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Design of broadband antenna matching networks

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The purpose of this paper is to provide the necessary background to design broadband antenna matching networks employing real the frequency techniques at radio frequencies. The importance of the topic stems from the recent advances taking place in the conceptual design and manufacturing technologies of the next generation wireless and mobile communication systems which will operate over ultra-wide frequency bands. In fact, for all communication systems, construction of wide-band power transfer networks is inevitable. Therefore, in this manuscript, modern methods, so called real frequency techniques to design broadband antenna matching network are covered, a design roadmap is given with relevant recommendations. Practical examples are presented to design antenna matching networks and microwave amplifiers for various applications.
Key Words-Antenna matching networks, Power transfer networks, Switches for antenna arrays, Microwave amplifiers.

1 Introduction

It is inevitable to deliver Radio Frequency (RF) signals to communication systems over a prescribed frequency band. With the advancement of the production technologies for mobile communication systems, carrier frequencies have been shifting beyond X-band and demands for ultra wideband systems have been increasing on the market place. Therefore, design of wide band power transfer networks is an essential problem for microwave engineers. Design of broadband antenna matching networks for communication systems may be understood in the following manner. Referring to Figure 1, at one end, per say, on the left, one has an idealized signal generator (perhaps, the Thévenin equivalent of a sub-system) with internal positive real impedance [Z_G] whose power will be delivered to a communication system which is placed on the right, over a wide frequency band. In this case, a proper power transfer network [N] will be inserted in between the signal generator [$E_G = E_m e^{j\omega t}$ *Volts*; $\omega_1 \leq \omega \leq \omega_2$] and the input port of the communication system. Obviously, power is money. Therefore, the ideal situation is "to be able to deliver the maximum or equivalently, the available power of the signal generator to the port under consideration over the band of operation".

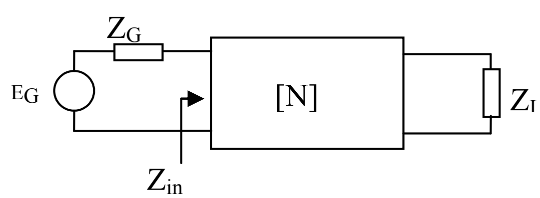

Fig. 1. Power transfer problem

At this point, it is meaningful to consider the input port of the communication system as Positive Real (PR) impedance [Z_L] which terminates the matching network [N]. Since it is desired to deliver the available power of the signal generator [E_G] to the load [Z_L] without any loss then, it is mandatory to design the power transfer network [N] as a lossless two-port. This practical fact tremendously simplifies the design

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problem. Over the real frequencies $(\omega = 2\pi f)$, the maximum power transfer is achieved when the generator's internal impedance $Z_G(j\omega) = R_G(\omega) + jX_G(\omega)$ sees its complex conjugate. In this manner, [N] has to be designed in such a way that, the load impedance $Z_L(j\omega) = R_L(\omega) + jX_L(\omega)$ is transformed to an input impedance $Z_{in}(j\omega) = R_{in}(\omega) + jX_{in}(\omega)$ yielding the complex conjugate of the generator impedance. Literally,

$$Z_{in}(j\omega) = Z_G^*(j\omega) = Z_G(-j\omega) \tag{1a}$$

or equivalently,

$$R_{in}(\omega) = R_G(\omega); \quad X_{in}(\omega) = -X_G(\omega) \tag{1b}$$

In this case, maximum transferable power which is defined as the available power of the generator is given by

$$P_A(\omega) = \frac{E_m^2}{4R_G(\omega)} \tag{2}$$

Let $P_L(\omega)$ designates the power delivered to the load. Based on the above discussion, ideally, $P_L = P_A$, or equivalently the transducer power gain $T(\omega) = P_L/P_A = 1$, over the band of interest. In this case, replacing $j\omega$ by the complex variable $p = \sigma + j\omega$, the lossless two-port [N] must yield the input impedance $Z_{in}(p) = Z_G(-p)$ over the specified frequencies. If this is the case, $Z_{in}(p)$ is forced to be non realizable since $Z_G(p)$ is positive real. It is concluded that there exist no ideal power transfer network which yields continuous flat gain $T(\omega) = P_L/P_A = 1$ over the entire frequency band. Therefore, one has to look into reasonable solutions.

