Highly-Selective Filters using Low-Q Components Suitable for MMIC Implementation

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Abstract

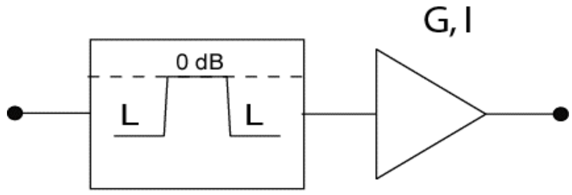
The losses and parasitics associated with MMIC technology prevent the realisation of integrated high-Q passive components, a necessity for highly-selective low-loss filters. For applications such as communications and radar, passive filters are preferred over active filters due to the latter's issues with linearity and stability. It is shown that equivalent system performance can be achieved with such a filter using a modified receiver architecture. A simplified lossy design technique using resistive-compensating networks is also presented, along with a prototype microstrip design.

Keywords — Passive filters, receivers, lossy circuits.

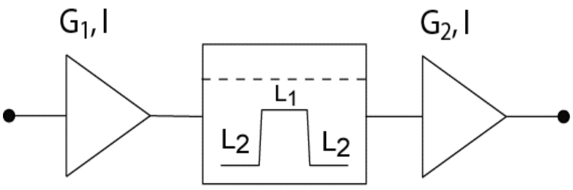
I. Introduction

The presence of loss in electrical filters results in decreased selectivity of the insertion-loss response. Classical predistortion [1] is a technique which compensates for the effects of finite-Q components at the expense of increased passband loss and poor return loss. Some recent work has provided alternative variations to predistortion which may be more applicable to present-day technology. Examples include reflection-mode filters [2], hybrid reflection-mode filters [3], and transmission filters designed using direct lossy synthesis [4].

The motivation behind this work arises from the strong need for the integration of a complete transceiver onto a single MMIC chip, which in turn is driven by the increasing miniaturisation of mobile communication devices. There is also a demand for on-chip filtering in military phased-array radar applications. The filters forming the diplexer in wireless handsets have yet to be successfully integrated, as most contemporary and future wireless standards require closely spaced transmit and receive bands with strict isolation specifications. Isolation between bands is especially important for wireless standards such as CDMA that require full-duplex transceiver operation. To achieve the necessary isolation and insertion-loss performance, filters are required to be both highly selective and low loss.



(a) Case 1: Conventional receive-path architecture using low-loss filter.
(b) Case 2: Receive-path architecture using lossy



filter.

Fig. 1. Receive path architectures using low-loss and lossy filters. Both cases can be shown to have equivalent intermodulation performance when *G*=40 dB, *G1*=10 dB, *G2*=36 dB, *L*=20 dB, *L1*=6 dB, *L2*=26 dB, and *I*=20 dBm.

As MMIC passive component Q factors are limited to around 40 at best, diplexer filters are now usually realised using relatively large off-chip components such as dielectric resonators or acoustic-wave resonators. This paper investigates an approach using highly selective but lossy passive filters in the receive path which may be implemented on MMIC, and a lossy filter design method using resistive-compensating networks.

Fig. 1a shows the receive path architecture in a typical transceiver. The receive LNA is directly proceeded by a highly-selective low-loss filter. As suggested in [2], the architecture shown in Fig. 1b can give equivalent intermodulation and noise figure performance. In this case, a highly-selective but lossy filter is used, which is proceeded by a LNA with just enough gain to maintain a decent noise figure.

