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Synthesis and Design of Suspended Substrate Stripline Filters for Digital Microwave Power Amplifiers

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***Abstract*—In this paper, a synthesis method for suspended substrate stripline filters for digital microwave power amplifier applications is presented. The synthesis method combines a lumped element and full-wave mixed approach in a very efficient way. In order to achieve high amplifier efficiency the filter must exhibit a high input impedance in the stopband. This has been implemented for the first time by using a capacitively end coupled filter combined with stepped impedance resonators. A third order filter was designed. Simulations show that the final stage drain efficiency of the power amplifier and suppression of out-of-band frequency components can be significantly improved when the new structure is used.**

I. Introduction

Today‘s and next-generation mobile communication infrastructure demands for high flexibility, low cost and high efficiency. As a consequence, the RF power amplifier (PA) as the most power consuming part in the transmitter architecture needs to be optimized further. There is trend towards complete digital transmitter architectures and in this regard, digital PA concepts like class-D/S have proven to be good candidates [1].

A particular challenge for these amplifiers is the realization of the reconstruction filter. Stringent requirements in terms of low insertion loss, high selectivity, broadband input impedance behavior and compactness must be fulfilled. Moreover, at the input, it is operated in a non-50 Ω environment, in order to present optimum termination impedance.

This paper investigates the synthesis and design of a suspended substrate broadside coupled stripline filter for this purpose with high-Q, that allows very short connection lines to the PA output. A suspended substrate stripline is a printed circuit technology, where a substrate is suspended in air and enclosed by a metal cavity. The filter will be built with H-shaped printed resonators, also called Stepped Impedance Resonators (SIR). H-shaped printed resonators have been used in the past for antenna and filter [2], [3] applications. In [3] a useful design technique for filters based on SIR resonators was presented. However, in that work only side coupled structures were considered. On the other hand, in [4] a broadside coupled suspended substrate stripline, with SIR resonators printed on both sides of the substrate, was also proposed. The structure is similar to the one we propose in this work. However, in that work the design was based only on full wave simulations and a detailed design approach was not addressed.

In this paper we present a novel synthesis procedure for broadside coupled filters based on SIR resonators printed on both sides of the substrate. This is the first time that a filter based on SIR resonators is used as reconstruction filter in PA systems. Its interest is in the reduction of the size achieved and in the large spurious free range exhibited, as compared to regular printed line resonators.

II. Synthesis Procedure

In this section, the synthesis and design of a suspended substrate stripline filter will be detailed. The procedure starts with the calculation of the inline coupling matrix of the desired transfer function, using [5]. It is well known that such a coupling matrix represents a lowpass circuit filter that we will express in its general parallel configuration, i.e. using admittance inverters and parallel capacitors. Once this lowpass circuit is formed, where all the lumped elements are known, it can be easily transformed into the bandpass domain by using well-known lowpass to bandpass transformation formulas. The obtained bandpass circuit with admittance inverters and parallel connected resonators, will be the starting point for the synthesis steps detailed next.

A. Synthesis of Resonators

Printed microstrip H-shaped resonators, as the one shown in Fig. 1 (right side), will be used in the suspended substrate filter presented in this paper. The H-shaped resonators are formed with two sections of low impedance Z_1 and an additional section with higher impedance Z_2 placed between them. As shown in the figure, the total electrical length of the resonator is 3θ.

To start with the synthesis, we will scale the admittance of each i-th resonator of the bandpass circuit by a constant value n_i. Thus, the inverters will also be accordingly scaled. This constant will allow to define the slope parameter of the H-shaped resonators.

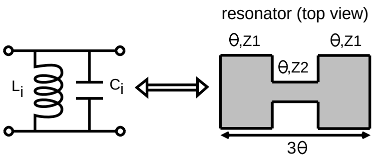

Fig. 1. Resonator equivalent circuit and printed microstrip pattern.