In the literature, design problems of power transfer networks are classified under three major headings [1]: Filter or insertion loss problem, Single Matching Problem, and Double Matching Problem. The filter or insertion loss problem is defined as the power transfer from a resistive generator $(Z_G = R_G =$ Constant$)$ to a resistive load $(Z_L = R_L =$ constant$)$ over the specified frequency band. The theory is well established to design RF filters in one kind of element namely, either with lumped elements or with equal length transmission lines or so called commensurate transmission lines or Unit Elements [UE]. Single Matching is the power transfer from a resistive generator to a complex load whereas Double Matching is the power transfer from a complex generator to a complex load. In theory, it has been shown that for single and double matching problems, over a prescribed band B, achievable, "maximum flat gain" is bounded by a level T_0. Over frequency band B, the upper bound T_0 is dictated by the generator and the load networks. For example, in a single matching problem where the load network consist of a single capacitor C in parallel with a resistance R then, the ideal flat gain level over the frequency band $(B = \omega_2 - \omega_1)$ is bounded by $T_0 = 1 - e^{-2\pi/RCB}$. For double Matching problems, let T_{0G} and T_{0L} be the ideal flat gain level imposed by the generator and the load networks, respectively then, the flat gain level T_0 of the doubly terminated system is given as the minimum of $\{T_{0G}, T_{0L}\}$.

Design of matching networks via analytic theory is not easy. Details are omitted here. For interested readers, references [1-2] are cited. It is well known that, beyond very simple problems, analytic theory is not accessible. If the solution exists, the resulting matching networks are highly complicated to be manufactured. However, all the above mentioned difficulties may be by passed using the proper CAD tools and Real Frequency Techniques (RFT).

2 CAD Tools and the Real Frequency Techniques for Designing Antenna Matching Networks and Amplifiers

Commercially available software packages such as AWR or Microwave Office of Applied Wave Research Inc. (www.appwave.com), EDS of Eagle Ware-Elanix (www.eagleware.com), EDL or Ansoft Designer of Ansoft Corp. (www.ansoft.com/products.cfm), ADS of Agilent Technologies (www.home.agilent.com) are excellent CAD tools to end up with the final design of matching networks and amplifiers. Once the matching network's topology with element values are provided; these software tools perfectly simulates the physical layout of the complete systems under consideration. Certainly, element values of the selected topology can be determined by means of non-linear optimization routines to reach to pre-defined targets. At this point it is crucial to know that optimization is heavily non-linear in terms of the element values of the selected topology. In many wide band design problems, even though the designer may select the circuit topology properly, optimization becomes almost impossible. In this regard, there are three major questions to be answered: What is the attainable maximum transducer power gain T_0 over the prescribed frequency band B? What is the optimum circuit topology to reach T_0? How can we initialize the element values of the selected topology to end up with the final design?

It has been experienced that real frequency modeling and design techniques can satisfactorily answer the above questions and provides excellent initials to the commercially available packages.

In the following, a road map is proposed to design wide-band matching networks.

3 Roadmap to Design Broadband Antenna Matching Networks

A. Design of Matching Networks for Antennas:

Step 1: Measure your antenna over the band of interest. Here, it is presumed that the measurements give the real normalized (50 ohms) reflection coefficient $S_L(j\omega) = \rho_L(\omega)e^{j\phi_L(\omega)}$ of the one port device under consideration.

Step 2: Normalize your band of operation with respect to upper end of the band. In this case, the upper edge of the band is fixed at 1 ($i.e.\omega_c = 1$).

