Reduction of Electromagnetic Interference Susceptibility in Small Signal Analog Circuits using Complementary Split-Ring Resonators

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Abstract—Low frequency analog and digital electronic circuits are susceptible to electromagnetic interference in the radiofrequency (RF) range. This disturbance is produced when the coupled RF signal is rectified by the non-linear behavior of the semiconductors used in the small signal analog input stage of the electronic system. Circuits based on operational amplifiers are usually employed for such input stages. These circuits present an AM demodulation produced by nonlinearity of internal transistors. Such a phenomenon generates demodulated signals in the low-frequency range. In this paper, this effect is suppressed by combining the conventional printed circuit board layout with complementary split ring resonators (CSRRs). CSRRs are constitutive elements for the synthesis of metamaterials with negative effective permittivity which are mainly excited to the host line by means of electric coupling. Electromagnetic simulations and experimental results show an effective rejection of the undesired RF demodulation effect with no-extra cost in terms of the device area or manufacturing process.

Index Terms—complementary split-ring resonators, electromagnetic interference, operational amplifier.

I. INTRODUCTION

Electronic systems are disturbed by high frequency radiofrequency (RF) electromagnetic interference (EMI) noise, in which the amplitudes change randomly in time. This out of frequency band electromagnetic compatibility (EMC) issue is produced when the RF signal is coupled to sensors and cabling of the system, conducting the EMI to signal-conditioning, and causes errors or malfunction due to the rectification by the non-linear behavior of the semiconductors used in the electronic systems. In fact, signal conditioners or transducers are typically based on operational amplifiers (OPAMPs) operating in the linear mode. The nonlinearity of amplifier transistors causes an AM demodulation of RF signals (Fig. 1) [1]. When this random AM modulated RF noise reaches the inputs of an OPAMP, three significant effects are produced at the output [2]: First, a demodulated low frequency signal, at the modulating frequency, $f_m$, whose amplitude is proportional to the amplitude of the random modulating noise. These lower frequency components are derived from the amplitude envelope of the original RF signal. Second, a DC signal due to the rectification effect (caused by non-linearity). This effect is provoked by the nonzero average value of signal after the nonlinearity. Finally, several RF components due to the original RF carrier signal, centered at carrier frequency, $f_c$, and its harmonics, containing also lateral bands at $f_c ± mf_m$ due to AM demodulation. These three different components are illustrated in Fig. 1. The components of the last group, usually present certain attenuation, since they are located at the rejection band of amplifiers. However, DC and low-frequency components fall into the pass band of OPAMPs and contribute to a worsening of the signal/noise ratio, i.e., the interference signal cannot be distinguished from useful signal.

To solve this problem, effective filtering can be implemented at the input stages of the system (i.e., input connectors) in order to avoid the non-desired rectification. In order to overcome the limitations of standard EMI filters in terms of signal delay, printed circuit board (PCB) area, and cost, new filtering techniques are required. Several solutions based on increasing design robustness [3] or layout shielding improvements [4-5] have been proposed. Solutions based on design robustness, can typically involve higher number of electronic components, whereas layout shielding can involve extra metal layers or area. Recently, specific multilayer layout techniques based on electromagnetic band gaps (EBGs) have appeared, in order to filter EMI in several applications with good performance [6-10]. EBGs belong to a broad family of artificial media with electromagnetic properties generally not found in nature, called metamaterials. EBGs correspond to metamaterials whose period, $p$, is comparable to the signal wavelength, $\lambda$. A second type of metamaterials, the so-called effective media (i.e., metamaterials satisfying the condition $\lambda << p$) are used in the present work. These metamaterials are divided into single negative media (SNG, i.e., effective media with negative magnetic permeability, $\mu < 0$, or electric permittivity, $\varepsilon < 0$) and double negative media (DNG, with simultaneous $\mu < 0$ and $\varepsilon < 0$), also called left-handed materials [11-13]. Physically, SNG can be implemented by using so-called split-ring resonators (SRRs) [14] and their dual
counterparts, the complementary split-ring resonators (CSRRs) [15]. Recently, CSRRs have been used as EMI reduction structures with a good behavior in terms of rejection band and signal integrity [16].

