Nonlinear Distortion in a 8-Pole Quasi-Elliptic Bandpass HTS Filter for CDMA System

Carlos Collado, Jordi Mateu, Oscar Menéndez, and Juan M. O'Callaghan

***Abstract*—High Temperature Superconductor materials are known to produce intermodulation and other nonlinear effects which may restrict their use in wireless communication systems. Quantifying this degradation is crucial in determining if the use of HTS filters in the receiver is beneficial. In this work we use an approach for representing the HTS filter nonlinear equivalent circuit and simulating its performance. We have tested the approach on a planar, 8-pole quasielliptic (QE) filter, showing agreement between simulations and measurements of intermodulation products (IMD). The nonlinear equivalent circuit is then used to predict the nonlinear behavior of the filter and its effects on system parameters when it is subject to broadband signals like those used in wireless communication systems.**

***Index Terms*—Equivalent circuit, HTS filters, nonlinearities, superconductor.**

I. Introduction

BANDPASS filters with High Temperature Superconductors (HTS) achieve superior sensitivity and frequency selectivity. These filters are used in base stations of cellular communication systems where they are placed at the front of a receiver to make it immune to interfering signals from other operators and services. However, there is concern about the possible degradation brought by the microwave nonlinearities of the HTS materials. Quantifying this degradation is crucial in determining if the use of the HTS filter in the receiver is beneficial.

Here we present an approach for obtaining a nonlinear equivalent circuit of the HTS filter and assessing its nonlinear performance. We have tested this approach using the microstrip eight pole QE filter described in [1].

In order to predict the feasibility of using this filter in a front-end receiver for wireless communication, where the filter is driven by wideband modulated signals, we first perform a wide range of intermodulation distortion (IMD) measurements extending the preliminary experiments described in [1]. The proposed circuit model is valid in the passband and transition bands of the filter. Also, the model gives good IMD results for different frequency spacings between input tones.

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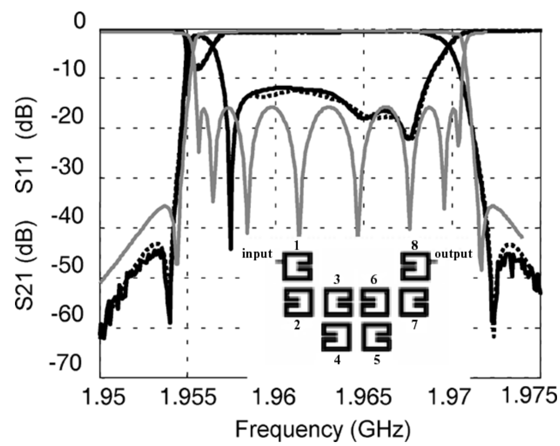

Fig. 1. Solid grey curve: synthesis response. Solid curve: measured filter response without tuning. Dotted curve: modeled filter response using the equivalent circuit.

To illustrate the usefulness of the equivalent circuit, we present a simulation example that predicts the performance of the filter when it is subject to several wideband signals: three weak in-band signals and one strong out-of-band interfering signal.

II. HTS Filter

The filter used in this work is an 8-pole QE filter consisting of eight open-loop resonators following the topology proposed in [2]. The layout of the filter is shown in the inset of Fig. 1 and details about its design can be found in [1].

The superconducting filter was fabricated using a 2 inch double sided 700 nm YBaCuO thin film deposited onto a 0.5 mm thick MgO substrate. The film was patterned using standard photolithography and wet chemical etching.

Fig. 1 shows the frequency response of the filter. Unlike [1] all experiments carried out in this work are done without tuning the filter. The measured response is outlined by the solid black curve. The filter performance exhibits a QE transfer function with a single pair of transmission zeros near the passband edges, showing high selectivity and low insertion loss. However such filter response does not match with the desired one (solid gray curve in Fig. 1). There are many effects that detune the filter response [3] by producing parasitic couplings, deviation in the synthesized couplings and changes of the resonant frequencies of the resonators.

