Dividing and filtering function integration for the development of a band-pass filtering power amplifier

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Abstract **— Integration of amplifying and filtering functions, based on transversal structure principles, is one solution to develop miniaturized mobile communication devices. An architecture using power combination has been developed, combining the filter and the three branches power divider. The system is dual-band (0.9GHz and 1.94GHz). The filter is constituted by two looped Stepped Impedance Resonators (S.I.R.), symetrically positionned on each side of the stopped feeding line. Power is equally divided thanks to coupling effects between output microstrip lines.**

I. Introduction

A band-pass filtering power amplifier based on a transversal structure (Fig. 1) has been studied. Incident power is equally separated into N branches. Each signal undergoes an identical and unilateral amplification ($s^i_{12}=0$) and different gain and phase frequency variations. In that case, global transfert response has a quite simple expression (Equ. 1).

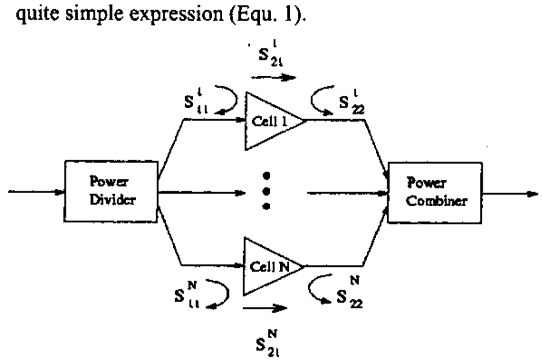

Fig. 1. Transversal structure constitution.

$$S_{21}=\frac{4}{N}\frac{\sum_{i=1}^{N}\frac{s^i_{21}}{(1+s^i_{11})(1+s^i_{22})}}{\left(1+\frac{1}{N}\sum_{i=1}^{N}\frac{1-s^i_{11}}{1+s^i_{11}}\right)\left(1+\frac{1}{N}\sum_{i=1}^{N}\frac{1-s^i_{22}}{1+s^i_{22}}\right)} \quad (1)$$

If power divider and combiner are matched to each cell ($s^i_{11}=s^i_{22}=0$), S_{21} is simplified (Equ. 2), but possibilities to modify global frequency response out of the band are reduced. So, the first solution has been kept.

$$S_{21}=\frac{1}{N}\sum_{i=1}^{N}s^i_{21} \quad (2)$$

To simplify realization and to respect size constraints, a three branches structure has been chosen. Therefore, transversal band-pass filtering usually requires a higher number of branches to respect bandwidth constraints. So, a filtering band-pass function must be integrated in the structure. Instead of setting filtering functions on each branch, we have prefered, for size reason, to integrate it into the power divider.

The global structure must work on different standards. So, filtering function has several preset bands, and the undesired ones have to be suppressed by reconfigurable zeroes of power combination on the global transfert response.

This introduction lights on the power divider characteristics :

1) Matched impedance input,
2) Three power-balanced mismatched outputs,
3) Dual-pass-band filtering transfer responses to switch on different telecommunication standards (ex : 0.9GHz and 1.94GHz).

II. Design Method

The final structure of the filtering power divider we have developed results from a sequence of transformations which are described in this second section.

A. Use of a S.I.R. resonator

The filtering function is first made with a S.I.R. non-conventional filter [1] (Fig. 2). The resonator geometry gives several degrees of freedom that allow to set resonant frequencies and bandwidths fitted with the application under consideration.

For size and simplification reasons, the following study is focused on structures based on a one-S.I.R. non-conventional filter (Fig. 2) with three degrees of freedom on frequencies ($K=Z_2/Z_1$, θ_{10} and θ_{20} where Z_1 and Z_2 are lines impedances and θ_{10} and θ_{20} electric lengths at center frequency) and four on bandwidth (K, θ_{10}, θ_{20} and Zoe that is even mode characteristic impedance). The first and second resonant frequencies (f_0 and f_{s1}) are fixed by the equations 3 and 4. It is interesting to give some general properties of this structure :

- Coupling areas must be equivalent to match the global filter,
- The more the resonator is coupled (length and distance), the greater the filter bandwidth is,
- If $K<1$, then $f_{s1}>2.f_0$ (for $K=1$, $f_{s1}=2.f_0$ and for $K<1$, $f_{s1}>2.f_0$),
- The resonator electrical length is close to 180° (180° if K=1).

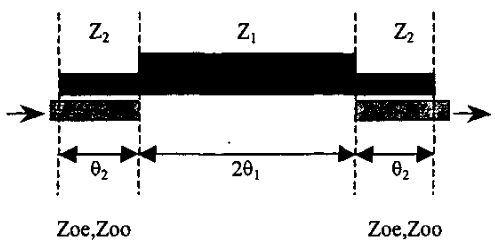

Fig. 2. One-S.I.R. filter.