Step 3: Assume that the load sees perfect match at the input of the matching network. In other words, input reflection coefficient of the equalizer $S(j\omega)$ is equal to complex conjugate of the load reflection coefficient yielding $S(j\omega) = [S_L(j\omega)]^*$ (see Fig. 2).

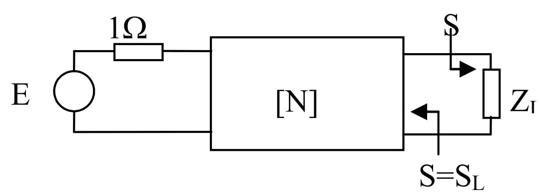

Fig. 2. Perfect match at the input of the matching network

Step 4: Since the input reflection coefficient of the matching network is assumed to be known over the specified frequencies, then the corresponding impedance data $Z(j\omega) = 1 + S/1 - S$ (or equivalently the admittance $Y = 1 - S/1 + S$ can be modeled by means of the linear interpolation technique given by [3], yielding the analytic form of the impedance $Z_M(p)$. On the other hand, the reflection coefficient $S(j\omega) = \rho_L(\omega)e^{-j\phi_L(\omega)}$ can also be modeled using the fixed point iteration technique given by [4] yielding the analytic form of the reflection coefficient $S_M(p)$. It should be mentioned that the fixed point iteration is a non-linear interpolation algorithm; and it requires initial start. Nevertheless, the method initializes itself properly with guaranteed convergence. Therefore, it is as efficient as linear interpolation of immitance functions with even better numerical stability.

Step 5: Using Darlington's procedure, synthesize the modeled impedance Z_M or the input reflection coefficient S_M. This process yields the matching network topology with element values.

Remark: Antenna matching networks consist of distributed elements, are synthesized using Richard Extraction procedure.

Step 6: Compute and plot the transducer power gain of the initially matched system. Determine the minimum T_{min} and the maximum T_{max} values of the transducer power gain within the operational band.

Then, set the maximum attainable flat gain level at the value of $T_0 \geq T_{max} + T_{min}/2$. It should be noted that "maximum attainable flat-gain level (MAFGAL)" or T_0 can also be approximately determined using the Real Frequency Line Segment Technique of Carlin [7]. In this technique, the highest value of the flat transducer power gain is obtained by sweeping MAFGAL starting from $T_{max} = 1$ down to an acceptable level T_0 for which TPG is reasonably flat over the band of operation. For many practical applications, fluctuations around the flat gain level of –1dB is acceptable. In this case, the transducer power gain (TPG) is expressed with the ripple factor ε(dB) over the pass band such that $10\log_{10}(TPG) = 10\log_{10}(T_0) \pm \varepsilon$ where $\varepsilon \leq 0.5$dB.

Step 7: Feed the resulting circuit topology to one of the above mentioned CAD tools with initial element values obtained as above then, using the target value T_0 found in Step 6, determine the final element values and the physical layout of the matched system via optimization.

B. Recommendations

Once the initial matching networks are obtained as in Step 6, the transducer power gain of the matched structure can further be improved utilizing the Simplified Real Frequency Technique (SRFT) [5, 6]. It has been experienced that SRFT provides excellent initial guess for the commercially available CAD packages.

In practice, it is always preferable to work with "easy to manufacture type of matching networks". If this is the case, most of the applications demand "low-pass" type circuit topologies. Therefore, if lumped elements are utilized in the matching network, then, the numerator polynomial of the transfer scattering coefficient $S_{21} = f/g$ is selected as $f(p) = 1$. If distributed elements are utilized, then, it is always preferable to work with cascaded connection of unit elements yielding $f(\lambda) = (1 - \lambda^2)^{n_\lambda/2}$; where $\lambda = \tanh(p\tau) = \Sigma + j\Omega$ is the conventional Richard variable.