II. Analysis

Out-of-band interferers with power P_{int} are assumed which create in-band third-order intermodulation products. All amplifiers are assumed to have the same output third-order intercept point, denoted by *I*. The power level of in-band intermodulation products created at the output of the Case 1 architecture is then:

$$P_{IM} = 3((P_{INT} - L) + G) - 2I \qquad (1)$$

where *L* is the stopband loss and *G* is the gain of the amplifier. The intermodulation product power level at the output created by the 1st amplifier in Case 2 is given by:

$$P_{IM1} = 3(P_{INT} + G_1) - 2I - L_1 + G_2 \qquad (2)$$

where G_1 is the gain of the 1st stage amplifier, L_1 and L_2 is the passband and stopband loss of the filter, respectively, and G_2 is the gain of the 2nd stage amplifier.
The power levels of the intermodulation products created by the 2nd stage amplifier is given by:

$$P_{IM2} = 3((P_{INT} + G_1 - L_2) + G_2) - 2I \qquad (3)$$

Setting P_{IM1} equal to P_{IM2} gives the following equation:

$$L_2 = \frac{2G_1 + L_1}{3} \qquad (4)$$

This equation can be used to select optimum design values. For the example values given in the caption of Fig. 1, the intermodulation products for both Case 1 and 2 are -100 dBm.

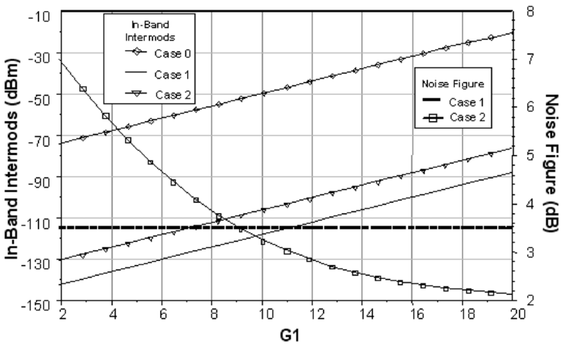


Fig. 2. Intermodulation performance and noise vs. *G1* for Case 1 (low-loss filter), Case 2 (lossy filter) and Case 0 (stand-alone amplifier).

To further investigate the performance tradeoffs, a simple system-level simulation was carried out. The system parameters are set equal to those stated above, with the addition of 1.5 dB passband loss assumed for the Case 1 filter. The filters have responses equivalent to 3rd-order elliptic functions.

Shown in Fig. 2 is a plot of intermodulation product power level and noise figure versus the gain of the first stage of Case 2, keeping the total gain of all three cases constant. Case 0 is a standalone amplifier identical to that of Case 1, included for the sake of comparison. This shows the trade-off between intermodulation performance and noise figure for Case 2. This also clearly demonstrates the merit of this approach -- with G_1 set to 10 dB the intermodulation power level of Case 2 is approximately 10 dBm more at -100 dBm, while the noise figure performance of Case 2 is slightly better.

III. Lossy Synthesis Techniques

The question now arises as to the best way to realise the highly-selective lossy filter. Reflection-mode filters provide excellent return loss, stopband rejection, and good peak Q values for a given selectivity, but require the use of a circulator. Hybrid reflection-mode filters are similar to conventional reflection filters with the one-port network split into even- and odd- mode networks, which are connected using a conventional 90-degree hybrid (Fig. 3). Transmission-mode filters designed using direct lossy synthesis give decent return loss with peak Q values equivalent to reflection-mode filters. A simplification of this technique presented in the next section gives the best performance in terms of minimum peak Q factor and even distribution of loss throughout the network.

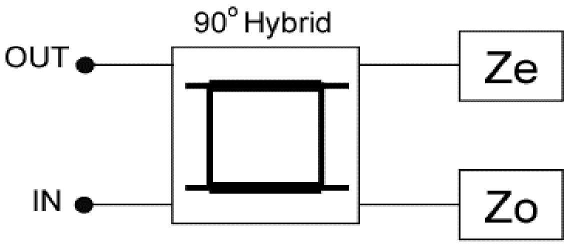


Fig. 3. Hybrid reflection-mode filter.

As shown in [4] it is possible to synthesise selective lossy filters by increasing the degree of the network.

For symmetrical networks:

$$S_{11} + S_{12} = S_e \qquad (5)$$

$$S_{11} - S_{12} = S_o \qquad (6)$$

Equations 5 and 6 give even- and odd-mode reflection coefficients which contain common factors in the numerator and denominator.

