Wideband Tuning of Impedance Matching Networks using Hierarchical Genetic Algorithms for Multistandard Mobile Communications

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Abstract—**An adaptive technique adopting hierarchical genetic algorithms (HGA) for wideband antenna impedance tuning is presented in the paper. Three simulation examples are given for GSM, UMTS and both standards, with different types of antenna impedance. The frequency range of the dual standards is from 1.7 to 2.2GHz. Simulation results show that the proposed tuning technique can achieve good accuracy of impedance matching and load power. The reflection coefficient and VSWR obtained are also very close to their ideal values. Comparison of the proposed HGA tuning method with conventional genetic algorithm based tuning method is also given, which shows that the HGA tuning algorithm is much faster. Moreover, the proposed method can be useful for software defined radio systems using a single antenna for multiple mobile and wireless bands.**
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Index Terms—**Impedance matching, antenna tuning, matching networks, mobile communications, genetic algorithms**

I. INTRODUCTION

Due to the maximum power transfer from power amplifiers to the antenna, impedance matching (IM) is one of the most significant techniques for energy or power efficient wireless and mobile communication systems. On the other hand, IM is as well necessary in other important front-end circuits such as low noise amplifiers, oscillators and mixers. Typically, most wireless applications require antennas to work in UHF frequency range. Thus antennas are "electrically small" with their dimensions being only a small fraction of their operating wavelength. The performance of these antennas is required to be acceptably designed so that the feed-point impedance may differ widely from the optimum value for good PA efficiency or LNA noise performance. In most communication system applications, the IM network is required to have the capability to transfer the maximum signal power and reject the sufficient unwanted interferences [1]. The low-pass Pi-LC network consists of a single series inductor and two parallel capacitors. It is ideal for such purpose, as well as the low-pass T-LC network with two series inductors and one parallel capacitor.

The Pi and T networks can precisely satisfy both the matching conditions and the requirement of harmonic rejection. The Q-based methods are extremely simple because only algebraic methods are involved in the design process. Moreover, they are more accurate than the graphical methods based on the Smith Chart. They have been documented in great details for both Pi and T networks in [1]. With the accurately formulated loaded Q, simple design formulas have been derived. The parasitic sensitivities and parasitic power loss in terms of the loaded Q and the unloaded Q have also been demonstrated.

In practical applications, antenna impedances always vary with frequency and environment. Thus the matching network must be tunable and have enough matching space [2] in order to transform such impedances to the required resistance. It can be shown that high-order networks such as Pi and T types have the capability of matching any impedance to the required resistance, provided that element values are variable from zero to infinity.

Unfortunately, variable elements can be adjusted only in a very limited range due to component material and structure and the stray effects. And the latter acts as often an important role in high frequency applications. Therefore the real matching area of a practical network only covers a small part of the theoretical matching area. In such cases it would be practical to know the maximum allowable changing region of the impedance by the network with the given limited tunable ranges of element

values. A systematic method for the matchable impedance domains with limited variable ranges of element values has been proposed and demonstrated both algebraically and graphically in [3]. Further, a Q-based design method to precisely determine the minimum adjustable ranges of element values of Pi and T networks for the changing region of the impedance to be transformed and the range of frequencies of operation has also been proposed [4].The Q-based design method [1] and tuning range solutions [4] have been used for reconfigurable MEMS impedance matching networks in the FM band in [5]. As an extension of the Q-based design method, [6] for band-pass impedance matching networks has been theoretically achieved.

Automatic tuning methods of IM networks have been investigated [2, 7-10]. Genetic algorithms have been used [7-13]. However, the tuning methods proposed so far are all single frequency based, not fast enough and mainly at HF range. All channels which have very narrow signal bandwidth, thus very high Q at RF frequencies can be tuned to, by sweeping the frequency range of interest. They involve switching between different frequencies, channels and bands/standards, as may be needed.

Here we consider a wideband tuning issue and propose a method based on hierarchical genetic algorithms (HGA). Using this method, the impedance matching networks can be tuned to cover a band of frequencies, for all channels of uplink, all channels of both uplink and downlink, and indeed multi-bands of several standards.

We describe the impedance matching network in Section II. We propose the wideband impedance tuning method using HGA in Section III. Simulations and results are presented for GSM, UMTS and both in Section IV. Conclusions are drawn in Section V.

II. IMPEDANCE MATCHING NETWORK ANALYSIS

The purpose of impedance matching is to maximize power transmission efficiency even if the impedance of the load changes. The impedance matching process is executed by controlling the impedance matching network parameters adaptively.