All these effects are taken into account when determining the linear circuit elements of the equivalent circuit described in the following section.

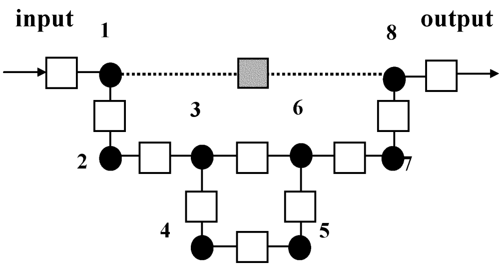

Fig. 2. Outline of the equivalent circuit model. The square white boxes represent the couplings between resonators and the input and output couplings. The black circles represent the halfwave resonators. The square grey box represents the parasitic coupling between resonators 1 and 8. Only this coupling is shown to preserve figure clarity, but other couplings are also considered.

III. EQUIVALENT CIRCUIT

The equivalent circuit, (see diagram in Fig. 2) follows the topology of the filter layout showed in the inset of Fig. 1. The black circles of Fig. 2 represent the resonators and square boxes represent couplings between resonators. White boxes account for the synthesized coupling between resonators and the input and output couplings, whereas the square gray box represents the parasitic coupling between the first and last resonator. In Fig. 2, we have only outlined one parasitic coupling, in order to preserve the clarity of the diagram. However, to fit the measured frequency response (solid line in Fig. 1), one needs to take into account many of them.

To convert this diagram to a circuit model we have taken into account that the nonlinear effects in superconductors are consequence of the nonlinear electrodynamics of the material, and thus, they occur over all the surface of the HTS, having more prominent effects in those areas where current density is highest [4]. As a consequence, resonator geometry and coupling between resonators have a strong influence on the nonlinear properties of the filter.

To facilitate the analysis without losing too much precision, we get a simple circuit model approach of the filter by modeling the couplings as impedance inverters and the open-loop resonators as straight half-wave microstrip lines [1]. This is a common approach to model transmission line filters [5] but it is not so evident that the current distribution along a straight microstrip resonator agrees with that in its corresponding open-loop due to wrapping field effects at its bends. However we think that this approach is reasonably good since the most important contribution to the nonlinear effects takes place where the current density is largest, that is, in the middle of the open-loop structure. This would make the nonlinear effects in a straight halfwave microstrip resonator similar to those in the open-loop resonator.

Note also that since the filter is a weakly nonlinear device it is not necessary to consider the nonlinear effects to adjust the parameters which fit the linear frequency response. Therefore for setting the parameters which define the whole equivalent circuit, we may split our analysis in a linear fitting plus a nonlinear fitting.

A. Linear Response Fitting

Prior to considering the nonlinearities of the HTS material, we tuned the couplings and resonant frequencies of the equivalent circuit to fit the low power measured scattering parameters of the prototype. We have used commercial tools [6] to perform a gradient optimization and we found that some of the designed couplings changed up to 30%, the resonant frequencies up to 0.15% and the parasitic couplings were about 100 times lower than the ones synthesized. The resulting matrix coupling is shown in the equation at the bottom of the page where the elements of the diagonal account the shift in frequency of the corresponding resonator. The external quality factor at input and output port result 197 and 174, respectively.

Fig. 1, dotted line, shows the result of the adjustment. Note the excellent agreement between measurements and circuit model simulations.

B. Nonlinear Circuit Model

Now, as commonly done [1], [4] and [7], the halfwave resonators may be modeled by concatenating many $RLCG$ elementals cells, where the inductance and resistance per unit of length depend on the total current flowing through the cross-section of the line:

$$L(i) = L_0 + \Delta L(i); \quad R(i) = R_0 + \Delta R(i) \tag{1}$$

The linear parameters -R, L, C, G- may be found by using commercial software, whereas the nonlinear parameters $\Delta L(i)$ and $\Delta R(i)$ have to be adjusted to fit the nonlinear effects produced by the filter.