When working with equal length transmission lines in the matching networks, it is recommended to fix the normalized delay length at $\tau \approx \pi/2/1.2.\omega_c$ which corresponds to a delay length of 90° at the frequency $\omega_e = 1.2\omega_c$. In this representation, ω_c designates the upper edge or the cut-off frequency of pass-band and it is usually set to unity by normalization (i.e. $\omega_c = 1$). In this case, normalized delay length is fixed at $\tau \approx 1.31$. However, delay length τ may as well be selected as a parameter to be determined via optimization to improve the transducer power gain. In this case τ may vary between 0.3 and 1.7.

If the matching problem demands a band-pass circuit configuration then, $f(p) = p^k$ or $f(\lambda) = \lambda^k(1 - \lambda^2)^{n_\lambda/2}$; is selected to design matching network with lumped or distributed elements. For many practical cases it sufficient to set $k = 1$ or 2. It should also be noted that if the production technology permits, matching networks may as well be designed with mixed lumped and distributed elements.

C. Design of Matching Networks for Active Devices:

Design of Front-End [FE] and Back-End [BE] matching networks for active devices can be carried out in a similar manner as explained in the above road map. In this case, first the front-end matching network is constructed while the out put port is terminated in its normalization resistance. Then, the back-end matching network is constructed while the input is terminated with the previously constructed front end matching network.

In the following section, three examples are given. In the first one, gain bandwidth limits of an 0.18μ C-MOS FET is investigated by designing proper front and back-end matching networks for a single stage amplifier. In the second one, front and back-end matching networks are designed for an antenna switch which will be used for MIMO applications over 17 GHz-23 GHz. Finally, in the third example, design of a single matching network for a dual band PIFA is presented. The antenna is utilized in cellular hand sets over the bands of 824 MHz-960 MHz and 1710 MHz-1990 MHz.

4 Examples

EXAMPLE 1: In this example, gain bandwidth limits of an 0.18µ Si-CMOS FET was investigated over the frequency bands of B_1 = 10 GHz – 2 GHz; B_2 = 10 GHz – 1 GHz; and B_3 = 10 GHz – 450 MHz; by designing single stage amplifiers employing the road map presented in this paper. The CMOS FETs was designed using the Si-VLSI processing technology introduced by the VDEC Foundry of Tokyo University. Performance summary of the amplifiers are given as follows.

Design 1: This amplifier covers the frequency band of B_1 = 10 GHz – 2 GHz. It was experimented that in the front and back-end matching networks, 5-element lowpass LC ladders yield almost optimum solution. Thus, the average transducer power gain of the amplifier is found as 10dB with ±1.57dB fluctuations.

Design 2: The frequency band of Design 1 was extended down to 1 GHz (i.e.B_2 = 10 GHz – 1 GHz). Employing the circuit topology with element values obtained in Design-1 as initials for AWR, new element values of the amplifier were determined for various flat gain targets. Eventually, it was converged to the average transducer power gain of 6.46dB with ±0.895dB fluctuations. As expected, extension of bandwidth down to 1GHz results in 3 dB gain loss as compared to Design 1.

Design 3: Using the result of Design 2 as initials for AWR, the frequency band of the new amplifier was shifted down to 450 MHz. Thus, the amplifier covers the frequency band of B_3 = 10 GHz – 450 MHz. The new design yields the average gain of 6.06 dB with a ripple of ±0.511dB.

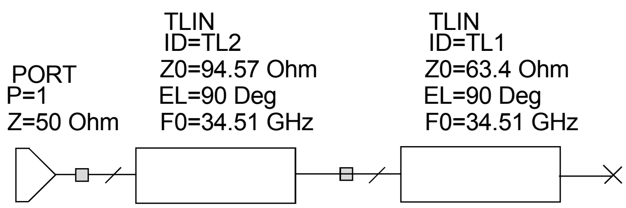

Fig. 3a. Final front end matching network

EXAMPLE 2: For this example, the measured 50 ohms normalized scattering parameters of an antenna switch which may be utilized to construct smart antenna system was provided by Ilmeanu Technical University of Germany. The switch is considered as a two-port passive device when it is on. In this state, connection loss varies between –1.72dB (at 17 GHz) to –4.5dB (at 23 GHz) which needs to be equalized over the band of 17 GHz-23GHz. When it is off, it yields approximately –25dB isolation loss which is reasonable for wireless communication antenna array applications. Utilizing the measured data, first, the front and the back end matching networks were constructed under the perfectly matched conditions. Then, employing SRFT, initial designs were improved. Eventually, using AWR, physical dimensions of the matched structure was obtained. Final layout is given in Figure 3 and performance of the matched switch is depicted in Figure 4.

