Visualization and Reduction of Mutual Coupling Between Antennas Installed on a Platform

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Abstract—Mutual coupling, or equivalently, the isolation between antennas, is a key parameter in antenna system design. In this work, the previously defined impedance density is generalized, and it is demonstrated how it can be used to obtain spatial information about the mutual coupling. The generalized impedance density is a real-valued scalar and can be visualized as a three-dimensional density in space. It is shown that there is a strong connection between regions with a positive (negative) generalized impedance density and a decrease (increase) of the coupling when an absorber is placed in that region. This predictive ability is a useful feature, which is tested for three numerical cases. The results are robust to the shape of the platform, and it can be compared across frequencies. By placing absorbers based on the generalized impedance density, it is possible to reduce the required amount of absorbers needed to obtain a certain reduction in mutual coupling. The visualization results and predictions of absorber positions are compared with a Poynting vector based method. Placing absorbers based on the generalized impedance density had a larger impact on the mutual coupling, compared to the predictions with the Poynting vector based method. By placing absorbers based on the generalized impedance density, it is possible to reduce the mutual coupling between antennas. Investigations of absorbers to mitigate the mutual coupling. A visualization method should describe all the above-mentioned mechanisms to be a useful tool for identifying regions with strong contributions to the mutual coupling. A path between a transmitting and receiving antenna that depicts the spatial contribution to the mutual coupling is referred to as a coupling path [10]–[13].

Several methods to visualize mutual coupling have been suggested in the literature, including ray-tracing and diffuse scattering [14], [15]. These methods are generally not applicable to antennas installed on platforms since the near-field contribution can substantially contribute to the coupling. A Poynting vector based method was defined [12], [13] and applied [16], [17] to mitigate electromagnetic interference. In these works, the coupling paths are visualized by computing streamlines through the power flow between the receiving and transmitting antennas. The Poynting vector based method, in its current form, does not distinguish between the incident and accepted power. Also, as seen in [17], the streamlines follow different paths depending on which antenna is transmitting and receiving, even though the setup is reciprocal, see [18].

In our previous paper [10], a normalized impedance density was defined using the reaction theorem [19]–[21]. It was proposed that paths of strong positive normalized impedance density are indeed coupling paths. A method to test this idea with a small absorber was also proposed. However, no such validation results were presented at the time. In the present paper, we follow up and improve on this idea. We here provide both a theoretical motivation between coupling paths and the impedance density and illustrate a numerical validation of coupling paths. We show that the real-part of a suitably normalized impedance density is directly related to the mutual impedance and that it indeed can be interpreted as coupling paths for weakly coupled antennas. It furthermore provides a predictive visualization method of important regions for weak mutual coupling between antennas.

Predictive here means, in a perturbative sense, that a small absorber placed on a positive region will reduce the mutual coupling between the antennas. Furthermore, we also show that an absorber placed on a negative region will increase mutual coupling. This predictability is compared with the ability of the method based on the Poynting vector [12], [13]. The comparisons are shown through numerical simulations in CST Microwave Studio for different setups.

It is well known that the distribution of electromagnetic fields over a platform can be sensitive to small changes in the geometry. Such geometry changes can either increase or decrease the mutual coupling between antennas. Investigations of fields and/or currents from one antenna usually tend to have
difficulties in predicting how geometry or material changes will impact the mutual coupling. This is natural since the electromagnetic behavior of both antennas is essential to obtaining the mutual coupling, and the method presented here solves this issue.

This method to determine coupling paths generalizes the normalized impedance density in [10] and the coupling density coefficient (CC) in [11], both of which are based on the reciprocity theorem. A novel and attractive feature in this generalization is that its amplitude can be compared across frequencies since the quantity connects to the mutual coupling as expressed by the scattering parameter \( S_{21} \). Another important extension in this work is that the proposed method handles arbitrary complex-valued self-impedances for the antennas, and also both well and poorly-matched impedances, which is directly linked to the mutual coupling as expressed by \( S_{21} \). Both features are of particular interest when investigating off-of-band coupling, where the receiving antenna often is badly tuned [1]. The work in [11] mainly focuses on identifying coupling paths, whereas the work presented in this paper focuses on identifying, validating, and affecting the coupling and its paths. In order to mitigate coupling, we propose and illustrate here a new visualization-based method that involves the comparison of the vector-field associated with mutual coupling across frequencies. Another unique aspect of the present quantity is its ability to identify regions that contribute to an increased coupling in addition to regions that reduce the coupling. See Section III for illustrations and utilization of this feature. This paper emanates from the master thesis [22].

