Filtenna Integration Achieving Ideal Chebyshev Return Losses

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Abstract. *This paper demonstrates that it is possible to find an ideal filter response (Chebyshev, Butterworth,..) considering the antenna as the last resonator of a filter under certain circumstances related with the antenna performance and the bandwidth of the filtenna device. If these circumstances are not accomplished, we can achieve excellent performance as well, by means of an iterative process the goal of which is defined by either a filter mask or a classical filter function itself. The methodology is based on the conventional coupling matrix technique for filter design and has been validated by fabricating a microstrip prototype using hairpin resonators and a rectangular patch antenna.*

Keywords

Filtenna, filter, antenna, return losses, Chebyshev, optimal bandwidth.

1. Introduction

One of the major topics in telecommunications and electronics had been for a long time how to match the impedance of a certain load to any circuit through a matching network. In particular, a widely studied example consists on matching the impedance of two of the most important parts of a RF/MW front-end: the antenna and the filter. The most straightforward way to do this is by introducing a matching network between both devices [1]. Nevertheless, in terms of compactness and overall subsystem performance, matching the load and the circuit, being the circuit itself the matching network, has advantages. This topic has been studied by several authors; in 1964 Matthaei [2] already detailed how to adapt a RLC load by means of synthesizing a filter matching network preceding the load. More recently, many authors have faced the problem of combining both the filter and the antenna by use of different techniques and technologies. In [3] a mutual-synthesis approach, that simultaneously optimizes the filtering and radiation functions to obtain an optimal matching, is presented. Similarly to [2], but specifically for UWB communications, [4] introduces a filtenna device which is designed by use of a filter as a matching network to match the antenna. References [5] and [6] propose two different techniques to synthesize the filtenna network making use of the coupling matrix and considering the antenna as the last resonating element of the filter. In [7] that last consideration is also assumed but they go further by proposing an automatic method to carry out a mutual-synthesis based on [3]. In addition, several other authors have proposed solutions that combine the antenna with active devices to provide the desired overall circuit response. Other authors have improved the miniaturization of radio front-ends by use of active integrated antenna (AIA) theories [8], [9].

Besides the techniques and concepts introduced on some of the previous references, this work also uses the antenna bandwidth as one of the parameter to account for the filter design. To this respect, when the filter bandwidth is used as synthesis parameter, this work demonstrates the existence of a unique filter bandwidth that offers an ideal filter response for any given shunt RLC antenna. This concept is fully detailed in Section 2. In addition to that and for those cases where the filter bandwidth and the given antenna is not selected by the designer, an optimization procedure has to be followed to achieve the best possible response. Sections 3 & 4 detail on the optimization procedure for the design of the filtenna when a certain patch antenna is given. Fabrication and measurements of the resulting design are reported in Section 5.

2. Theoretical Background

2.1 Antenna Integration as the Last Filter Stage

The filter-antenna subsystem can be analyzed as the case where the antenna acts as the last stage of a filter network given its intrinsic resonant behavior. The inverse process that we follow here is transforming the last stage of a conventional filter in our antenna in order to keep the load matched and achieve, at the same time, the desired overall performance response. The following analysis is valid as long as our antenna can be accurately characterized as a shunt RLC circuit with values R_{ant}, L_{ant}, C_{ant}.

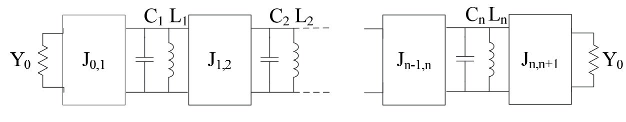

Fig. 1. Classical n-order pass-band inline filter topology.

Now we consider a classical inline topology of a *n*-order pass-band filter network (Fig. 1), with identical resonators whose impedance can be defined as:

$$Z_{0,i} = \sqrt{\frac{L_i}{C_i}}\,. \tag{1}$$

As in conventional filter design techniques the values of the impedance inverters are defined by the filter response [10]. Since we are interested in replacing the last stage of the filter in Fig. 1 by the antenna, we can scale by a factor X the last impedance $Z_{0,n}$ in such a way its value is equal to that from the antenna model (2).