Zoo is the odd mode characteristic impedance.

$$K = \tan \theta_{10} . \tan \theta_{20} \tag{3}$$

$$K . \tan\left(\frac{f_{s1}}{f_0}\theta_{10}\right) + \tan\left(\frac{f_{s1}}{f_0}\theta_{20}\right) = 0 \tag{4}$$

Equ. 3 and 4. Equations corresponding with first and second resonant frequencies

A S.I.R. filter has been realized fitting to resonant frequencies 0.9GHz and 1.94GHz (Fig. 3). From this filter, a three-balanced-port filtering power divider has been imagined (Fig. 3). S_{21}, S_{31} and S_{41} are not so different from previous simple filter frequency response on condition that coupled length on port 3, which have two-side coupled lines, be shorter than ports 2 and 4 ones. Using an electromagnetic simulator (IE3D, Zeland) and without losses, the structure has been optimized and the results are presented on Fig. 4.

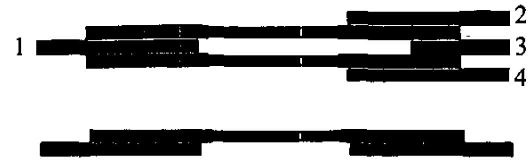

Fig. 3. Three-port S.I.R. power divider and its corresponding S.I.R. filter

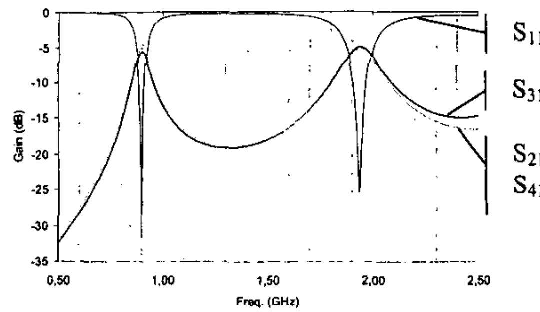

Fig. 4. Simulated S.I.R. power divider response

Without losses, each output must provide a third part of input power. This means that S_{21}, S_{31} and S_{41} must theoretically be closed to –4.8dB. Here, there is a 0.5dB loss when no dielectric and ohmic losses were declared for simulation and when the feeding line is correctly matched (-32dB). So, the structure is radiating (8% of input power is radiating at 0.9GHz).

B. Loop-filter configuration

In order to reduce the previous structure length, it could be interesting to loop resonators. So, it is necessary to show how non-conventional S.I.R. filters can be transformed into loop-filters.

The main difference in their constitution is coupling accesses. In the S.I.R. filter, output is the isolated access of the coupler when it is the coupled access in loop-filters. A link between these two structures is presented in Fig. 5.

In opposition to the constitution in series of the S.I.R. filter (Fig.5 bottom), in which each lines length is independent of the others, the loop-filter (Fig. 5 top) imposes conditions on lengths that lead to a maximum coupling length Lm. This maximum is equivalent to a Lm/2 coupling length on S.I.R. filters. As length coupling is linked to bandwidth, loop-filter bandwidth is much smaller than S.I.R. filter one.

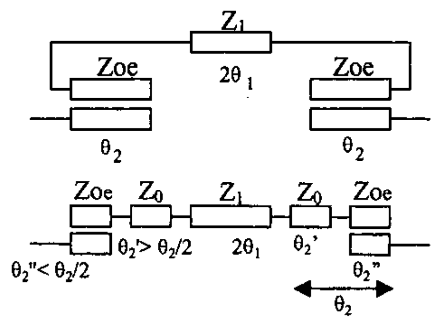

Fig. 5. Correspondence between loop and S.I.R. filters.

It is possible to evaluate how geometry influences bandwidth, considering three different filters : S.I.R. filter, semi-loop-filter and loop-filter (Fig. 6).

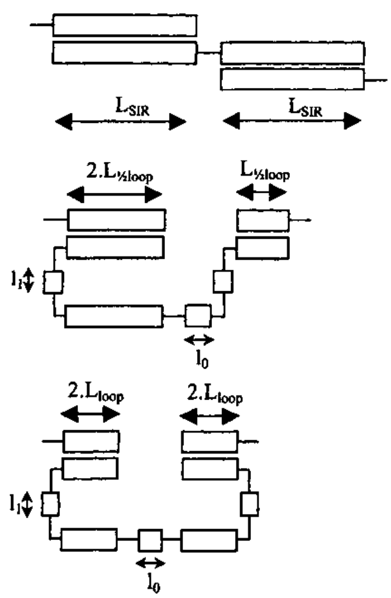

Fig. 6. Different filter configurations

l_0 and l_1 are chosen to limit undesired coupling effects. Each resonator has a length closed to L_R which squares to a 180° electric length at f_0.