This paper is organized as follows. In Section II, the theory of the generalized impedance density and its connection to mutual coupling is explained. In Section III, the possibility to use the generalized impedance density to predict absorber placement that reduces the mutual coupling is investigated. This is done through numerical simulations for three different cases of increasing complexity. The predictability of the generalized impedance density is compared to the Poynting vector based method. Finally, our conclusions are presented in Section IV.

II. THEORY

A. Mutual Coupling and Impedance Parameters

Mutual coupling, \( S_{21} \), between two antenna ports is defined as the received power in one port, normalized with the transmitted power in the other port. The isolation between two antennas is defined as \( 1/|S_{21}| \). The corresponding impedance parameters connecting the two antenna ports are defined as

\[
\begin{bmatrix}
    V_{11} \\
    V_{21}
\end{bmatrix} = \begin{bmatrix} Z_{11} & Z_{12} \\
Z_{21} & Z_{22} \end{bmatrix} \begin{bmatrix} I_{11} \\
I_{21}\end{bmatrix}.
\]

(1)

Here, \( Z_{11} \) and \( Z_{22} \) are the self-impedances of Antenna 1 and Antenna 2, respectively, and \( Z_{12}, Z_{21} \) are the mutual impedances between the two antennas. The influence of each impedance parameter and the voltages \( V_{ij} \) and currents \( I_{ij} \) for \( i, j = 1, 2 \) is shown in Fig. 1 for a reciprocal two-port, where \( Z_0 \) is the reference impedance. From (1) and Fig. 1 it is seen that when no load is connected to Port 2, the mutual impedance is given by

\[
Z_{21} = \frac{V_{21}}{I_{11}} \bigg|_{I_{21}=0} = \frac{V_{21}^{oc}}{I_{11}^{oc}}.
\]

(2)

and represents the coupled voltage in Port 2 due to a current in Port 1. The superscript “oc” refers to an open circuit condition.

Expanding the relation between scattering and impedance parameters [23] in a Taylor series, in the limit of small \( Z_{21} \), \( Z_{12} \) gives the first-order approximation

\[
S_{21} \approx \frac{Z_{21}}{(Z_{11} + Z_0)(Z_{22} + Z_0)},
\]

(3)

which is valid when

\[
|(Z_{11} + Z_0)(Z_{22} + Z_0)| \gg |Z_{21}Z_{12}|.
\]

(4)

Here, \( Z_0 \) is the transmission line reference impedance, assumed to be real-valued and identical in the transmitter and receiver. For the case of antennas with typically reference impedance \( Z_0 = 50 \Omega \) and a small mutual coupling, this approximation is useful. For antennas on a platform, it is often desired that the mutual impedance \( Z_{21} \) is several orders of magnitude below the reference impedance \( Z_0 \). It is seen from (3) that the mutual impedance \( Z_{21} \) is to leading order proportional to \( S_{21} \) and it is therefore of interest to study \( Z_{21} \) when solving problems with weak antenna coupling.

B. Mutual Impedance and the Reaction Theorem

The mutual impedance is proportional to the reaction [21]. Since this relation is a key result in the derivation of the coupling-paths, the main steps are outlined below. The reciprocity theorem for electromagnetic fields [19] relates two sets of sources \( J_1, M_1 \) and \( J_2, M_2 \) that produce the fields \( E_1, H_1, E_2, H_2 \) in a linear and isotropic surrounding medium by

\[
\int_{v_1} (E_2 \cdot J_1 - H_2 \cdot M_1) dv = \int_{v_2} (E_1 \cdot J_2 - H_1 \cdot M_2) dv.
\]

(5)

For the case of antennas, \( J_i, M_i \) represents Antenna \( i \) that generates the fields \( E_i, H_i \) for \( i = 1, 2 \). The volume \( v_1 \) includes Antenna 1 and \( v_2 \) includes Antenna 2. The right-hand side of (5) has been defined as the reaction

\[
\langle 2,1 \rangle = \int_{v_2} (E_1 \cdot J_2 - H_1 \cdot M_2) dv.
\]

(6)
of Port 1 on Port 2 [20], [21]. Utilizing Lorentz’ reciprocity theorem [23], see Appendix, the reaction (6) can be expressed in terms of the fields as
\[
\langle 2, 1 \rangle = \oint_{s_2} (E_2 \times H_1 - E_1 \times H_2) \cdot \hat{n} \, ds,
\]
where \( s_2 \) is a surface that encloses Antenna 2 but not Antenna 1, and \( \hat{n} \) is an outward pointing normal to \( s_2 \).