$$Z_{ant} = X \cdot Z_{0,n} = \sqrt{X^2 \cdot \frac{L_n}{C_n}} = \sqrt{\frac{X \cdot L_n}{C_n / X}} = \sqrt{\frac{L_{ant}}{C_{ant}}}\,. \tag{2}$$

The change of the impedance of the last resonator is equivalent to scaling the coupling matrix of the inline network (3).

$$\begin{matrix} & \downarrow\left(\sqrt{X}\right)^{-1} \\ \left(\sqrt{X}\right)^{-1} \rightarrow & \begin{bmatrix} G_0+jB_0 & jM_{0,1} & 0 & \cdot\cdot & \cdot\cdot & 0 \\ jM_{0,1} & sC_1+jB_1 & jM_{1,2} & 0 & \cdot\cdot & \vdots \\ 0 & jM_{1,2} & \cdot\cdot & \cdot\cdot & 0 & \vdots \\ \vdots & 0 & \cdot\cdot & \cdot\cdot & jM_{n-1,n} & 0 \\ \vdots & \vdots & 0 & jM_{n-1,n} & sC_n+jB_n & jM_{n,n+1} \\ 0 & \cdot\cdot & \cdot\cdot & 0 & jM_{n,n+1} & G_{n+1}+jB_{n+1} \end{bmatrix} \end{matrix} \tag{3}$$

where G_i and B_i are respectively the conductance and the susceptance values related to the *i* node of the network and $M_{i,j}$ the coupling value between the nodes *i* and *j*. Therefore it will not only affect the last resonator but both adjacent impedance inverters (4).

$$J'_{n-1,n} = \sqrt{X} \cdot J_{n-1,n}\,, \qquad J'_{n,n+1} = \sqrt{X} \cdot J_{n,n+1}\,. \tag{4}$$

Once the required transformations on the resonator and adjacent couplings have been done, we obtain the network in Fig. 2. At resonance, the conductance G_{in}, related to the last stage admittance $Y_{in} = G_{in} + jB_{in}$ in Fig. 2, will be:

$$G_{in} = \frac{X \cdot J^2_{n,n+1}}{Y_0} \tag{5}$$

and this conductance may not be equal to the antenna radiation resistance, that is in general $G_{in} \neq 1/R_{ant}$.

If we were able to design an antenna fulfilling the constraint in (5), then we would only need replacing the last stage of our filter network with our antenna. With that, the filtenna return losses would follow the required mathematical description of an *n*-order ideal filter. On the other hand, if the antenna cannot be modified, additional transformations are required.

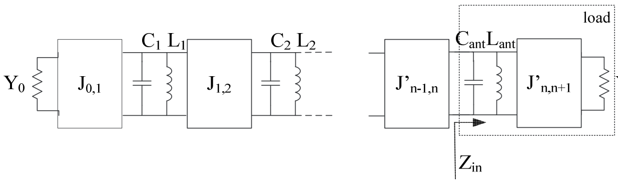

Fig. 2. The network from Fig. 1 where the last filter stage is framed representing the load antenna, after scaling the last impedance.

2.2 Optimal Bandwidth

The impedance inverters of a pass-band filter [11] are related to the parameters of the normalized low-pass filter g_i, cut-off frequency Ω_c and fractional bandwidth *FBW* of the pass-band filter by:

$$J_{0,1} = \sqrt{\frac{FBW \cdot \omega_0 \cdot C_1 \cdot Y_0}{\Omega_c \cdot g_0 \cdot g_1}}\,;$$

$$J_{i,i+1} = \frac{FBW \cdot \omega_0}{\Omega_c} \sqrt{\frac{C_i \cdot C_{i+1}}{\cdot g_i \cdot g_{i+1}}}, \forall i < n\,, \tag{6}$$

$$J_{n,n+1} = \sqrt{\frac{FBW \cdot \omega_0 \cdot C_n \cdot Y_0}{\Omega_c \cdot g_n \cdot g_{n+1}}}\,.$$

From (6), it is obvious that we can also scale all the impedance inverters, keeping the impedance of the resonators equal, just by modifying the fractional bandwidth.

The modification on the bandwidth introduces the required degree of freedom to match R_{in} to R_{ant}. That is, for a certain optimal bandwidth, we will obtain an *n*-order filter network with the desired classical filter function shape.

