method accounts for active impedance mismatch and finds the feed vector that is the best tradeoff between the impedance mismatch and array gain in a given direction. The method significantly extends the beam steering range and frequency band. To the authors' knowledge, no prior active impedance matching or antenna array beam control methods take the varying output impedance of the PA into account. There has not yet been a method that maximizes realized gain when the impedances of the feeding PAs vary dynamically.

II. FORMULATION OF THE PROBLEM

A four-channel vector modulator with 4 bit gain control and 6 bit phase control exhibits over 17 billion feed combinations. When driven antenna elements are mutually coupled, there is no straightforward way to solve the best state for a given beam-steering angle other than evaluating all the cases one by

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Fig. 1. Antenna system consisting of a transmitting array (TX) and a receiving antenna (RX) represented with (a) a system S-matrix and (b) separate S-matrices.

one. However, the large number of cases makes this approach impractical already with four antenna elements and impossible for larger arrays.

We next derive an analytical, S-parameter-based formulation for the realized gain, and later use the formulation to find the optimal feed vector. The formulation takes into account the active impedance mismatch between the amplifiers and the antenna array together with gain of the antenna array. The formulation leads to a system of nonlinear equations, that are solved by iteration.

A simple transmission system consisting of a transmitting array (TX) including the feeding PAs and a receiving antenna (RX) is illustrated in Fig. 1. Sections of the system and interactions between them can be described with scattering matrices S_{TX} for TX, S_{RX} for RX, and the transmission vector S_{TR} between them. These can be combined to form the system matrix S_{SYS} as

$$b_{SYS} = \begin{bmatrix} b_{TX} \\ b_{RX} \end{bmatrix} = \begin{bmatrix} S_{TX} & S_{TR}^T \\ S_{TR} & S_{RX} \end{bmatrix} \begin{bmatrix} a_{TX} \\ a_{RX} \end{bmatrix} = S_{SYS} a_{SYS}.$$

When $a_{RX} = 0$, the transmission efficiency η_{TR} , i.e., the ratio of the power output from RX P_{rec} to the power input to TX P_{in} , can be formulated with the feeding vector a_{TX} and S_{TR} to be

$$\eta_{TR} = \frac{P_{rec}}{P_{in}} = \frac{b_{RX}^H b_{RX}}{a_{TX}^H a_{TX}} = \frac{a_{TX}^H S_{TR}^H S_{TR} a_{TX}}{a_{TX}^H a_{TX}} \quad (1)$$

where H denotes the conjugate transpose of a matrix. Assuming that the amplifiers are matched to the given reference impedance, the feed vector a_{TX} maximizing the transmission efficiency η_{TR} is the eigenvector corresponding to the largest eigenvalue of $S_{TR}^H S_{TR}$ [21].

The output impedance Z_{amp} of a PA, however, varies, among other things, with the dynamic gain, i.e., the values of a_{TX} . As a consequence, any solution where the values of a_{TX} deviate from

Fig. 2. Amplifier output impedance Z_{amp} as a function of gain. The change in the impedance is caused by the variation of supply voltage of the transistor controlling the PA power.

the nominal value corresponding to the nominal impedance, the assumption of matched amplifiers fails, and the obtained result is wrong. Also, the input impedance of antenna Z_{af} depends on a_{TX} , i.e., antenna exhibits an active impedance. In order to maximize the system efficiency η_{TR} , we should take into account the variation of both impedances.

In this letter, we use the simulated impedance behaviour of the PAs of the beam-forming transmitter chip described in [22]. Fig. 2 shows the amplifier impedance as a function of the normalized gain. When power is maximized, the impedance is close to 70 Ω but increases over 200 Ω when the transceiver is driven at very low power. The impedance is purely real at all PA gain levels. The behavior causes increased mismatch at the amplifier-antenna interface at low gain levels when the active antenna impedance is around 50 Ω. In order to overcome this, the impedance change should be taken into account to achieve the desired power variation in antenna elements.