The reaction of a two-port circuit is derived from (6) with sources \( J_{2}dv = I_{22}dl_2 \), \( M_{2}dv = V_{22}dl_2 \), analogously to [24, Ch. 3], as
\[
\langle 2, 1 \rangle = V_{22}I_{21} - V_{21}I_{22}.
\]

In Port 2. The generalized reaction theorem [21] states that the reaction \( \langle 2, 1 \rangle \) is independent of the load \( Z_0 \) connected to Port 2. This follows even though both \( V_{21} \) and \( I_{21} \) on the right-hand side of (8) are dependent on \( Z_0 \). The reason is that changes in \( V_{21} \) compensates any changes in \( I_{21} \) to cancel the effects of the value of \( Z_0 \). This property implies that letting \( I_{21} = 0 \) (open circuit at Antenna 2) will not affect the reaction. As an easy case to explicitly show this with the aid of the Thevenin theorem [23], consider the two-port circuit in Fig. 1. The two-port circuit and the current source connected to Port 1 are represented as a Thevenin equivalent as seen in Fig. 2(b), with the equivalent voltage \( V_{21}^{oc} \) and impedance \( Z_{in2} \), given by (9).

From Fig. 2 it is clearly seen that
\[
I_{22} = \frac{V_{22}}{Z_{in2}}, \quad V_{21} = V_{21}^{oc} \frac{Z_0}{Z_{in2} + Z_0}, \quad I_{21} = -\frac{V_{21}^{oc}}{Z_0}.
\]
Inserting (9) and (10) into (8) and rearranging gives
\[
\langle 2, 1 \rangle = -\frac{V_{21}^{oc}}{Z_0} I_{22},
\]
as the generalized reaction theorem states [21]. The mutual impedance \( Z_{21} \) is calculated with the generalized reaction theorem by inserting (7) and (11) into (2) as
\[
Z_{21} = \frac{-\langle 2, 1 \rangle}{I_{11}^{oc}I_{22}^{oc}} = \frac{-1}{I_{11}^{oc}I_{22}^{oc}} \int_{s_2} (E_2 \times H_1 - E_1 \times H_2) \cdot \hat{n} \, ds,
\]
which is the formulation used in [10], [21], [25], [26] for mutual impedance calculations. Note that the relation (12) contains an implicit constraint that the receiving antenna must be open circuit.

To overcome this constraint, a correction factor is derived for the case of non-ideal current sources or a load connected to the receiving antenna, i.e. \( I_{21} \neq 0 \). Fig. 3(a) shows the case when Port 2 is an open circuit and Fig. 3(b) shows the case when Port 2 is connected to a load. By expressing \( I_0 \) from Fig. 3(a,b) in terms of the loads and the driving current \( I_{11} \) and the open circuit driving current \( I_{11}^{oc} \), the following identity is obtained
\[
I_{11}^{oc}(Z_{11} + Z_0) = I_{11}(Z_{in1} + Z_0).
\]
The factor \( \eta \) is defined from the identity in (13) as
\[
\eta = \frac{I_{11}}{I_{11}^{oc}} = \frac{Z_{11} + Z_0}{Z_{in1} + Z_0}, \quad Z_{in1} = Z_{11} - \frac{Z_{21}Z_{21}^{oc}}{Z_{22} + Z_0}.
\]
Inserting the factor \( \eta \) to (12) gives
\[
Z_{21} = \frac{V_{21}}{I_{11}} = \frac{\eta}{I_{11}I_{22}^{oc}} \int_{s_2} (E_1 \times H_2 - E_2 \times H_1) \cdot \hat{n} \, ds.
\]
Note that (15) does not contain any open-circuit constraints, but can be used with arbitrary port loads. This simplifies the simulation to determine the required fields in (15) and follows from the fact that the excitation of the antennas can be done sequentially with a non-ideal source and arbitrary loads for the same antenna geometry.

C. Coupling Visualization Using the Reaction Theorem

To determine \( Z_{21} \), it suffices to calculate the integral in (15) over a surface \( s_2 \) that encloses Antenna 2 but not Antenna 1, which has been reported in [10], [21], [25]–[27]. A finite truncated plane can be used with a negligible loss of precision as an integration surface [10], which corresponds to one of the red planes in Fig. 4 that separates the two antennas into two disjoint regions. The normalized impedance density
\[
\hat{\theta}_{21} = \frac{E_1 \times H_2 - E_2 \times H_1}{Z_{21}I_{11}I_{22}} \cdot \hat{n}
\]
was introduced and applied in [10] to identify coupling paths by visualizing \( \hat{\theta}_{21} \) on separation planes between the antennas.

