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Broadband Microwave Amplifier Design with Lumped Elements

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Abstract: This study introduces a broadband microwave amplifier design that utilizes the measured scattering parameters of active devices without assuming an initial topology for the matching networks or an analytic form of the system transfer function. The algorithm can be extended to design multistage broadband microwave amplifiers. An example is given to illustrate the application of the proposed method. It was found that the proposed method provides very good initials for CAD tools to further improve amplifier performance by working on the element values.

Keywords: microwave amplifiers, broadband, lumped elements, matching networks

1 Introduction

In the design of broadband microwave amplifiers, a fundamental problem is how to realize lossless front-end and back-end matching networks so that the transfer of power from source to load is maximized over a prescribed frequency band. In this case, the overall amplifier structure consists of cascaded lossless matching networks and active two-port.

For the characterization of cascaded structures, the scattering description is especially suitable. Since Simplified Real Frequency Technique (SRFT) employs the scattering parameters to optimize the transducer power gain (*TPG*) of a lossless matching system, it provides an easy and efficient tool for the design of matching two-ports in amplifier problems [1]–[4].

The lossless front-end and back-end matching networks for microwave amplifiers can be designed and optimized by a CAD tool. 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 proper choices of the matching network topologies are not known.

Different approaches have been proposed for the design of broadband amplifiers in the literature. In Refs. [5], [6], first the optimum input and output termination values for the active device are produced. Then, these termination values are modeled utilizing the proposed immitance modeling method to synthesize the front-end and back-end matching networks [7].

In Ref. [8], a genetic algorithm based method has been proposed. In Refs. [9–[11], lossless front-end and back-end matching networks have been designed via the proposed algorithms based on parametric approach and line segment method, respectively.

Also in Refs. [12] and [13], simplified real frequency technique has been adapted for the design of mixed lumped and distributed element and symmetrical mixed lumped and distributed element matching networks, respectively.

Now let us consider the classical double matching problem which can be defined as the power transfer from a complex generator to a complex load shown in Figure 1. Transducer power gain (*TPG*) can be expressed in terms of the real and imaginary parts of the load impedance $Z_L = R_L + jX_L$ and those of the back-end impedance $Z_2 = R_2 + jX_2$, or in terms of the real and imaginary parts of the generator impedance $Z_G = R_G + jX_G$ and those of the front-end impedance $Z_1 = R_1 + jX_1$ of the matching network as follows

$$TPG(\omega) = \frac{4R_\alpha R_\beta}{(R_\alpha + R_\beta)^2 + (X_\alpha + X_\beta)^2} \tag{1}$$

Here if $\alpha = 1$, $\beta = G$, and if $\alpha = 2$, $\beta = L$.

The objective in broadband matching problems is to design the lossless matching network in such a way that the *TPG* given by (1) is maximized inside the interested frequency band. So the matching problem in this formalism is reduced to the determination of a realizable impedance function Z_1 or Z_2. Once Z_1 or Z_2 are determined properly, the lossless matching network can be easily synthesized.

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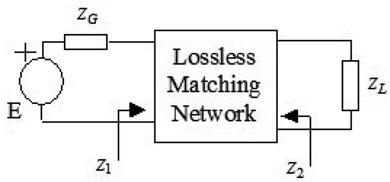

Figure 1: Double matching arrangement.

Based on (1), a new approach for the design of broadband matching networks was proposed in Ref. [14]. For cascaded lossless matching networks and active two-port, if a transducer power gain expression based on impedances similar to (1) can be found, the approach proposed in Ref. [14] can be extended to design broadband amplifiers. So in the next sections, firstly the *TPG* expression based on impedances is given and then the algorithm for the design of broadband amplifiers with lumped elements is explained.

2 Broadband amplifier design

Here A is the active device, and N_1 and N_2 are the front-end and back-end matching networks, respectively (Figure 2).

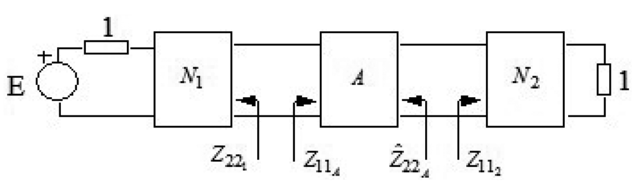

Figure 2: Single stage amplifier.