For simplicity we will assume that all the resonators can be modeled with the same parameters $\Delta L(i)$ and $\Delta R(i)$. This approach implies the assumption that the profile of the current density distribution in a cross-section of the resonators is the same that would be in a isolated structure. In this case, this simplification is reasonable since the coupling between resonators are

$$10^{-3}\begin{bmatrix} 3.49 & 5.3 & -0.17 & 0 & 0 & 0.08 & 0 & 0 \\ 5.3 & 2.59 & 3.8 & -0.2 & 0 & 0 & 0 & 0 \\ 0.17 & 3.8 & 2.32 & 0.16 & -0.16 & -1.7 & 0 & 0.04 \\ 0 & -0.2 & 0.16 & 3.43 & 5.8 & -0.12 & 0 & 0 \\ 0 & 0 & -0.16 & 5.8 & 2.0789 & 0.17 & -0.19 & 0 \\ 0.08 & 0 & -1.7 & -0.12 & 0.17 & 1.59 & 4.6 & 0.15 \\ 0 & 0 & 0 & 0 & -0.19 & 4.6 & -0.73 & 6.8 \\ 0 & 0 & 0.04 & 0 & 0 & 0.15 & 6.8 & -2.96 \end{bmatrix}$$

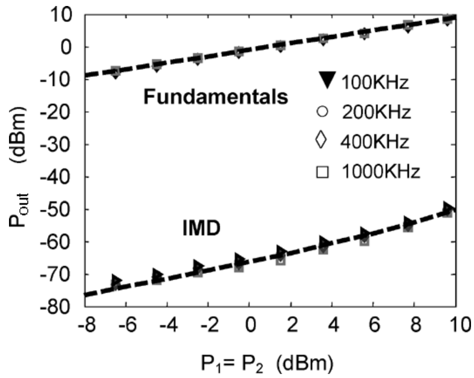

Fig. 3. Results of measurements and simulations. Marks indicate the measured power of IMD (2f2-f1) and the power at the output of the filter at fundamental frequencies as a function of the source power when the power of both tones are increased simultaneously.

very weak [5]—the filter is narrow band—so the fringing field between different resonators is not too strong.

To analyze the whole equivalent circuit, we split the circuit in its linear part (containing the linear parameters of the transmission line and the impedance inverters), and its nonlinear part, which contains the nonlinear elements that model $\Delta L(i)$ and $\Delta R(i)$. Then we apply numerical techniques based on Harmonic Balance (HB). We have used home made HB routines [7] and commercial tools [6], both performing the same simulated results.

IV. Nonlinear Parameter Extraction

The nonlinear distributed parameters $\Delta L(i)$ and $\Delta R(i)$ should be able to model the nonlinear effects when the filter is driven by a wideband modulated signal. That is, a signal with many frequency components having various frequency spacings between each component and placed through the whole band of the filter, even in the transition band.

We propose two sets of two-tone IMD experiments to extract and check $\Delta L(i)$ and $\Delta R(i)$.

The first set considers four different experiments. In all of them the two input tones are centered at 1.96 GHz (in-band of the filter), and spaced (Δf) 100 KHz, 200 KHz, 400 KHz and 1000 KHz, respectively. Both tones are kept balanced (i.e., $\mathrm{P}_1 = \mathrm{P}_2$) and the power is swept from -8 dBm to 10 dBm. The results of these experiments are depicted at Fig. 3. They show that the nonlinear effects do not depend on the frequency spacing between input tones, as is assumed by our circuit model.

We have used this set of experiments to extract the nonlinear parameters to fit the measurements of Fig. 3. To do that we have applied an iterative procedure based on HB [6], [7].

Since scalar IMD measurements do not provide information about the phase of the outcoming IMD, we have assumed that the nonlinear effects are dominated by the nonlinear reactive term, i.e. $\omega\Delta L(i) \gg \Delta R(i)$, as it would be expected for a HTS microstrip transmission lines with intrinsic nonlinearities [4]. Under this hypothesis, the function $\Delta L(i) = 7.9 \cdot 10^{-12}|\mathrm{i}|^{0.2} + 5 \cdot 10^{-12}|\mathrm{i}|^2\ \mathrm{H/m}$ fits the IMD measurements as outlined in Fig. 3, dashed line.