Changes in Z_{amp} can be taken into account by re-normalizing the transmission vector S_{TR} when Z_{amp} change. The re-normalized S-matrix $\tilde{S}$ can be calculated with port impedances Z_{pn} , $i = 1, \dots, n$ and the system Z-matrix Z with

$$\tilde{S} = F(Z - Z_p)(Z + Z_p)^{-1}F^{-1} = \begin{bmatrix} \tilde{S}_{TX} & \tilde{S}_{TR}^T \\ \tilde{S}_{TR} & \tilde{S}_{RX} \end{bmatrix} \quad (2)$$

where

$$Z_p = \begin{bmatrix} Z_{p1} & \dots & 0 \\ \vdots & \ddots & \vdots \\ 0 & \dots & Z_{pn} \end{bmatrix}$$

$$F = \begin{bmatrix} \frac{1}{2\sqrt{Z_{p1}}} & \dots & 0 \\ \vdots & \ddots & \vdots \\ 0 & \dots & \frac{1}{2\sqrt{Z_{pn}}} \end{bmatrix}$$

from which the new $\tilde{S}_{TR}$ can be extracted [23]. The changing impedance makes $\tilde{S}_{TR}$ effectively a function of a_{TX} , i.e., $\tilde{S}_{TR}(a_{TX})$. Since the PAs are directly connected to the array elements, the port impedances are equal to the impedances of the PAs, i.e., $Z_{pn} = Z_{ampn}$.

III. FEEDING METHODS

We compare three different feeding methods. The first method, the traditional progressive phase shift method μ_0 , has the weights

$$\mathbf{a}_{\mu_0} = \begin{bmatrix} 1 & e^{-j\phi_0} & \dots & e^{-j(n-1)\phi_0} \end{bmatrix}^T \quad (3)$$

where ϕ_0 is the phase shift between elements in an array, in linear array the phase shift being

$$\phi_0 = kd \sin \theta \quad (4)$$

where k is the propagation constant, d is the element spacing in the array, and θ is the scan angle [24]. Method μ_0 provides the highest possible realized gain when antenna elements are uncoupled and inherently matched to the feed.

Method μ_1 , the second method, uses exact phases from the transmission vector $\mathbf{S}_{\text{TR}}$ for maximum power transmission from TX to RX. The individual phase shifts ϕ_n with respect to the phase of port 1 for the elements in the array are

$$\phi_n = \arg(\mathbf{S}_{\text{TR}n}) - \arg(\mathbf{S}_{\text{TR}1}) \quad (5)$$

making the weights for μ_1

$$\mathbf{a}_{\mu_1} = \begin{bmatrix} 1 & e^{-j\phi_2} & \dots & e^{-j\phi_n} \end{bmatrix}^T \quad (6)$$

Our novel iterative algorithm μ_2 , the third method, uses eigenvalue analysis on (1). The rightmost side has the form of a Rayleigh quotient, and thus η_{SYS} can be maximized by choosing $\mathbf{a}_{\text{TX}}$ to be the eigenvector corresponding to the largest eigenvalue of $\mathbf{S}_{\text{TR}}^H \mathbf{S}_{\text{TR}}$ [21], [25], [26]. These weights give the global maximum η_{TR} for a given $\mathbf{S}_{\text{TR}}$ vector under the assumption of 50 Ω amplifiers at all signal levels.

The solution for the weights calculated with eigenvalue analysis has typically quite a varied amplitude distribution. However, the solution ignores the changing port impedance of the PA, which we want to account for in our study. Since PA impedances $Z_{\text{amp}i}$ dynamically vary as a function of gains (see Fig. 2), the $\mathbf{S}$ -matrix needs to be re-normalized with (2) in order to calculate the real η_{TR} .

After re-normalization, the previously calculated weights are no longer optimal for the new re-normalized $\tilde{\mathbf{S}}_{\text{TR}}$. μ_2 combines the optimal solution given by eigenanalysis with (2) forming the novel iterative algorithm to improve η_{TR} over that of method μ_0 . The algorithm is given as follows.

- 1) Calculate the eigenvector $\mathbf{a}_0$ corresponding to the maximum eigenvalue of $\mathbf{S}_{\text{TR}}^H \mathbf{S}_{\text{TR}}$ from (1).
- 2) Calculate $\tilde{\mathbf{S}}$ from (2) with new port impedances $Z_{\text{amp}i}$ corresponding to $\mathbf{a}_k$ calculated in step 1) or 3).
- 3) Calculate the eigenvector $\mathbf{a}_{k+1}$ corresponding to the maximum eigenvalue of $\tilde{\mathbf{S}}_{\text{TR}}^H \tilde{\mathbf{S}}_{\text{TR}}$ from (1).
- 4) Calculate the change between $\mathbf{a}_k$ and $\mathbf{a}_{k+1}$. When a predetermined error in convergence is reached, terminate the algorithm. Otherwise, go to 2)

At its core, μ_2 finds the eigenvector of $\tilde{\mathbf{S}}_{\text{TR}}^H(\mathbf{a}_{\text{TX}})\tilde{\mathbf{S}}_{\text{TR}}(\mathbf{a}_{\text{TX}})$ corresponding to the maximum eigenvalue.

Fig. 3. (a) Antenna array, (b) the element geometry top view, and (c) the element geometry side view.