Next, each parallel LC i-th resonator of the bandpass circuit must be related to its equivalent microstrip printed circuit (see Fig. 1). The transfer matrix of each LC i-th resonator scaled by a constant factor n_i is:

$$[T]=\begin{bmatrix} A & B \\ C & D \end{bmatrix}=\begin{bmatrix} 1 & 0 \\ n_i^2 Y_i & 1 \end{bmatrix} \tag{1}$$

where A, B, C and D are given by the transfer matrix, and Y is the admittance of the LC i-th parallel resonator. Besides, for the H-shaped microstrip resonator, we may write:

$$[T]=\begin{bmatrix} c & jsZ_1 \\ js\frac{1}{Z_1} & c \end{bmatrix}\begin{bmatrix} c & jsZ_2 \\ js\frac{1}{Z_2} & c \end{bmatrix}\begin{bmatrix} c & jsZ_1 \\ js\frac{1}{Z_1} & c \end{bmatrix} \tag{2}$$

where $c = cos\theta$ and $s = sin\theta$. At the resonant frequency it is true that $Y = 0$ in (1), thereby it is also true that parameter C in(1) is $C = 0$.

Then, equating at the resonant frequency the parameter C of matrixes (1) (i.e. $C = 0$) and (2), and after some manipulations, we have the following resonant condition:

$$\tan^2\theta=\frac{2/Z_1+1/Z_2}{Z_2/Z_1^2} \tag{3}$$

Besides, since the resonators are symmetrical and reciprocal, the following identities are true:

$$A = D \tag{4}$$

$$AD - BC = 1 \tag{5}$$

Thus, at the resonant frequency it is true that $A = D = 1$, and (3) can be used to calculate impedance Z_2 once impedance Z_1 and angle θ are fixed.

At this point, the slope parameter of each ideal LC i-th resonator can be calculated and equated to the slope parameter of a H-shaped printed resonator. By doing this , constant n_i is found to be:

$$n_i=\frac{\omega^2 L_i C_i+1}{\omega L_i \theta c\left(c^2(\frac{2}{Z_1}+\frac{1}{Z_2})+s^2(\frac{4}{Z_1}+\frac{2}{Z_2}+\frac{Z_2}{Z_1^2})\right)} \tag{6}$$

B. Synthesis of Input/Output Couplings

Now, the input and output inverters can be transformed into equivalent circuits formed with a capacitor connected to a transmission line, as shown at the bottom of Fig. 2 (left side, labeled as input coupling). For the input coupling circuit on the top, the input admittance may be written:

$$Y_{in,1}=\left(\frac{J_{S1}}{n_1}\right)^2 \tag{7}$$

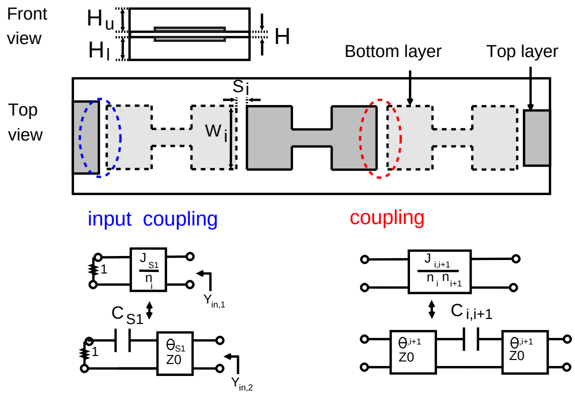

Fig. 2. Input inverter equivalent circuit (top) and microstrip pattern (bottom).

