A Dual-Band MMIC Low frequency 180° Hybrid

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Abstract — **In this paper design formulas, for the development of an unusual 180° hybrid working in two distinct frequency bands, are provided. In particular the design formulas of a lumped-element rat race, starting from a conventional transmission-line rat-race, are found out. In order to generalize the matching problem the input and output port impedances are considered to be different .**

I. Introduction

Hybrid couplers are key components in the design of microwave circuits such as mixers and modulators [1]. Moreover, they can be used as power dividers and combiners in array antennas. The rat-race, one of the simplest 180° hybrid to be designed, is normally realized, also in the MMIC version, using microstrip lines, at least at frequencies over 20 GHz. In the lower microwave frequency range the conventional transmission-line rat-race has too large dimensions to be effectively integrated in a MMIC, so that the most suitable realization becomes the lumped element one.

As it is known, in order to realize a rat race using lumped elements (L.E.), the transmission-line segments are replaced, in the design procedure, by equivalent pi and tee networks, by equating the ABCD transmission matrices of the line segments to the corresponding ABCD matrices of the L.E. pi and tee networks [2].

This design approach normally results in a very effective realization both in terms of isolation and transmission performances among the relevant ports, but normally in a narrow frequency bandwidth, so limiting a widespread utilization of the L.E. solution [3].

On the other hand, there are many important applications where it is necessary to implement a 180° hybrid that has to work in two frequency ranges, very apart from each other, as in the case, for instance, of mixers working with RF (or LO) frequency much greater than LO (or RF) frequency. This problem is usually approached by carrying out a broadband hybrid solution, trying to overcome the rat-race frequency limitations which normally does not exceed two octaves in bandwidth.

However a completely different approach can be followed designing a Dual Band Rat-Race, i.e. a hybrid having the requested performances *only* in the relevant frequency bands without trying to maintain the same level of performances in the bandwidth *beetween* the operating ones, as demonstrated in [4].

Moreover, in some cases the level of input and output port impedances could be different from 50Ω. The problem, usually solved by using large impedance

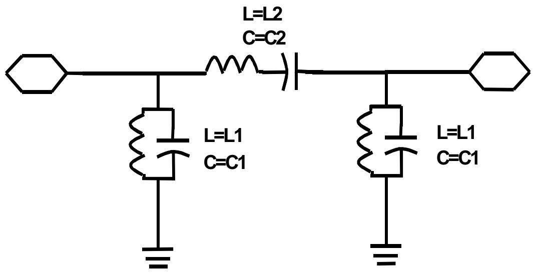

Fig.1. Bandpass pi network.

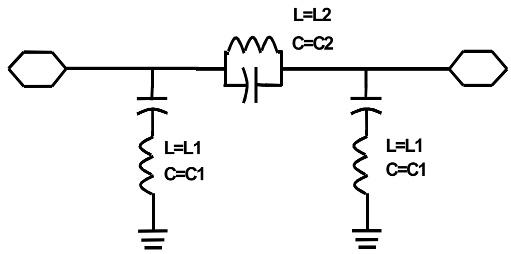

Fig.2. Band-stop pi network.

transformers at the relevant ports, can be better faced through a direct synthesis procedure. This procedure, that results in new synthesis formulas, is described in the following.

Taking into consideration a rat race hybrid with different input and output impedances, in fact it is possible to directly synthesise a L.E. circuit, which assures the requested level of impedances, while enabling a perfect match at the ports and the best isolation conditions.

II. Design of the 180° Hybrid

The design procedure starts from a conventional transmission-line rat-race design , carried out with three 90° line section and one 270° line section, finding the element values of ABCD matrix for the two line section kinds. Then the line segments are modelled in terms of pi networks by imposing the corresponding ABCD matrix elements to assume the same values.

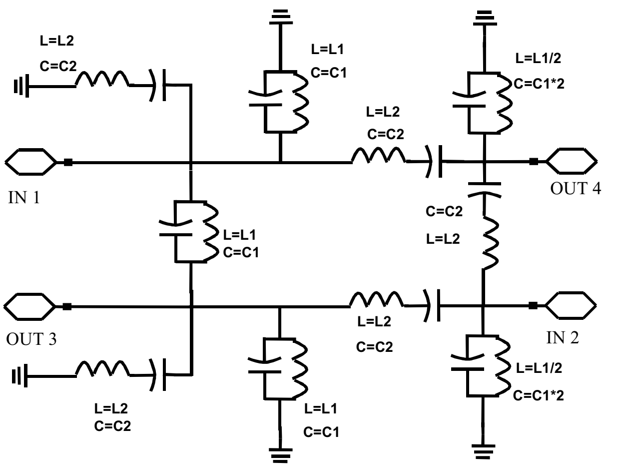

Fig.3. Double-frequency 180°hybrid circuit.

