Multiport Multiband Decoupling Optimization for Miniature Antennas

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Abstract. *Multiband multiport antennas are increasingly used for wireless communications and sensing miniature devices. The equations governing the multiport multiband antennas are analyzed in this paper with the objective of drawing the design guidelines for low coupling small antennas. Those guidelines have been applied in the design and optimization of a two-port dual band small antenna of size around $\lambda_0/13$ x $\lambda_0/13$ at the lowest frequency. Certain coupling conditions are applied to the port loads achieving a coupling reduction of 8 dB when having a simple two-element real load. A reduction of 27 dB can be obtained when having ideal loads composed by a higher number of elements. The antenna geometry is shown together with coupling minimization results.*

Keywords

Multiport multiband antennas, miniature antennas, decoupling, S-parameters, efficiency.

1. Introduction

The evolution towards minimization due to increase of mobility of communication devices and their pervasive incorporation in the environment leads to high degree of integration of the antenna with the rest of the circuitry. New applications require multiband operation, implying multiband multiport antenna designs. All this imposes specific restrictions to the antenna: it needs to be thin enough to be compatible with the multilayer integration technologies; its coupling with the circuitry must be low and the total size of the antenna must be kept as small as possible. In addition to the worsening of the radiation properties (radiation efficiency and bandwidth) when reducing the size of an antenna, as it has been formulated in different publications, such [1] and [2], one of the backbones of the small multiport antennas is the strong port coupling. The port coupling is directly related to the total efficiency of the multiport antennas, being very unfavorable for certain applications. There have been a lot of studies focused on the coupling minimization techniques for single band multiantennas composed by radiating elements working at the same frequency, which are positioned at very short distance with symmetry in any of its axis [3], [4]. However, those techniques are quite complicated to be applied when dealing with very small antennas with no symmetry between the radiating structures or when the different structures share the same radiating metallic layer. Taking the advantage of the multiband multiport operation, we can think of other techniques to minimize coupling by imposing certain operation conditions to the antenna ports not only at its proper resonating frequency (at which the condition would be the perfect matching), but also at the rest of the operation frequencies.

Given a multiport (ports i,j) structure working at its corresponding operation frequencies, f_i, f_j, the coupling between ports will depend on the antenna S-parameters and the reflection coefficients at the ports. At a certain port i, matching requirements have to be fulfilled at its operating frequency, f_i, and some coupling condition could be imposed to the port loads to achieve those reflection coefficient needed for an optimum coupling.

The aim of this work is the deep study of those equations governing the multiport multiband small antennas, in order to have designs criteria for optimum antennas.

2. Analytic Analysis

Considering the multiband multiport antenna as a multiport network (Fig. 1), each port i, will be required to radiate at a certain frequency f_i. In multiport multiband antennas, the total efficiency at f_i, η^{fi}_{total}, can be divided in three main contributions: reflections (due to mismatching, η^{fi}_{ref}, ohmic losses, η^{fi}_{l}, and coupling to other ports, η^{fi}_{c}:

$$\eta^{fi}_{total} = \frac{P_{rad}}{P_{av}} = \eta^{fi}_{ref} . \eta^{fi}_{l} . \eta^{fi}_{c} . \quad (1)$$

In this way, one of the conditions for a high efficiency antenna, is a good matching at port i at its corresponding operating frequency f_i (approaching η^{i}_{ref} to 1). In the same way, the power which is transferred from port i to port j must be minimized, achieving high η^{i}_{c}. Therefore, the objective is to find a condition at the ports i, j to minimize the coupling between them.

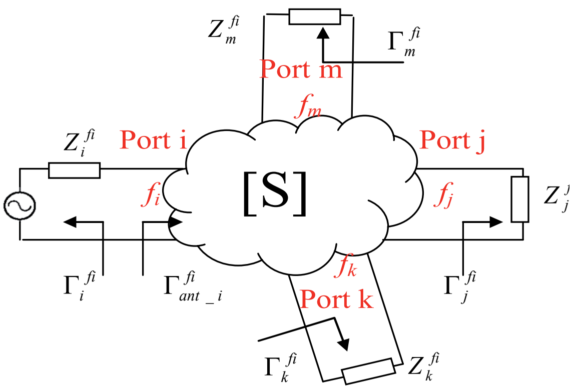


Fig. 1. Multiport multiband antenna model.