In this paper, a filter developed by means of effective media metamaterials based on CSRRs is used to reduce RF interference in a conventional differential amplifier based on an OPAMP. These kind of sub-wavelength resonators are able to achieve a good compromise between stop-band rejection level, cost, and PCB area consumption by taking into account that the design of the filter must cover the frequency separation between interference and useful signals. Specifically, the rejection frequency band has been designed at 2.4-GHz (Industrial, Scientific and Medical, ISM), due to the current multiple radiofrequency interference sources in this frequency range. The considered modulated frequency corresponds to 10 kHz. The paper is organized as follows: Section II describes the complementary split ring resonators geometry and its frequency behavior. Section III introduces the proposed layout (i.e., metamaterial filter combined with the differential amplifier) and the performed electromagnetic simulations. In Section IV, experimental results are discussed. Specifically, the filter attenuation, the demodulated signal reduction and the impact in terms of offset and modulation index have been analyzed. Finally, Section V summarizes the main conclusions of this work.

II. COMPLEMENTARY SPLIT RING RESONATORS

The sub-wavelength resonators used in this work in order to mitigate EMI at RF band are the CSRRs (the dual particle of the SRRs). Fig. 2 shows the basic topologies of both particles and their equivalent circuit models coupled to transmission lines. Essentially, SRRs (Fig. 2a) consist of a pair of metal rings etched on a dielectric slab with apertures in opposite sides which can be mainly excited by means of a parallel magnetic-field along its axis. If an array of SRRs is located close to a transmission line (i.e., a microstrip line), current loops can be induced in the rings and, at resonance, they reflect the incident host signal. Therefore they behave as an LC tank (described by $L_S$ and $C_S$ parameters) magnetically coupled (by using a mutual inductance, $M$) to the host line (defined by $L$ and $C$, Fig. 2b) [17]. From duality arguments based on an approximation of the Babinet’s principle for dielectric boards, it is demonstrated that the CSRRs (Fig. 2c), the negative image of SRRs, roughly behave as their dual counterparts (i.e., their resonance frequency is approximately equal to that of the corresponding equivalent SRRs). Therefore, CSRRs are mainly excited by means of an axial time-varying electric field (magnetic field contribution is significantly minor) [18]. Since the main coupling contribution of these particles is electric, it can be basically described by means of a coupling capacitance ($C'$). Therefore, an efficient way to achieve stop-band frequency responses is to etch CSRRs in the ground plane of a structure such as a microstrip line (or similar) or even in the conductor strip. The synthesis of the filter is performed by using the LC equivalent model of CSRRs (described by $L_C$ and $C_C$ parameters) coupled by means of a capacitor (whose value, $C'$, is slightly different to the corresponding to the capacitance of the transmission line) with the host line (Fig. 2d). According to the equivalent circuit model, the zero transmission frequency, $f_Z$, of each CSRR coupled to the victim line is given by,

$$f_Z = \frac{1}{2\pi\sqrt{L_C \cdot C_C + C}}$$  \hspace{1cm} (1)

The main advantages of these sub-wavelength particles are their low implementation cost (no extra components or layers are required and they are implemented by means of conventional etching techniques) and the quasi-non-area consumption, since they are located in the ground plane. Moreover, key to the application of arrays of CSRRs (or SRRs and derived particles) to the synthesis of SNG is their small electrical size (periods involved are lower than $\lambda$) and, therefore, it implies an intrinsic compactness factor.