If we define β as the relative variation of the bandwidth $FBW_{opt} = \beta \cdot FBW$ and we impose that R_{in} must be equal to R_{ant}, it is straightforward to obtain the optimal bandwidth β from (5) and (6):

$$\beta = \frac{Y_0}{R_{ant} \cdot X \cdot J^2_{n,n+1}}\,. \tag{7}$$

Finally we will have the network of Fig. 3 with all its parameters described in Tab. 1.

Obviously, for a given application the bandwidth cannot be arbitrarily changed. However, for a certain bandwidth and antenna, the optimal bandwidth provides useful information on how far we are from a mathematical ideal response. As an example, Fig. 4 shows the return losses of a 4-pole filtenna sweeping the parameter β.

The RLC parameters of the utilized antenna are $R_{ant} = 61\ \Omega$, $L_{ant} = 61.2$ pH, $C_{ant} = 67.9$ pF and the ideal filter response, Chebyshev in this example, is plotted with black line.

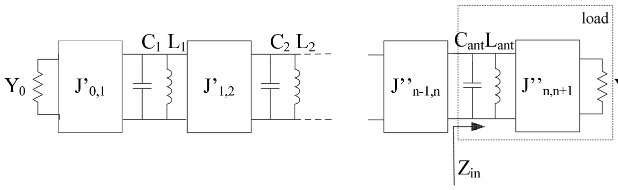

Fig. 3. The network from Fig. 2 where all the impedance inverters have been scaled in order to obtain an ideal filter response.

$J'_{0,1} = \sqrt{\beta} \cdot J_{0,1}$	$L_{ant} = X \cdot L_n$
$J'_{i,i+1} = \beta \cdot J_{i,i+1}, \forall i < n-1$	$C_{ant} = C_n / X$
$J''_{n-1,n} = \sqrt{X} \cdot \beta \cdot J_{n-1,n}$	
$J''_{n,n+1} = \sqrt{X\beta} \cdot J_{n,n+1}$	$\mathrm{Re}\{1/Y_{in}\} = R_{ant} = \dfrac{Y_0}{X \cdot \beta J^2_{n,n+1}}$

Tab. 1. Expressions for the network elements in Fig. 3.

We would like to outline that, for a given bandwidth not equal to the optimal one, using the transformation given in Tab. 1 does not guarantee than we can obtain the best return-losses performance of the filtenna since they are directly constrained to the parameters J_{ij} of the ideal low-pass circuit.

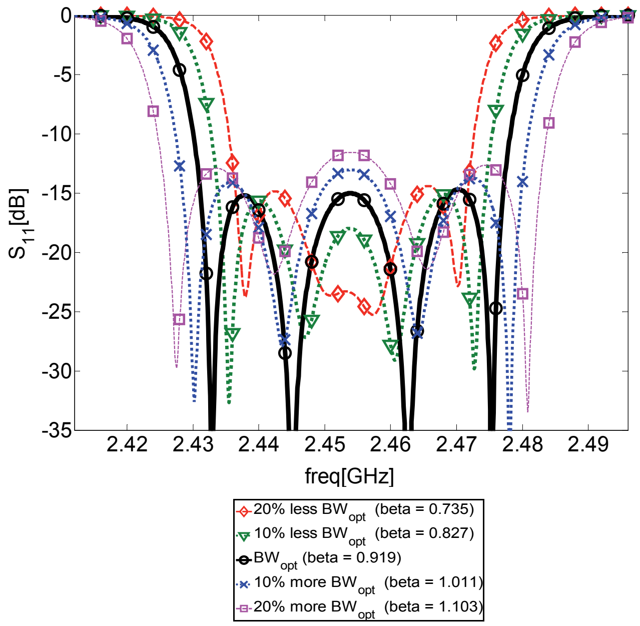

Fig. 4. Return Losses of a 4-pole filtenna vs. β for a given RLC antenna.

If we are interested in getting the best overall performance of the filtenna we need to perform an optimization process as shown in the next section.

3. Optimization Procedure

As stated in the introduction, the optimization procedure allows obtaining the best filtenna response when a prescribed bandwidth and antenna are given to the designer [6].

The approach used in this paper is based on the optimization of the characteristic polynomials [11]. This has the advantage of not being restricted to any particular filter topology or technology. Therefore, the process from the synthesized characteristic polynomials to the circuit configuration can be either the classical element extraction procedure [11] or obtaining the filter coupling matrix. Also, once we have a certain topology, circuit transformation could be applied to obtain additional networks satisfying the same characteristic polynomials and therefore the same filtenna response.