The relation between the power coupled from port i to port j at f_i (P^{fi}_{ji}) with respect to the input power at port i $P^{fi}_{in\_i}$ can be expressed as in (2), taken from [5] and [6].

$$C^{fi}_{ji} = \frac{P^{fi}_{ji}}{P^{fi}_{in\_i}} = \frac{\left|S^{fi}_{ji}\right|^2}{\left|1-S^{fi}_{jj}.\Gamma^{fi}_{j}\right|^2} \cdot \frac{\left(1-\left|\Gamma^{fi}_{j}\right|^2\right)}{\left(1-\left|\Gamma^{fi}_{ant\_i}\right|^2\right)}, \forall j \neq i. \quad (2)$$

Neglecting indirect coupling between ports j and k for $k \neq i$, Γ^{fi}_{i} the reflection coefficient looking into the port i at f_i, $\Gamma^{fi}_{ant\,i}$ the input reflection coefficient looking towards antenna port i at f_i and Γ^{fi}_{j} the reflection coefficient of the load at port j at f_i.

When injecting power at port i, at its working frequency, f_i, coupling and efficiency are related according to the expression:

$$\eta^{fi}_{c} = 1 - \sum_{\forall j \neq i} C^{fi}_{ji}. \quad (3)$$

The trivial solution to minimize the coupled power is to make $|S^{fi}_{ji}|$ close to 0, which will be extremely difficult to achieve, in case of very close radiating elements, or small multiband antennas, where the resonant structures share the same radiating metallic element.

Assuming matching conditions at port i at working frequency f_i, $Z^{fi}_{i} = Z_{0i}$ (Z^{fi}_{i} is the input impedance at port i at f_i, shown in Fig. 1, and Z_{0i} is the characteristic impedance), then the expression for the coupling is simplified to (4):

$$C^{fi}_{ji} = \frac{P^{fi}_{ji}}{P^{fi}_{in\_i}} = \frac{\left|S^{fi}_{ji}\right|^2}{\left|1-S^{fi}_{jj}.\Gamma^{fi}_{j}\right|^2} \cdot \frac{\left(1-\left|\Gamma^{fi}_{j}\right|^2\right)}{\left(1-\left|\Gamma^{fi}_{ant\_i}\right|^2\right)}. \quad (4)$$

The maximum power transference condition has to be fulfilled at the port i: $\Gamma^{fi}_{i} = (\Gamma^{fi}_{ant\,i})^*$.

The condition to minimize expression (4) is expressed as follows:

$$S^{fi}_{jj}.\Gamma^{fi}_{j} = -1 \quad (5)$$

Therefore, when the product $S^{fi}_{jj}\,\Gamma^{fi}_{j}$ equals to -1, the transferred power from port i to port j at f_i is minimized.

There are two design options: either the antenna S parameters [S] or the load RF circuitry Γ_j have to be adjusted to fulfil the coupling minimization condition:

- If we choose to modify the antenna, a first design condition will be imposed to, the S^{fi}_{jj} parameter phase, θ_{Sjj} to fulfil the expression (6).

  $$\theta_{S^{fi}_{jj}} + \theta_{\Gamma^{fi}_{j}} = 180^{\circ} \quad (6)$$

  being $\theta_{\Gamma^{fi}_{j}}$ the phase of Γ^{fi}_{j}.

Once condition (6) is assured, the maximum reduction will be obtained when the S^{fi}_{jj} amplitude, $|S^{fi}_{jj}|$ equals to 1. However, this would imply having reactive load (radiating element) at port 2. Having $|S^{fi}_{jj}| = 1$ would lead again to the trivial solution $|S^{fi}_{ji}| = 0$, since there will not be dissipated power in the load at port j. The condition for $|S^{fi}_{jj}|$ to minimize the coupling in (4) will be bounded by the power balance expression (7) in case of lossy networks:

$$\left|S^{fi}_{ji}\right|^2 + \left|S^{fi}_{jj}\right|^2 \leq 1 \quad (7)$$

meaning that, considering lossless antennas, we can assume $|S^{fi}_{ji}|^2 + |S^{fi}_{jj}|^2 + P^{fi}_{radj} = 1$. If we achieve a high $|S^{fi}_{jj}|$, this will imply either a reduction of $|S^{fi}_{ji}|$ or a reduction of the radiated power at f_i at port j (P^{fi}_{radj}), which is also advantageous, since port j is required to work only at f_j.