The second set of measurements considers three different experiments sweeping the two input tones through the frequency band of the filter, from 1.95 GHz to 1.975 GHz. The frequency spacing and the input power are kept constant during each experiment:

- Experiment 1: $\Delta\mathrm{f} = 100\ \mathrm{KHz}; \mathrm{P}_1 = \mathrm{P}_2 = 9.5\ \mathrm{dBm}$.
- Experiment 2: $\Delta\mathrm{f} = 400\ \mathrm{KHz}; \mathrm{P}_1 = \mathrm{P}_2 = 9.5\ \mathrm{dBm}$.
- Experiment 3: $\Delta\mathrm{f} = 400\ \mathrm{KHz}; \mathrm{P}_1 = \mathrm{P}_2 = 0.5\ \mathrm{dBm}$.

The results of these experiments (Exp. 1, Exp. 2 and Exp. 3) are shown in Fig. 4(a), 4(b), and 4(c), respectively. The squares and crosses indicate the fundamental and IMD at upper side band, respectively.

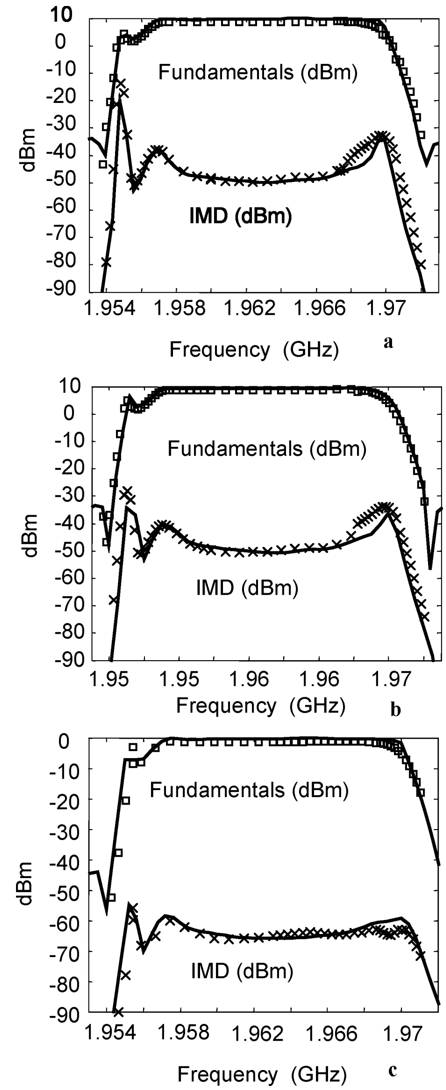

Fig. 4. Measurements and simulations. Squares and crosses represent the measured fundamental and IMD at upper side band, respectively. Lines represent the results of the simulation. a), b) and c) correspond to Experiment 1, Experiment 2 and Experiment 3, respectively.

In Fig. 4 the dashed lines represent the simulated results obtained from the equivalent circuit that fits the measurements of Fig. 3. We see that the fundamentals and IMD measurements in Fig. 4 are simulated correctly. In other words, the nonlinear equivalent circuit allows predicting the nonlinear behavior in the whole band of the filter and for different frequency spacing between input tones. Note also that this is true even at the edges of the passband where the circulating power is higher and in the transition band where the power of the fundamental tones

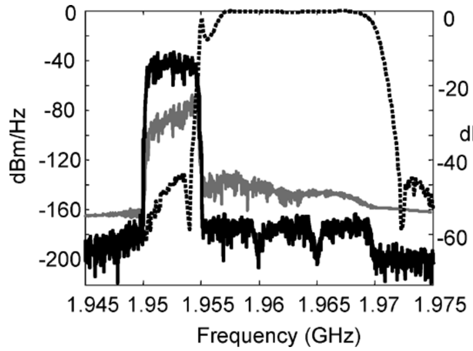

Fig. 5. Interference scenario in a UMTS uplink with the pre-select 8-pole QE filter of Fig. 1. Dotted line represents the transmission response of the filter; the units are in the right axes. Black trace indicates input power spectral density. Grey trace indicates output power spectral density.

might be very unbalanced. As mentioned at the introduction this is very relevant since, in a receiver, the interfering signals may be close to the edges of the passband.