The transmission matrix of a line segment of length L is shown below:

$$\begin{bmatrix} A & B \\ C & D \end{bmatrix} = \begin{bmatrix} \cos(\beta L) & jZ_0 \sin(\beta L) \\ jY_0 \sin(\beta L) & \cos(\beta L) \end{bmatrix} \tag{1}$$

and for $\beta L = 90°$ and $\beta L = 270°$ the transmission matrix becomes, respectively,

$$\begin{bmatrix} A & B \\ C & D \end{bmatrix} = \begin{bmatrix} 0 & jZ_0 \\ jY_0 & 0 \end{bmatrix} \text{ and } \begin{bmatrix} A & B \\ C & D \end{bmatrix} = \begin{bmatrix} 0 & -jZ_0 \\ -jY_0 & 0 \end{bmatrix} \tag{2}$$

Passing to the L.E. approach, a bandpass pi network (Fig.1) is modelled, in order to obtain, at the chosen frequencies two different behaviours: at the frequency ω_l the pi network will perform like a transmission line with $\beta L = 90°$ and characteristic impedance Z_1, while at the other frequency ω_h this one will operate like a transmission line with $\beta L=270°$ and characteristic impedance Z_2.

$$\begin{bmatrix} A & B \\ C & D \end{bmatrix} = \begin{bmatrix} 1 & 0 \\ jB & 1 \end{bmatrix} \cdot \begin{bmatrix} 1 & jX \\ 0 & 1 \end{bmatrix} \cdot \begin{bmatrix} 1 & 0 \\ jB & 1 \end{bmatrix} = \begin{bmatrix} 0 & jZ_2 \\ jY_2 & 0 \end{bmatrix}$$
$$\begin{bmatrix} A & B \\ C & D \end{bmatrix} = \begin{bmatrix} 1 & 0 \\ jB & 1 \end{bmatrix} \cdot \begin{bmatrix} 1 & jX \\ 0 & 1 \end{bmatrix} \cdot \begin{bmatrix} 1 & 0 \\ jB & 1 \end{bmatrix} = \begin{bmatrix} 0 & -jZ_2 \\ -jY_2 & 0 \end{bmatrix} \tag{3}$$

where B is the shunt L_1-C_1 susceptance and X is the series L_2 -C_2 reactance. So the following design formulas are easily achieved:

$$L_1 = \frac{10 \cdot \sqrt{R_l \cdot R_h} \cdot \left(\omega_h^2 - \omega_l^2\right)}{\omega_h \cdot \omega_l \cdot \left(\omega_l \cdot \sqrt{R_l} + \omega_h \cdot \sqrt{R_h}\right)} \tag{4}$$

$$C_1 = \frac{\omega_l \cdot \sqrt{R_h} + \omega_h \cdot \sqrt{R_l}}{10 \cdot \sqrt{R_l \cdot R_h} \cdot (\omega_h^2 - \omega_l^2)} \tag{5}$$

$$L_2 = \frac{10 \cdot \left(\omega_l \cdot \sqrt{R_l} + \omega_h \cdot \sqrt{R_h}\right)}{\omega_h^2 - \omega_l^2} \tag{6}$$

$$C_2 = \frac{\omega_h^2 - \omega_l^2}{10 \cdot \omega_l \cdot \omega_h (\omega_l \cdot \sqrt{R_h} + \omega_h \cdot \sqrt{R_l})} \tag{7}$$

where R_l and R_h are the impedances at the output port, at the frequencies ω_l and ω_h respectively, while the level of the impedances at the input ports are equal to 50Ω.

As a consequence the following relations stand:

$$Z_1 = \sqrt{2 \cdot 50 \cdot R_l} \quad Z_2 = \sqrt{2 \cdot 50 \cdot R_h} \tag{8}$$

Obviously when $R_l = R_h = 50\Omega$, the formulas (4),(5),(6),(7) result into the simpler expressions given in [4].

In a similar way the band-stop pi network, Fig.2, can be modelled. As it is well known, in fact, with this approach changing only the topology, while maintaining the same values for the used L.E., the bandpass behaviour is changed into the corresponding band-stop one. For this reason the pi network at Fig.2 behaves like a transmission line with $\beta L=270°$ at ω_l and like a transmission line with $\beta L=90°$ at frequency ω_h.

The resulting 180° L.E. hybrid is shown in Fig. 3.