Assume that the scattering parameters of the active device and lossless two-ports N_1 and N_2 are denoted by A_{ij}, S_{ij_1} and S_{ij_2}, respectively. Then transducer power gain of the configuration can be written as follows;

$$TPG(\omega) = |S_{21_1}|^2 \frac{|A_{21}|^2 |S_{21_2}|^2}{X(\omega)} \tag{2}$$

where

$$X(\omega) = X_1(\omega) \cdot X_2(\omega) = \frac{4|Z_{22_1} + Z_{11_A}|^2}{|Z_{22_1} + 1|^2 \cdot |Z_{11_A} + 1|^2} \cdot \frac{4|\hat{Z}_{22_A} + Z_{11_2}|^2}{|\hat{Z}_{22_A} + 1|^2 \cdot |Z_{11_2} + 1|^2} \tag{3}$$

Here Z_{11_A} denotes the input impedance of the active device when its output is terminated by a virtual one ohm resistor, Z_{22_1} is the output impedance of the lossless two-port N_1 when its input is terminated by a one ohm source resistor, Z_{11_2} is the input impedance of the lossless two-port N_2 when its output is terminated by a one ohm load resistor and $\hat{Z}_{22_A}$ is the output impedance of the active device when its input is terminated by the lossless two-port N_1.

Let us rewrite the eq. (2) in the following form:

$$TPG(\omega) = T_1(\omega) \frac{|S_{21_2}|^2}{X_2(\omega)} \text{ and, } T_1(\omega) = |S_{21_1}|^2 \frac{|A_{21}|^2}{X_1(\omega)} \tag{4}$$

In this form, T_1 represents the transducer power gain of the structure shown in Figure 3(a), when the output of the active device is terminated by a one ohm resistor. So, this is a single matching problem which can be defined as the power transfer from a purely resistive generator to a complex load, where the load is the active device. *TPG* in (4) denotes the transducer power gain of the structure seen in Figure 3(b). The problem is again a single matching problem which can be described as the design of the matching network N_2 between a complex generator (output of the active device) and a resistive load.

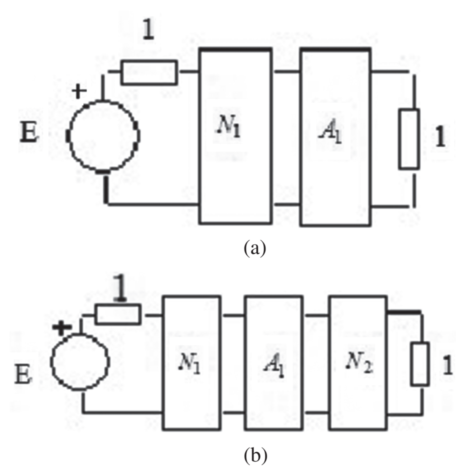

Figure 3: Computation steps for designing broadband single stage amplifier. (a) Design of front-end matching network. (b) Design of back-end matching network.

This procedure can easily be extended to the design of multistage amplifiers (Figure 4), where it remains basically unchanged. In this case, the designer needs to apply the design steps sequentially to each stage of the amplifier.

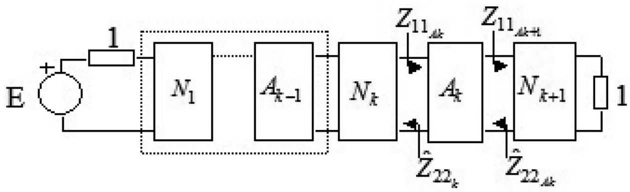

Figure 4: Computation steps for designing a broadband multistage amplifier.

As depicted in Figure 4, for the first k-stages of the multistage amplifier configuration, TPG can be written in the same form of (2), i. e.,

$$TPG_k(\omega) = T_{k-1}\frac{|A_{21_k}|^2|S_{21_{k+1}}|^2}{X_k(\omega)} \tag{5}$$

where

$$X_k(\omega) = X_{1_k}(\omega)\cdot X_{2_k}(\omega) = \frac{4|\hat{Z}_{22_k} + Z_{11_{Ak}}|^2}{|\hat{Z}_{22_k}+1|^2\cdot|Z_{11_{Ak}}+1|^2}\cdot\frac{4|\hat{Z}_{22_{Ak}} + Z_{11_{k+1}}|^2}{|\hat{Z}_{22_{Ak}}+1|^2\cdot|Z_{11_{k+1}}+1|^2}$$