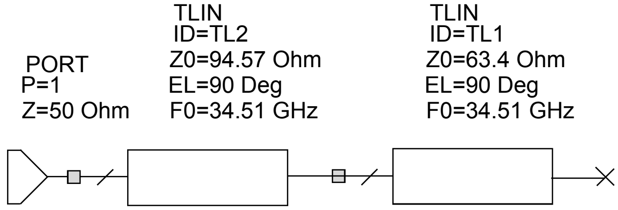

Fig. 3b. Final back end matching network

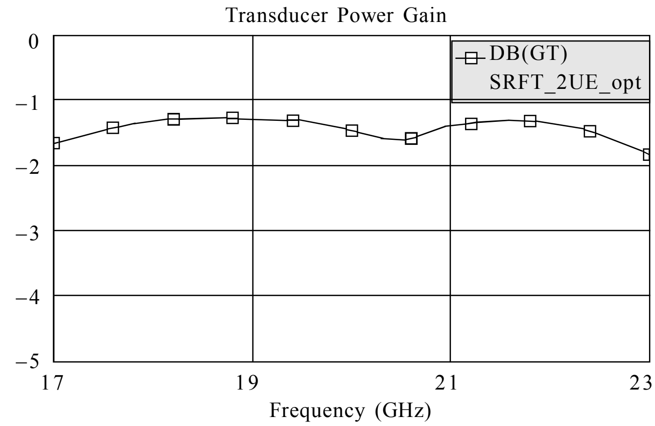

Fig. 4. Plot of the transducer power gain of the finally matched antenna switch

It is found that the maximum of the TPG is T_{max} = – 1.263 dB and the minimum is T_{min} = –1.584 dB yielding the average value of T_0 = –1.4235 ± 0.160 dB.

Clearly, matching networks designed for the switch significantly improves the loss over the pass-band from –4.5 dB to –1.4235 dB with a small ripple factor of ε = ±0.160 dB.

*EXAMPLE 3***:** In this example, the above outlined process is implemented to design a double band PIFA to be matched to 50 ohm over the bands of B_1 = (824 – 960) MHz and in B_2 = (1710 – 1990) MHz.

Measured data for the PIFA is given in Table 1.

Table 1. Measured Input Reflection Coefficient of the PIFA Normalized with respect to R_0 = 50 Ohm: $S_{22} = \rho_{22} e^{\phi_{22}}$ Over the bands of B_1 and B_2

	B_1: 0.824-0.960 GHz					B_2: 1.710-1.990 GHz						
F(GHz)	0.80	0.85	0.90	0.95	1.0	1.70	1.75	1.80	1.85	1..90	1.95	2.00
ρ_{22}	0.916	0.752	0.289	0.3618	0.671	0.57	0.36	0.289	0.447	0.608	0.716	0.785
ϕ_{22} (Degree)	33.9	10	–39.9	130.3	98	–44.7	–86.9	–165	137.3	109.8	93.2	81.9

Utilizing ***SRFT***, it was possible to design the matched antenna system with an average gain of –1.65 ± 0.4 dB in B_1 and –2.2 ± 0.5 dB in B_2. Thus, employing ***SRFT,*** a seven element L-C ladder matching network was constructed as shown in Fig. 5.

