# MICROSTRIP DUPLEXERS WITH SINGLE-SIDED FILTERS

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## Abstract

*This paper describes a novel approach to the design of high selectivity duplexers in microstrip technology for applications in mobile communications systems. The configuration proposed makes use of two single-sided filters (devices characterized by requirements in one passband and one stopband, both of finite extension). Using this class of filters it is possible to obtain high selectivity and small insertion losses at the same time (even with low unloaded Q microstrip resonators).*

## INTRODUCTION

In the synthesis of diplexers it is often required to achieve high selectivity in limited frequency bands (i.e. TX and RX bands in RX and TX branches); using classical bandpass filters configuration (even with suitably located transmission zeros) many resonators are however required, thus precluding the possibility of using low unloaded-Q resonators (i.e. microstrip technology). Recently some works have appeared in the literature reporting design procedures for single-sided filters, which are characterized by requirements in one passband and one stopband, both of finite extension; these filters require a smaller number of resonators with respect to classical bandpass filters and they are then particularly suited for realizing duplexers in microstrip technology.

In this paper a design approach for duplexers using single-sided filters in microstrip technology is presented; the filters are characterized by a quasi-equiripple response both in the passband and in the stopband. In fact, transmission zeros are suitably introduced in the stop band, in order to improve selectivity and minimize the number of resonators required.

## THE DESIGN PROCEDURE FOR SINGLE-SIDED FILTERS

It is assumed that the single-sided filters here considered are characterized by the usual equivalent network in fig.1, where *n* shunt resonators are cascaded through *n+1* ideal admittance inverters.

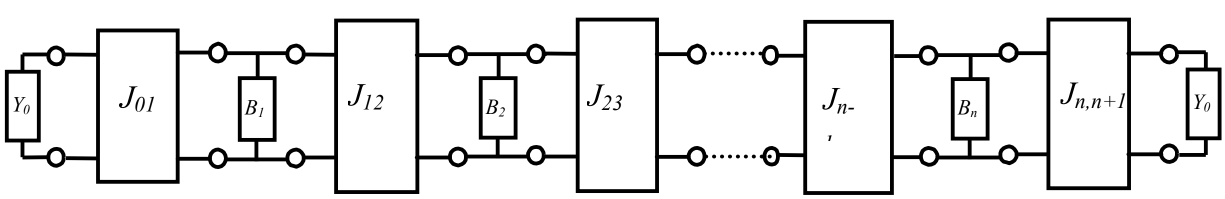


**Fig. 1**: General configuration of a single-sided filter

All the admittance inverter parameters $J_{i,i+1}$ are equal to 1 but for n even, where $J_{n/2,n/2+1}$ is given by $J_{n/2,n/2+1} = \{\varepsilon^2+1\}^{1/2} \pm \varepsilon$ (lower sign for n=4,8,12,.. upper sign for n=2,6,10,...), where $\varepsilon$ is defined by the maximum attenuation in the filter passband $A_m = 10\log(1+\varepsilon^2)$. The resonators susceptance $B_k$ of the *k-th* resonator is assumed to have the following frequency dependence:

$$B_k = A_k \frac{f - f_{pk}}{f - f_{sk}} \tag{1}$$

where $f$ is a suitable normalized frequency variable, which depends on the kind of resonator considered; $A_k$, $f_{pk}$ and $f_{sk}$ are parameters independent on $f$, which completely characterize the susceptance $B_k(f)$. A synthesis procedure has been developed, which allow to determine the coefficient $A_k$, $f_{pk}$ and $f_{sk}$ , once the single-sided filter specifications are given (that is the initial and final frequencies of passband ($f_{pa}$, $f_{pb}$) and stopband ($f_{sa}$, $f_{sb}$), the number $n$ of resonators, the required passband return loss). This procedure, originally developed for waveguide filters, is described in detail in [2] and it allows to obtain a quasi-equiripple response both in pass-band and in stop-band; from a computational point of view only the Chebycheff low pass prototype parameters $g_n$ are required, making the procedure very easily implementable (network synthesis based on elliptic functions is not required).