Fig. 1 shows the lowpass pi impedance network between the load/antenna and source/transmitter, where Rs represents the transmitter source resistance (typically 50Ω). And Z_L is the complex load impedance, Z_A, Z_B and Z_{in} represent the impedances of the network at each transformation stage from the load to the source, respectively. In order to achieve conjugate impedance matching, and hence maximum real power transfer to the load, Z_{in} must equal the complex conjugate of the source impedance, Rs.

We may easily obtain that

$$Z_A = \frac{Z_L}{1 + j\omega C_2 Z_L} \qquad (1)$$

where $Z_L = R_L + jX_L$ is the load impedance, ω is the operating angular frequency. Let $Z_A = R_A + jX_A$, we have
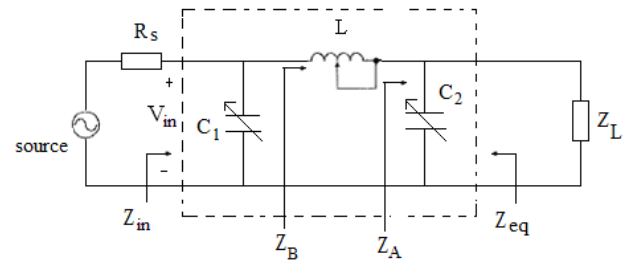

Fig.1. Tunable impedance matching network

$$R_A = \frac{R_L}{(\omega C_2 R_L)^2 + (1 - \omega C_2 X_L)^2}$$

$$X_A = \frac{X_L - \omega C_2 (R_L^2 + X_L^2)}{(\omega C_2 R_L)^2 + (1 - \omega C_2 X_L)^2} \qquad (2)$$

Then we can write

$$Z_B = Z_A + j\omega L = R_A + j(X_A + \omega L) \qquad (3)$$

Let $Z_B = R_B + jX_B$, from (3) we have

$$R_B = R_A, X_B = X_A + \omega L \qquad (4)$$

Further we get the input impedance

$$Z_{in} = \frac{Z_B}{1 + j\omega C_1 Z_B} \qquad (5)$$

Let $Z_{in} = R_{in} + jX_{in}$, it may be obtained that

$$R_{in} = \frac{R_B}{(\omega C_1 R_B)^2 + (1 - \omega C_1 X_B)^2} \qquad (6)$$

$$X_{in} = \frac{X_B - \omega C_1 (R_B^2 + X_B^2)}{(\omega C_1 R_B)^2 + (1 - \omega C_1 X_B)^2} \qquad (7)$$

So the input power would be

$$P_{in} = \frac{|V_{in}|^2}{R_{in}} = 4 \frac{R_s}{R_{in}} | \frac{Z_{in}}{Z_{in} + R_s} |^2 P_{ava} \qquad (8)$$

where $P_{ava} = |V_s|^2 / (4R_s)$ is the maximum power and V_s is the amplitude of the source voltage. Let Γ_s be the reflection coefficient, we have

$$\Gamma_s = (R_s - Z_{in})/(R_s + Z_{in}) \qquad (9)$$

Voltage standing wave ratio (VSWR) can be written as

$$VSWR = (1 + |\Gamma_s|) / (1 - |\Gamma_s|) \qquad (10)$$

In the ideal lossless case, the power transfer gain will be maximum at the matching frequency and we have $\Gamma_s = 0$, $VSWR = 1$ and $Z_{eq} = Z_L^*$. For wideband tuning, we have to cover a whole frequency band of interest, for which perfect matching is not possible. In practice, however, we may achieve if Z_{eq} is approximately equal to Z_L^* across the whole band as accurately as possible. So our goal is to minimize the function $Z_{eq} - Z_L^*$ over the

band of interest. The optimized tuning may not be easy to achieve. Traditional gradient based methods may not be able to find the global minimum and involve time consuming derivative calculations. Evolutionary optimization methods may thus have to be used.