III. PROPOSED STRUCTURE

The test circuit consists of a conventional differential amplifier, as depicted in Fig. 3. The configuration corresponds to a basic instrumentation amplifier based on 4 resistors and one OPAMP. The resistors are chosen equal to provide an output voltage, $V_{Io}$, approximately equal to the input voltage difference, $V_i = V_{IN+} - V_{IN-}$. In order to avoid RF EMI reaching the OPAMP input, a 4-CSRR filter has been designed and combined in the final prototype. Fig. 4 shows the PCB designed layout. As can be observed, two CSSR arrays have been located in the ground plane, underneath the input transmission lines carrying the signal of interest to the differential input, before it reaches the non linear part of the circuit. Notice that this distribution implies effective non-area consumption with respect to a conventional 1-layer design. Moreover, no extra lumped circuitry is needed and no series stage filter is required. Other CSRRs topologies, such as rectangular shape, can be used in order to compact the effective area of CSRRs with respect to the metal strip. Obviously, a trade-off between rejection level and number of stages appears. The design has been carried out by means of the Agilent ADS and Momentum software. The resonance frequency of the CSRRs has been designed in the vicinity of 2.4 GHz in order to prevent susceptibility in the ISM band. The dimensions of the CSRRs have been calculated by developing full-wave electromagnetic simulations of single CSRRs coupled to a microstrip line section and a sweeping algorithm. The notch peak in the transmission frequency response reveals the presence of the CSRR. In order to compute the CSRRs dimensions, we determine $a$ and $b$ parameters to the minimum resolution of the used drilling machine (200 $\mu$m). Concerning the initial computed frequency, it is possible to determine the dimensions of a circular SRR by following the algorithm detailed in [17]. According to theory the resonance frequency of the CSRR is approximately equal to the corresponding to its dual counterpart (SRR). Therefore, by considering the equivalent area of the circular CSRR we can compute an initial value for
a square CSRR (by determining \( c \) and \( d \) dimensions). In fact, the involved single resonator dimensions have been slightly detuned in order to achieve a wider stop-band bandwidth (i.e., single CSRRs with close resonance frequencies). Therefore, a final simulation step based on a multiple tuning procedure has been developed. EMI filter simulations have been performed between external input ports (P1 and P3, Fig. 4) and internal ports (P2 and P4, Fig. 4) before OPAMP and circuitry stage (output port corresponds to P5, Fig. 4). CSRRs dimensions correspond to \( a=b=5.2 \) mm for each stage, \( c_1=d_1=7.0 \) mm, \( c_2=d_2=6.9 \) mm, \( c_3=d_3=6.8 \) mm, where \( i=1,2,3,4 \) corresponds to the stage number. The inter-resonator distance is 0.5 mm. Electromagnetic simulations have been performed by considering a \( MC \ 100 \ FR4 \) substrate which will be described in next section.

Fig. 5a shows a detailed layout of the proposed 4-stage filter, whereas Fig. 5b depicts the corresponding full lumped circuit model by taking into account the electrical model illustrated in Fig. 2d. The inter-resonator coupling between adjacent CSRRs has been modelled by means of capacitances \( C_s, C_a, C_f \). The extraction of the different parameters is directly related to the design methodology in order to implement these structures from some given specifications (typically the expected notch frequency, rejection level and the actual dimensions of the involved transmission line). The design methodology is summarized in Fig. 6 and consists of a sequence of steps described below. Since the model consists of several variables, the extraction of the parameters must be done sequentially.

First, a single microstrip line is considered. The per-section inductance and capacitance of the line, \( L \) and \( C \), can be calculated from a transmission line calculator. The system is assumed lossless for simplicity (this is a reasonable assumption as a first step approximation). Second, the coupled RF interference frequency determines the dimensions of the CSRRs (their resonance frequency is determined by the \( L_cC_c \) tank.). Basically, the CSRRs are electrically coupled to each microstrip line by means of the distributed capacitance, \( C' \). By analyzing a single CSRR coupled to the line (i.e., no inter-resonator coupling), two resonance frequencies arise: the frequency that nulls the shunt impedance (i.e., transmission zero frequency), given by (1) and the resonance frequency of the CSRR given by:

\[
f_o = \frac{1}{2\pi \sqrt{L_cC_c}} . \tag{2}
\]

On the other hand, according to [19], the periodic structure under study by considering \( C'A=CM=0 \), satisfies (3).

\[
\cos \varphi = 1 + \frac{Z_s(j\omega)}{Z_p(j\omega)} . \tag{3}
\]