The matrix in (8) recalls the relation between the characteristic polynomials and the filter frequency response, as scattering parameters [11]. Here F(s) refers to the reflection zeros, P(s) refers to the transmission zeros and E(s) accounts for the poles of the transmission and reflection response, being ε_R the ripple coefficient [10]. Note also that s corresponds to the Laplace variable which is related to the normalized frequency variable as $s = j\Omega$.

$$[S] = \begin{bmatrix} \dfrac{F(s)}{\varepsilon_R \cdot E(s)} & \dfrac{P(s)}{\varepsilon \cdot E(s)} \\ \dfrac{P(s)}{\varepsilon \cdot E(s)} & \dfrac{(-1)^{N+1} F^*(s)}{\varepsilon_R \cdot E(s)} \end{bmatrix} \tag{8}$$

Since the characteristic polynomials are defined in the low pass domain and the prescribed bandwidth and antenna are set in the bandpass domain, a frequency transformation [11] should be applied to evaluate the overall filtenna response. The frequency transformation is also recalled in (9) below, where Ω is the transformed frequency, Ω_c is the cut-off frequency in the normalized domain and f_0 is the center frequency:

$$\Omega = \frac{\Omega_C}{FBW}\left(\frac{f}{f_0} - \frac{f_0}{f}\right). \tag{9}$$

Therefore, the methodology is as follows:

1) Propose the designing characteristic polynomials. Note that the characteristic polynomials should also be consistent with the requirement of passive and reciprocal network in order to ensure the feasibility of the resulting filter. This conditions are well bounded by the equations below:

$$S_{11} \cdot S_{11}^* + S_{21} \cdot S_{21}^* = 1, \tag{10}$$

$$S_{22} \cdot S_{22}^* + S_{12} \cdot S_{12}^* = 1,$$

$$S_{11} \cdot S_{12}^* + S_{21} \cdot S_{22}^* = 0. \tag{11}$$

2) From the characteristic polynomials we obtain the frequency response of the pre-filtering stage by means of (8), which are then transformed at the operating frequency, f_0, and prescribed FBW using (9).

3) The S parameters resulting from the previous step are cascaded with the prescribed antenna. The resulting

S_{11} parameter is then compared with our goal response. Note that the goal response corresponds to the reflection coefficient of a filter of order n, where $n - 1$ is the order of the proposed characteristic polynomials. Or eventually, one can use a defined specifications mask.

4) From step 3 an error function is obtained which allows to propose new values for the characteristic polynomials by using a gradient function and with the constraints set in (10) and (11).

The above procedure is illustrated in the flow chart of Fig. 5.

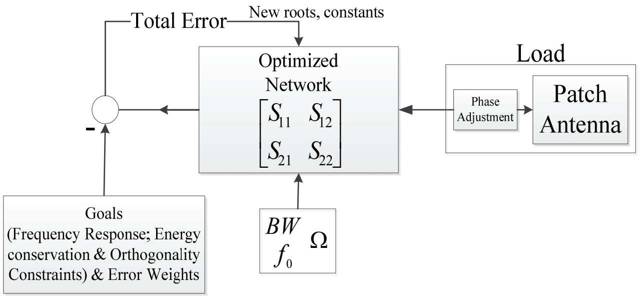

Fig. 5. Flowchart describing the optimization procedure.

The optimum characteristic polynomials are then used to obtain the network of the prefiltering stage of the antenna.

The following section applies the described procedure above to an order 3 filter connected to a patch antenna.

4. Synthesis and Design of a Filtenna

This section illustrates the procedure above by designing a filtenna based on a pre-filtering stage of order 3 with a connected patch antenna, being therefore the overall goal a reflection coefficient comparable to a 4th order filter. In this case we set a Chebyshev 4th order response with return losses of 15 dB.

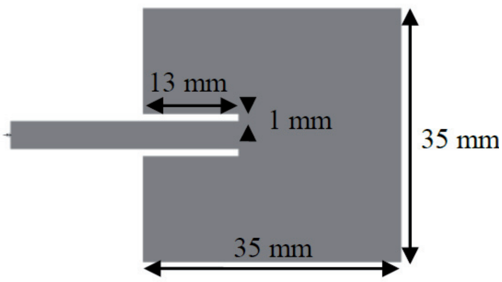

Fig. 6. Microstrip patch antenna.

