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SAR Reduction With Antenna Cluster Technique

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Abstract—An efficient and straightforward antenna design method is presented that maximizes the ratio of total efficiency and specific absorption rate (SAR). This goal is achieved with a multiport antenna cluster technique where several ports are excited collectively with appropriate feeding weights. These weights are found as an eigenvalue solution formulated in terms of the radiated and near-field power of an antenna. The method relies on both the near-field (SAR) and far-field (radiation) physics of an antenna and it can be applied to various design cases. The importance of the maximization of total-efficiency-SAR ratio and the feasibility of the proposed approach are demonstrated with a simple multiple-dipole antenna example and with a more realistic antenna design where a metal-rimmed antenna is held in the user’s hand.

Index Terms—Antenna cluster, mobile antenna, Rayleigh quotient (RQ), specific absorption rate (SAR).

I. INTRODUCTION

The number of antennas used in close proximity to the user is continuously increasing emphasizing the importance of exposure-aware antenna design. Safety limits are determined with a parameter called the specific absorption rate (SAR). This parameter measures the absorbed electromagnetic power per unit mass of tissue. Both the American [1] and the European standards [2] have defined limits for maximum SAR values. These values have to be taken into account in the design of mobile devices [3], [4], [5], [6] as well as in different wearable antenna solutions [7], [8], [9].

Several detailed studies have been performed in the past to predict the SAR of an antenna (see, e.g., [10], [11]). In these studies, the focus is more on the estimation of the SAR than on reducing it. In addition, SAR computation is typically performed as a postprocessing step once the antenna has been designed [12], rather than taking SAR minimization as a design goal.

The easiest way to achieve the SAR safety limits is to lower the transmit power, but the reduction in transmit power leads to weaker transmitted fields and reduced coverage. Another straightforward way to reduce SAR is to locate antennas as far as possible from the user, for example, to the bottom of the device. This approach is not suitable for modern multiantenna mobile devices since all antennas cannot locate at the bottom of the device. Other widely used methods are insulating the antenna with wave-absorbing material, or shielding it with conductive material [13], [14], [15]. The downside of these methods is that they typically affect negatively the gain, efficiency, and bandwidth of the antenna [4]. Metamaterial structures [16], [17], [18] and electromagnetic bandgaps [19] have also been studied in SAR reduction. In practice, however, implementation of these structures into modern thin and small ground clearance devices is difficult due to their large size [3].

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SAR is defined as the power absorbed by the unit mass of tissue [23], [24]. The dielectric body is divided into small cubes \( D_i \) with mass \( m \), and at a fixed time-harmonic frequency, so SAR is associated with the cube center \( c_i \) as

\[
\text{SAR}_i(c_i) = \frac{1}{2V_i} \int_{D_i} \frac{\sigma(r) ||E(r)||^2}{\rho(r)} dV.
\]

(1)

Here, \( V_i \) is the volume of \( D_i \), \( \sigma \) and \( \rho \) are the conductivity and density of the tissue, respectively, and \( E \) is the electric field.

As mentioned above, it may not be sufficient to study solely SAR. Rather, for a realistic antenna, for example, a handset or wearable, both the total efficiency and SAR must be taken into account during the design process. We define an FoM that describes the ratio of the total efficiency, \( \eta_{\text{tot}} \), and the maximum SAR

\[
\text{FoM} = \xi \frac{\eta_{\text{tot}}}{\text{max}(\text{SAR}_i)}, \quad \eta_{\text{tot}} = \frac{P_{\text{rad}}}{P_{\text{in}}},
\]

(2)

Here, \( P_{\text{rad}} \) and \( P_{\text{in}} \) are the radiated and input power of an antenna, and \( \xi = P_{\text{in}}/m \) is a normalization factor. Mass \( m \) is either 1 or 10 g depending on the used SAR standard.

The benefit of the low SAR is that it enables higher input power without exceeding SAR safety limits. Naturally, higher input power means higher transferred power. Since the FoM takes into account both the SAR and the total efficiency, the antenna with a high FoM can transfer more power while the SAR value is still below the safety limits.

### B. RQ and Near-Field Power

From the numerical optimization point of view, the definition of maximum SAR poses two challenges. First, SAR depends on the position, and the optimization process can be sensitive to the choice of the cube center \( c_i \). Second, the definition of a cube where SAR is evaluated is rather complicated as the cube is near the surface of the body [25].

