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LETTER

Design of broadband microwave amplifiers with mixed-elements via reflectance data modeling

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Abstract

A practical method is introduced, to design single-stage broadband microwave amplifiers with mixed lumped and distributed elements via modeling the reflectance data obtained from lumped-element input and output matching network prototypes. The same transducer power gain level is obtained by using less number of lumped-elements in the mixed-element amplifier than that of the lumped-element amplifier prototype. A mixed-element amplifier design is presented, to exhibit the utilization of the method. It is expected that the method will be employed, to design microwave amplifiers for broadband communication systems.

Keywords: Microwave amplifiers; Broadband; Modeling; Mixed lumped and distributed elements

1. Introduction

For many communications engineering applications, design of broadband microwave amplifiers are essential. Lumped-element amplifiers are preferable because of their small dimensions. However, interconnections of lumped-elements may be considered as transmission lines. These unavoidable connections destroy the performance of the lumped-element amplifiers. But these connection lines can be used as circuit components. In this case, the circuits must be in two-dimensional, namely the circuits must be composed of mixed lumped and distributed elements. A simple way to design mixed-element amplifiers may be to select the input (front-end) and output (back-end) matching network topologies. In the topologies, interconnections between lumped-elements are regarded as transmission lines. Then, values of the lumped-elements and characteristic impedances of the transmission lines are determined by means of the optimization of the gain performance of the amplifier. Although this approach is very simple, it presents some difficulties. First, the optimization is strongly nonlinear in terms of element values that may result in local minima or prevent convergence at all. Secondly, there is no established process, to initialize the element values of the chosen network topologies. Worst of all *the optimum choices of the matching network topologies are not known*. Fortunately, these problems are overcome employing the design technique introduced in this paper. The new amplifier design technique includes two phases. In Phase I, lumped-element amplifier prototype is constructed employing the well-established amplifier design methods. Then, output and input reflection coefficients of the front-end and back-end matching network prototypes are evaluated point by point over the passband, respectively. In Phase II, data

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generated from the reflection coefficients are modeled in two complex variables (one for lumped- and one for distributed-elements), which in turn results the desired amplifier in two kinds of elements. In practice, commensurate transmission lines or equal length lines (Unit Elements, UEs) are used as distributed-elements, to connect lumped-elements of the amplifier.

In the following sections, first the characterization of two-variable networks is introduced. Then, the modeling approach is explained, and the algorithm is presented. Finally, a microwave amplifier is built with mixed lumped and distributed elements, to exhibit the utilization of the proposed method.

2. Characterization of two-variable networks

In this paper, the modeling problem is defined as the generation of a realizable, two-variable bounded real (BR) reflectance function that best fits the given data. Eventually, this BR reflectance function describes the lossless front-end/back-end matching network in two kinds of elements in resistive termination which is called the Darlington representation of the input reflection function (Fig. 1).

Let $S(\mathrm{j}\omega_i) = S_R(\omega_i) + \mathrm{j}S_X(\omega_i)$ designate the given data obtained form the lumped-element front-end/back-end matching network prototype over the angular frequencies ω_i. Let $\{S_{kl}; k, l = 1, 2\}$ designate the scattering parameters of the lossless matching networks which consist of two kinds of elements. For a mixed lumped and distributed element, reciprocal, lossless two-port, the scattering parameters may be expressed in Belevitch form as follows [1–4]

$$S(p,\lambda) = \begin{bmatrix} S_{11}(p,\lambda) & S_{12}(p,\lambda) \\ S_{21}(p,\lambda) & S_{22}(p,\lambda) \end{bmatrix} = \frac{1}{g(p,\lambda)} \begin{bmatrix} h(p,\lambda) & \mu f(-p,-\lambda) \\ f(p,\lambda) & -\mu h(-p,-\lambda) \end{bmatrix}, \quad (1a)$$

where $\mu = f(-p,-\lambda)/f(p,\lambda)$.

The polynomials $g(p,\lambda)$, $h(p,\lambda)$ and $f(p,\lambda)$ satisfy the following properties [2]:

- $g(p,\lambda)$, $h(p,\lambda)$ and $f(p,\lambda)$ are real polynomials of the complex variables p and $\lambda = \tanh p\tau$, where τ is the commensurate one-way delay of the distributed elements.
- $g(p,\lambda)$ is a scattering Hurwitz polynomial [5–7], i.e.
  - $g(p,\lambda) \neq 0$ for $\mathrm{Re}\{p,\lambda\} > 0$.
  - $g(p,\lambda)$ is relatively prime with $g(-p,-\lambda)$.
- $f(p,\lambda)$ is a monic polynomial.