This equation allows the analysis of periodic circuits based on the cell depicted in Fig. 2(d) with the help of the dispersion relation and Bloch impedance. \( \varphi \) denotes the phase shift of the elemental cell, and \( Z_p \) and \( Z_s \) correspond to the shunt and series impedance, respectively, of the T-circuit model. By using an electromagnetic simulator, it is possible to obtain the phase of the insertion loss transmittance, \( \psi_{SL}(\omega) \). Since \( Z_p \) and \( Z_s \) are known expressions, by obtaining a given \( \psi_{SL} \) at its corresponding frequency, \( \omega_0 \), it is possible to solve the equation system (1-3) in order to determine each \( L_c, C_c \) and \( C' \). Concerning the rejection level, the maximum number of subwavelength resonators guarantees the highest rejection level. Obviously, this fact implies a tradeoff with the available area and in case of close located CSRRs, the inter-resonator coupling must be taken into account in the model. Inter-resonator coupling is modeled by a mutual capacitance in a similar way to [20]. Therefore, by considering a second CSRR close to the original one, the inter-resonator coupling mutual capacitance, \( C_M \), can be extracted as a fitting parameter. Table I summarises the values of the parameters included in the equivalent model depicted in Fig. 5b.

<table>
<thead>
<tr>
<th>Component</th>
<th>Value</th>
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<tr>
<td>L1, L2, L3, L4, L5</td>
<td>1.5907 nH</td>
</tr>
<tr>
<td>L6, L7, L8</td>
<td>1.5907 nH</td>
</tr>
<tr>
<td>Lc1</td>
<td>0.116 nH</td>
</tr>
<tr>
<td>Lc2</td>
<td>0.112 nH</td>
</tr>
<tr>
<td>Lc3</td>
<td>0.122 nH</td>
</tr>
<tr>
<td>Lc4</td>
<td>0.147 nH</td>
</tr>
<tr>
<td>C1</td>
<td>3.887 pF</td>
</tr>
<tr>
<td>C2</td>
<td>4 pF</td>
</tr>
<tr>
<td>C3</td>
<td>3.846 pF</td>
</tr>
<tr>
<td>C4</td>
<td>2.188 pF</td>
</tr>
<tr>
<td>C5, C6, C7</td>
<td>0.05 pF</td>
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<tr>
<td>Cc1</td>
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</tr>
<tr>
<td>Cc2</td>
<td>33.136 pF</td>
</tr>
<tr>
<td>Cc3</td>
<td>33.130 pF</td>
</tr>
<tr>
<td>Cc4</td>
<td>33.798 pF</td>
</tr>
<tr>
<td>Term1, Term2</td>
<td>50 ( \Omega )</td>
</tr>
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</table>

Fig. 7a illustrates the electromagnetic simulation corresponding to the discussed filter in comparison with the single frequency response of each CSRR. It is observed a significant enhancement in the bandwidth and rejection level by including the 4-stages together due to the coupling between them. Moreover, Fig. 7b shows the detailed fitting between the proposed equivalent circuit frequency response and the EM simulation. Fig. 8a depicts the CSRRs filter electromagnetic response by considering the overall implementation. As can be observed, a significant rejection level is obtained (\( S21<40 \) dB) at the frequency band of interest with an enhanced bandwidth with respect to a peak notch response. The enlargement of the operational bandwidth concerning the case of CSRRs with the same dimensions depends on the number of used stages. Obviously, the rejection level can be increased by adding extra CSRRs stages and electric coupling, but a trade-off exists with respect to the dimensions of input transmission lines. Regarding the impact of PCB tolerances on filtering characteristics, the dielectric constant tolerance \( (\varepsilon_r=4.5\pm0.1) \) on the central frequency and rejection level corresponds to a variation of 0.8% and 1 dB, respectively, whereas the corresponding to the drilling machine resolution (\( \pm0.01 \) mm, typical for the PCB manufacturing) corresponds...
higher in CSRR filtered prototype, since the best offset level for the conventional device is approximately -4 mV. Fig. 14 shows the impact of the modulation index on the output level, \( V_0 \), at frequency of interest (10 kHz) for both prototypes. By fixing \( A_C = -5 \text{ dBm} \), the \( m \) sweep again reveals a constant behavior for the CSRRs implementation (\( V_0 = -90 \text{ dBm} \)) whereas the conventional one dramatically increases the output level of the disturbance. \( V_0 \) has also been tested by fixing \( m = 50\% \) and sweeping the EMI voltage level, \( A_C \) (Fig. 15). In this case the differences are produced from \( A_C > -20 \text{ dBm} \). Obviously, at high interference voltage levels, the total level rejection of the filter is highlighted (45 dB) in good agreement with the simulation results (Fig. 8a).