In (5), T_{k-1} is the gain of the first $(k-1)$ stages with resistive terminations. A_{ij_k} and $S_{ij_{k+1}}$ denote the scattering parameters of the *kth* active device and the next matching network, respectively. It must be noted that $\hat{Z}_{22_k}$ and $\hat{Z}_{22_{Ak}}$ are the output impedances of the *kth* matching network and the *kth* active device, respectively, when the previous stages are all connected. They can be calculated using the information obtained from the previous stages.

In this method described above, the stability considerations of the amplifier are not taken into account. But in actual amplifier designs, high gain active devices may have high input reflection coefficients, which may cause an unstable region of operation. Therefore it is necessary to use lossy sections or feedback circuits to stabilize the active device. In the light of this explanation, the active device A_i (Figures 2–4) can be assumed to represent a stabilized transistor module, including the transistor as well as the feedback circuitry.

As a result, the following algorithm can be proposed to design broadband amplifiers with lumped elements. But the same algorithm can easily be adapted to design distributed or mixed element broadband amplifiers as well.

3 Proposed algorithm

Inputs:

A_{ij}: Scattering parameters of the active device.

$\omega_{i(measurement)}$: Measurement frequencies, $\omega_{i(measurement)} = 2\pi f_{i(measurement)}$.

f_{norm}: Normalization frequency.

R_{norm}: Impedance normalization number in ohms.

$h_{0_1}, h_{1_1}, h_{2_1}, \ldots, h_{n_1}$ and $h_{0_2}, h_{1_2}, h_{2_2}, \ldots, h_{m_2}$: Initial real coefficients of the polynomial $h_1(p)$ and $h_2(p)$ describing the lossless two-ports N_1 and N_2, respectively. Here n and m are the degrees of the polynomials which are equal to the number of lossless lumped elements in the lossless two-ports N_1 and N_2, respectively. The coefficients can be initialized as ± 1 in an ad hoc manner, or the approach explained in Ref. [15] can be followed.

$f_1(p)$ and $f_2(p)$: Monic polynomials constructed on the transmission zeros of the lossless two-ports N_1 and N_2, respectively. For practical problems, the designer may use the following form of $f_i(p)$

$$f_i(p) = p^{m_1}\prod_{k=0}^{m_2}(p^2 + a_k^2) \tag{6}$$

where m_1 and m_2 are nonnegative integers and a_k's are arbitrary real coefficients. This form corresponds to ladder type minimum phase structures, the transmission zeros of which are on the imaginary axis of the complex p-plane.

Shortly, the user must supply only the transmission zeros of the front-end and back-end matching networks, it is not necessary to completely define the matching network topologies, it is a natural consequence of the proposed algorithm.

T_0: Desired flat transducer power gain level which can be estimated as the mean value of the power gain that can be achieved under perfect match assumption at the input of the resistively terminated active device, i. e. $S_{22_1} = A_{11}^*$. Thus an approximate value of the gain level is obtained as

$$T_0 \approx \left\{\frac{|A_{21}|^2}{1-|A_{11}|^2}\right\}. \tag{7}$$

Outputs:

Analytic forms of the input reflection coefficients of the lossless matching networks N_1 and N_2 given in the Belevitch form of $S_{11_1}(p) = h_1(p)/g_1(p)$ and $S_{11_2}(p) = h_2(p)/g_2(p)$, respectively. It should be pointed out that this algorithm determines the coefficients of the polynomials $h_1(p)$, $g_1(p)$, $h_2(p)$ and $g_2(p)$, which in turn optimizes system performance.

Circuit topologies of the lossless matching networks with element values: The circuit topologies and element values are obtained as the result of the synthesis of $S_{11_1}(p)$ and $S_{11_2}(p)$. Synthesis is carried out in the Darlington sense. That is, $S_{11_i}(p)$ is synthesized as a lossless two-port which is the desired matching network [16]. Also the synthesis process can be carried out by using impedance based Foster or Cauer methods via $Z_{11_i}(p)=(1+S_{11_i}(p))/(1-S_{11_i}(p))$ as explained in Ref. [17].