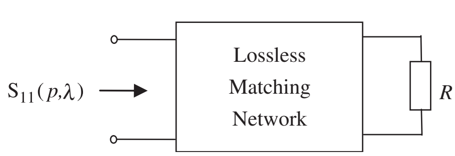

Fig. 1. Darlington representation of the modeled input reflectance function $S_{11}(p,\lambda)$.

- $g(p,\lambda)$, $h(p,\lambda)$ and $f(p,\lambda)$ are related by

$$g(p,\lambda)g(-p,-\lambda) = h(p,\lambda)h(-p,-\lambda) + f(p,\lambda)f(-p,-\lambda). \quad (1b)$$

- Let us designate the lumped and distributed subsections by $[L]$ and $[D]$, respectively. If the two-port $[D]$ includes cascaded UEs, then $f(p,\lambda)$ is defined in product separable form as $f(p,\lambda) = f_L(p)f_D(\lambda) = f_L(p)(1-\lambda^2)^{n_\lambda/2}$, n_λ is the number of UEs.

As far as the modeling problem is concerned, one has to generate the two-variable, realizable, BR scattering parameters of the lossless two-port of Fig. 1 in such a way that the input reflection coefficient $S_{11}(p,\lambda)$ is fit the given data $S(\mathrm{j}\omega)$ at each frequency point under consideration. In the following section, a practical approach is presented, to build the models which guarantees the realizibility of the two-variable scattering parameters specified by Eq. (1).

3. A practical modeling approach

First, let us consider the generic form of a lossless matching network formed with cascade connections of series inductances, transmission lines and shunt capacitances as shown in Fig. 2. In this figure, distributed elements are all equal length transmission lines (Unit Elements, UEs) with constant delay τ. Since Fig. 2 presents a lossless two-port network constructed with simple **L**ow **P**ass **L**adder elements connected with **U**nit elements, it is called an **LPLU** structure.

An LPLU structure can fully be described in terms of the real coefficients of the boundary polynomials $h(p,0) = \sum_{i=0}^{n_p} h_{i0}p^i$ and $h(0,\lambda) = \sum_{i=0}^{n_\lambda} h_{0i}\lambda^i$ as detailed by [1–4]. In short, once the real coefficients h_{i0} and h_{0i} are initialized, then strictly Hurwitz polynomials $g(p,0) = \sum_{i=0}^{n_p} g_{i0}p^i$ and $g(0,\lambda) = \sum_{i=0}^{n_\lambda} g_{0i}\lambda^i$ can be computed by means of the explicit factorization of Eq. (1b).

In this case, input reflection coefficients defined by $S_{11}(p,0) = \frac{h(p,0)}{g(p,0)} = S_L(p) = \frac{h_L(p)}{g_L(p)}$ and $S_{11}(0,\lambda) = \frac{h(0,\lambda)}{g(0,\lambda)} = S_D(\lambda) = \frac{h_D(\lambda)}{g_D(\lambda)}$ completely describe the matching network constructed in two kinds of elements.

Synthesis of these networks can separately be carried out using classically known methods or by means of the decomposition algorithm of Fettweis [8]. Then, by mixing the elements of $[L]$ and $[D]$ in sequential order, the desired matching network is obtained. Eventually, complete

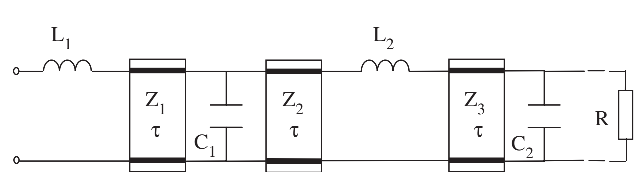

Fig. 2. A practical model topology.

scattering parameters of the matching network are derived from the final topology as shown in Fig. 2.

It should be noted that LPLU structure can easily be generalized by selecting a desirable form for $f_L(p)$. For example, a generic form for a simple band-pass (BP), lumped-element ladder connected with commensurate transmission lines can be obtained by setting $f_L(p) = p^k$. Then, the rest of the procedure follow as described above.

The above clarification leads us to propose the following numerical approach, to build the mixed-element matching networks.