Therefore, it is demonstrated that, for high frequency, CSRRs can be a good solution to reduce EMI susceptibility in circuits based on OPAMPs. In fact, this method implies quasi-non-cost, since CSRRs can be implemented by conventional etching process in the ground plane, and no extra area or electronic devices are required in comparison with conventional EMI filters or shielding solutions.

V. CONCLUSIONS

In summary, it has been demonstrated that EMI effects due to random RF noise signals reaching the OPAMP input circuits, which present an inherent non-linear behavior, can be significantly reduced by means of filters based on CSRR. Basically, the demodulated low frequency signal is attenuated by these conveniently tuned sub-wavelength resonators. A significant increase of device RF immunity as well as DC offset minimization has been demonstrated, both by simulation and experimentally. In fact, ground loaded CSRR transmission lines can be a compact-low-cost method in order to significantly decrease the PCB RF coupling interference at the ISM band. Simulated and experimental results show a 45 dB coupling reduction at 2.4 GHz. The authors are confident about the application of these structures for EMI reduction in planar electronic circuits operating at high frequencies/data-transmission rates.

REFERENCES


to 1% and 1.5 dB, respectively. Fig. 8b shows the group delay of the proposed filter in terms of frequency. As can be observed, the maximum value corresponds to 12 ns at a frequency close to 2.3 GHz.

IV. EXPERIMENTAL RESULTS

Two experimental prototypes have been fabricated and tested. Both present the same top level metal layer layout and are differentiated by a conventional ground plane in the first case and a ground plane disturbed by etched CSRR arrays in the second. Fig. 9 shows the prototype setup consisting of a 3-port 4-stage CSRR loaded parallel transmission lines, which have been designed to obtain a stop band filter around 2.4 GHz. The OPAMP used in the prototypes is a UA741CD, supplied by two voltage regulators, a MC78M15BDTG (15V) and a MC79M15CDTG (-15V). All resistor values are 1 kΩ. Decoupling capacitors (100 nF and 10 μF) have been also used for the supply lines. The substrate corresponds to the commercial MC 100 FR4 (dielectric constant \( \varepsilon_r \approx 4.5 \), thickness \( h \approx 1.53 \text{ mm} \)), whereas metal layers correspond to copper (thickness, \( t \approx 35 \mu \text{ m} \)). Specifically, 50 Ω microstrip access lines are considered with dimensions: width \( W = 2.84 \text{ mm} \), length \( l = 4 \text{ cm} \), and separation \( s = 6.55 \text{ mm} \). The CSRRs dimensions (detailed in the previous section) have been etched by means of a LPKF S62 drilling machine. The total circuit area is 6.9x4.8 cm².