Computational steps

Step 1: Normalize the measurement frequencies with respect to f_{norm} and set all the normalized angular frequencies

$$\omega_i = f_{i(measurement)}/f_{norm}.$$

Step 2: Calculate the desired transducer power gain level (T_0) via (7).

Step 3: Obtain the strictly Hurwitz polynomials $g_1(p)$ and $g_2(p)$ from the Feldtkeller equation;

$$g_i(p)g_i(-p)=h_i(p)h_i(-p)+f_i(p)f_i(-p).$$

Then calculate the scattering parameters via

$$S_{11i}(p)=h_i(p)/g_i(p), S_{12i}(p)=\mu_i f_i(-p)/g_i(p),$$
$$S_{21i}(p)=f_i(p)/g_i(p), S_{22i}(p)=-\mu_i h_i(-p)/g_i(p),$$

Step 4: Calculate the $X(\omega)$ values via (3). Here $Z_{22_1}=\frac{1+S_{22_1}}{1-S_{22_1}}$, $Z_{11_A}=\frac{1+A_{11}}{1-A_{11}}$, $Z_{11_2}=\frac{1+S_{11_2}}{1-S_{11_2}}$ and $\hat{Z}_{22_A}=\frac{1+\hat{A}_{22}}{1-\hat{A}_{22}}$ where $\hat{A}_{22}=A_{22}+\frac{A_{21}A_{12}S_{22_1}}{1-A_{11}S_{22_1}}$.

Step 5: Calculate the transducer power gain ($TPG(\omega)$) via (2).

Step 6: Calculate the error via $\varepsilon(\omega)=T_0-TPG(\omega)$, then $\delta=\sum|\varepsilon(\omega)|^2$.

Step 7: If δ is acceptably small, stop the algorithm and synthesize $S_{11_1}(p)$ and $S_{11_2}(p)$. Otherwise, change the initialized coefficients of the polynomials $h_1(p)$ and $h_2(p)$ via any optimization routine and return to step 3.

4 Example

In this example, the design of a single stage FET amplifier is considered. The active device is HFET2001. The magnitude (mg) and phase (ph) data for the scattering parameters of FET are given in Table 1.

The front-end and back-end matching networks are assumed to be of low-pass type, so the polynomials $f_1(p)$ and $f_2(p)$ are selected as $f_1(p)=1$ and $f_2(p)=1$, respectively.

Table 1: Scattering parameters of HFET2001.

Freq.GHz	S_{11}		S_{21}		S_{12}		S_{22}	
	mg	ph	mg	ph	mg	ph	mg	ph
6	0.88	−65	2.00	125	0.05	60	0.71	−22
7	0.86	−75	1.91	117	0.06	57	0.70	−26
8	0.83	−85	1.81	109	0.06	53	0.68	−30
9	0.81	−93	1.73	102	0.06	52	0.67	−34
10	0.78	−101	1.64	95	0.06	51	0.66	−37
11	0.78	−107	1.56	89.5	0.06	52	0.66	−40
12	0.76	−113	1.48	84	0.06	52	0.66	−43
13	0.75	−120	1.44	78.5	0.06	53	0.65	−46
14	0.73	−126	1.39	73	0.06	54	0.64	−48
15	0.72	−134	1.35	67	0.07	55	0.65	−52
16	0.71	−141	1.32	61	0.07	55	0.63	−56

The initialized polynomials $h_1(p)$ and $h_2(p)$ are $h_1(p)=-p^3+p^2-p$ and $h_2(p)=-p^3+p^2-p$. So there are three elements in both the front-end and back-end matching networks. The desired transducer power gain level is $T_0=7.72$ (or 8.87dB) from (7). f_{norm} and R_{norm} are selected as $f_{norm}=16\,GHz$ and $R_{norm}=50\ \Omega$. The error $\delta=\sum|T_0-TPG(\omega)|^2$and the total number of iterations are selected as zero and 1,500, respectively. After running the proposed algorithm, all iterations are completed in a few seconds and the following descriptive polynomials are obtained:

$$h_1(p)=0.5394\,p^3-0.5334\,p^2+0.0035\,p$$
$$g_1(p)=0.5394\,p^3+1.4568\,p^2+1.7069\,p+1$$
$$f_1(p)=1$$
$$h_2(p)=2.4124\,p^3+0.8281\,p^2+1.9002p$$
$$g_2(p)=2.4124\,p^3+2.3180\,p^2+2.8717p+1$$
$$f_2(p)=1$$