Our goal is to formulate the optimization problem as an RQ. The optimal solution to the problem can then be found by solving an eigenvalue problem

\[
Ax_n = \lambda_n B x_n
\]

and time-consuming numerical optimization can be avoided. The wanted optimal solution is the eigenvector \( x_n \) corresponding to the maximum (or minimum) eigenvalue \( \lambda_n \). Obviously, FoM defined in (2) cannot be expressed as an RQ. Therefore, we need to find another quantity that describes FoM as well as possible and can be expressed as an RQ.

In [23], it is demonstrated that the maximum electric field appears on the surface of the body block. This means that in minimizing maximum SAR, it is sufficient to study the maximum electric field on the body surface and minimize it. However, using a field value defined at a single point is usually numerically unstable and sensitive even to very small numerical errors. To avoid this problem, we consider the near-field surface power defined here as

\[
P_{\text{near}} = \frac{1}{2\eta_0} \int_S ||E(r)||^2 dS
\]

(4)

where the electric field is integrated over a surface \( S \) and \( \eta_0 = \sqrt{\mu_0/\epsilon_0} \) is the free space wave impedance.

The optimization goal is then formulated as the ratio of the radiated (far-field) power and the surface (near-field) power. In Section II-C, it is shown that this ratio can be expressed as an RQ in terms of port input currents.

### C. Formulation With Port Input Currents

Let us next assume that the antenna structure includes \( P \) feeding ports and let \( I = [I_1, \ldots, I_P]^T \) denote the complex-valued input currents of the ports. These port input currents are the unknowns to be optimized. First, the radiated power is expressed utilizing the far-field matrix \( F^\text{far} \) [26] with elements

\[
F^\text{far}_{pq} = \frac{1}{\eta_0} \int_{S_\infty} E_p^* (r) \cdot E_q (r) dS,
\]

(5)

Here, \( S_\infty \) is a spherical surface in the far-field region, \( E_p \) is the electric field due to unit input current at port \( p \), and \((\cdot)^*\) denotes complex conjugate. Using \( F^\text{far} \), we have

\[
P_{\text{rad}} = \frac{1}{2} I^H F^\text{far} I
\]

(6)

where \((\cdot)^H\) denotes Hermitian transpose. Analogously, we define a near-field matrix with elements

\[
P^\text{near}_{pq} = \frac{1}{\eta_0} \int_S E_p^* (r) \cdot E_q (r) dS,
\]

(7)

and express the near-field surface power as

\[
P_{\text{near}} = \frac{1}{2} I^H F^\text{near} I
\]

(8)

Obviously, both matrices \( F^\text{far} \) and \( F^\text{near} \) are positive definite, and the ratio of the far-field radiated power and the near-field surface power can be formulated as an RQ. Consequently, the generalized eigenvalue equation to be solved reads

\[
F^\text{far} I_n = \lambda_n F^\text{near} I_n
\]

(9)

and the optimal solution (port input currents) is the eigenvector \( I_n \) corresponding to the largest eigenvalue \( \lambda_n \).

### D. Numerical Implementation

The far-field and near-field matrices, needed in the eigenvector equation (9), are implemented using an in-house MoM code for combined PEC and lossy dielectric bodies [27], [28] with multiple feeding ports. For given inputs, the program finds the equivalent electric surface current on the antenna and the equivalent electric and magnetic surface currents on the surface of the dielectric block. Matrices \( F^\text{near} \) and \( F^\text{far} \) are constructed by using unit inputs at the ports, one by one, and by computing the corresponding electric field in the near and far-field regions.

While the assembly of the far-field matrix requires numerical integration, the near-field matrix \( F^\text{near} \) can be obtained directly from the MoM solution. Let \( J = n \times H \) and \( M = -n \times E \) denote the equivalent electric and magnetic surface currents on the surface \( S \) of the dielectric body. The surface electric field at point \( r \in S \) is given by

\[
E(r) = n(r) \times M(r) - \frac{n(r)}{j\omega \epsilon} \nabla_s \cdot J(r).
\]

(10)

Here, \( \nabla_s \cdot J \) is the surface divergence of \( J \), and \( n \) is the exterior unit normal vector of \( S \). In the numerical solution, integral (7) is evaluated at the center points of the triangles of the surface mesh and the integration weights are the areas of the triangles. Thus, the elements of the near-field matrix can be obtained as

\[
F^\text{near}_{pq} \approx \frac{1}{\eta_0} \sum_{n=1}^{N} A_n E_p^* (r_n) \cdot E_q (r_n)
\]