The S-parameters depend on the antenna geometry, so, the way of modifying S^{fi}_{jj} is to play with the geometry or the feeding points, which will have also certain effects on the rest of the parameters (S_{ii} amplitude and phase and S_{ji} phase at all the frequency bands) that are difficult to analyze and may not be advantageous for other antenna characteristics such us radiation efficiency or port matching.

- However, if we choose to impose the coupling minimization condition to the load RF circuitry, Γ^{fi}_{j}, the problem is reduced to the design of a port load fulfilling certain conditions at f_j: According to (5), the reflection coefficient is tuned to approximate -1:

  $$\theta_{\Gamma^{fi}_{j}} = 180^{\circ} - \theta_{Sjj} \text{ and } \left|\Gamma^{fi}_{j}\right| \approx 1 \quad (8)$$

These coupling minimization criteria must be applied keeping $\Gamma_{ant\,i}$ as low as possible.

a) In sum, each port (i.e. port i) should fulfill the design criteria: at its operating frequency f_i has to present the proper impedance to maximize matching conditions $\Gamma^{fi}_{i} = (\Gamma^{fi}_{ant\,i})^*$,

b) at the rest of the operating frequency bands (f_j), the impedance at port i has to fulfill $\Gamma^{fj}_{i} = -1$, ($\forall j \neq i$).

In order to manage and meet the different conditions at different frequency bands for the ports reflection coefficients, $\Gamma_i^{fi,fj}$, optimization methods based on genetic algorithms may be used [7]. Eventually, genetic algorithms can be also applied in the antenna design process, if we chose to impose the coupling conditions to the S parameters (S^{fi}_{jj}).

3. Miniature Two-Port Dual Band Antenna

The previous analysis has been applied to a dual band dual port miniature antenna. Its design has been done following the minimization coupling requirements.

3.1 Minimum Coupling Two-Port Dual Band Antenna Application

The miniature antenna is thought to work for self-powered miniature wireless sensor nodes, which means that the node must generate its own energy to keep functioning without any external battery. In other words, the node must be able to harvest the sufficient energy from the ambient sources to feed all the subsystems of which is composed. From all the possible energy sources, the ambient RF power is a convenient option in case of indoor short range wireless applications. In this way, when using RF ambient energy to power the node circuitry, two radiating structures are needed: one of the structures will be in charge of collecting energy from the external source, at a certain frequency band f_1, and the other structure is used for the data communication at frequency f_2. In this kind of applications, the energy received to be harvested is very low, and therefore, the scavenging circuitry and the antenna should be designed for the maximum efficiency. In this context, a high coupling between both ports would mean that a significant amount of the received power is not used for the powering of the node system but it is lost through the communication port.

3.2 Design

The miniature antenna system, which should fit in 25 x 25 mm, since it is the space occupied by the sensor node die, is required to collect energy at f_1= 950 MHz and transmit and receive data at f_2= 2.45 GHz. The 950MHz antenna port is connected to a 50Ω single-ended port rectifier, whereas the 2.45GHz resonating structure is directly fed by a 50Ω balanced output.

Having chosen a patch type antenna as radiating structure at 950 MHz, the need for using downsizing techniques is obvious, since the resonant length at 950 MHz is 157.9 mm, which equals half wavelength in free space ($\lambda_1/2$). The proposed antenna has transversal dimensions (*l, w*) of 25 x 25 mm ($\approx\lambda_1/13$ x $\lambda_1/13$, at 950 MHz) and 1.5 mm thickness, *t*, ($\approx\lambda_1/200$ at 950 MHz). The use of low loss substrate (ε_r = 3.55, tg(δ) = 0.0027), a shorting edge on one side of the patch antenna (w_{sc}), which lowers the dimension to $\lambda/4$, and the lateral folding of the structure, creating a slot of length l_f, leads to sufficient size reduction of the antenna for on-chip integration with an acceptable cost in efficiency and bandwidth. The active layer of the patch is fed by a conductive via which, passing through the ground plane, is connected to the rectifier that is placed on the circuitry layers, which are located bellow. In Fig. 2 a complete 3D view of the whole node architecture is shown. In Fig. 3 a 3D view of the 950 MHz resonant structure is drawn.