The test prototypes have been measured by means of a Rohde & Schwarz ESPI spectrum analyzer and Tektronix TDS 5104 oscilloscope. The test setup consists of emulating the RF EMI coupling by means of a signal modulated in AM (whose carrier frequency is \( f_c \approx 2.4 \text{ GHz} \)) with a low frequency sinus signal (whose modulated frequency has been chosen as \( f_m = 10 \text{ kHz} \)). The direct power injection carrier amplitude, \( A_C \), corresponds to \( A_C = -10 \text{ dBm} \) and the modulation index, \( m \), is \( m = 50\% \). Fig. 10 illustrates the described experimental setup. Fig. 11 shows the comparison between the circuit output modulated spectrum including or not the CSRRs filter. As can be seen, the intrinsic frequency response of the OPAMP at high frequency implies an attenuation of -25 dB for the conventional structure. By using the CSRRs implementation the attenuation is improved significantly, and a rejection level higher than 20 dB is achieved. The experimental output spectrum at the low operation frequency (10 kHz) reveals a disturbance on the order of 20 dB for the conventional case (Fig. 12), which is produced by the non-linear behavior of OPAMP, as explained in Section I. However, the CSRRs prototype completely removes this EMI effect, since the impact of resonators notably filters the undesired noise signal at this frequency. In order to evaluate the effectiveness of the proposed implementation, several parameters have been tested. Fig. 13 illustrates the experimental DC offset voltage in terms of interference amplitude. When the noise signal is injected, a significant increase of the offset with EMI amplitudes higher than -5 dBm is observed in the non-filtered circuit, whereas in the same conditions, the offset of the prototype equipped with the CSRR filter remains constant (-1.19 mV). Notice also that the offset level is almost 4 times


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Figure Captions

Fig. 1. Response of linear ICs to EMI at the input. (a) EMI source. (b) OPAMP response.

Fig. 2. (a) Topology of the SRR with its relevant dimensions, and (b) lumped-element equivalent circuit for the unit-cell of the SRR coupled to a transmission line. (c) Topology of the CSRR with its relevant dimensions. Metallization zones are depicted in grey. (d) Lumped-element equivalent circuit of the CSRR coupled to a transmission line.

Fig. 3. Schematic of the test circuit.

Fig. 4. Designed layout. The black layer corresponds to the up side metal, whereas the grey layer corresponds to ground, where CSRRs have been etched.

Fig. 5. (a) Layout detail of the implemented filter. (b) Equivalent circuit model.

Fig. 6. Design methodology synthesis.

Fig. 7. (a) Electromagnetic simulation of the proposed 4-stage filter in comparison with the single frequency response of each CSRR. (b) Equivalent circuit frequency response (black) and electromagnetic simulation (green).

Fig. 8. (a) Full-wave electromagnetic simulation Fabricated 4-port prototype device. Insertion losses, |S21|, and return losses, |S11|, are depicted. (b) Group delay of the proposed filter vs frequency.

Fig. 9. Fabricated 3-port prototype device. (a) Top side. (b) Bottom side including CSRRs.

Fig. 10. Experimental setup.

Fig. 11. Experimental AM modulation output device spectrum for conventional and CSRRs prototypes.

Fig. 12. Experimental demodulation output device spectrum for conventional and CSRRs prototypes.

Fig. 13. Experimental DC offset vs. RF interference amplitude for conventional and CSRRs prototypes (m=50%).

Fig. 14. Experimental output level vs. modulation index at the interest operation frequency of both prototypes (A_C=-5 dBm).

Fig. 15. Experimental output level vs. EMI voltage level at the interest operation frequency of both prototypes (m=50%).
Fig. 1
Fig. 2
Fig. 3
Fig. 5

(a)

(b)
### Fig. 6

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<th>Specification</th>
<th>EM Simulation</th>
<th>Extracted parameters</th>
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<tr>
<td><strong>Transmission Lines’ dimensions</strong></td>
<td>single TL</td>
<td>L, C</td>
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<td><strong>Undesired coupling frequency</strong></td>
<td>single CSRR and TL</td>
<td>L, C, C’, C”</td>
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<td><strong>Rejection level</strong></td>
<td>multiple CSRR and TL</td>
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**Determination of overall specifications**

- EM Simulation: TL’s system (original)
  - Extracted specs: dimensions, undesired coupling frequency, rejection level
Fig. 7
Fig. 8
Fig. 9
Fig. 10
Fig. 11
Fig. 12
Fig. 13
Fig. 14
Fig. 15

![Graph showing the comparison between Conventional implementation and CSRRs implementation for Ac vs. Vo. The y-axis represents Vo [dBm] ranging from -90 to -20, and the x-axis represents Ac [dBm] ranging from -25 to 15. The graph compares the performance of two implementations with a clear distinction between the Conventional implementation and the CSRRs implementation.]