3.1. A numerical approach to design mixed-element matching networks via modeling

Over the angular frequencies ω, let $\varepsilon(\mathrm{j}\,\omega) = S(\mathrm{j}\,\omega) - S_{11}(\mathrm{j}\,\omega, \mathrm{j}\tan(\omega\tau))$ be the error function defined as the difference between the given data and the analytic form of the input reflection coefficient of the front-end/back-end matching network which will be constructed in two kinds of elements. Obviously, $|\varepsilon(\mathrm{j}\,\omega)|^2$ is the function of both $h_L(p)$ and $h_D(\lambda)$. This functional relation can be expressed as

$$|\varepsilon|^2 = \varepsilon \cdot \varepsilon^* = F(S_{11}); \quad S_{11} = F(h_L, h_D), \tag{2}$$

where "*" represents the complex conjugate of a complex number.

Referring to Eq. (2), one can minimize the error $|\varepsilon|^2$ in the directions of the partial derivatives given by

$$\frac{\partial \varepsilon\varepsilon^*}{\partial h_L} = \frac{\partial|\varepsilon|^2}{\partial h_L} = \frac{-\varepsilon^*}{g(p, \lambda)}, \tag{3a}$$

$$\frac{\partial \varepsilon\varepsilon^*}{\partial h_D} = \frac{\partial|\varepsilon|^2}{\partial h_D} = \frac{-\varepsilon^*}{g(p, \lambda)}. \tag{3b}$$

In this case, an iterative method, perhaps the gradient method, may be employed, to minimize the error function $|\varepsilon|^2$ which in turn yields the polynomials h_L and h_D from the initialized coefficients as follows:

$$h_L^{(r)} = h_L^{(r-1)} - \frac{\partial|\varepsilon|^2}{\partial h_L}\Big|_{h_L^{(r-1)}}, \tag{4a}$$

$$h_D^{(r)} = h_D^{(r-1)} - \frac{\partial|\varepsilon|^2}{\partial h_D}\Big|_{h_D^{(r-1)}}. \tag{4a}$$

In Eq. (4), the subscript r designates the iteration index starting at $r = 1$, and, $h_L^{(0)}(p)$ and $h_D^0(\lambda)$ are the initialized polynomials stem from the polynomials $h_L(p)$ and $h_D(\lambda)$, respectively. Thus, the following algorithm is proposed, to design amplifiers in two kinds of elements via modeling.

3.2. Algorithm: Generation of mixed-element amplifier from lumped-element prototype

Inputs:

- ω_i; $i = 1, 2, \ldots, N_\omega$: Sample frequencies.
- N_ω: Total number of sample frequencies.
- $S(\mathrm{j}\,\omega_i) = S_R(\omega_i) + \mathrm{j}\,S_X(\omega_i)$; $i = 1, 2, \ldots, N_\omega$: Sample points generated from the output/input reflection coefficient of the lumped-element front-end/back-end matching network prototype, respectively.
- n_λ: Total number of elements in $[D]$.
- $f_D(\lambda)$: A monic polynomial constructed on the transmission zeros of $[D]$. It is noted that for cascaded connection of commensurate transmission lines, $f_D(\lambda) = (1 - \lambda^2)^{n_\lambda/2}$ is selected.
- n_p: Total number elements in $[L]$.
- k: Total number of transmission zeros of $[L]$ at DC.
- $f_L(p)$: A monic polynomial constructed on the transmission zeros of $[L]$. In our modeling approach, all the transmission zeros are imbedded into the lumped ladder section $[L]$ by choosing $f_L(p) = p^k$.
- $h_{0D}^{(0)}, h_{1D}^{(0)}, h_{2D}^{(0)}, \ldots, h_{n_\lambda D}^{(0)}$ and $h_{0L}^{(0)}, h_{1L}^{(0)}, h_{2L}^{(0)}, \ldots, h_{n_p L}^{(0)}$: Initialized coefficients of the polynomials $h_D^{(0)}(\lambda)$ and $h_L^{(0)}(p)$, respectively. The gradient method determines a local minimum. So to be able to reach the global minimum, suitable initial values must be generated. But, unfortunately, there is no way to obtain the suitable values for two-variable modeling.
- δ: The stopping criteria for the sum of the square errors. Usually, it is sufficient to choose $\delta = 10^{-3}$.

Computational Steps

Step 1: Set $r = 1$ and start the iterations for the gradient method.