In a first approach, we can consider that β is constant along the resonator, where $\beta=2\pi/\lambda_g$ is the propagation constant and λ_g the guided wavelength. Its electric length at f_0 is $\beta_0.L_R=\pi$ (actually, β is different for coupling areas and for microstrip lines). Then, an estimation of maximum equivalent coupling lengths for each structure and a numerical application corresponding to a 0.9GHz filter in FR4 substrate and with 0.2mm spacing between coupled lines is given by equations 5, 6 and 7.

$$L_{SIR}=\frac{L_R}{2} \qquad B_{max}=14\% \qquad (5)$$

$$L_{\frac{1}{2}loop}=\frac{L_R}{5}-\frac{2l_1+l_0}{5} \qquad B_{max}=2.5\% \qquad (6)$$

$$L_{loop}=\frac{L_R}{8}-\frac{2l_1+l_0}{8} \qquad B_{max}=1\% \qquad (7)$$

Equ. 5, 6 and 7. Maximum coupling lengths for different filter configurations and corresponding maximum bandwidths for a 0.9GHz filter with FR4 substrate.

III – SEMI-LOOP-FILTERING POWER DIVIDER

A- Selection criterions

In section II-B, three different filters have been presented. S.I.R. filters, in spite of their high bandwidth possibilities, have a too long-shaped structure. Loop-filters have the reverse problem : their geometry is compact but their bandwidth is very reduced (<1%). So, the semi-loop-filter is a compromise which is interesting in this power-divider : it is easier to couple three ports in this structure than in loop-filters. As a matter of fact, output configuration is absolutely alike to 3-ports S.I.R. power divider one (Fig. 3 and Fig. 7).

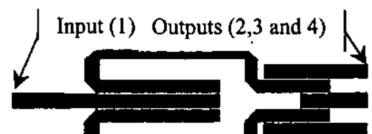

Fig. 7. Layout of the three-port semi-loop-resonator power divider

B- *Global structure realization*

Final structure has been realized on Epoxy FR4, with dielectric permittivity of 4.4 and loss tangent of 0.02. As seen on Fig. 8, results are fitting correctly with the simulation.

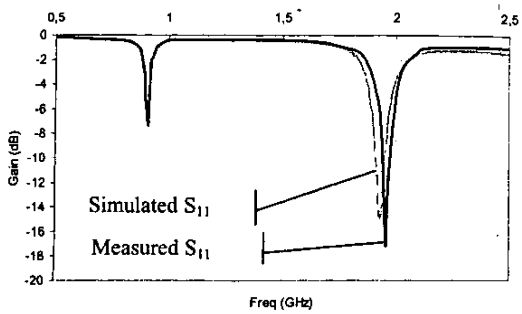

Fig. 8. Simulated and measured return loss

Device behavior presents two resonant frequencies (0.91GHz and 1.96GHz) with reduced bandwidths (about few percents) according to previous analyses in section II-B. But, considering the chosen substrate and its loss tangent value, return loss is not sufficient (-7dB) at 0.9GHz. However, these levels are strictly according to simulation.

By simulation with reduced losses ($<10^{-3}$), we have verify that return loss levels improve significantly (until –12dB). So one solution is to choose a low-loss substrate or to modify feed line width between the two resonators. This last point presents the particularity of still reducing bandwidth.

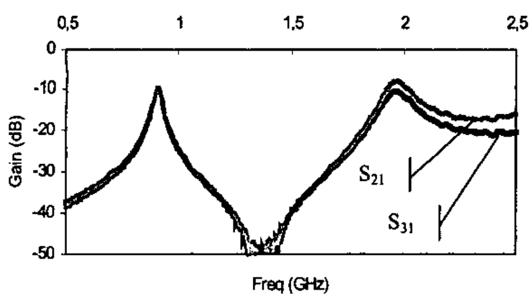

Fig. 9. Measured isolation loss

In transmission, power levels on the three outputs are almost identical (Fig. 6), with always a shift of 5dB respect to the theoretical –4.7dB. This difference has been identified by numerical simulation and is also due to dielectric losses.

V. Conclusion

A dual-band structure (0.9GHz and 1.94GHz) which integrates power divider and filter has been designed. The design method used allows to obtain a compact structure, based on the coupling between microstrip lines and semi-loop SIR resonators, while controling frequencies, bandwidths and output power levels. This work aim is the realization of a filtering amplifier with transversal structure which can be used in mobile communication receiver design. Afterwards, considering loss return levels obtained, it will be necessary to use a low loss substrate (like Duroïd). An other solution is to modify input coupling at the risk of reducing bandwidth. Actually, the microstrip feed line turns into a coplanar one, which creates an abrupt impedance change. A third improvement is to reduce coupling lengths and then device size, by inserting interdigited structure into resonator input and ouput. This avoids constraints on resonator geometry especially on line lengths (cf. Fig. 6).

References

[1] S. Denis, « Caractérisation théorique et expérimentale de structures de propagation multicouches – Application aux filtres plaqués microondes à hautes performances », PhD Thesis Bretagne Occidentale 1997.

[2] L. Billonnet, B. Jarry and P. Guillon, « Design concept for microwave recursive and transversal filters using Lange couplers », IEEE MTT-S Digest, pp. 925-928 1992.