As a prescribed antenna we have a patch microstrip antenna designed on a $\varepsilon_r = 3$ dielectric. Details on the antenna appear in Fig. 6 above. The antenna operates at a central frequency of $f_0 = 2.45$ GHz and has a bandwidth of BW = 50 MHz.

Simulations depicted in Fig. 7 show the simulated overall efficiency in a logarithmic scale. The loss of efficiency, around 3 dB, is mainly due to losses of the filtering stage. Note that on one hand the filtenna efficiency gives a better selectivity and a flatter performance all over the frequency band. On the other hand, the conventional setup gives a worse selectivity and poorer performance within the frequency band.

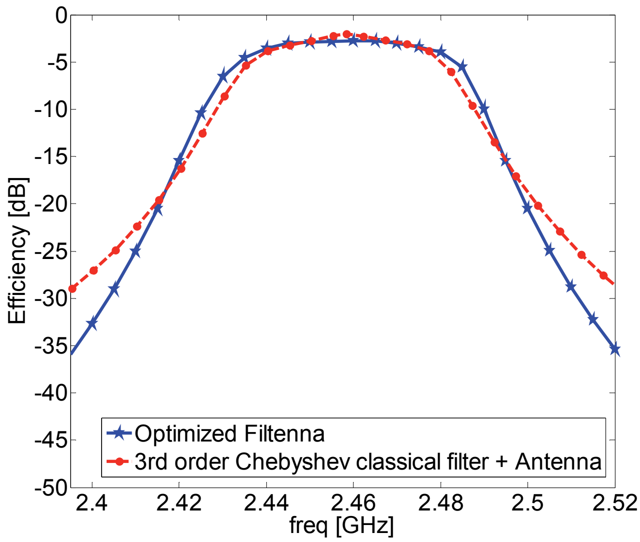

Fig. 7. Efficiency of the filtenna in blue solid line and efficiency of a 3rd order Chebyshev filter (BW = 50 MHz & RL = 15 dB) + antenna in red dashed line.

By applying the optimization procedure explained above we obtain the reflection coefficient depicted in Fig. 8. Note that the figure shows the reflection coefficient corresponding to the optimal bandwidth for our patch antenna, in blue, and the optimized value for a BW = 50 MHz, in red.

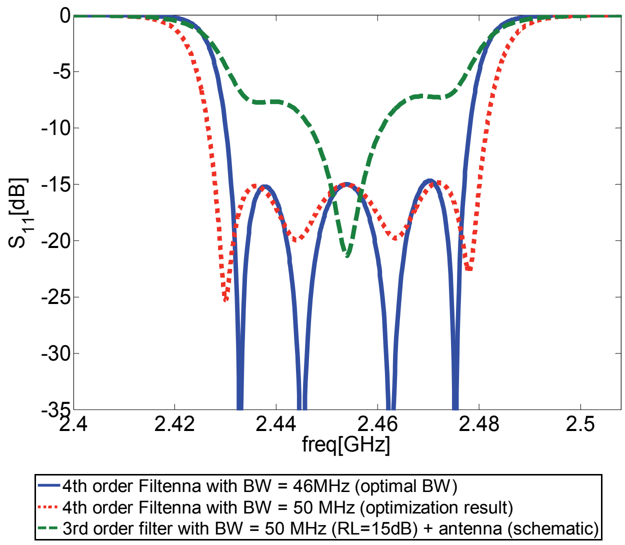

Fig. 8. 50 MHz bandwidth optimization – red dotted line – vs. Optimal BW response found for our microstrip patch antenna (BW = 46 MHz) – blue solid line - vs. 3rd order filter (BW = 50 MHz & RL = 15 dB) + Antenna.

Results show that the optimized response satisfies the 15 dB return losses at expenses of larger bandwidth. Note that this is consistent with the fact that the 50 MHz do not correspond to the optimum bandwidth defined in Section 2. From this figure we can also observe the existence of 4 reflection zeros which indicate that the antenna acts as a filtering resonator as well.