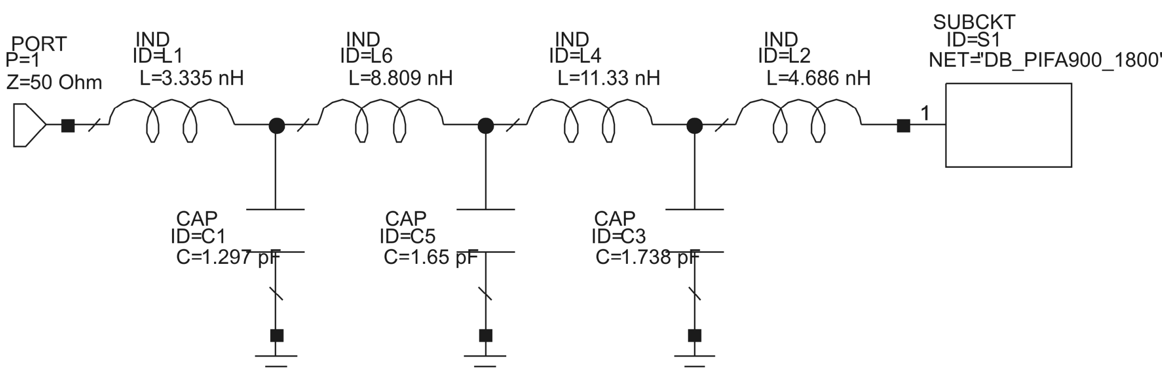

Fig. 5. Matching Network for the PIFA

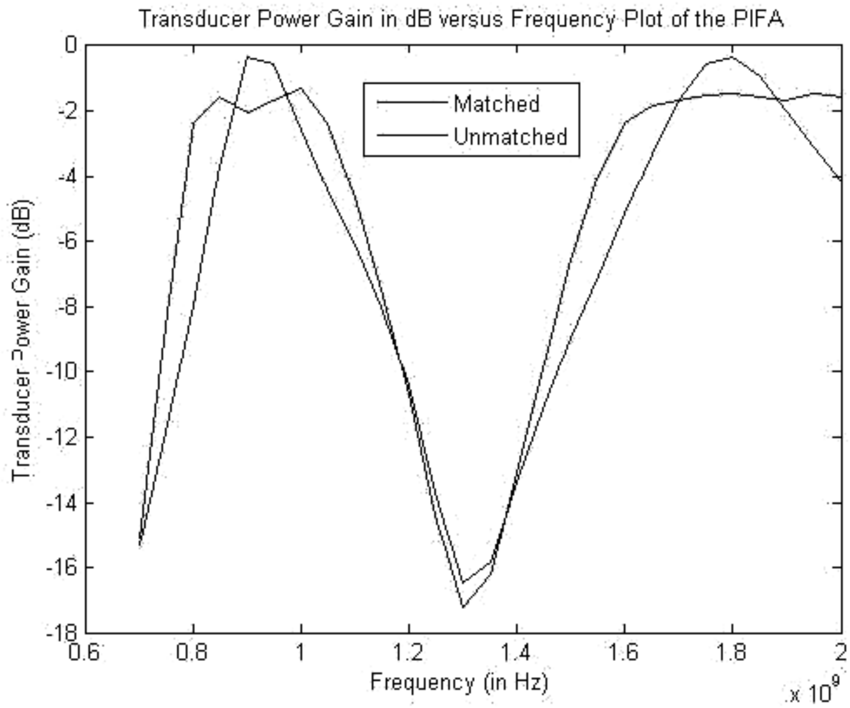

Fig. 6. Comparison of the Transducer Power Gain Characteristics of the matched and unmatched PIFA