(11)

where \( A_n \) and \( r_n \) are the area and the center point of a triangle \( T_n \) on \( S \), respectively. The electric field is evaluated using (10).
E. Other RQ-Based Optimization Goals

Above we showed how the ratio of far-field power and near-field power can be expressed as an RQ. To compare this approach with other design goals, we next shortly review the formulation of other antenna design parameters as RQs. Let $S$ be the S-parameter matrix of a multiport antenna. Total active reflection coefficient (TARC), defined as

$$\text{TARC} = \frac{a^H S^H S a}{a^H a}$$  \hspace{1cm} (12)

can be formulated as an eigenvalue problem [20]

$$\lambda_n = \min \text{eig} \left(S^H S\right)$$  \hspace{1cm} (13)

and the “TARC-optimal” solution is the eigenvector corresponding to the eigenvalue $\lambda_n$. Also radiation efficiency

$$\eta_{\text{rad}} = \frac{P_{\text{rad}}}{P_{\text{rad}} + P_{\text{diss}}} = \frac{I^H F_{\text{far}} I}{I^H R I}$$  \hspace{1cm} (14)

and total efficiency

$$\eta_{\text{tot}} = \frac{P_{\text{rad}}}{P_{\text{in}}} = \frac{I^H F_{\text{far}} I}{a^H a}$$  \hspace{1cm} (15)

can be formulated as RQs, since all involved matrices are Hermitian and positive definite. Here, $a = (Z_0)^{-1/2}(Z + Z_0) I$, with a diagonal matrix $Z_0$ including the characteristic impedances of the ports, and $P_{\text{diss}}$ is the dissipated power due to dielectric losses. The dissipated power can be expressed with the real part of the port input impedance matrix, $Z = R + j X$, and the far-field matrix as

$$P_{\text{diss}} = \frac{1}{2} I^H \left(R - F_{\text{far}}\right) I.$$  \hspace{1cm} (16)

If the antenna is lossy, radiation efficiency should include the contribution of metallic losses, too. In most practical situations, dielectric losses due to human tissue dominate over metallic losses and the assumption of a lossless antenna are reasonable.

III. EXAMPLES

In this section, we demonstrate the feasibility of the proposed method in maximizing FoM and investigate solutions with different input weight optimization goals. The first example considers a cluster of three dipole antennas on top of a dielectric (fat) block. The second one involves a more practical metal-rimmed mobile antenna device in a human hand. The optimal feeding weights for the multiport antenna are solved with in-house codes, and the corresponding SAR, S-parameter, and efficiency results are computed with CST Studio Suite.

A. Three Dipoles

The goal of this section is to compare different input weight optimization goals to show that traditional approaches may lead to prohibitively high SAR values. We consider a cluster of three dipoles of lengths (46, 48, and 50 mm) placed above a fat block of dimensions 40 mm $\times$ 72 mm $\times$ 30 mm. Due to the dissenting lengths of the dipoles, the resonance frequencies of the dipoles are different. This enables effective utilization of the antenna cluster technique and achieves wider bandwidth. The simulation setup is shown in Fig. 1. The distance between the dipoles is 11 mm, and the distance between the dipoles and the fat block is 5 mm. The size of the dielectric block is chosen large enough to model an antenna close to a human body so that edges do not have a significant effect on the results. In this example, the parameters of the fat block ($\epsilon_r = 10.6$, $\delta = 0.389$ S/m, and $\rho = 911$ kg/m$^3$) are assumed frequency-independent.

The feeding weights of the dipoles are found by using four different optimization goals: TARC, $\eta_{\text{rad}}$, $\eta_{\text{tot}}$, and $P_{\text{rad}}/P_{\text{near}}$. The optimal solutions are based on the RQ formulations introduced in Sections II-C and II-E. In addition, we show results also for a nonoptimized case, that is, when all three dipoles have equal feeding weights (equal in legends).

Fig. 2(a) shows TARC with different feeding weights. Naturally, the TARC-optimized weights give the best TARC. Differences between other optimization goals are rather small but $\eta_{\text{tot}}$ and equal weights lead to slightly better matching than $P_{\text{rad}}/P_{\text{near}}$ or $\eta_{\text{rad}}$.