Here, a generalized form of the normalized impedance density is derived from (15) with the correction factor \( \eta \) that handles non-ideal current sources and an arbitrary load connected to the receiving antenna. Multiplying both sides of (15) with \( Z_{21}^{oc}/|Z_{21}| \) gives
\[
|Z_{21}| = \frac{\eta}{|Z_{21}|} \frac{Z_{21}^{oc}}{I_{11}I_{22}} \cdot \hat{n}.
\]
From (21), it is clear that the generalized impedance density and hence, according to (3), under the approximation (4), the quantity \( \frac{\eta Z_{21}}{Z_{21}} \) is a real-valued, and time-invariant quantity with diagonal coupling paths across frequencies, see Section III-C. The left-hand side of (17) is real-valued, which implies that the integral of the imaginary part must be zero and the only contribution to \( |Z_{21}| \) comes from the real-part. Furthermore, the integral is strictly positive, i.e., the integrand is dominantly non-negative on the surface \( s_2 \). Omitting the imaginary part that will cancel out gives the following vector integrand

\[
\delta_{21} = \text{Re} \left\{ \eta \frac{Z_{21}}{Z_{21}} \left( \frac{E_1 \times H_2 - E_2 \times H_1}{I_{11} I_{22}} \right) \right\}. \tag{18}
\]

Note that (18) does not contain the projection on the separation plane normal \( \hat{n} \). Hence, \( \delta_{21} \) is invariant of the separation surface \( s_2 \). The vector-field of \( \delta_{21} \) in \( \mathbb{R}^3 \) provides a direct visualization of the coupling that is straightforward to calculate, since \( \eta \) and (18) contain only fields, excitation currents and impedances. Projecting \( \delta_{21} \) on an arbitrary separation plane, as in Fig. 4, results in the scalar quantity

\[
\delta_{21} = \delta_{21} \cdot \hat{n}, \tag{19}
\]

which is a real-valued, and time-invariant quantity with dimension \( \Omega / m^2 \) that varies with position.

Compared to the normalized impedance density \( \tilde{\delta}_{21} \) in (16), the quantity \( \delta_{21} \) defined in (19) handles non-ideal current sources. With the introduced changes, \( \delta_{21} \) will be referred to as a generalized impedance density. The normalization of \( \delta_{21} \) makes it possible to compare the relative importance of coupling paths across frequencies, see Section III-C.

The relation between \( \delta_{21} \) and \( Z_{21} \) can now be written as

\[
|Z_{21}| = \int_{s_2} \delta_{21} \, ds. \tag{20}
\]

According to (3), under the approximation (4), \( S_{21} \propto Z_{21} \), and hence,

\[
|S_{21}| \propto \int_{s_2} \delta_{21} \, ds. \tag{21}
\]

From (21), it is clear that the generalized impedance density\(^1\)

\(^1\)Note that the result can be extended to the strong coupling regime by replacing \( Z_{21} / |Z_{21}| \) in (18) with

\[
\frac{2Z_{21}}{(Z_{11} + Z_{0})(Z_{22} + Z_{0}) - Z_{21} Z_{12}} |S_{21}| \tag{22}
\]

\( \delta_{21} \) describes the spatial distribution of \( |S_{21}| \) over any separation surface \( s_2 \).

A 3D representation of the spatial variation of the mutual coupling contributions is constructed by calculating \( \delta_{21} \) on several of the separation planes, from which it is possible to visualize the coupling on any plane, see e.g. the green plane \( s_0 \) in Fig. 4. It is important to remember that integrating (20) over a visualization plane \( s_0 \) will not yield \( Z_{21} \). Nevertheless, it is useful for determining the distribution of how different regions in each vertical plane contribute to the mutual coupling and to provide a visualization of this information.

Note that \( \delta_{21} \) depends on the normal \( \hat{n} \), associated with a separation surface \( s_2 \). Thus, in using \( \delta_{21} \), it is important to state which sequence of separating surfaces is used to determine the coupling paths. Any such sequence works since the integration over each \( s_2 \) gives \( |Z_{21}| \). In the investigated cases in Section III, the separation planes (red) and visualization planes (green) similar to Fig. 4 are used.

**D. The Poynting Vector Based Method**

One method to generate coupling paths is to draw streamlines following the vector field described by the time-averaged Poynting vector \( |\mathbf{S}|, |\mathbf{H}| \). When Antenna 1 radiates, the power-flux goes from Antenna 1 to Antenna 2, and the streamline is thus the trace of the vector-field that connects back the power-flux that hits Antenna 2. A Poynting vector based coupling path can be determined from the relations

\[
\mathbf{d}r_s \times \mathbf{S}_{\text{avg}} = 0, \quad \mathbf{S}_{\text{avg}} = \frac{1}{2} \text{Re} \{ \mathbf{E} \times \mathbf{H}^* \} \tag{23}
\]

starting from the location of the receiving antenna port. Fig. 5 shows the notation and illustrates a streamline from Antenna 2 when Antenna 1 transmits.

**III. NUMERICAL SIMULATIONS**

The goal of this section is to evaluate the ability of the generalized impedance density to predict coupling paths and absorber positions that decrease the mutual coupling. Results are compared to the Poynting vector based method.

