Real Frequency Design of Broadband Microwave Amplifiers with Mixed Lumped and Distributed Element Equalizers for MMIC's

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***Abstract* - In this paper, a computer–aided real frequency technique is presented for treating the design of mixed lumped and distributed element equalizer networks in broadband microwave amplifiers for wireless communication systems. The work aims to bring together the available real frequency design techniques for wideband matching and the multidimensional modelling approach for accurate description of MIC realisations. The scattering based two-variable description of lossless equalizers with mixed lumped and distributed elements will be discussed and a practical two-stage microwave amplifier design example will be studied to exhibit the application and the potential benefits of the algorithm.**

I. INTRODUCTION

One of the major issues in wireless communication is appropriate design of wideband RF subsytems. In order to ensure the success of the design and analysis algorithms, accurate modelling of physical components and implementation process is essential. In this contex, MMIC design of wideband RF subsytems, such as amplifiers, filters and matching networks require that precise device and device to circuit medium interface models incorporating parasitic effects are developed and the circuit-design approach is well disciplined. In this regard, utilisation of mixed lumped and distributed multidimensional circuit models to represent the interstage equalizers and interconnects of RF sytems in MMIC's would offer advantages for accurate simulation of MIC layouts, where the physical sizes, parasitics and discontinuities are embedded in the design process.

In this paper, the characterisation and the design of active and passive RF subsytems will be based on multidimensional electrical network models. In particular, utilisation of mixed lumped and distributed circuits to model the front-end, back-end and interstage equalizers of microwave amplifiers will be discussed.

It is well known that, in the design of lossless matching equalizers, the real frequency technique is a very efficient approach, since it is numerical and utilizes experimental load, generator and active device data. Furthermore, no analytic form of the transfer function is required. For the design of broadband multistage microwave amplifiers with mixed lumped-distributed interstage equalizers, the generalisation of the well known scattering based real frequency technique will be used. As an application of the proposed approach, an example is included to illustrate the synthesis mixed element equalizers in the design of a two stage broadband FET amplifier for a mobile communication MIC.

II. TWO-VARIABLE SCATTERING DESCRIPTION OF MIXED ELEMENT LOSSLESS TWO-PORTS

Lossless two-ports, constructed with mixed-lumped and distributed elements can be described in terms two variable scattering matrix $S=S(p,\lambda)$ (Fig.1). Here, $p=\sigma+j\omega$ is the conventional complex frequency variable associated with lumped elements and $\lambda=tanh(p\tau)=\Sigma+j\Omega$ is the Richard variable, associated with the equal length transmission lines, also called Unit Elements (UE). In Belevitch representation, the scattering parameters are given by,

$$S_{11}=\frac{h(p,\lambda)}{g(p,\lambda)} \qquad S_{12}=\sigma\frac{f(-p,-\lambda)}{g(p,\lambda)}$$

$$S_{21}=\frac{f(p,\lambda)}{g(p,\lambda)} \qquad S_{22}=-\sigma\frac{h(-p,-\lambda)}{g(p,\lambda)} \tag{1}$$

where $\sigma=\pm1$ and the two variable real polynomials $g(p,\lambda)$ and $h(p,\lambda)$ can be expressed as

$$g(p,\lambda)=[P]^T\Lambda_G[\lambda]; \quad h(p,\lambda)=[P]^T\Lambda_H[\lambda]$$

where $[P]^T = [1 \quad p \quad p^2 \dots p^n]$ and $[\lambda] = [1 \quad \lambda \quad \lambda^2 \dots \lambda^n]^T$

$$\Lambda_G = \begin{bmatrix} g_{00} & g_{01} & \cdots & g_{0n_\lambda} \\ g_{10} & g_{11} & \cdots & g_{1n_\lambda} \\ \vdots & \vdots & \ddots & \vdots \\ g_{n_p0} & g_{n_p1} & \cdots & g_{n_pn_\lambda} \end{bmatrix} \quad \Lambda_H = \begin{bmatrix} h_{00} & h_{01} & \cdots & h_{0n_\lambda} \\ h_{10} & h_{11} & \cdots & h_{1n_\lambda} \\ \vdots & \vdots & \ddots & \vdots \\ h_{n_p0} & h_{n_p1} & \cdots & h_{n_pn_\lambda} \end{bmatrix}$$

In this representation, Λ_H and Λ_G are called connectivity matrices and they are formed by the corresponding real polynomial coefficients.