For an nth-degree all-pole network with lossy transfer function:

$$S_{12} = \frac{k}{(p+r_1)(p+r_2)\cdots(p+r_n)} \qquad (7)$$

with factor $k < 1$gives the passband loss. The general form of the reflection coefficient can then be given by:

$$S_{11} = \frac{p^n + x_{n-1}p^{n-1} + \cdots + x_1 p + x_0}{(p+r_1)(p+r_2)\cdots(p+r_n)} \qquad (8)$$

where r_i may be complex and represent the common roots occurring throughout. The coefficients x_i are obtained by substituting the values of r_i between the numerators of S_e and S_o as follows:

$$Num\{S_{11} + S_{12}\} = 0\Big|_{p=r_1,\ldots,r_{(n/2)}} \qquad (9)$$

$$Num\{S_{11} - S_{12}\} = 0\Big|_{p=r_{(n/2+1)},\ldots,r_{(n)}} \qquad (10)$$

When the overall network degree n is odd the even-mode network is of a higher degree than the respective odd-mode network. In this case the extra cancellation occurs in (10).

For networks higher than 2nd degree, choosing $k < 1$ simply adds loss to the input and output resonators. This loss can be distributed throughout the network by adding a finite complimentary pole/zero pair to the lossy transfer function:

$$S_{12} = \frac{N(p)}{D(p)}\frac{(p-\delta)}{(p+\delta)} \qquad (11)$$

This effectively creates additional transmission paths which allow for an increase in selectivity around the band edge to be produced.

This synthesis technique produces networks with selectivities identical to the desired ideal response, with superior return-loss performance compared to other loss-compensation synthesis techniques such as classical predistortion. Drawbacks include a complex synthesis procedure for higher-order filters, as all possible permutations for allocating the common roots to the even-and odd-mode equations (17) and (18) need to be considered to ensure a realisable network. For higher-order networks it is also difficult to distribute the loss evenly throughout the network.

When a lossy transfer function with the added complimentary pole/zero pair(s) is synthesised, extra transmission paths are created between 3rd-order sections. As shown by the example in [4], these extra transmission paths consist of extremely low-Q resonators, where the reactive part can be omitted with little change to the insertion-loss characteristic.

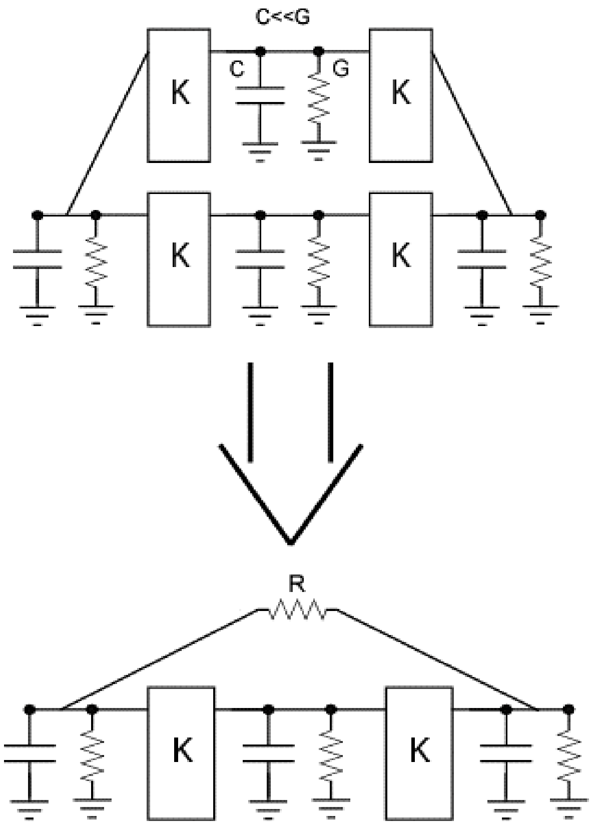


Fig. 4 Approximation of the low-Q network created by exact lossy synthesis technique -- reactive part is eliminated leaving a real impedance.