**A. Perforated Screen**

The first configuration on which the generalized impedance density is tested consists of two dipole antennas separated resulting in that the proportionality in (21) becomes an equality. However, all considered examples in this work are in the weak coupling regime, and the results therefore use (18).
by a perfect electrical conductor (PEC) screen as shown in Fig. 6(a). The screen has two slots of dimension $\lambda/4$ and $\lambda/2$, respectively, both with height $\lambda/10$, as shown in Fig. 6(b). The slots are separated by a distance $\lambda$.

The widths and orientations of the slots were chosen such that the larger slot has a larger contribution to the mutual coupling than the smaller one. The frequency considered is 300 MHz. The dipoles are identical with length $448$ mm, radius $3$ mm, resulting in the self-impedances $Z_{11} = Z_{22} = 65.7 - j12.1\,\Omega$. The reference impedance is set to $Z_0 = 50\,\Omega$. The mutual impedance $Z_{21} = 0.67 - j0.92\,\Omega$, which satisfies the weak-coupling relation in (3).

This case is used for an initial comparison of the two visualization methods described in Section II. A question here is in which way one can compare the correctness of the predicted coupling paths or regions that contribute to a larger coupling. The approach to answering this question is to use a small absorber as a tool to determine the impact of a spatial region’s local contribution to the mutual coupling. The absorber used here is a cube with side $0.1\lambda$ and material parameters $\varepsilon_r = 1 - j2$, $\mu_r = 1 - j2$. This electrically small absorber provides a perturbation of the initial setup to directly indicate the regional sensitivity on the mutual coupling.

The coupling is visualized in the green $xy$-plane seen in Fig. 6(a). Fig. 7 depicts the generalized impedance density $\delta_{21}$ in the background calculated with (19). The surface normal of the $yz$-separation planes is $\hat{n} = -\hat{x}$. On the $xy$-visualization plane, $\delta_{21}$ has a positive (red) area that connects the two antennas through the larger slot and a negative (blue) region that connects the antennas through the smaller slot.

Poynting vector streamlines, based on (23), are also visualized in Fig. 7 as purple and green lines. The purple (green) lines connecting the two antennas through the larger slot are 20 streamlines when Antenna 1 (Antenna 2) transmits. The starting points of the streamlines are uniformly distributed on a circle located in the green $xy$-plane depicted in Fig. 6(a). The circle has a radius of $\lambda/100$ and is centered at the port location of the receiving antenna. Fig. 7 shows that the streamlines starting from Antenna 1 do not follow the same path as the streamlines starting from Antenna 2. This behavior is expected [18].

Now, all the tools required to evaluate the accuracy of the predicted regions with dominant contributions to the coupling are in place. The mutual coupling of the original problem is perturbed by moving the above-described absorber along the dashed line $x = 0.3\lambda$, $z = 0$ shown in Figs. 6 and 7. The orange curve with circles in Fig. 8 shows the mutual coupling $S_{21}(y)$ as a function of the position $y$ of the absorber. The gray curve is $\delta_{21}$ at $x = 0.3\lambda$, $z = 0$. The dashed horizontal line is $|S_{21}|$ for the unperturbed case. The purple and green vertical bars indicate where the streamlines in Fig. 7 pass.

The first thing to notice in Fig. 8 is the agreement between the generalized impedance density and the mutual coupling. The black dashed vertical line highlights the correlation between the positive peak of the generalized impedance density (indicating strong coupling) and the maximal reduction region of the mutual coupling. An absorber placed in this region will have the best effect on reducing the mutual coupling.

Furthermore, Fig. 8 shows that when the cube is located in the most negative generalized impedance density, the mutual coupling...
coupling is increased, as highlighted with the black vertical dash-dotted line.

Finally, placing the cube on the Poynting vector streamlines, i.e., the purple and green vertical lines in Fig. 8, only has a minor impact on the mutual coupling in this configuration.

It has thus been demonstrated that it is possible to predict regions with strong coupling. The method used shows that it is possible to predict absorber locations that will either decrease or increase the mutual coupling by inspecting \( \delta_{21} \). Visualizing \( \delta_{21} \) on a plane that consists of several separation planes, as proposed in Section II, provides insight into how the coupling is distributed between the antennas. By using \( \delta_{21} \), as seen from Fig. 8, the contribution through the smaller slot partially cancels the coupling through the larger slot at the investigated frequency.

\[ \delta_{21} \]

\[ \frac{S_{\text{avg}}}{S_{\text{ref}}} \]