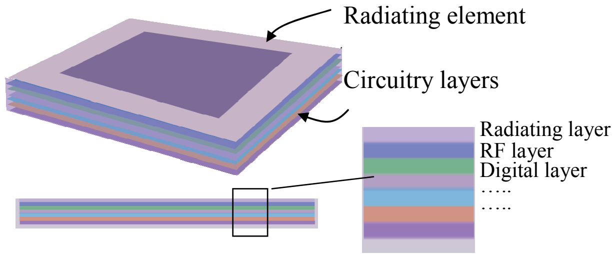


Fig. 2. Node integrated architecture.

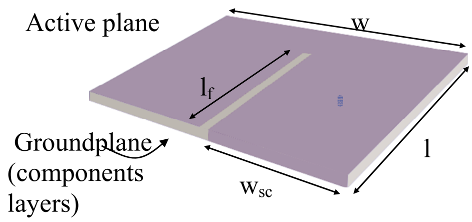


Fig. 3. 950 MHz antenna design.

The resonant structure at 2.45 GHz is to be connected to a balanced port, which leads to think of a structure with a balanced feeding, such as dipoles or slots, which are suitable for direct integration with differential transceivers. Taking advantage of the active plane of the 950MHz patch antenna, the most convenient solution seems to be to etch a slot on it, of the 2.45GHz resonant length. This resonant length in free space is 61.2 mm ($\lambda_2/2$ at 2.45 GHz), which is too large to fit in 25 x 25 mm, consequently downsizing techniques need to be also applied in this case. The symmetry presented by the field distribution of a $\lambda/2$ a slot antenna, of certain length l_{slot} and width w_{slot} allows to place a PMC in the symmetry axis, which is right in the middle, at $\lambda/4$. Therefore, an open circuited $\lambda/4$ has the same performance as the $\lambda/2$ slot. Considering the dielectric constant of the patch substrate, a sufficient length reduction is achieved ($\lambda_2/4$ is around 17 mm at 2.45 GHz) fitting on 25 x 25 mm. The location of the 2.45GHz slot on the top conductive plate follows isolation optimization criteria, being placed on a foreseeable low current zone of the 950MHz patch (surface density current at 950 MHz shown in Fig. 4). Fig. 5 presents the complete structure for the two-port dual band miniature antenna of thickness *t*, including the feeding positions at 950 MHz (being *d* the distance to the shortcircuit) and at 2.45 GHz (being d_{slot} the distance to the open edge of the resonating $\lambda_2/4$ slot)

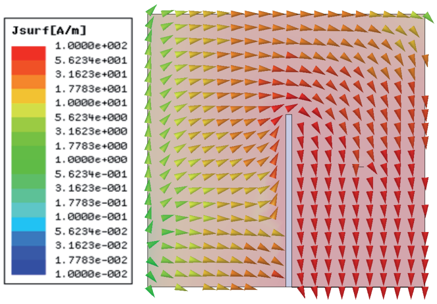
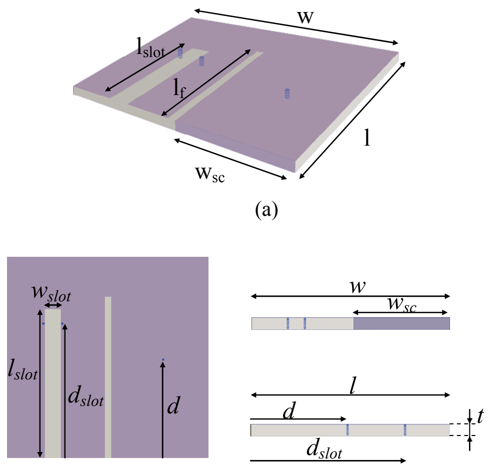


Fig. 4. Surface density current for the one-port patch antenna.



(a)

(b)

Fig. 5. Two-port dual band small antenna: (a) 3D view, (b) top and lateral views.