Besides, for the circuit at the bottom:

$$Y_{in,2}=\frac{C+D}{A+B} \tag{8}$$

where A, B, C and D are given by the transfer matrix of the circuit:

$$[T]=\begin{bmatrix} c & jsZ_0 \\ js\frac{1}{Z_0} & c \end{bmatrix}\begin{bmatrix} 1 & 1/j\omega C_{S1} \\ 0 & 1 \end{bmatrix} \tag{9}$$

Note that Z_0 will be given by the impedance of the low section impedance of the resonators, i.e. Z_1 in Fig. 1. Equating (7) and (8), and setting $Z_0 = 1$ in the circuit at the bottom of Fig. 2 (left side), the design equations are found for the input inverter:

$$\theta_{S1}=\tan^{-1}\left(\frac{J_{S1}}{n_1}\right) \tag{10}$$

$$C_{S1}=\frac{-1}{2\omega\cot(2\theta_{S1})} \tag{11}$$

Applying a similar procedure, the output coupling element values θ_{NL} and C_{NL} can be calculated.

C. Synthesis of Couplings between Resonators

The couplings between resonators can be modeled with a capacitor between two sections of transmission lines, as shown at the bottom of Fig. 2 (right side, labeled as coupling). For the circuit on the top of the two circuits labeled as coupling on Fig. 2, the transfer matrix can be written as:

$$[T]=-j\begin{bmatrix} 0 & n_i n_{i+1}/J_{i,i+1} \\ J_{i,i+1}/n_i n_{i+1} & 0 \end{bmatrix} \tag{12}$$

Besides, for the circuit at the bottom of Fig. 2 (right side):

$$[T]=\begin{bmatrix} c & jsZ_0 \\ js\frac{1}{Z_0} & c \end{bmatrix}\begin{bmatrix} 1 & \frac{1}{j\omega C_{i,i+1}} \\ 0 & 1 \end{bmatrix}\begin{bmatrix} c & jsZ_0 \\ js\frac{1}{Z_0} & c \end{bmatrix} \tag{13}$$

Equating now (12) and (13), the following design equations are found for the couplings between resonators:

$$\theta_{i,i+1}=\tan^{-1}\left(-Z_0\frac{J_{i,i+1}}{n_i n_{i+1}}\right) \tag{14}$$

$$C_{i,i+1}=\frac{-1}{2\omega Z_0\cot(2\theta_{i,i+1})} \tag{15}$$

where Z_0 will be given by the impedance of the low section impedance of the resonators, i.e. Z_1 in Fig. 1, which will be set to $Z_1 = 1$ as it will be shown in the next section.

D. Synthesis Example

A third order filter centered at $f_s = 1.8$ GHz with a BW of 100 MHz and 20 dB return loss, has been synthesized following the described procedure. The synthesis starts with the calculation of the inline coupling matrix, by using [5]. The symmetric coupling matrix is given by $M_{s1} = M_{3l} = 1.0824$ and $M_{12} = M_{23} = 1.0302$.

The coupling matrix allows to obtain the equivalent circuit with admittance inverters and parallel connected resonators. Next, we can fix $\theta = 45^o$ and $Z_1 = 1\ \Omega$, and using the design formula (3), the impedance $Z_2 = 2.4142\ \Omega$ can be calculated. The constant related to the slope parameter of the resonators calculated with (6) results $n_i = 3.6640$. After that, design equations (10), (11), (14) and (15) can be applied to directly transform the ideal bandpass lumped element circuit into a mixed lumped-distributed circuit. Note that transmission line sections with negative lengths created by the input/output couplings as well as by the couplings between the resonators are absorbed by the transmission line sections of the resonators. The obtained values for the capacitors and the transmission lengths parameters are $C_1 = 28.617$ pF, $C_2 = 6.825$ pF, $\theta_1 = 28.542^o$, $\theta_2 = 45^o$ and $\theta_3 = 40.612^o$.

If the ideal response of the filter given by the coupling matrix is compared to the S-Parameters calculated from the mixed lumped-distributed circuit, it can be observed that the frequency of the central resonance has been shifted. However, this can be easily solved with a simple fine tuning of the electrical length of the central resonator. The low impedance sections of the central resonator after the fine tuning process, have an electrical length of $\theta'_3 = 40.86^o$.

Finally, to normalize the filter to a 50 Ω environment, we need simply to divide all the capacitors by the impedance value, and multiply all the impedances of the transmission line sections by the same value. In general, we can point out that the input impedance of the synthesized filter exhibits the required out-of-band behavior (i.e. high impedance values) for proper power amplifier operation.