**B. Cylinder model**

To investigate how the presence of a platform impacts the coupling visualization, a PEC cylinder is used as a platform with two monopole antennas mounted on it, one antenna on the top and one at the bottom, as shown in Fig. 9. The cylinder is \( 6\lambda \) long and has a \( 4\lambda \) circumference at 18 GHz. Perfectly matched layer (PML) boundary conditions are applied to the ends of the cylinder in the \( x \)-direction to reduce edge effects. The two monopoles are identical with length 3.6 mm, and radius 0.05 mm. Both antennas are located equally far from the edges of the cylinder. Antenna 2 is rotated an angle \( \alpha = 11.25^\circ \) along the cylinder axis from the bottom to introduce a length difference between the power flowing on the left- and right-hand side of the cylinder. The self-impedances \( Z_{11} = Z_{22} = 37.2 + j1.5 \Omega \) and the reference impedance \( Z_0 = 50 \Omega \) with a mutual impedance \( Z_{21} = 0.35 - j0.37 \Omega \), which satisfy the weak-coupling requirement in (3).

The coupling is visualized both with the generalized impedance density and the Poynting vector based method. The case illustrated here is for the \( yz \)-visualization plane that contains both the antennas, where Fig. 10(a) depicts the generalized impedance density \( \delta_{21} \) computed with (19), using the \( xy \)-separation planes with \( \hat{n} = \hat{z} \). Fig. 10(b) depicts the magnitude of \( S_{\text{avg}} \) computed with (23) when Antenna 2 at the bottom transmits. The green lines in Fig. 10(b) are 40 streamlines computed with (23) that start from the port location of Antenna 1. The starting points are uniformly distributed over two lines parallel to the port of Antenna 1, i.e. along the \( z \)-axis. The lines are located at a distance \( \lambda/20 \) from the port on the left- and right-hand side respectively.

Fig. 10(a) shows that the right- (left-) hand side of the cylinder has a positive (negative) contribution. From the conclusions in Section III-A and the relation (3), it is therefore expected that placing an absorber on the left (right) hand side would increase (decrease) the mutual coupling. The streamlines of the Poynting vector are flowing on both sides of the cylinder as seen in Fig. 10(b). The inset figure shows that 9 of the 40 streamlines are flowing on the left-hand side. Thus, the streamlines indicate a high coupling between the antennas on both sides of the cylinder, whereas the generalized impedance density indicates that only the right-hand side contributes to an increased coupling. To validate these two different predictions, we add an absorber to the left- or right-hand side, respectively, to one side of the cylinder at the time. Fig. 11(a) depicts the absorbers on the left- and right-hand side located equally far from the cylinder edges with length \( L = \lambda \), width \( w = 0.5\lambda \), and thickness \( t = 1 \text{ mm} \). The absorber used in this case is a Laird Q-Zorb 2338 surface wave absorbing material (SWAM) [28] that is intended for attenuating surface waves with material

![Fig. 9. A PEC cylinder with two monopoles located in the \( yz \)-plane and equally far from the edges of the cylinder. Antenna 2 is rotated \( \alpha = 11.25^\circ \) from the bottom.](image)

![Fig. 10. Visualization in the \( yz \)-plane that contains both the antennas, at 18 GHz. (a) Generalized impedance density \( \delta_{21} \), (b) Magnitude of \( S_{\text{avg}} \) in the background with green streamlines when Antenna 2 transmits.](image)
parameters $\varepsilon_r = 14.4 - j0.26$, $\mu_r = 1.44 - j1.01$ at 18 GHz. The complete frequency behavior of this absorber can be found in [22] and the material library of CST Microwave Studio. The weak-coupling relation in (3) is satisfied.

Fig. 11(b) shows that placing the absorber on the right-hand side decreases the coupling, and that placing the absorber on the left-hand side increases the coupling at 18 GHz, as predicted by the generalized impedance density, $\delta_{21}$ in Fig. 10(a). The effects on the antenna self-impedances $Z_{11}$, $Z_{22}$ are below 0.8 Ω (2.2%) in magnitude and 1.2° in phase when introducing the absorbers. The main contribution to the change in $|S_{21}|$ hence comes from the reduction in $Z_{21}$. Thus, similar to the previous case in Section III-A, we see that the generalized impedance density can be used to locate regions which contribute to high coupling. In the present case, we note that the streamlines, using (23) could only partly predict regions of strong coupling.

In Fig. 12(a), $\delta_{21}$ is visualized with an absorber placed on the left-hand side, as highlighted with the black dashed line. The region close to the surface on the left-hand side of the cylinder has turned from a negative generalized impedance density region, depicted as blue in Fig. 10(a), to a positive region, as depicted with red and white. The contribution from this area is now positive, leading to an increased mutual coupling. Fig. 12(b) illustrates $\delta_{21}$ when the absorber is placed on the right-hand side. The large red area close to the surface on the right-hand side in Fig. 10(a) is reduced, resulting in a decreased coupling.