If the structure consists of only series inductor and shunt capacitor type of lumped elements connected with unit element (UE) then,

$$f(p,\lambda) = \left(1-\lambda^2\right)^{n_\lambda/2} \tag{2}$$

In the above formulation n_p stands for the total number of lumped circuit elements, n_λ designates the total number of unit elements. Since the two-port is lossless,

$$S(p,\lambda)S^T(-p,-\lambda) = I \tag{3}$$

where I is the unity matrix. Employing the Belevitch form of (1) and the losslessness condition of (3) we have,

$$g(p,\lambda)g(-p,-\lambda) = h(p,\lambda)h(-p,-\lambda) + f(p,\lambda)f(-p,-\lambda) \tag{4}$$

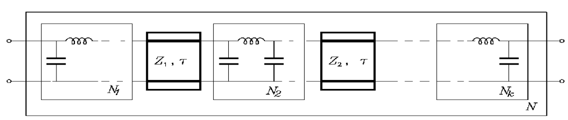

Fig.1. Cascade with mixed lumped-distributed elements

For the cascade topology under consideration, the following boundary cases may be remarked:

a) When the transmission lines are removed from the cascade structure, one would end up with a lumped network whose scattering matrix can fully be described independently in terms of the canonical real polynomials $h(p,0), g(p,0)$ and $f(p,0)$ where $g(p,0)$ is strictly Hurwitz and

$$g(p,0)g(-p,0) = h(p,0)h(-p,0) + f(p,0)f(-p,0) \tag{5}$$

b) When the lumped elements are removed from the cascade structure, one would obtain cascade connection of transmission lines. For the particular case of low-pass type lumped sections the boundary polynomials $h(0,\lambda), g(0,\lambda)$ and $f(0,\lambda)$ define the cascade of UEs which take place in the composite structure, where $g(0,\lambda)$ is strictly Hurwitz and

$$g(0,\lambda)g(0,-\lambda) = h(0,\lambda)h(0,-\lambda) + (1-\lambda^2)^{n_\lambda} \tag{6}$$

The more general cases with finite transmission zeros in lumped sections can also be treated following a similar reasoning.

The single variable boundary polynomials $h(p,0), g(p,0)$ and $h(0,\lambda)$, $g(0,\lambda)$ satisfying the relations (5) and (6) define the first row and the first column entries of Λ_H and Λ_G matrices. Now the problem is to generate the remaining unknown entries, which carry the cascade connection information so that the two-variable paraunitary relation (4) is satisfied together with the boundary conditions introduced in (a) and (b).

Based on the above discussion, we end up with a semi analytic procedure to construct two-variable canonical polynomials [3]: Assuming a regular cascaded structure as in Fig. 1, select the number of lumped and distributed elements (n_p, n_λ) and the polynomial $f(p,\lambda)$. Then choose the coefficients of the polynomials $h(p,0)$ and $h(0,\lambda)$ as the independent parameters and generate the Hurwitz polynomials $g(p,0)$ and $g(0,\lambda)$. Utilizing the boundary conditions in paraunitary relation, obtain the unknown coefficients.

For practical low-pass or high-pass type lumped element regular structures the proposed semi-analytic approach has been successfully applied and two-variable characterization for a variety of practical mixed element circuits have been obtained [2]-[3].

III. REAL FREQUENCY TECHNIQUE TO DESIGN AMPLIFIERS WITH MIXED ELEMENT EQUALIZERS

It is well known that the scattering based Real Frequency Technique provides an efficient tool for the design of microwave amplifiers [4] , [5]. The extension of the real frequency technique for designing equalizers with mixed elements require a unique characterization of the matching network in terms of a number of independent free parameters. The proposed procedure in the previous readily leads us to the two-variable generalization of the scattering based Real Frequency Technique, which can be outlined as follows:

- In the two-variable description of cascaded lumped and distributed two-ports, the real normalized scattering parameters are generated from partially defined numerator polynomial $h(p,\lambda)$ of the input reflection function $S_{11}(p,\lambda) = h(p,\lambda)/g(p,\lambda)$. In doing so, the mixed element structure is assumed to be separable into its lumped and distributed parts which can in turn completely be defined in terms of the corresponding h polynomials $h(p,0)$ and $h(0,\lambda)$, provided that the polynomials $f(p,0)$ and $f(0,\lambda)$ are defined.

- Starting from the arbitrary, unconstraint coefficients of $h(p,0)$ and $h(0,\lambda)$, we generate the remaining unknown coefficients of the polynomials $h(p,\lambda)$ and $g(p,\lambda)$ by utilizing the connectivity information supplied for the cascade structure applying the procedure given above.
- Then, the coefficients of the polynomials $h(p,0)$ and $h(0,\lambda)$ are chosen as the independent unknowns of the problem and determined to optimize the gain of the system by means of a nonlinear search routine.