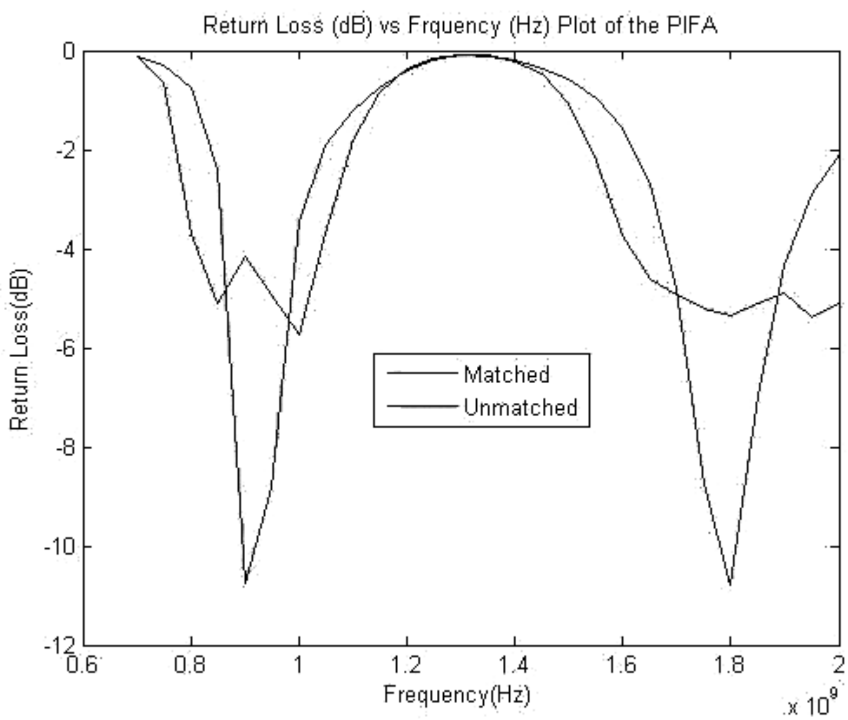

Fig. 7. Comparison of the return loss characteristics of the matched and unmatched PIFA

For comparison purposes, matched and unmatched system transducer power gains and return losses are depicted in Fig. 6 and Fig. 7 respectively. Measurement results are given in Fig. 7. If band width of the antenna is taken over frequencies for which TPG is over –3dB, then careful examination of Fig. 6 reveals that matching network design by using SRFT has increased the bandwidth of PIFA more than 25% in B1. Furthermore, gain variations have been significantly reduced by using a properly designed matching network. For example, in B_1, matched PIFA yields an average gain of –1.65dB with less than ± 0.4 dB fluctuations. In the same band, unmatched antenna gives an average gain of –5.25dB with huge fluctuation of ±2.65dB. In other words, PIFA does provide the desired bandwidth. As it is seen in Figures 6 and 7, expected and measured performances of the matched PIFA are in good agreement. Thus, we can confidently state that matching network designed employing SRFT significantly improves both the bandwidth and the transducer power gain performance of the PIFA under consideration.

5. Conclusion

This manuscript provides a road-map to design practical matching networks for antennas, antenna switches and microwave amplifiers utilized for smart wireless communication systems. It is recommended that, the designer first generates an initial matching network topology with element values by using the modeling techniques based on the perfect matched assumption of the measured data. Then, utilizing real frequency techniques performance of the matched system is improved. Eventually, employing commercially available CAD packages, physical dimensions of the matching networks is determined via optimization to reach the predefined targets. It has been exhibited that the “Road Map to RF-Designs”, presented in this manuscript yields excellent results to design practical matching networks for antennas and amplifiers. As a matter of fact, in the first example, gain bandwidth limits of an 0.18 μ C-MOS FET is investigated over 0.45 GHz to 10 GHz by designing proper matching networks. In the second one, the matching networks designed for an RF antenna switch results tremendous improvement in the transducer power (–4.5 dB loss vs. –1.4235 dB loss) with small ripple factor of $\varepsilon = \pm 0.1605$ dB over the pass band of 17 GHz-23 GHz. Similarly, as it has been shown in the third example, a single matching network designed for a PIFA-900 can easily cover the dual frequency band of B_1 = (824 – 960) MHz and in B_2 = (1710 – 1990) MHz.

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