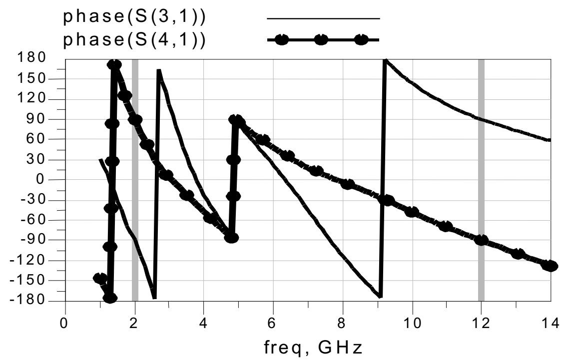

Fig.4. Phase Shift ports 3-1 and 4-1.

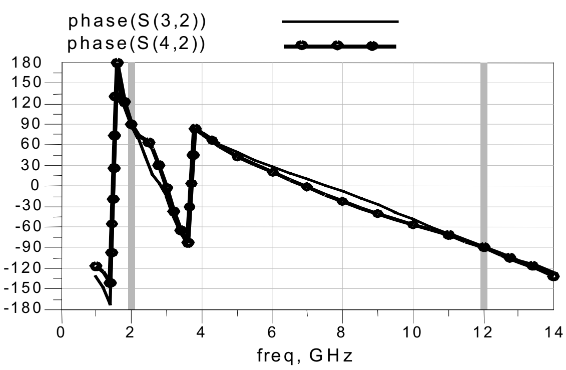

Fig.5. Phase Shift ports 3-2 and 4-2.

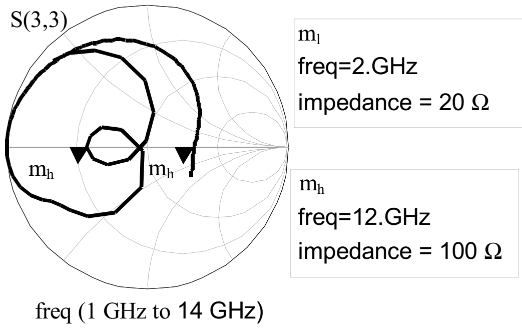

Fig.6. Impedance at the ouput port 3.

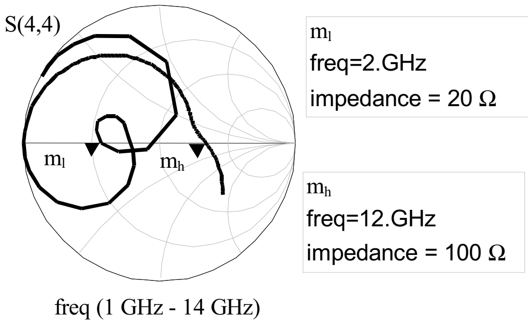

Fig. 7. Impedance at the ouput port 4.

The effectiveness of the given formulas is demonstrated by designing a 180° hybrid with f_l=2GHz, f_h=12GHz, demanding an impedance level of R_l=20Ω and R_h=100Ω respectively. The corresponding electrical behaviour at the relevant ports is shown in Fig.4,5,6 and 7.

III. RESULTS

In order to validate the proposed design approach, a 180° hybrid has been designed assuming f_l=2GHz, f_h=5GHz, R_l = R_h = 50Ω and has been fabricated on a 100-µm thick GaAs substrate (ε_r=12.8, conductivity =5.5e7 S/m, T=3.3 µm, tanδ=6.5e-3, roughness=0.15).

The ideal lumped elements have been replaced with rectangular spiral inductors and MIM capacitors. The final monolithic layout (2.9 X 1.9mm) of the 180° hybrid

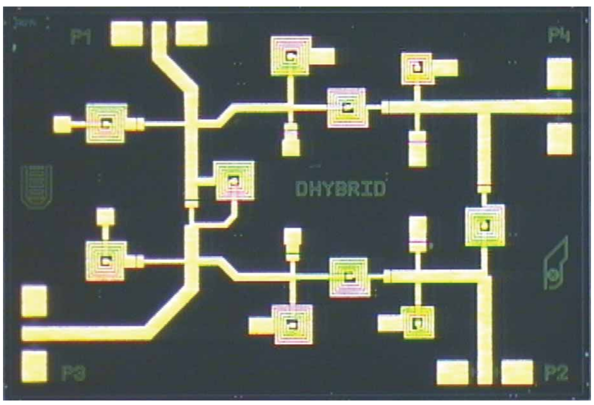

Fig.8. Monolithic layout of the 180° hybryd.