3.3 Coupling Minimization

In order to adjust the physical parameters for the proper performance of the antenna, a series of simulations have been done using the simulation tool Ansoft HFSS10.1.2, which is based on the finite element method.

First, the S parameters (Fig. 6 presents the two-port network) are calculated and, as it was foreseeable and can be seen in Fig. 7, the S_{21} parameter is too high to achieve low coupling considering that both ports are load with 50 Ω at both frequencies, then, in expression (2) $|\Gamma_i^{fi}|=|\Gamma_j^{fi}|=0$.

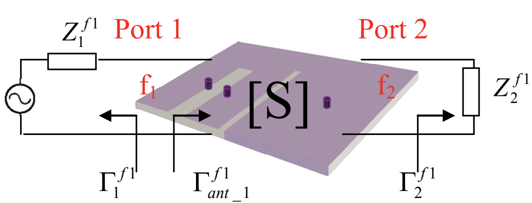
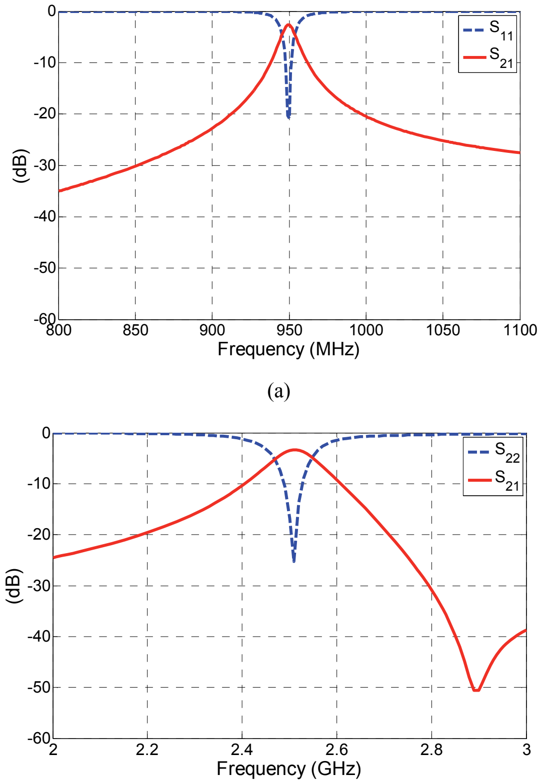


Fig. 6. Two-port dual band antenna model.

However, having already the antenna S-parameters, we can play with the load reflection coefficient in order to achieve low coupling. The efficiency at the scavenging port (port 1) will be maximized, reducing the power which is coupled to the communication port (port 2) at f_1=950 MHz.



(a)

(b)

Fig. 7. Dual band two-port small antenna. S-parameters at (a) 950 MHz and (b) 2.45 GHz.

According to the previous discussion, Γ_2^{f1} should fulfill the conditions defined in (8) : Γ_2^{f1}≈-1 Fig. 8 shows the variation of coupling with $\theta_{\Gamma_2^{f1}}$ for different amplitudes $|\Gamma_2^{f1}|$ at frequency f_1= 950 MHz, achieving lower coupling for $\theta_{\Gamma_2^{f1}} \approx 29°$, which is remarked with the vertical line in Fig. 8 (θ_{S22} at f_1=950 MHz is 151° approx.) and for $|\Gamma_2^{f1}|$ close to 1.

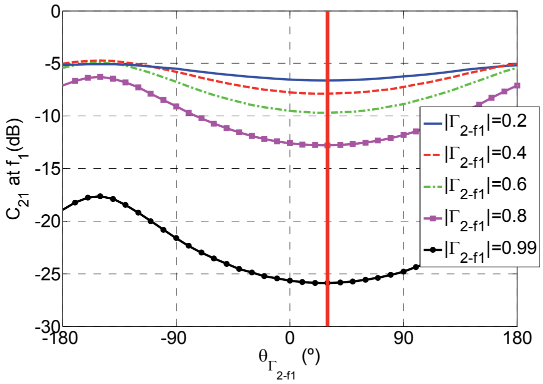


Fig. 8. Variation of coupling with $\theta_{\Gamma_2^{f1}}$ for dual band two-port small antenna at f_1=950 MHz for different values of $|\Gamma_2^{f1}|$.