III. Final Design and Manufacturing

In this section, the final steps for the design and manufacturing of a suspended substrate stripline filter will be detailed. Also, it will be tested in simulation with a typical digital PA configuration, i.e., a voltage-mode class-D/S power amplifier circuit.

A. Design Issues

After the synthesis procedure, the physical dimensions of the transmission line sections must be calculated. This can be easily done once the electrical length θ_i and the impedance Z_i of the line sections are known, and the physical dimensions of the cavity enclosing the suspended substrate as well as the ones from the substrate are fixed. For our design, a substrate with a relative dielectric constant $\varepsilon_r = 2.23$ and thickness $H = 0.78$ mm has been selected. The height from the substrate to the top of the cavity and from the substrate to the lower ground plane spacing (see H_u and H_l in Fig. 2) is 5 mm. Then, if the total height of the structure is defined as $b = H_u + H_l + H$, the width of each transmission line section W_i can be calculated as [6]:

$$W_i = \frac{b}{4}\left(\frac{377}{Z_i} - 1.84\right) \tag{16}$$

Finally, the length of each transmission line section L_i is:

$$L_i = \frac{\theta_i \lambda_0}{360\sqrt{\varepsilon'_r}} \tag{17}$$

where θ_i is the electrical length of the section in degrees, and ε'_r is the effective dielectric constant of the suspended substrate.

After calculating the width and length of each transmission line section, only a small tuning in the length of the central transmission line sections of the resonators is needed to recover the perfect filter response. Once this final adjustment is performed, full-wave simulations must be used to finalize the design.

At this point, the filter couplings are represented by capacitors.Each capacitor needs to be translated into an overlap (or a gap if negative) of area $W_i \cdot S_i$ in the final circuit. An initial estimated value of the length of the overlap S_o can be calculated by using the parallel plate capacitors formula:

$$C = \frac{\varepsilon_0 \cdot \varepsilon_r \cdot W_i \cdot S_o}{H} = \frac{10^{-9}}{36\pi} \cdot \frac{\varepsilon_r \cdot W_i \cdot S_o}{H} \tag{18}$$

ε_r and H are the relative dielectric constant and the thickness of the substrate respectively, W_i the width of the overlap and S_o the initial estimated length of the overlap. However, fringing capacitances are not considered in this formula, and the initial length of the overlaps S_o are only approximated and usually much larger than the final correct values S_i. Thereby, full wave simulations must be used at this point to get the final optimized dimensions of the filter. Note that the previous use of a circuit-level simulator has provided us with a design relatively close to the final one, which makes the design process very efficient.

Once the filter design has been optimized, with full wave simulations a final step is still required. The whole design has been realized for a 50 Ω input impedance. However, at the input, the filter must be operated in a non-50 Ω environment at the center frequency, in order to present optimum termination impedance to the final stage of the power amplifier. In our case, the optimum input impedance (Z_{opt}) at the signal frequency of $f_s = 1.8$ GHz is found to be 40 Ω. To this end, the filter must be made asymmetric by adjusting the width of the input port to 40 Ω, and a new optimization must be performed to find the right value of the first overlap between. After the final optimization, the S-Parameters with $40\,\Omega$ reference impedance for the input port and $50\,\Omega$ reference impedance for the output port, are those shown in Fig. 3 (continuous line). The input impedance of the final design is plotted in Fig. 4 (continuous line). It can be observed that 40.5 Ω input impedance are achieved at the center frequency, which is very close to the optimum value for the PA.