It has been demonstrated that the generalized impedance density $\delta_{21}$ is a useful tool when predicting absorber positions that reduce the mutual coupling as well as increase the coupling. Visualizing $\delta_{21}$ is also a useful for larger perturbations of the platform, in this case with larger absorbers. We conclude that the visualization method can be used to obtain accurate information for reducing the mutual coupling, contrary to what is claimed in [11].

C. Integrated Antennas

The third case is a more complex configuration consisting of a ground plane with two cavities for integrating antennas, as seen in Fig. 13. The configuration could represent a section of an integrated mast of a naval ship, as well as a wingtip on an aircraft or a part of its fuselage.

The possibility of using the generalized impedance density to predict where to place absorbers to reduce the mutual coupling between two wideband antennas integrated in the platform is investigated here. A 6×8 element body-of-revolution (BOR) array antenna [29], [30], depicted in Fig. 13(b), is used and referred to as Antenna 1. The other antenna is a cavity-backed spiral antenna that is shown in Fig. 13(c) and is referred to as Antenna 2. The center of the spiral, depicted in the orange inset in Fig. 13(c), is designed to intrinsically generate left-hand circular polarization. Both antennas are placed in their respective cavities seen in Fig. 13(a) with the top of the antennas located 1.2 mm below the surface of the platform.

The mutual coupling between a center element in the array, which is highlighted red in Fig. 13(a), and the spiral antenna is studied. The other array elements are terminated with a matched load. Potential absorber locations have been restricted to the flat section between the two cavities. The considered frequency range is 10–14 GHz. For the described configuration, the condition (3) is satisfied.
Predicting good absorber positions to reduce the mutual coupling between two antennas is a common and challenging task. The initial mutual coupling between the antennas is shown in Fig. 14(a) as the black dashed curve, with the highest coupling at 10.5 GHz and two other peaks at 11.25 GHz, and 14 GHz. One can approach this problem by investigating the surface currents. However, studying the surface currents of the spiral antenna on the flat surface of the platform at 10.5 GHz shows that the entire region contains strong surface currents, see Fig. 15. Similar behavior is observed in other frequencies in the band. From this type of information, it is difficult to determine if there is a preferred region to place absorbers on to reduce the mutual coupling.

In contrast, if the generalized impedance density $\delta_{21}$ is visualized, clear regions for absorber placement emerge. Fig. 16 shows $\delta_{21}$ for $z = 1$ mm above the platform, at the frequencies of the aforementioned peaks in the coupling. The black dashed lines highlight the location of the cavities and the antennas. The surface normal is $\hat{n} = -\hat{x}$, corresponding to a sequence of $yz$-separation planes that are used to construct the $xy$-visualization plane seen in Fig. 16. This choice of separation-planes is natural since the platform is modeled as a good conductor.

Note that for 10.5 GHz, the dominant positive (red) region is at $y = 40$ mm and below $y = 8$ mm. The same regions in 11.25 GHz are also dominantly positive. A close look at 14.0 GHz shows that a smaller lower region 8 to -38 mm has mainly positive generalized impedance density with some weakly negative $\delta_{21}$. For $y = 40$ mm, there is both a weak negative and a strong positive region.

Based on the visualization results in Fig. 16, absorbers are placed according to Fig. 14(b). The upper absorber has a width $w_1 = 8$ mm and is centered at $x = 0$ mm, $y = 40$ mm. The lower absorber has a width $w_2 = 46$ mm, and is centered at $x = 0$ mm, $y = -15$ mm. Both absorbers have a thickness of 1 mm and length $L_c = 49$ mm. The type of absorber used is the same SW AM [28], which is used in Section III-B. The red solid curve in Fig. 14(a) represents a simulation of the transmission coefficient with the absorbers placed according to Fig. 14(b). The mutual coupling is reduced in the frequency range 10.2–14 GHz when the absorbers are placed based on the visualization of the generalized impedance density in Fig. 16.

By comparing the intensity (color) of the generalized impedance density in the regions $y = 40$ mm and $y = [-38, 8]$ mm in Fig. 16(ab), it is clear that $\delta_{21}$ is much more positive in these regions at 10.5 GHz than at 11.25 GHz. From this fact, it is straightforward to understand the 9 dB versus 5 dB reduction differences between 10.5 GHz and 11.25 GHz.

Fig. 14. (a) Transmission coefficient between the array element and the spiral without absorbers (black dashed), with two absorbers according to Fig. 14(b) (red solid), and with one absorber according to Fig. 14(c) (blue dash-dotted). (b) Absorbers placed based on $\delta_{21}$. (c) An absorber covering the surface between the cavities.

Fig. 15. Magnitude of the instantaneous surface currents on the platform generated by the spiral antenna at 10.5 GHz.