Therefore, the design solution will resides in the load circuitry at port 2, requiring the above-mentioned values at f_1 and perfect matching at f_2=2.45 GHz ($\Gamma_2^{f2} \approx 0$) as it is summarized in Tab. 1.

Optimum load for minimum coupling	f_1=950 MHz		f_2=2.45 GHz	
	$\theta_{\Gamma 2}$	$\vert\Gamma_2\vert$	$\theta_{\Gamma 2}$	$\vert\Gamma_2\vert$
	29°	1	29°	1

Tab. 1. Optimum load at port 2 for minimum coupling.

According to the criteria described previously, we have to design a frequency dependent load for port 2, $Z_2^{f1,f2}$, fulfilling the conditions fixed in Tab. 1 at f_1 and f_2. There are multiple solutions to implement the optimum load, depending on the number of elements used for it. Fig. 9 shows the conditions for the amplitude and phase of the reflection coefficient of the optimum load (big dots) and two different solutions for its implementation: the one composed by 12 elements and other one composed by 2. These loads behave like filters, meaning that $Z_2^{f1,f2}$ will be composed by some extra lumped components.

The more elements the load has, the broader bandwidth is achieved.

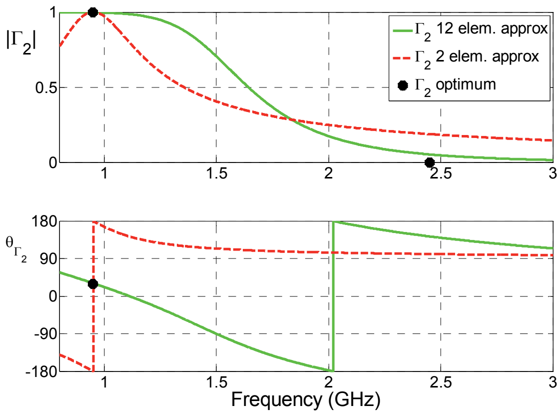


Fig. 9. Solutions for the optimum load for minimum coupling.

4. Results

In order to test our two-port dual band small antenna system design, once optimized for minimum coupling, some results are shown hereby.

Due to the straightness and ease of the implementation of a 2-elements load, this antenna load solution was fabricated and connected to the antenna prototype.

The antenna system was characterized using a measurement method based on the Modulated Scattered Technique (explained in [6]). A perturbation is created in the balanced port (port 2) modulating its impedance (Z_2), so the coupling can be extracted by solving a simple equation system. The coupling reduction achieved is around 8 dB which corresponds to an improvement of the efficiency of 13%, according to (3). Fig. 10 shows the measured and simulated coupling for the two different working frequencies and the improvement achieved at the scavenging frequency. Fig. 11 shows the measurement setup for the coupling characterization and Fig. 12 presents the two-port dual band small antenna prototype.

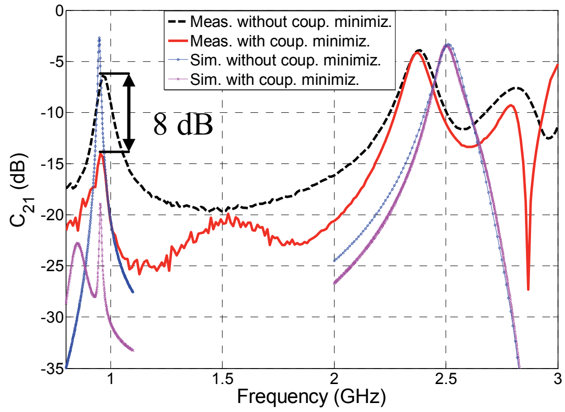


Fig. 10. Coupling improvement at f_1= 950 MHz.

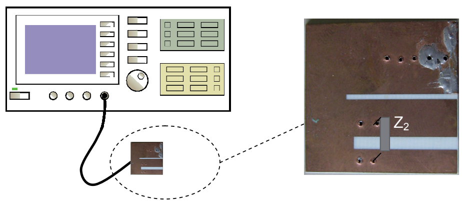
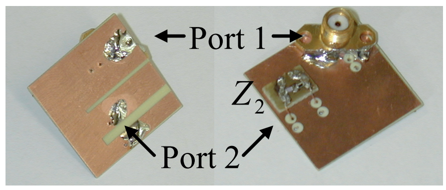


Fig. 11. Coupling characterization setup.