Step 2: By using the $(r - 1)$th initial coefficients $h_{0D}^{(r-1)}, h_{1D}^{(r-1)}, h_{2D}^{(r-1)}, \ldots, h_{n_\lambda D}^{(r-1)}$ and $h_{0L}^{(r-1)}, h_{1L}^{(r-1)}, h_{2L}^{(r-1)}, \ldots, h_{n_p L}^{(r-1)}$, compute the strictly Hurwitz polynomials $g_D^{(r-1)}(\lambda)$ and $g_L^{(r-1)}(p)$ employing the energy conservation conditions of $[D]$ and $[L]$.

Step 3: Synthesize lumped-element two-port $[L]$, and distributed-element two-port $[D]$, to obtain the component values.

Step 4: By using the component values, form the scattering transfer matrix for each element.

Step 5: According to the connection order, multiply the scattering transfer matrices of the components, and obtain scattering transfer matrix of the mixed model, and then obtain two-variable polynomials $g(p, \lambda)$, $h(p, \lambda)$ and $f(p, \lambda)$.

Step 6: Compute the error $\varepsilon^{(r-1)}(\mathrm{j}\,\omega_i) = S(\mathrm{j}\,\omega_i) - \frac{h^{(r-1)}(\mathrm{j}\,\omega_i, \mathrm{j}\tan(\omega_i\tau))}{g^{(r-1)}(\mathrm{j}\,\omega_i, \mathrm{j}\tan(\omega_i\tau))}$ over the given frequencies.

Step 7: Compute the sum of the square errors $\delta^{(r-1)} = \sum_{i=1}^{N_\omega} |\varepsilon^{(r-1)}(\mathrm{j}\,\omega_i)|^2$. If $\delta^{(r-1)} \leqslant \delta$, set $S_{11}(p, \lambda) = \frac{h^{(r-1)}(p,\lambda)}{g^{(r-1)}(p,\lambda)}$ and stop. Otherwise go to the next step.

Step 8: Compute the complex quantities over the sample frequencies for the given $(r - 1)$th initials, $h_L^{(r)}(\mathrm{j}\,\omega_i) = h_L^{(r-1)}(\mathrm{j}\,\omega_i) + \frac{\varepsilon^{(r-1)}(-\mathrm{j}\,\omega_i)}{g^{(r-1)}(\mathrm{j}\,\omega_i, \mathrm{j}\tan(\omega_i\tau))}$ and $h_D^{(r)}(\mathrm{j}\tan(\omega_i\tau)) = h_D^{(r-1)}(\mathrm{j}\tan(\omega_i\tau)) + \frac{\varepsilon^{(r-1)}(-\mathrm{j}\,\omega_i)}{g^{(r-1)}(\mathrm{j}\,\omega_i, \mathrm{j}\tan(\omega_i\tau))}$ point by point.

Step 9: Using these complex quantities , find the coefficients of the polynomials $h_D^{(r)}(\lambda)=\sum_{i=1}^{n_\lambda} h_{iD}^{(r)}\lambda^i$ and $h_L^{(r)}(p)=\sum_{i=1}^{n_p} h_{iL}^{(r)}p^i$ by means of any linear interpolation or curve fitting routine [9,10].

Step 10: Set $r=r+1$ and go to Step 2.

By using the same algorithm defined above, front-end and back-end mixed-element matching networks of the amplifier are designed. In the following, an example is worked out, to generate the mixed-element amplifier for a given lumped-element amplifier prototype.

4. Example: mixed-element microwave amplifier

In this example, a lumped-element microwave amplifier prototype designed in [2] was transformed to a mixed-element counterpart via proposed modeling algorithm. Throughout the computations normalized elements were used. Lumped-element amplifier prototype and scattering parameters of the transistor are given in Fig. 3 and Table 1, respectively. In Fig. 3, f_N is normalization frequency and R_N is normalization resistance.

The generated data ($S_{22}^{(F)}(j\omega)$ and $S_{11}^{(B)}(j\omega)$) from the output/input reflection coefficients of lumped-element front-end/back-end matching networks are given in Table 2, respectively.

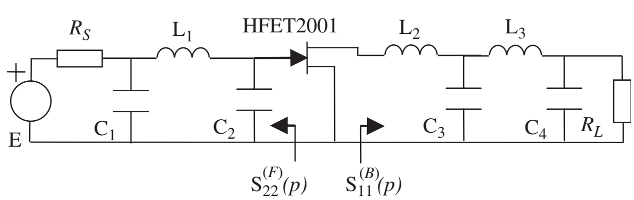

Fig. 3. Lumped-element microwave amplifier ($L_1=0.98$, $L_2=3.23$, $L_3=1.93$, $C_1=1.27$, $C_2=0.06$, $C_3=1.04$, $C_4=0.57$, $R_S=R_L=1$, $f_N=16\,\text{GHz}$, $R_N=50\Omega$).