## MICROSTRIP IMPLEMENTATION OF SINGLE-SIDED FILTERS.

In order to realize a duplexer, two complementary single-sided microstrip filters must be designed (i.e. one filter with the stopband above the passband and one filter with the stopband below the passband). A possible lumped elements equivalent circuit for the resonators to be used in the single-sided filters here considered is shown in fig. 2a: $B_k$ and $X_k$ represent frequency invariant immitances and $L_k$ is an inductor. In order to have the series resonance above or below the shunt resonance (passband), the sign of $B_k$ must be suitably selected; in fact there are the following relationships between the elements of the equivalent circuit and the parameters $A_k$, $f_{pk}$ and $f_{sk}$:

$$L_k = \frac{1}{2\pi A_k (f_{pk} - f_{sk})}, \qquad X_k = -2\pi \cdot f_{sk} L_k, \qquad B_k = \frac{1}{2\pi L_k (f_{pk} - f_{sk})} \tag{2}$$

Note that, being $L_k$ positive, $X_k$ is always negative; moreover, filter one ($f_{sk}>f_{pk}$) requires $B_k < 0$ and filter 2 ($f_{sk}<f_{pk}$) requires $B_k > 0$. Possible implementations of these resonators with transmission line sections are reported in fig. 2b) and 2c)

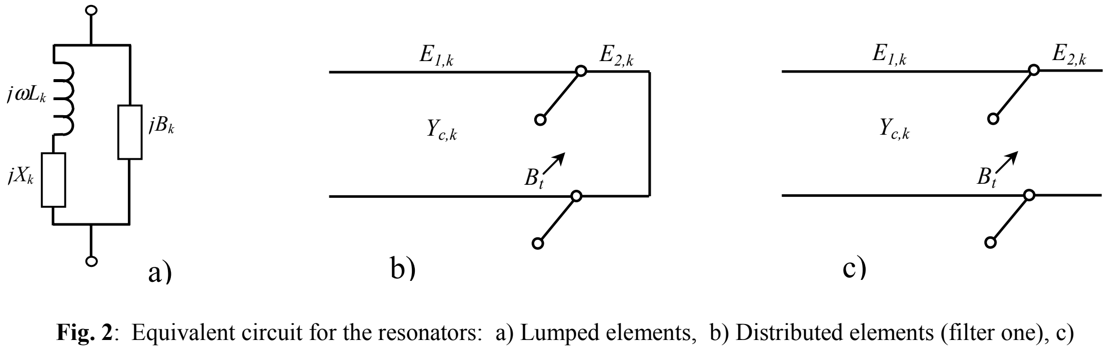


**Fig. 2**: Equivalent circuit for the resonators: a) Lumped elements, b) Distributed elements (filter one), c) Distributed elements (filter two)

The susceptance slope parameter for the lumped resonator around the passband resonance is given by:

$$B_{eq,k} = \frac{1}{2}\frac{\partial B_t}{\partial \omega}\bigg|_{\omega=\omega_{p,k}} = \frac{B_k}{2\pi\left(f_{pk} - f_{sk}\right)} \tag{3}$$

The parameters of the distributed-elements resonators ($Y_{c,k}$ $E_{1,k}$, $E_{2,k}$) are computed by imposing series and shunt resonances at $f_{s,k}$ and $f_{p,k}$ respectively, and by requiring the same $B_{eq}$ at $f_{p,k}$. For filter one it has:

$$E_{1,k} = \beta_{s,k} l_{1,k} = \frac{\omega_{s,k}}{v} l_{1,k} = \frac{\pi}{2}, \qquad E_{2,k} = \beta_{p,k} l_{2,k} = \frac{\omega_{p,k}}{v} l_{2,k} = \frac{\pi}{2}\left(1 - \frac{\omega_{p,k}}{\omega_{s,k}}\right)$$

$$Y_{c,k} = \frac{\omega_{p,k} B_k}{\omega_{p,k} - \omega_{s,k}}\left[\frac{\cos^2\left(\frac{\omega_{p,k}}{\omega_{s,k}} E_{1,k}\right)\sin^2\left(E_{2,k}\right)}{\left(\frac{\omega_{p,k}}{\omega_{s,k}} E_{1,k}\right)\sin^2\left(E_{2,k}\right) + E_{2,k}\cos^2\left(\frac{\omega_{p,k}}{\omega_{s,k}} E_{1,k}\right)}\right] \tag{4}$$

For filter 2:

$$E_{1,k} = \beta_{s,k} l_{1,k} = \frac{\omega_{s,k}}{v} l_{1,k} = \frac{\pi}{2}, \qquad E_{2,k} = \beta_{p,k} l_{2,k} = \frac{\omega_{p,k}}{v} l_{2,k} = \frac{\pi}{2}\left(2 - \frac{\omega_{p,k}}{\omega_{s,k}}\right)$$

$$Y_{c,k} = \frac{\omega_{p,k} B_k}{\omega_{p,k} - \omega_{s,k}}\left[\frac{\cos^2\left(\frac{\omega_{p,k}}{\omega_{s,k}} E_{1,k}\right)}{\left(\frac{\omega_{p,k}}{\omega_{s,k}} E_{1,k}\right) - \sin\left(\frac{\omega_{p,k}}{\omega_{s,k}} E_{1,k}\right)\cos\left(\frac{\omega_{p,k}}{\omega_{s,k}} E_{1,k}\right)}\right] \tag{5}$$

For what concern the admittance inverters, they could be realized through λ/4 line sections of suitable characteristic impedance; however, in order to reduce their length, we have introduced open circuited stub at the center of the line sections.