V. Example

In this example we show how the circuit model can predict the nonlinear effects in a dense electromagnetic environment, like those expected for UMTS communications systems [8]. To do so we analyze interferences caused by an adjacent band signal, when the leakage power at the channel of interest passes through an HTS filter.

The scenario for this example is shown in Fig. 5. The input signal (Fig. 5, black trace) consists of three received signals (-120 dBm) at the band of interest and a fourth one at an adjacent channel at 10 dBm. From this scenario we calculate the increase of signal power in the band of interest due to the spectral regrowth of the adjacent band signal when it goes through a HTS bandpass filter. As a consequence of the spectral regrowth the system performance would undergo a sharp degradation. Fig. 5, shows the result of this simulation, where the grey trace shows the spectrum signal going out of the HTS filter. In the filter passband, this trace (filter output power) is significantly higher than the black solid trace (filter input power), indicating that, at these frequencies, power comes predominantly from the interfering signal and not from the signals that the receiver is supposed to pick up.

Note that the simulation presented in this section only attempts to give a preliminary study of the effects that the nonlinearities may produce when high power signals are located at adjacent channels. Many different interference scenarios may arise in wireless base station receivers depending, among other factors, on the technical specifications of the communication system used. The equivalent circuit developed here may be useful in assessing the performance of an HTS receiver in many of these scenarios.

VI. Conclusion

We have experimentally verified that the circuit model of the HTS bandpass QE filter and the numerical techniques based on HB are valid to both characterize and predict the IMD of an 8-pole QE planar HTS filter at the passband and transition band and for several spacings between input tones. This circuit model is also useful to predict the nonlinear performance of real systems subject to broadband signals whose spectrums might partially leak to adjacent channels. This might happen in wireless base station receivers where the interference signals can appear in or out of the reception band of base station receivers.

References

[1] J. Mateu, C. Collado, O. Menendez, and J. M. O'Callaghan, "Nonlinear performance characterization in an eight-pole quasielliptic bandpass filter," *Supercond. Sci. Technol.*, vol. 17, pp. S359–S362, 2004.

[2] J.-S. Hong and M. J. Lancaster, "Design of highly selective microstrip bandpass filters with a single pair of attenuation poles at finite frequencies," *IEEE Trans. Microw. Theory Tech.*, vol. 48, no. 7, Jul. 2000.

[3] J.-S. Hong, M. J. Lancaster, D. Jedamzik, and R. B. Greed, "On the development of superconducting microstrip filter for mobile communication applications," *IEEE Trans. Microw. Theory Tech.*, vol. 47, no. 9, Sep. 1999.

[4] T. Dahm and D. J. Scalapino, "Theory of intermodulation in a superconducting microstrip resonator," *J. Appl. Phys.*, vol. 81, pp. 2002–2009, 1997.

[5] G. L. Matthaei, L. Young, and E. Jones, *Microwave Filters, Impedance-Matching Networks and Coupling Structures*. New York: McGraw-Hill.

[6] "Agilent Eesoft EDA Product Documentation," ADS.

[7] C. Collado, J. Mateu, and J. O'Callaghan, "Analysis and simulation of the effects of distributed nonlinearities in microwave superconducting devices," *IEEE Trans. Appl. Supercond.*, to be published.

[8] H. Holma and A. Tostala, *WCDMA for UMTS Radio Access for Third Generation Mobile Communications*. New York: Wiley, 2000.