III. HGA-BASED METHOD FOR WIDEBAND IMPEDANCE TUNING

Let $\omega = [\omega_1, \omega_2, \cdots \omega_M]$ be a set of M discrete frequencies spaced over a band of $[\omega_{\min}, \omega_{\max}]$, where $\omega_{\min}$ and $\omega_{\max}$ are the minimum and maximum angular frequency of the considered frequency interval. As mentioned in Section II, our objective is to minimize the reflection coefficient magnitude $\Gamma_s(\omega_i), i = 1,2, \cdots M$. We assume that the load impedances have been measured at each of these M frequencies. Let Γ_{s0} be a target constant, ideally equal to zero, for which $\Gamma_s(\omega_i), i = 1,2, \cdots M$ is to be minimized in our proposed method. Let $chrom = [C_1, C_2, L]$, the cost function can be chosen as:

$$f_1(\omega) = \Gamma_s(\omega_i, chrom) - \Gamma_0$$
$$f_2(\omega) = Z_{in}(\omega_i, chrom) - Z_0$$
$$f(\omega) = \frac{1}{2M} \sum_{i=1}^{M} \left[|f_1(\omega)|^2 + |f_2(\omega)|^2 \right] \qquad (11)$$

where Γ_0 is the ideal value of the reflection coefficient, and Z_0 is the target value of the input impedance, which is ideally Rs. Then the fitness of the proposed technique would be written as

$$fitness = -f(\omega) \qquad (12)$$

Our aim is to maximize the cost function $fitness$. As $f(\omega)$ is a highly nonlinear function of the components C_1, C_2 and L, a direct maximization of the cost function is a very difficult task.

Conventional Genetic Algorithm (CGA) has been extensively used to solve any NP problem because of its parallel characteristics and global optimal search[12-13]. However, its partial comparability of chromosome produced by the genetic operators leads to poor climbing ability and immature (premature) convergence. Moreover, the search process may stagnate at the local optimal preference [12-13]. So a hierarchical genetic algorithm (HGA) is proposed here to overcome the difficulty of CGA. The method strives to minimize the drawbacks of the stochastic search method in exploring the search over the entire space. Firstly, the whole space S is partitioned into m subspaces, named $S_i, i = 1,2, \cdots m$, where m is dependent on the number of hyper-planes required to represent the function to be solved. So we can run m CGAs in m subspaces concurrently. Thus the genetic process can be conducted concurrently in various segments over the entire space, and each segment is hierarchically processed in parallel. Then we may extract the global optimum solution after a best solution is found within each subspace.

A. Representation

A decimal floating-point encoded scheme is adopted here. We encode C_1, C_2, and L as decimal chromosomes, which is shown in Fig.2. Working directly with decimal fractions, we can avoid the rounding errors that otherwise typically occur when converting between decimal fractions and binary fractions.

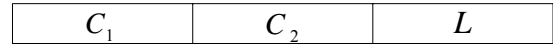| C_1	C_2	L

Fig.2.   Gene representation

Moreover, it is beneficial to simplify the process of numerical coding and calculation of the fitness numerically, and further minimize the errors of the evolution operations such as selection, mutation and crossover.

B. Space Partition and Subspace Search

The objective of our work is to optimize expression (12) subject to $C_1, C_2 \in [C_{\min}, C_{\max}], L \in [L_{\min}, L_{\max}]$, where $C_{\min}, C_{\max}$, $L_{\min}$ and $L_{\max}$ are predefined constants. Denote the problem as $S \to fitness$, where S is the search space. To simplify the search process, S is initially partitioned into m subspaces $S_i, i = 1,2, \cdots m$, that is, $S = \{S_1, S_2, \cdots S_m\}$. Then the populations with different size are then distributed to each segment, which helps to spread the search over a broader space. The research begins in each subspace $S_i, i = 1,2, \cdots m$ in parallel. And the algorithm behaves as CGA in each subspace: $S_i \to fitness, i = 1,2, \cdots m$. And the best solutions are available when the system convergent to its balanced points by performing the evolution operations.

In order to obtain better performance, in each subspace, the individuals are selected as parents according to their fitness. The better the chromosomes are, the more chances to be selected they have. In each iteration, all the individuals are sorted as the fitness function. The probability to select the jth individual is $prob(j) = r(1-q)^{j-1}$, $j = 1,2, \cdots size$, where $size$ is the population size, $r = q / (1 - (1-q)^{size})$, and $q \in (0,1)$ is a predefined constant.

In the crossover phase, let chromosomes of the individual with best fitness value in the current generation be $chrom_1 = [C_1^1, C_2^1, L^1]$, and it is selected as a parent. In order to choose the best individuals, as well as to obtain the diversity, the other parent (other than $chrom_1$) with chromosome $chrom_2 = [C_1^2, C_2^2, L^2]$ is randomly selected in the population. Then the corresponding offspring may be written as $chrom = a(chrom_1 - chrom_2) + chrom_1$, where $a \in (-1,1)$ is a random.