The values detailed in the coupling matrix are set on the low pass prototype. Conventional procedures are then used for the design of the filtering stage [10].

Table 2 shows the resulting characteristic polynomials of the optimization. In the same table we also show the synthesized coupling matrix of the pre-filtering order 3 section.

Optimized Characteristic Polynomial's Roots and Ripple Constants
$F(s)=s^3-0.367s^2+0.431s+0.071$; $P(s)=1$ $E(s)=s^3+1.781s^2+1.953s+1.004$; $\varepsilon=0.99$; $\varepsilon_R=1$
Synthesized Coupling Matrix
$[M]=\begin{bmatrix} 0 & 1.034 & 0 & 0 & 0 \\ 1.034 & 0 & 0.869 & 0 & 0 \\ 0 & 0.869 & 0 & 0.661 & 0 \\ 0 & 0 & 0.661 & 0 & 0.844 \\ 0 & 0 & 0 & 0.844 & 0 \end{bmatrix}$

Tab. 2. 2-port network results from the optimization process. Characteristic polynomials and coupling matrix.

To illustrate the advantage of the optimization procedure, Fig. 8 also shows in dashed green the reflection coefficient of 3rd order Chebyshev filter (BW = 50 MHz, RL = 15 dB) directly connected to the prescribed patch antenna.

5. Prototype and Measurements

Once the optimization is done we will synthesize the two-port network shown in Fig. 9 using hairpin microstrip resonators with RO3003 substrate $\varepsilon_r = 3$ and thickness $h = 1.52$ mm). This two-port network will be then connected to the microstrip patch antenna described in the previous section.

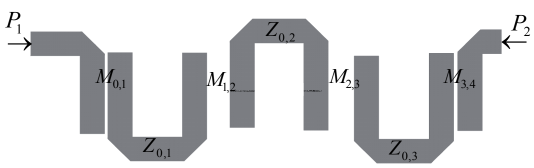

Fig. 9. Optimized network designed in microstrip technology.

The resulting non symmetrical two-port network (without antenna) is initially designed considering lossless transmission lines using a circuit simulator [12] by conventional filter design techniques [10], [11]. The obtained circuit is the starting point to optimize the two-port network, including losses, by means of an electromagnetic (EM) simulator [12].

Figure 10 shows the S_{11} and S_{22} by use of circuital and EM simulations of the 2-port network. We have also included the target given by the coupling matrix, which almost overlap the circuital simulations.

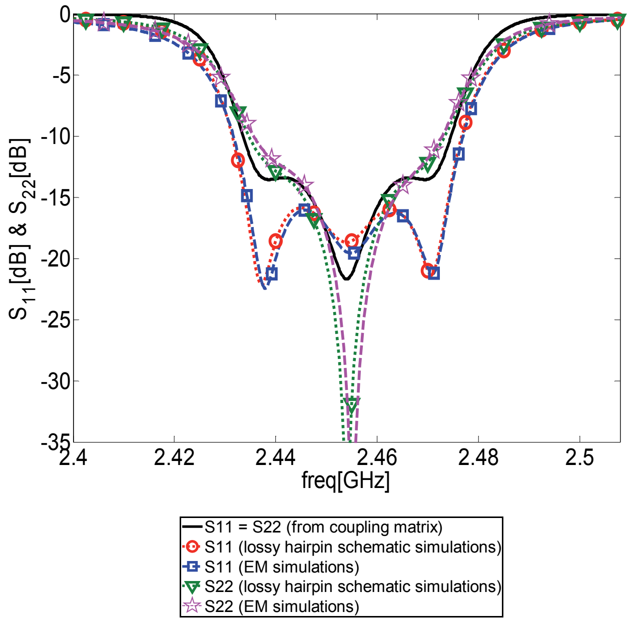

Fig. 10. 3rd order 2-port network placed before the antenna. S_{11} & S_{22} parameters.