Fig. 4. Array matching and coupling. Gray lines are for coupling between elements.

IV. RESULTS

The test system, shown in Fig. 3, consists of a patch antenna array as a TX and a simple dipole as a RX. The patches are fed with probes located on the negative y -axis edges, and isolated patches are matched to 50 Ω . Fig. 4 shows the $\mathbf{S}$ -parameters of the TX array. RX is placed at coordinates $(r, \theta, \phi) = (1.5 \text{ m}, 0^\circ, 0^\circ)$ in the far-field broadside of the arrays on z -axis in a polarization matched orientation. RX is then moved along a circumference centered at TX with 10 degree steps. RX was tilted at each individual location to maintain polarization matched orientation with TX.

From the simulated $\mathbf{S}_{\text{SYS}}$, the transmission vector $\mathbf{S}_{\text{TR}}$ is extracted and used to calculate weights for methods μ_1 and μ_2 , while weight for traditional progressive phase shift method μ_0 is calculated from the array geometry. The weights are used to calculate the transmission efficiency η_{TR} for each method with (1). Calculations are done for point frequencies in MATLAB and the presented results are envelopes encompassing the best

Fig. 5. Change in array η_{TR} using μ_2 with respect to η_{TR} using μ_0 .

Fig. 6. Change in array η_{TR} using μ_2 with respect to η_{TR} using μ_1 .

performance of each method at a given steering angle and frequency.

Fig. 5 shows the improvement of η_{TR} for μ_2 in respect to μ_0 . Average improvement over the calculated region is 3.5 dB. Apart from couple of regions at extreme steering angles, the algorithm improves η_{TR} over the whole calculated region. The results of the algorithm are confirmed by calculating 100 000 random samples of different feeding weights to assess the validity of μ_2 converging toward the best possible result to maximize η_{TR} .

Fig. 6 shows the improvement of η_{TR} for μ_2 in respect to μ_1 over the same region as Fig. 5. The improvement of η_{TR} is smaller compared to the results in Fig. 5 being 0.9 dB on average. This is understandable, as μ_1 takes into account the distortions in the element patterns whereas μ_0 assumes similar ideal patterns. Interelemental coupling affects element patterns quite drastically in small arrays, as edge elements exhibit coupling differently compared to inner elements. The pattern distortion directly affects coupling of TX elements to RX, which is used to calculate the weight of μ_1 .

As is customary, the designed patch antenna in this letter is matched to 50 Ω . The PAs of the IC however exhibit a minimum output impedance of 70 Ω . Because of this the elements of the

Fig. 7. Change in array η_{TR} using μ_2 normalized with respect to η_{TR} using μ_0 at 3.1 GHz.

array are never perfectly matched, albeit close. As the impedance of the PAs further increases with decreasing gain, the individual elements are further mismatched if an amplitude variation is used to feed the array.

The novel situation raises a question of how the impedance of the antenna should be selected with a dynamically varying PA impedance. The question was studied by employing the original impedance curve in Fig. 2 of the PA shifted by a constant impedance value ΔZ_0 which is between $\pm 60 \Omega$.

Fig. 7 shows the improvement in η_{TR} with method μ_2 compared to μ_0 for the array, when the impedance curves of the PA are adjusted with a constant ΔZ_0 . The x-axis corresponds to ΔZ_0 and the y-axis to the steer-angle. Calculations were done at the resonant frequency 3.1 GHz. When ΔZ_0 is increased the improvement of η_{TR} at large beam steering angles decreases, whereas at small angles the improvement in η_{TR} stays almost the same. Only when $\Delta Z_0 < -40 \Omega$ the improvement increases at small steering angles. This is because the individual element matching improves between $-40 \Omega < \Delta Z_0 < 0 \Omega$ at maximum PA gain. When $\Delta Z_0 < -40 \Omega$ the matching actually becomes worse. Method μ_2 takes this into account and lowers the PA gain to improve matching, while μ_0 and μ_1 do not.

V. CONCLUSION

A novel method to calculate feeding weights of an antenna array to maximize transmission efficiency, was presented. The method takes into account the output impedance of the feeding PAs. The previously known technique is used to facilitate the specific problem in this letter and can be used for any antenna array. The simulated example case achieves improved transmission efficiency when compared to traditional phase shifting. On average, a 2.1 dB improvement is achieved in this study, which could be further increased with higher array coupling, as the method takes the coupling into account. Furthermore, the effect of the PA impedance was demonstrated to have a quite drastic effect on the transmission efficiency of the example case. Additional design targets, e.g., to achieve certain sidelobe level, could be introduced in future to extend the method.

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