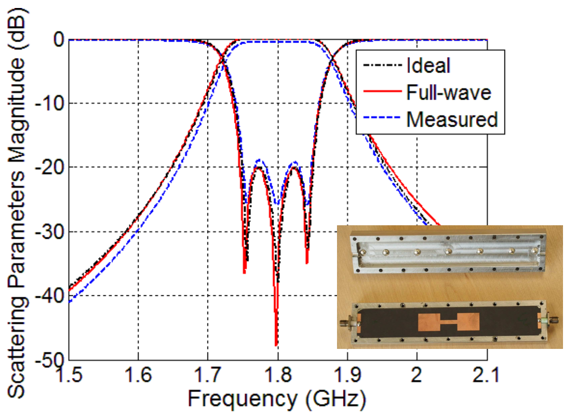

Fig. 3. Ideal third order circuit frequency response versus full-wave simulation and measurement results for comparison.

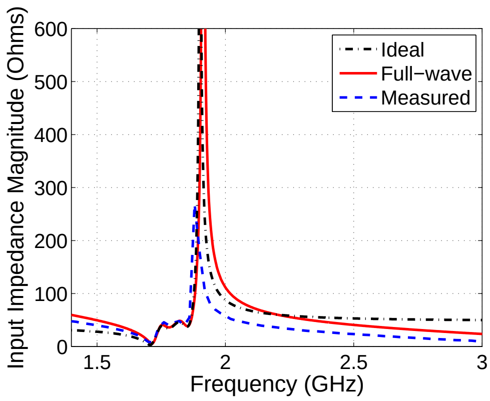

Fig. 4. Ideal third order circuit Input impedance (Z_{in}) versus full-wave simulation and measurement results for comparison.

B. Manufacturing example

The designed filter has been manufactured and measured. The prototype is shown in the inset of Fig. 3. Fig. 3 and Fig. 4 (dashed line) present the measured response of the manufactured filter, after performing a tuning process, as compared to the circuit response as well as to the full-wave simulated response. All the S-parameters are again given under $40/50\,\Omega$ reference impedance values at the input/output ports respectively. The measurements agree well with the simulation. The insertion loss at the center frequency of the filter is -0.46 dB, which is suitable for a low-loss operation of the targeted voltage-mode class-D/S PA.

The input impedance behavior of the measured filter, after de-embedding the connectors to avoid any phase shift during the measurements is plotted in Fig. 4 (dashed line). It can be observed that the measured and the designed characteristic fit well. More importantly, however, the curve reveals that the filter network provides values close to the optimum Z_{in} of 40 Ω within the passband. In the out-of-band range high impedances beyond 100 Ω are achieved as required for proper PA operation.

TABLE I

SIMULATED FINAL STAGE DRAIN EFFICIENCY AND OUT-OF-BAND REJECTION OF DIGITAL PA WITH SUSPENDED SUBSTRATE STRIPLINE (SSS) AND LUMPED ELEMENTS INLINE (LE) OUTPUT FILTER .

-	sss-filter	3rd order le-filter
η (%) Final stage drain efficiency	66.5	58.1
Spectral rejection out-of-band (dB)	57	32

In order to benchmark the filter performance in terms of the digital PA, the fabricated filter has been tested in simulation at the output of a voltage-mode class-D/S digital microwave power amplifier. The simulation results are compared to the ones obtained when using simple third-order lumped elements inline filter, as applied in this type of PA so far.

The simulations have been performed by using the software ADS from Agilent Technologies with an envelope delta-sigma modulated (EDSM) input signal. It consists of a single carrier of 5 MHz bandwidth with a peak-to-average power ratio (PAPR) of 7.5 dB encoded. Table I compares the simulation results in terms of final stage drain efficiency η (%) and spectral rejection of out-of-band frequency components which are important figures of merit. The filter high Q-factor and low insertion loss leads to an improvement of 8.41 % on the final stage drain efficiency η.

IV. CONCLUSION

In this paper a synthesis and design method has been described for suspended substrate bandpass filters, to be used as reconstruction filters in power amplifiers. Since the first part of the design is performed at circuit-level and only in the final part of the design the time-consuming full-wave simulations are needed, the approach results in high design efficiency. A third order prototype filter for a signal frequency of $f_s = 1.8$ GHz has been designed, manufactured and measured to validate the proposed procedure.

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