After synthesizing the obtained scattering parameters or the corresponding impedance functions, the broadband amplifier seen in Figure 5 is obtained. If the element values are denormalized by using the selected normalization frequency ($f_{norm}=16$ GHz) and impedance normalization number ($R_{norm}=50\ \Omega$), the following real element values are obtained: $L_1=0.2696$ nH, $C_1=0.3389$ pF, $L_2=0.58106$ nH, $L_3=1.6106$ nH, $C_2=0.19327$pF, $L_4=0.76275$ nH.

The same example is solved in Ref. [18] via SRFT, where three and four lumped elements are used in front-end and back-end matching networks, respectively.

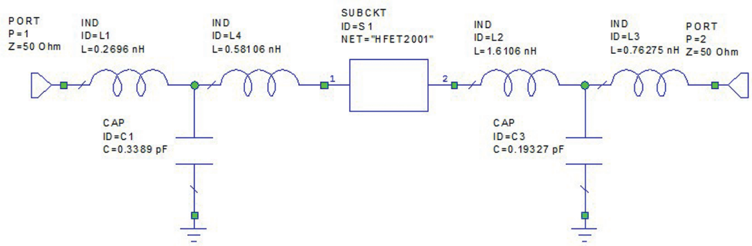

Figure 5: Design of broadband amplifier with front-end and back-end matching networks.

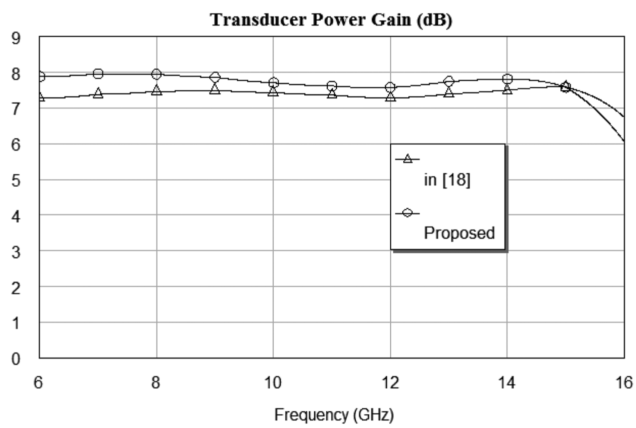

Figure 6: Performance of the broadband amplifier designed with lumped elements.

The performance of the designed amplifier seen in Figure 5 has been simulated via Microwave Office by AWR [19] as seen in Figure 6. Usually the transducer power gain of the amplifier is further improved via optimization utilizing commercially available design packages [19], [20]. But in this example no further improvement has been obtained, since the initial performance of the designed amplifier is very close to the performance that can be achieved via the CAD tool employed. For comparison purposes, the performance of the amplifier and the performance obtained in Ref. [18] are depicted in Figure 6. It is seen that via the proposed algorithm, a higher transducer power gain level is obtained (except the last 1 GHz region) by using fewer elements in the matching networks.

5 Conclusion

A real frequency technique has been proposed for the design of broadband microwave amplifiers and with this approach, the front-end and back-end matching networks have been designed simultaneously. While designing the front-end and back-end matching networks, the output of the active device is assumed to be matched and the front-end matching network is connected to the input of the active device, respectively.

Lastly, the front-end and back-end matching networks are synthesized as lossless two-ports. The actual performance of the amplifier may be improved by means of a commercially available CAD tool.

The advantages of the proposed method can be explained as follows: The polynomials $f_i(p)$ are constructed by using the transmission zeros of the matching networks, so they are under the control of the designer. Also as explained, the algorithm can be extended for the design of multistage broadband microwave amplifiers.

This method can be used to generate proper matching network topologies and initial element values for real amplifier designs, since it does not consider parasitic elements. An example was presented in this study for the construction of a broadband amplifier with lumped elements. It was shown that the proposed method generates very good initials. Therefore, it is expected that the proposed algorithm can be used as a front-end for commercially available CAD tools to design practical broadband amplifiers for microwave communication systems.

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