The phase shift across the third-order sections is zero at the passband edge, and 180 degrees at band centre. This technique thus effectively creates resistors within the network which dissipate a maximum amount of power at mid-band frequencies and a minimum amount at the passband edge. This is similar to a technique discussed by Belevitch [5]. Mathematically, a purely resistive transmission path adds a transmission zero on the positive real axis:

$$S_{12} = \frac{N(p)}{D(p)}(p-\delta) \qquad (12)$$

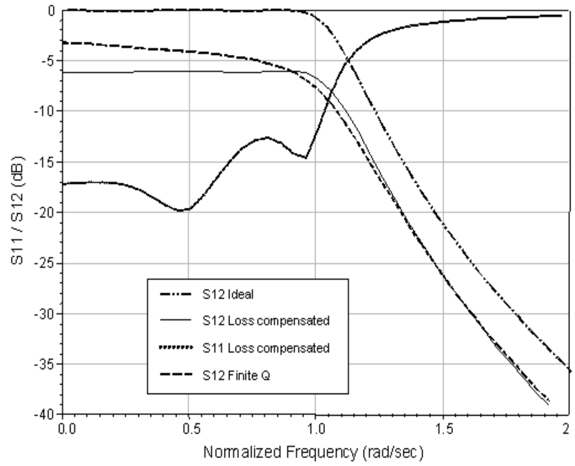


Fig. 5 Resistive-compensated 5th-degree Chebyshev filter with 6-dB passband loss designed compared to ideal response and finite-Q response.

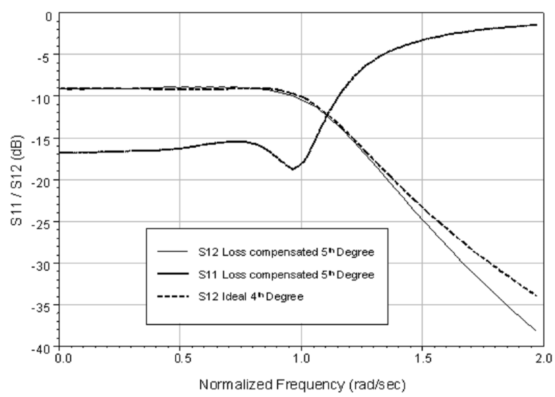


Fig. 6 Resistive-compensated 5th-degree network with response equivalent to 4th-degree Chebyshev with 9-dB passband loss.

With this approximation it becomes clear that the loss of a filter network with the topology shown in Fig. 4 may be compensated for by placing resistors between every 3rd resonator and optimising using circuit simulation software. These compensating networks may be realised with series or shunt resistors with inverters, depending on what is most convenient for physical realisation. Exact synthesis of such a network with multiple compensating resistors would be extremely difficult. This method also easily allows for loss to be distributed evenly throughout the network.

Shown in Fig. 5 is the insertion loss and return loss of a 5th degree Chebyshev filter with 6-dB passband loss, designed using resistive compensation. Also shown is the ideal lossless response, and the same using finite-Q components. As an example, a bandpass filter with 10% bandwidth at 10 GHz designed using this prototype network would give resonators with Q factors of 114. A hybrid reflection-mode filter designed to give the same response would require resonators with a peak Q of 256.

When lower Q resonators are used, reasonable selectivities can be achieved with the compromise of increased passband loss and filter degree. Shown in Fig. 6 is the response of a 5th degree prototype network with a passband response equivalent to a 4th degree Chebychev with 9-dB loss.

IV. Microstrip Prototype

Shown in Fig. 7 is 3rd-order Chebychev prototype with loss compensation. This network was used to design a microstrip filter with a centre frequency of 1.028 GHz and a bandwidth of 50 MHz. The low-Q cross-coupling network is implemented using a low-Q resonator as shown in Fig. 8. As loss in a microstrip line is inversely proportional to its width, a microstrip low-Q resonator is realised as a very thin transmission line. As a 0.5 mm linewidth gives a Q of approximately 30 on the chosen substrate, the couplings of the lossy resonator to the 1st and 3rd resonator are reduced instead of lowering the Q further and making fabrication difficult. A low-Q factor can alternatively be realised by placing a resistor in the cross-coupling resonator. As discussed in Section III, lossy cross-coupling can also be realised as a series resistor directly linking the 1st and 3rd resonators.