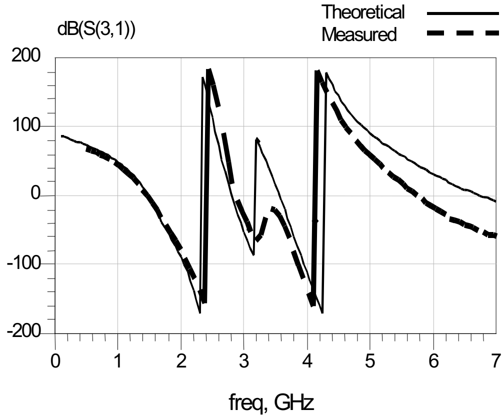

Fig.9. Phase Shift ports 3-1.

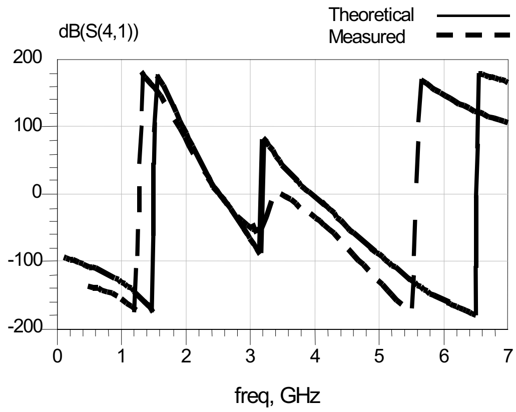

Fig.10. Phase Shift ports 4-1.

is shown in fig. 8.

Measurements have been performed by mounting the circuit on the probe station RF-1 Cascade Microtech. As usually, two of the four ports were terminated with a 50Ω load and the S-parameters of the other two ports were measured with an HP8510C network analyser.

The theoretical and the measured performances, vs frequency, of the 180° L.E. hybrid are plotted in figures 9-13: in particular, fig 9, 10 show the phase shift behaviour between port 3 and port 1 and between port 4 and port 1 respectively, figure 11 shows the isolation, and figure 12 e 13 describe the return losses.

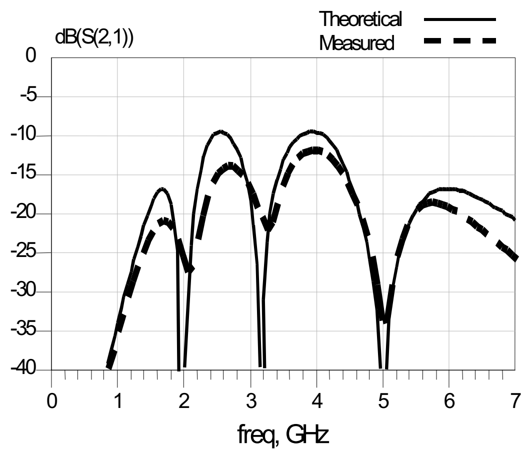

Fig.11. Isolation between input ports.

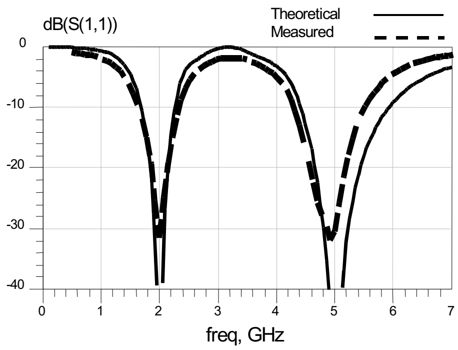

Fig.12. Return Loss at Port 1.

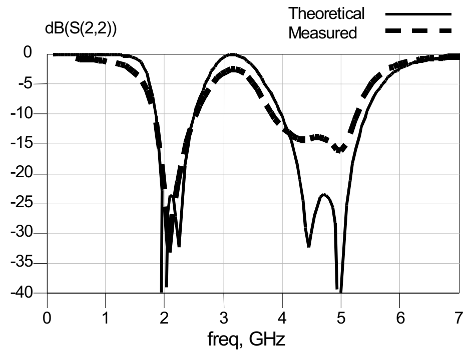

Fig.13. Return Loss at Port 2.

IV. CONCLUSION

In this paper a technique for the design of 180° lumped element hybrid at two frequencies has been described. This particular network can be employed to develop mixers with the RF (or LO) frequency much greater than LO (or RF) frequency. Besides in order to get over general matching problems the required synthesis formulas have been achieved assuming the 180° hybrid ouput and input impedances different.

An effective circuit has been realized, measured performances have been reported and a good agreement between measurements and simulations has been pointed out.

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