Referring to Fig.2, assume that the active two-ports are indicated by the real normalized scattering parameters A_{ij}, and the equalizer networks real normalized scattering parameters are indicated by S_{ij}. Multistage amplifier design procedure with the corresponding gain expressions in front-end, interstage and back-end equalizer design steps can be summarised as follows:

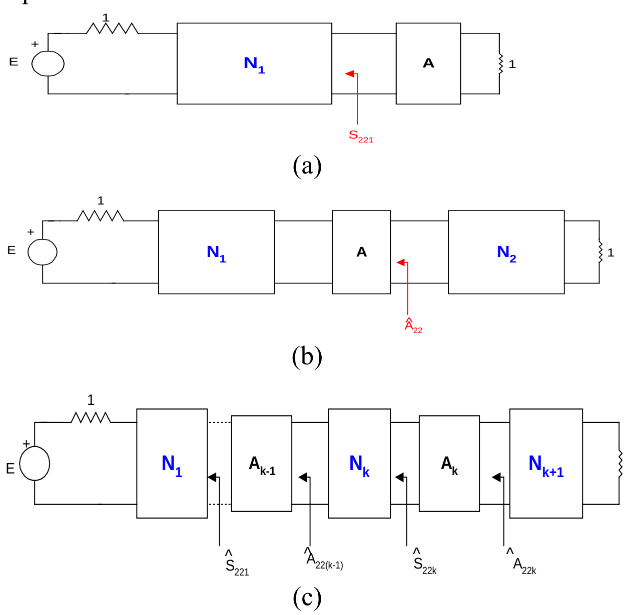

Fig.2 Sequential design of multistage amplifier; a) Front-end equalizer design, b)Back-end equalizer design, c) Interstage equalizers in multistage case

Step 1: Front-end Equalizer Design can be obtained assuming a unit termination after the transistor as shown in Fig. 2a, where the gain can be written as

$$T_1(w) = \frac{\left|S21_1\right|^2 \left|A21\right|^2}{\left|1 - S22_1 A11\right|^2}$$

Step 2: For the Back-end Equalizer Design assume the configuration shown in Fig.2b where the gain is defined as

$$T_2(w) = T_1(w)\frac{\left|S21_2\right|^2}{\left|1-\hat{A}22\,S11_2\right|^2} \qquad \hat{A}22 = A22 + \frac{A21A12S22_1}{1-S22_1A11}$$

Step 3: Inter-stage Equalizers in Multistage case can easily be obtained by extending the sequential gain expression as

$$T_k(w)=T_{k-1}(w)\frac{\left|A21_k\right|^2\left|S21_{k+1}\right|^2}{\left|1-\hat{S}22_k\,A11_k\right|^2\left|1-\hat{A}22_k\,S11_{k+1}\right|^2}$$

In the above procedure, there is no restriction on the unknown real coefficients of the polynomials $h(p,0)$ and $h(0,\lambda)$. Therefore any unconstraint optimization routine can be employed. The numeric in the optimization is well behaved and the convergence becomes much faster than the direct optimization of the element values.

IV. DESIGN EXAMPLE

In this section, a double stage FET amplifier is designed for 50 Ohm terminations. The proposed mixed lumped and distributed circuit models are utilized as front-end, back-end and intersatage matching equalizers. The measured scattering data available for the active device pair are directly processed to obtain an octave band amplifier over the GSM band of 0.9GHz to 2 GHz.

In the design process, the sequential optimization scheme described in the previous section is employed. The flat power gain is achieved by employing the least square error optimization where the nitial values are selected in an ad hoc manner. The active device pair in the amplifier is chosen as AM012MXQF for which the scattering parameters are given by the manufacturer as in Table1.

The amplifier is designed for obtaining an average gain level of 21dB . For both the input and interstage equalizer circuits we choose a four elements LPLU section ($n_p = n_\lambda = 2$), and a five elements LPLU section for the output equalizer circuit ($n_p = 3, n_\lambda = 2$).

As a result of design process, the following polynomials characterizing the scattering functions of the front-end, interstage, and back-end matching networks are obtained.

Front-End Equalizer:

$$\Lambda_H = \begin{bmatrix} 0 & -1.4721 & -0.0816 \\ -0.1279 & -0.3177 & 0.1248 \\ -0.0322 & 0.0156 & 0 \end{bmatrix} \quad \Lambda_G = \begin{bmatrix} 1 & 2.4847 & 1.0033 \\ 0.2842 & 0.5178 & 0.1248 \\ 0.0322 & 0.0156 & 0 \end{bmatrix}$$

TABLE I

SCATTERING PARAMETERS OF AM012MXQF(V_{DS}=5V,I_D=0.5I_{DSS})