Fig. 16. Visualization of $\delta_{21}$ in the $xy$-plane for $z = 1$ mm above the platform at (a) 10.5 GHz, (b) 11.25 GHz, and (c) 14.0 GHz.
The more standard approach of covering the whole surface between the two cavities with an absorber is also investigated. An absorber with length $L_c = 49$ mm, width $w_3 = 88$ mm, and thickness $1$ mm is placed between the cavities as shown in Fig. 14(c). The blue dash-dotted curve in Fig. 14(a) is the simulated mutual coupling with the absorber in Fig. 14(c) present. The mutual coupling is $5$ dB higher at $10.5$ GHz when the whole surface is covered by an absorber, compared to placing absorbers according to Fig. 14(b), which is based on the generalized impedance density. Covering the whole surface between the cavities with an absorber reduces the negative generalized impedance density region centered at $y = 20$ mm in Fig. 16(a). This explains why the mutual coupling is higher compared to placing absorbers according to Fig. 14(b), as predicted by the generalized impedance density. At $11.25$ GHz and $14$ GHz, it is seen in Figs. 16(b) and (c) that an absorber covering the whole surface will reduce both red and blue regions, which describes why the difference between the two absorber configurations are small in terms of mutual coupling at $11.25$ GHz and $14$ GHz.

The absorber placement described above successfully reduces the coupling at the investigated frequencies. However, it results in an increased mutual coupling at $10$ GHz, as shown in Fig. 14(a). This can be explained by inspecting the generalized impedance density at $10$ GHz, see Fig. 17. The regions where the absorbers are placed are mainly negative (blue) at this frequency, leading to an increased mutual coupling.

Thus, the generalized impedance density has been used to systematically reduce the mutual coupling between two wideband antennas placed on a common platform with a satisfying result. By visualizing the generalized impedance density $\delta_{21}$ at several frequencies, it is possible to predict absorber positions that reduce the mutual coupling over a band of frequencies.

IV. Conclusion

In this paper, we generalize and evaluate a visualization method for identifying regions around a platform with large contributions to the mutual coupling between installed antennas. The presented real-valued time-invariant quantity $\delta_{21}$ (19) is named the generalized impedance density. The direct connection between the reciprocity theorem and the mutual coupling is revisited, including a derivation of the generalized impedance density.

The generalized impedance density captures the interaction between fields and represents all the mechanisms that contribute to the coupling between antennas. It can be used to visualize how the mutual coupling is spatially distributed between the antennas.

The generalized impedance density has here been tested in numerical simulations to calculate coupling paths and predict absorber positions in three different cases. With the generalized impedance density, it is possible to separate the positive and negative coupling paths. This novel feature has been used to systematically reduce the mutual coupling between two antennas. The visualization method presented in this paper predicts good absorber locations, as expected from the derivation. Indeed, placing a small absorber at the strongest predicted generalized impedance density position gave a larger reduction of the mutual coupling than when the same absorber was placed at the position predicted by the Poynting vector based method. It is also clearly demonstrated that it is possible to achieve a lower mutual coupling with less absorber material when placing absorbers based on predictions by the generalized impedance density, compared to covering all available space with absorbers.

From the here presented examples, it is clear that the generalized impedance density provides important coupling information, which is robust to the platform shape. The proposed quantity offers the possibility to compare coupling paths at different frequencies. This useful feature makes it possible to reduce the coupling over a band of frequencies. It is furthermore easy to implement as a post-processing step in full-wave electromagnetic software.

The ability to distinguish between regions where the placement of absorbers would reduce or increase the mutual coupling, together with the straightforward interpretation of the results, makes the generalized impedance density a useful tool to visualize and systematically reduce the mutual coupling.

APPENDIX

For a set of sources $J_2, M_2$ in the volume $\nu_2$, the associated fields $E_2, H_2$ satisfy Maxwell’s equations

$$\nabla \times E_2 = -j \omega \mu H_2 - M_2, \quad \nabla \times H_2 = j \varepsilon E_2 + J_2. \quad (24)$$

The fields $E_1, H_1$ are source-free in $\nu_2$, thus

$$\nabla \times E_1 = -j \omega \mu H_1, \quad \nabla \times H_1 = j \varepsilon E_1. \quad (25)$$

Recall the divergence of a cross-product:

$$\nabla \cdot (E \times H) = \mathbf{H} \cdot \nabla \times E - E \cdot \nabla \times H. \quad (26)$$

Using (24), (25), and the divergence identity (26) gives

$$\nabla \cdot (E_2 \times H_1 - E_1 \times H_2) = E_1 \cdot J_2 - H_1 \cdot M_2. \quad (27)$$

from where (7) follows by the divergence theorem.
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