## DESIGN OF THE DIPLEXER

The topology of the diplexer is the usual one schematically depicted in fig. 3.

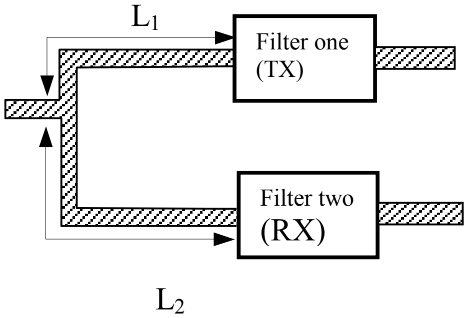


**Fig. 3**: Scheme of the microstrip diplexer

The two line sections lengths L1 and L2 are computed as usual (each filter must present an open circuit at the center passband frequency of the other filter).

In order to verify the novel diplexer configuration using single-sided filters, a prototype with the following specifications has been designed:

Filter one (TX): Passband: 850 – 950 MHz; Stopband: 1.08 – 1.18 GHz; number of resonators: 3; Return Loss: 20 dB.
Filter two (RX): Passband: 1.08 – 1.18 GHz; Stopband: 850 – 950 MHz; number of resonators: 3; Return Loss: 20 dB.

After computing the parameters of the two prototypes, a preliminary design of the microstrip filters has been obtained using eqs. 4 and 5 (the considered substrate is a Duroid with the following parameters: $\varepsilon_r$=2.55, $h$=0.8mm, $t$=35µ; the diameter of vias is 0.8 mm). This design has been then optimized with a commercial circuit simulator: the final layout of the diplexer is reported in fig 4 (the overall size is about 185 x 185 mm; the smallest line width is 0.6 mm)

The duplexer design has been then verified using a planar electromagnetic simulator; the response obtained, reported in fig 5 (input return loss and attenuation to the output ports), shows a satisfactory agreement between requirements and simulations (note that all the dimensions obtained from the design have been rounded to 0.2 mm, which is the cell size used in the simulator). For what concerns losses, their evaluation was not included in the em simulations due to the exceedingly long computation time required; however, circuit simulations show insertion losses for both filters lower than 1 dB.

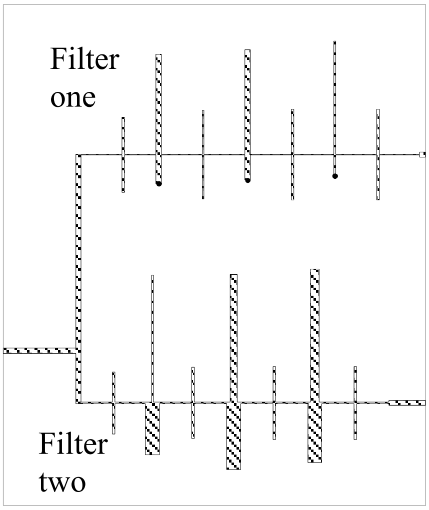


**Fig. 4**: Layout of the designed and optimized duplexer

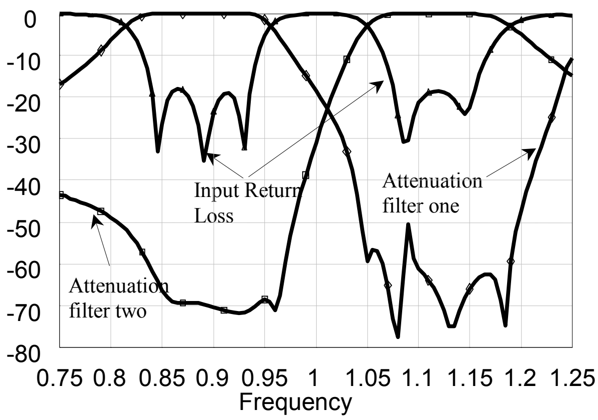


**Fig. 5** Response of the test duplexer obtained from electromagnetic simulation.

## CONCLUSIONS

A microstrip duplexer configuration using single-sided microstrip filters has been presented, which allows high selectivity with a relatively small number of resonators (so producing low losses). A design procedure for this device has been presented. A test duplexer has been designed according this procedure; electromagnetic simulations of the test device have confirmed the effectiveness of the design procedure.