Frq.	S_{11} (Mag)	S_{11} (Ang.)	S_{21} (Mag)	S_{21} (Ang.)	S_{12} (Mag.)	S_{12} (Ang.)	S_{22} (Mag.)	S_{22} (Ang.)
900	.920	-82.891	7.764	126.457	.053	42.404	.299	-83.543
1000	.912	-89.512	7.420	121.981	.056	38.498	.301	-89.695
1100	.897	-96.695	7.119	117.758	.058	34.885	.312	-93.973
1200	.884	-102.121	6.782	113.949	.059	32.135	.309	-97.695
1300	.875	-107.207	6.473	110.379	.061	30.073	.310	-101.172
1400	.868	-111.996	6.189	106.984	.063	27.780	.311	-104.535
1500	.862	-116.535	5.926	103.770	.065	25.321	.311	-107.828
1600	.857	-120.793	5.684	100.652	.066	23.124	.312	-110.758
1700	.852	-124.816	5.456	97.734	.067	20.910	.311	-113.762
1800	.847	-128.609	5.239	94.906	.068	18.836	.309	-116.566
1900	.843	-132.070	5.043	92.188	.068	17.014	.308	-119.039
2000	.839	-135.375	4.859	89.555	.069	15.353	.307	-121.449
2100	.837	-138.523	4.688	89.555	.070	13.667	.306	-123.871
2200	.833	-141.438	4.526	87.043	.070	12.030	.305	-126.367

Interstage Equalizer:

$$\Lambda_H = \begin{bmatrix} 0 & 1.0097 & -1.2753 \\ 0.3613 & 0.4974 & 2.0620 \\ 0.2147 & 0.5064 & 0 \end{bmatrix} \quad \Lambda_G = \begin{bmatrix} 1 & 2.5021 & 1.6206 \\ 0.7483 & 1.5075 & 2.0620 \\ 0.2147 & 0.5604 & 0 \end{bmatrix}$$

Back-End Equalizer:

$$\Lambda_H = \begin{bmatrix} 0 & 5.0555 & -1.8360 \\ 3.7690 & -2.2674 & 5.7196 \\ -0.6340 & 5.6164 & 0 \\ 1.0620 & 0 & 0 \end{bmatrix} \quad \Lambda_G = \begin{bmatrix} 1 & 5.6338 & 2.0907 \\ 4.0164 & 3.5733 & 5.7196 \\ 0.9630 & 5.6164 & 0 \\ 1.0620 & 0 & 0 \end{bmatrix}$$

Transducer power gain performance and finalized amplifier circuit with real element values is presented in Fig. 3 and Fig.4 respectively.

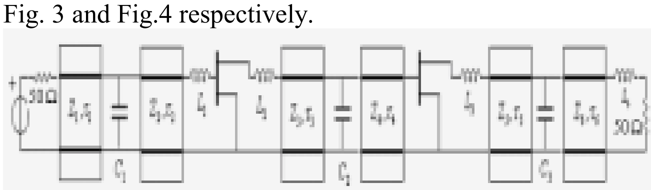

Fig.3 Two-stage amplifier design : (L_1=7.5nH,C_1=20.6nF, L_2=55.5nH,C_2=19.3nF,L_3=66.5nH,C_3=12.4nF,L_4=322.7nH, Z_1=24.3, Z_2=26.5, Z_3=45.3,Z_4=130.5,Z_5=108.5,Z_6=426, $\tau_1= \tau_2$=0.2, τ_3=0.1)

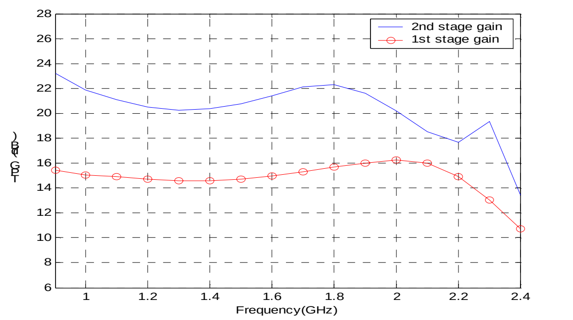

Fig.4 Performance of two-stage amplifier

V. CONCLUSION

This work is concerned with the development of a practical design tool to assist in the modelling and design process of broadband equalizers for MMIC amplifiers. A computer aided design algorithm for broadband microwave amplifiers employing distributed equalizers with lumped discontinuity models is presented. The method is based on the two variable description of lossless matching equalizers on a scattering basis and extended to the real frequency design of lossless equalizers of broadband amplifiers. Use of the method is illustrated by a multistage broadband amplifier design example. The modelling tool and the results of the work will find practical applications in the behaviour characterisation, simulation and design of high speed/high frequency analog mobile communication subsytems manufactured on VLSI and MMIC chips.

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