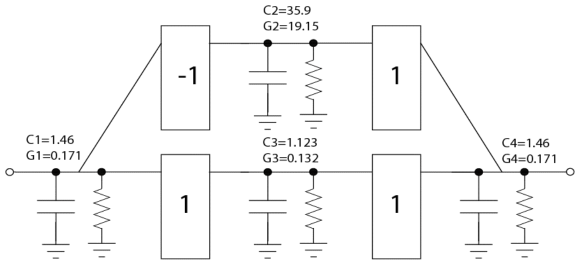


Fig. 7 Third-order Chebychev lowpass prototype with loss compensation.

This is difficult to implement in distributed-element filters as the finite electrical length of the resistive path must be taken into account. This approach is most suitable for lumped or quasi-lumped element designs.

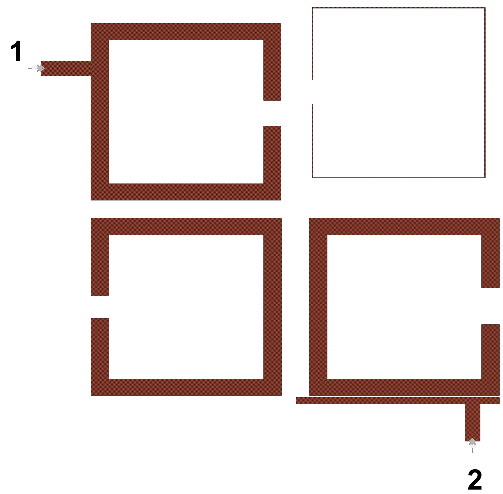


Fig. 8 ADS Momentum layout of 3rd-order loss-compensated microstrip filter. Substrate is 1.27 mm thick, ε_r=10.8.

The folded half-wavelength resonator design shown in Fig. 8. is based on work by Hong and Lancaster [6]. This topology allows for both magnetic and electric coupling (or a combination thereof) between resonators. As shown in Fig 7., two couplings opposite in sign are necessary to realise the cross-coupling network. The required negative coupling is accomplished by taking advantage of the fact that electrical and magnetic couplings are opposite in sign.

Shown in Fig. 9 is the insertion loss and return loss of the microstrip prototype as simulated in ADS Momentum. Also shown is a conventional 3rd-order Chebychev filter with an equivalent finite Q factor.

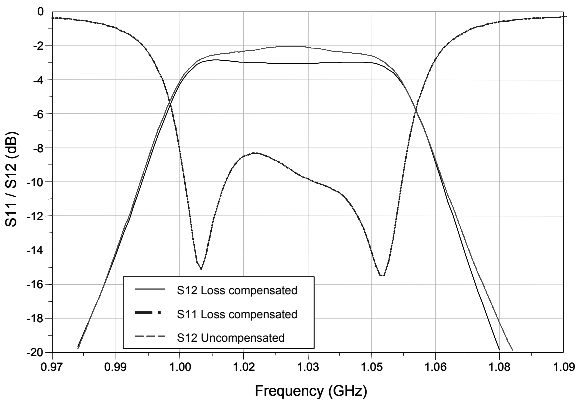


Fig. 9 EM simulation of microstrip prototype and response of an uncompensated filter with equivalent finite Q-factor.

V. Conclusion

The analysis of a receiver architecture using lossy filters with low-Q resonators within an amplifier subsystem is presented. It is clearly shown that it is possible to achieve intermodulation and noise performance comparable to a conventional architecture. A new method of filter design is presented which allows a 10% bandwidth filter to be created with a resonator Q of only 39. A prototype microstrip filter with loss compensation was successfully designed and simulated.

VI. Future Plans

Low-frequency prototype filters will be built and tested. The intention is then to proceed to MMIC filter designs, along with a full receiver system.

VII. Acknowledgements

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