The FoM, defined as in (2), with 1 g of SAR and with different feeding weights are rather small, except in the TARC optimal weights which lead to significantly lower efficiency than others. Fig. 3(b) represents the 1-g maximum SAR when input power is 22.4 dBm (175 mW). Defining the input weights so that TARC is optimized produces clearly the highest SAR, almost
three times as high as with \( \eta_{\text{rad}} \) and \( P_{\text{rad}}/P_{\text{near}} \) weights. Also equal and \( \eta_{\text{tot}} \) weights produce about 30%–80% higher SAR than \( \eta_{\text{rad}} \) and \( P_{\text{rad}}/P_{\text{near}} \) weights.

SAR with both TARC and \( \eta_{\text{tot}} \) input weights varies rather much as a function of frequency, while the other optimization methods achieve less frequency-variant SAR. Both the high SAR and its frequency-dependent behavior can be explained with SAR patterns shown in Fig. 4. A high SAR value is obtained when the SAR maximum is concentrated on a small area as shown in Fig. 4(b) with TARC weights. Equal feeding achieves lower SAR than TARC-optimized weights since its SAR has distributed on a wider area near the center dipole. Accordingly, the SAR of both \( \eta_{\text{rad}} \) and \( P_{\text{rad}}/P_{\text{near}} \) weights obtain the lowest values since their patterns do not exhibit hot spots, like TARC, but spread more constantly as shown in Fig. 4(c). An SAR pattern with \( \eta_{\text{tot}} \) weights behaves similarly to the one of TARC weight except for around 2.85 GHz, where it obtains a local minimum. Around that local minimum, the SAR distribution is similar as with \( \eta_{\text{rad}} \) and \( P_{\text{rad}}/P_{\text{near}} \) weights.

Fig. 5 shows the amplitudes and phases of the TARC-, \( P_{\text{rad}}/P_{\text{near}} \)-, \( \eta_{\text{rad}} \)-, and \( \eta_{\text{tot}} \)-optimized input weights. The input weights obtained with \( P_{\text{rad}}/P_{\text{near}} \) and \( \eta_{\text{rad}} \) optimization goals are similar especially in the middle of considered frequency band. Therefore, their FoM and other results are similar. Clearly, \( P_{\text{rad}}/P_{\text{near}} \) weights have only moderate frequency dependence, while TARC and \( \eta_{\text{tot}} \) weights are strongly frequency-dependent. Since the radiation properties of an antenna cluster are not unique, different feeding weights can give the same far-field result. This also explains why at 2.85 GHz the results with \( \eta_{\text{tot}} \)-, \( \eta_{\text{rad}} \)-, and \( P_{\text{rad}}/P_{\text{near}} \) input weights are almost identical.

**B. Metal-Rimmed Mobile Antenna Design**

As a more realistic antenna design, we consider a metal-rimmed handset antenna held in a human hand as shown in Fig. 6. The overall dimensions of the device are 71 mm × 150 mm × 5 mm and the ground clearance is 3 mm all around the device. The minimum distance between the rim and the hand is 1 mm. We compare two antenna designs. In the first one, a 30 mm long element is placed on the rim (port 2 in Fig. 6). In the second one, we have a “three-element stack” of equal length 30 mm elements (ports 1, 2, and 3 in Fig. 6). The elements are overlapping so that element 2 locates on the rim (3 mm from the ground plane), and elements 3 and 1 are at the distance of 1 and 2 mm from the rim, respectively. The excitation for element 2 is located at the center of the element, element 1 is excited from the right end, and element 3 from the left end. The heights of elements 1 and 3 are 4 mm and the height of element 2 is 5 mm. The equal length of the elements enables a compact design, and the asymmetrical location of the feeding ports is favorable for the cluster design.

The same feeding weight optimization methods are used as in the case of three dipoles: TARC, \( \eta_{\text{rad}} \), \( \eta_{\text{tot}} \), and \( P_{\text{rad}}/P_{\text{near}} \). The optimization results are compared with the nonoptimized case (equal) and with the one-element case (1-element).

Fig. 7 shows TARC and total efficiency for the metal-rimmed antenna with different optimization methods. The 5G frequency band n77 (3.3–4.2 GHz) has been marked with a black dashed line in the figure. Similar behavior as in the case of three dipoles can be noticed, and all optimization methods obtain adequate matching and total efficiency in the n77 band.