Table 2. Reflection coefficient data of the lumped-element matching networks

Frequency (GHz)	$S_{22}^{(F)}(j\omega)$	$S_{11}^{(B)}(j\omega)$
6	−0.1021 − 0.0216i	0.3273 + 0.3870i
7	−0.1338 − 0.0072i	0.3854 + 0.3859i
8	−0.1668 + 0.0151i	0.4317 + 0.3737i
9	−0.1993 + 0.0458i	0.4644 + 0.3543i
10	−0.2296 + 0.0850i	0.4815 + 0.3313i
11	−0.2555 + 0.1324i	0.4810 + 0.3097i
12	−0.2753 + 0.1872i	0.4607 + 0.2971i
13	−0.2871 + 0.2482i	0.4222 + 0.3066i
14	−0.2894 + 0.3135i	0.3789 + 0.3549i
15	−0.2815 + 0.3811i	0.3638 + 0.4469i
16	−0.2631 + 0.4486i	0.4092 + 0.5505i

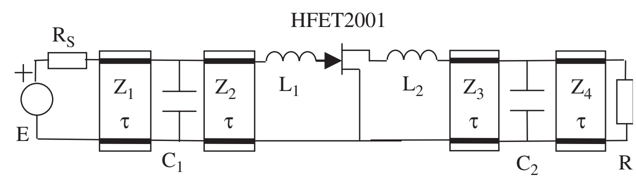

Fig. 4. Mixed-element microwave amplifier ($L_1=0.6088$, $L_2=1.2333$, $C_1=0.7766$, $C_2=0.9540$, $Z_1=1.1062$, $Z_2=0.7806$, $Z_3=2.8900$, $Z_4=2.2740$, $\tau=0.1042$, $R_S=1.0092$, $R_L=0.7618$, $f_N=16\,\text{GHz}$, $R_N=50\Omega$).

In the mixed-element matching networks, equal length of the transmission lines were fixed as 45° (half of the quarter wavelength) at the normalized frequency $f_0=1.2$, i.e., normalized delay length was fixed at $\tau=0.1042$.

For the front-end matching network, since the lumped-element prototype had a low-pass nature, $f^{(F)}(p,\lambda)$ was selected as $f^{(F)}(p,\lambda)=f_L^{(F)}(p)f_D^{(F)}(\lambda)=1\cdot(1-\lambda^2)=1-\lambda^2$. Then, by ad hoc choice of the initials, the complex quantities $h_L^{(F)}$ and $h_D^{(F)}$ were determined by means of gradient

Table 1. Scattering parameters of HFET2001

Frequency (GHz)	S_{11}	S_{12}	S_{21}	S_{22}
6	0.3719−0.7976i	0.0250 + 0.0433i	−1.1472+1.6383i	0.6583 − 0.2660i
7	0.2213−0.8259i	0.0304 + 0.0459i	−0.8649+1.6974i	0.6247 − 0.3047i
8	0.0723−0.8268i	0.0361 + 0.0479i	−0.5893+1.7114i	0.5889 − 0.3400i
9	−0.0424−0.8089i	0.0369 + 0.0473i	−0.3586+1.6873i	0.5587 − 0.3698i
10	−0.1507−0.7755i	0.0378 + 0.0466i	−0.1429+1.6338i	0.5271 − 0.3972i
11	−0.2266−0.7411i	0.0374 + 0.0470i	0.0136+1.5599i	0.5056 − 0.4242i
12	−0.2970−0.6996i	0.0369 + 0.0473i	0.1547+1.4719i	0.4827 − 0.4501i
13	−0.3669−0.6484i	0.0361 + 0.0479i	0.2861+1.4062i	0.4556 − 0.4636i
14	−0.4291−0.5906i	0.0353 + 0.0485i	0.4064+1.3293i	0.4282 − 0.4756i
15	−0.4956−0.5223i	0.0377 + 0.0529i	0.5294+1.2473i	0.3971 − 0.5083i
16	−0.5518−0.4468i	0.0402 + 0.0573i	0.6399+1.1545i	0.3523 − 0.5223i