(a) (b)

Fig. 12. Two port dual band small antenna prototype.(a) Active layer. (b) Bottom layer with 2 element load $Z_2^{f1,f2}$ at port 2 and SMA connector at port 1.

If a higher improvement of the coupling is needed, a solution with higher number of elements could be used. As an example, a solution with 12 elements has been designed and simulated. The design of the variable load has been done with the software tool ADS (Advanced Design System) from Agilent. Simulation results, in Fig. 12, foresee a port coupling of -30dB and a relative improvement in efficiency of around 63%.

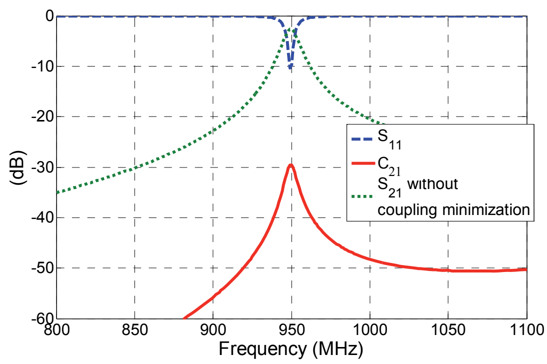


Fig. 13. Two-port dual band small antenna optimized for minimum coupling. Results for 950 MHz.

5. Conclusions

Coupling between ports in multiport multiband antennas increases significantly when the size of the antenna shrinks. However, by analyzing the equations governing the port coupling, we have found the conditions to be applied to the port loads in order to minimize the amount of coupled power from one port to the others.

The minimum coupling conditions have been applied to a two-port dual band small antenna. The improvement of the coupling depends on the complexity of the load: The higher number of elements composing the load, the broader bandwidth will have the antenna system.

We can foresee an efficiency improvement of 63% when the load is composed by 12 elements. Measurement results show a 8dB coupling reduction when having a 2 element load, which leads to an efficiency increase of 13%. Here we confront a compromise between the space we have for extra components if needed, and the improvement of the coupling.

The use of Genetic Algorithms to find the optimum impedances may improve the optimization process, reducing the number of extra components and making it more accurate.

The design criteria have been proved for a two-port dual band miniature antenna.

Further improvement could be done by reducing also the coupling from port 2 (communication data at 2.45 GHz) to port 1 (scavenging at 950 MHz) following the same procedure: a frequency dependent load would be designed for port 1 (Z_1).

Acknowledgements

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References

[1] CHU, L. J. Physical limitations of omni-directional antennas. *Journal of Applied Physics*, 1948, 19, p. 1163-1175.

[2] MCLEAN, J. S. The radiative properties of electrically-small antennas. *IEEE International Symposium on Electromagnetic Compatibility*, p. 320 – 324, Aug. 1994.

[3] SERRANO, R., CAPDEVILA, S., DIALLO, A., BLANCH, S. LUXEY, C., LE THUC, P., STARAJ, R. S., ROMEU, J., JOFRE, L. Optimal multiantenna structures for maximum capacity in wireless compact systems. In *EUCAP*, Edinburgh (UK), Nov 2007.

[4] DIALLO, A. LUXEY, C., LE THUC, P., STARAJ, R., KOSSIAVAS, G., FRANZEN, M., KILDAL, P-S., Evaluation of the performances of several four-antenna systems in a reverberation chamber. In *Proc International Workshop on Antenna Technology 2007 (iWAT07)*. Cambridge (UK), 21st-23rd March 2007, p. 166-169.

[5] POZAR, D. M. *Microwave and RF Design of Wireless Systems*. Chapter 11. John Wiley & Sons, Inc., 2001.

[6] BEATTY, R. W. Insertion loss concepts. *Proceedings of the IEEE*. June 1964, vol. 52, p. 663-671.

[7] RAHMAT-SAMII, Y., MICHIELSSEN, E. *Electromagnetic Optimization by Genetic Algorithms.* Ed. Wiley, 1999.

[8] SERRANO, R., BLANCH, S. ROMEU, J., JOFRE, L. Evaluation of two novel non-invasive measurement methods for small antennas. In *EUCAP*, Nice (France), Nov 2006.

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