Finally we connect the microstrip antenna to the network by means of a 50 ohms transmission line with the required length to load the 2-port network as a shunt RLC resonator. The whole filtenna device is simulated again with the EM simulator. Figure 12 shows the return-losses by use of circuital and EM simulations. We can observe great results from the optimization process, where the optimized network is considered lossless. As expected, worse selectivity and less "depth" in the poles of the S_{11} are seen in the EM simulations due to the addition of the substrate losses. Furthermore, the measurements show even worse performance due to losses underestimation when simulating. Figure 12 also shows measurements of the fabricated prototype (Fig. 11). The central frequency has been shifted 17.5 MHz in order to make easier the comparison and re-

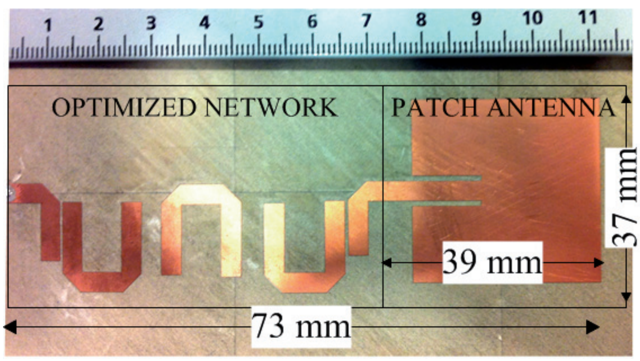

Fig. 11. Filtenna prototype fabricated in microstrip technology.

turn losses are better than 12 dB in the 50 MHz bandwidth. This is a better performance than that achieved using the 3rd order filter + antenna set-up (RL = 7 dB as shown in Section 4), whose simulations has been included in Fig. 12 as well.

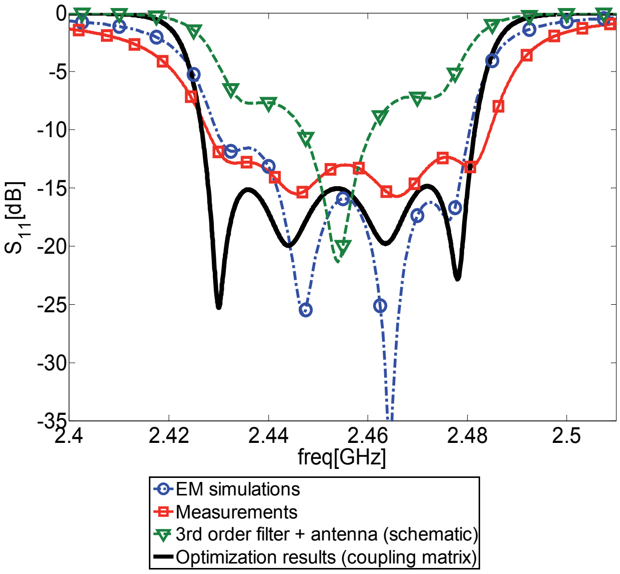

Fig. 12. 4th order filtenna (optimization results & EM simulations & measurements) vs 3rd order filter + antenna.

6. Conclusions

In this work a mathematical step-by-step description process has been followed to show that there is an optimal overall system bandwidth which allows integrating both the filter and the antenna sub-networks, as a unique higher order filtenna network achieving an optimal performance in terms of return losses.

The basic premise for the approach to be valid is that the antenna has to be modeled as a shunt RLC circuit due to its intrinsic resonant behavior. As a result of this, we can combine both subsystems in such a way that the load antenna finally performs as an additional stage of the initial filter, thus increasing by one its order. Then, it is possible to achieve any classical polynomial filter response, as done with Chebyshev filtering functions.

In this way, the return losses of some conventional set-up, where an n-order filter is cascaded to an antenna by means of a feeding line, can be improved up to a $n+1$ order filter performance. We have to keep in mind we should demand a system performance consistent with the actual network order, which will be one order higher than the filter by itself.

Besides, whenever we are not able to use the optimal bandwidth in our system, it is also possible under the same circumstances to fit a certain filter mask through optimization as well. The optimization method has been presented in this paper to simulate a filtenna in order to compare its performance with a conventional filter followed by the same antenna in cascade. Since we were aiming to fit a one order higher filter mask, the results of our optimized network have clearly improved. The methodology success relied on both the fulfillment of constraints and the convergence of the well-known characteristic polynomials that define a 2-port filtering network.

Acknowledgements

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Jordi VERDU was born in Sabadell, Spain, in 1980. He received the Telecommunications Engineering degree in 2006, the M.S. degree in 2007 and the Ph.D in 2010 from the Universitat Autònoma de Barcelona. His research topics are focused in the design of microwave devices and also filters based on acoustic resonators.

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