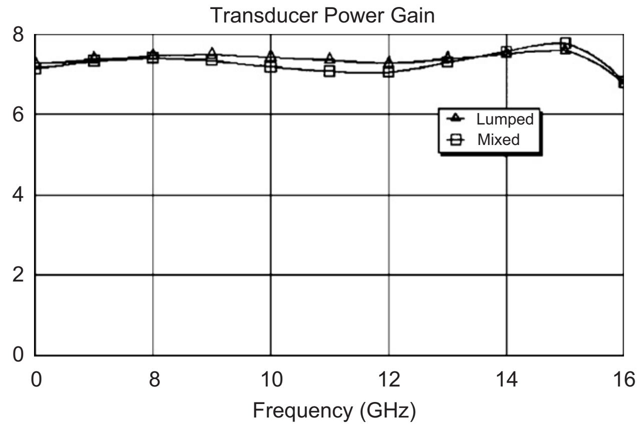

Fig. 5. *TPG* plots of lumped- and mixed-element amplifiers.

method, and by linear interpolation techniques, coefficients $\{h_{0L}^{(F,0)}, h_{1L}^{(F,0)}, \ldots, h_{n_pL}^{(F,0)}\}$ and $\{h_{0D}^{(F,0)}, h_{1D}^{(F,0)}, \ldots, h_{n_\lambda D}^{(F,0)}\}$ were determined which in turn yields the coefficients of the denominator polynomials $g_L^{(F)}(p) = \sum_{i=0}^{n_p} g_{iL}^{(F)} p^i$ and $g_D^{(F)}(\lambda) = \sum_{i=0}^{n_\lambda} g_{iD}^{(F)} \lambda^i$ by the energy conservation condition. Eventually, $S_{11}^{(F)}(p,\lambda) = S_{22}^{(F)}(p)$ was generated by synthesis of $S_L^{(F)}(p)$ and $S_D^{(F)}(\lambda)$ and by connecting the elements of $[L]$ and $[D]$ in sequential order as shown in Fig. 4. Thus, $S_{11}^{(F)}(p,\lambda) = \frac{h^{(F)}(p,\lambda)}{g^{(F)}(p,\lambda)}$ was computed with the coefficient matrices

$$\Lambda_h^{(F)} = \begin{bmatrix} 0.0046 & -0.1585 & -0.3511 \\ -0.0040 & 0.0357 & 0.0480 \\ 0.0009 & 0.0010 & 0 \end{bmatrix},$$

$$\Lambda_g^{(F)} = \begin{bmatrix} 1 & 2.0367 & 1.0599 \\ 0.0433 & 0.0876 & 0.0480 \\ 0.0009 & 0.0010 & 0 \end{bmatrix}.$$

In a similar manner, for the back-end matching network, $S_{11}^{(B)}(p,\lambda) = S_{11}^{(B)}(p) = \frac{h^{(B)}(p,\lambda)}{g^{(B)}(p,\lambda)}$ was computed with the coefficient matrices,

$$\Lambda_h^{(B)} = \begin{bmatrix} -0.1364 & 2.6155 & 0.1036 \\ 0.0205 & 0.0345 & 0.2451 \\ 0.0019 & 0.0054 & 0 \end{bmatrix},$$

$$\Lambda_g^{(B)} = \begin{bmatrix} 1.0093 & 3.3013 & 1.0053 \\ 0.0698 & 0.1707 & 0.2451 \\ 0.0019 & 0.0054 & 0 \end{bmatrix}.$$

After synthesizing the obtained reflection functions and connecting front-end matching network, active element (HFET2001) and back-end matching network in cascade, the single-stage microwave amplifier seen in Fig. 4 was obtained.

Consequently, the transducer power gain (*TPG*) performances of the lumped- and mixed-element amplifiers are shown in Fig. 5. Close examination of this figure reveals that the mixed-element amplifier constructed via modeling exhibits a similar gain performance.

The total number of lumped-elements in the lumped-element amplifier prototype is seven. Although it is four in the mixed-element amplifier, approximately the same transducer power gain curve is obtained.

5. Conclusion

In this paper, a practical method is proposed, to construct amplifiers in two kinds of elements via modeling. The proposed method consists of two phases. In Phase I, a lumped-element amplifier prototype is designed using the classical techniques. Then, the output and input reflection coefficient data of the prototype front-end and back-end matching networks are generated over the passband frequencies. In the second phase, the generated reflectance data are modeled as bounded real-input reflection functions in two complex variables, which in turn results in the desired mixed-element front-end and back-end matching networks.

It is exhibited that the proposed method provides very good initials, to further improve the amplifier performance by working on the element values. Therefore, it is expected that the proposed design procedure is used as a front-end for the commercially available CAD packages, to design practical broadband microwave amplifiers for wireless or in general microwave communication systems.

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