Similarly, as in the 3-dipole case, a clear advantage of multiport feeding with properly optimized weights, either \( P_{\text{rad}}/P_{\text{near}} \) or \( \eta_{\text{rad}} \) case, can be seen in the maximum SAR shown in Fig. 8. In the considered frequency band, the maximum SAR of the 1-port design is two to four times higher than that of the best three-port case. We may also observe that the choice of the optimization goal has a
significant effect on the results. For example, optimizing the TARC of the 3-port design can lead to an even higher SAR than that in the 1-port design. In addition, the relative differences between the results of the optimization goals are larger in the case of 1-g SAR than in the case of 10-g SAR. SAR for a limb is defined for a 10-g tissue volume [1], [2] but we also show 1-g SAR for comparison. Obviously, in 1-g SAR, the tissue volumes are much smaller, since the maximum SAR values are significantly higher and more focused on the surface of the body than for 10-g SAR.

The SAR patterns with TARC and $P_{\text{rad}}/P_{\text{near}}$ weights are shown in Fig. 9. Similar behavior as in the case of three dipoles can be observed. The SAR pattern with TARC weights has clearly stronger local maxima than with $P_{\text{rad}}/P_{\text{near}}$ weights. Actually, one element and all other considered feeding methods, except $P_{\text{rad}}/P_{\text{near}}$ and $\eta_{\text{rad}}$, show similar high local maxima as TARC weights. Hence, $P_{\text{rad}}/P_{\text{near}}$ and $\eta_{\text{rad}}$ weights lead to clearly the lowest SAR values.

Fig. 10 shows the FoM with various input weights. In computing FoM, both 1-g and 10-g maximum SAR are used. The results demonstrate that defining the input weights, so that the far-near field power ratio is maximized, leads to the best efficiency-SAR ratio on a wide frequency range.

To get more insights into the function of the multiport antenna feeding design with different optimization goals in SAR reduction, we next study both the input weights and the surface currents. Fig. 11 shows the amplitudes and phases of the $P_{\text{rad}}/P_{\text{near}}$-optimized input weights. These results show that on the n77 band, the input weights of both methods are near constants, and their amplitudes are almost equal. The main difference appears in the phase difference of the feeding signal of port 3 compared to the other ports. In the $P_{\text{rad}}/P_{\text{near}}$ weights, this difference is about $-60^\circ$, while in the TARC weights, it is about $+20^\circ$.

In Fig. 12, we display the real part of the surface current on the antenna elements calculated with TARC and $P_{\text{rad}}/P_{\text{near}}$ input weights at 3.7 GHz. The currents on the two outermost elements, elements 1 and 2, are akin due to similar amplitude and phase, while significant differences in the currents on element 3 can be observed. Since with $P_{\text{rad}}/P_{\text{near}}$ weights, the currents on the element furthest from the hand (element 3) are stronger and the currents on the nearest element...
Fig. 13. Realized gain of the metal-rimmed mobile antenna with TARC and $P_{\text{rad}}/P_{\text{near}}$ input weights at 3.7 GHz. (a) $yz$ and (b) $xy$ planes.

(element 2) are smaller than with TARC weights, SAR is lower with $P_{\text{rad}}/P_{\text{near}}$ input weights.

Fig. 13 illustrates the realized gain in $xy$ and $yz$ planes with TARC and $P_{\text{rad}}/P_{\text{near}}$ input weights at 3.7 GHz. Clearly, these patterns are nearly identical and thus the antenna radiation pattern is almost independent of the used input weights. This result further verifies that with the proposed method low SAR values can be obtained without sacrificing antenna radiation.

IV. CONCLUSION

A novel antenna design method based on the multiport cluster technique that simultaneously provides high efficiency and low SAR is introduced. The optimal weights of the feeding ports are found as a solution to a generalized eigenvalue problem expressed with the radiated and near-field power port matrices of the antenna. The feasibility of the proposed approach is demonstrated with a metal-rimmed mobile antenna held in the user's hand. More than 50% reduction in maximum SAR values can be obtained compared to a conventional single-port antenna design, without significantly sacrificing the efficiency. It is also shown that solely maximizing matching (minimizing TARC) can lead to prohibitively high SAR values.

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