In mutation phase, after an individual is randomly selected in the population, the mutation exchanges the positions of two randomly chosen genes in a permutation.

C. Extraction of the global optimum solution

In each subspace S_i , the m genetic programs $S_i, i = 1, 2, \cdots m$ simultaneously run in all subspaces in parallel. The algorithm in each S_i terminates after a best solution found within the corresponding subspace although the best solution in S_i may not be the global optimum of the optimizing function. Since the entire space S is divided into m subspaces. Let the best solution of the ith subspaces S_i be $chrom_b^i$, $i = 1, 2, \cdots m$. So the global solution can be achieved by optimizing the following function:

$$chrom = \max[chrom_b^1, chrom_b^2, \cdots chrom_b^m]$$
$$\forall chrom \subseteq S : fitness(chrom) \leq fitness(chrom_b^i) \quad (13)$$
$$i = 1, 2, \cdots m$$

The genetic operations are the same as the sublevel program.

IV. SIMULATION RESULTS

Example 1: GSM-1800 uses 1710–1785 MHz for uplink and 1805–1880 MHz for downlink. Here we consider the frequency range/band from 1.71GHz to 1.88GHz. A sample of 100 points is used in the simulation. The source resistance is 50Ω. The load impedance is capacitive, Zload=20-j30 Ω. The program is convergent to its balanced point after trained for 474 epochs. And the obtained matching component values are C1=2.27954pF, C2=0.10000pF, and L=4.72761nH. The curves of input impedance Zin, reflection coefficient, load power and VSWR are shown in Fig.3.
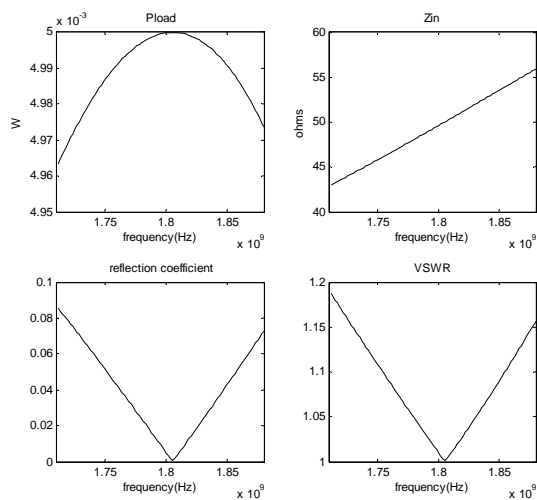

Fig.3.   Load power, input impedance, reflection coefficient, and VSWR for GSM-1800

Example 2: UMTS specifies the bands 1900-2025 MHz and 2110-2200 MHz for 3G transmission. Here we consider the band from 1.9 to 2.025GHz (we will consider even wider band to include 2.11-2.2GHz in Example 3). The source resistance is 50Ω. The load impedance is resistive, Zload=20Ω. A set of 100 sampled points are considered in the test. The obtained matching component values are C1=1.98496pF, C2=0.11721pF and

L=2.02896nH. The curves of load power, Zin, reflection coefficient, and VSWR are shown in Fig.4.
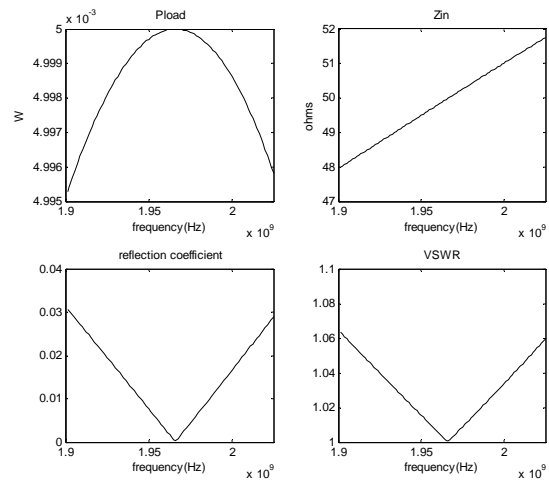

Fig.4.   Load power, input impedance, reflection coefficient, and VSWR for UMTS

Example 3: We now consider dual modes GSM-1800 and UMTS. The frequency ranges from 1.7 to 2.2GHz for the dual standards. The source resistance is 50Ω. The load impedance is assumed to be 60 Ω. A set of 500 sampled points are considered in this experiment. The obtained matching component values are C1=0.10000pF, C2=0.60851pF, and L=2.06678nH. The simulated load power, input impedance, reflection coefficient and VSWR are given in Fig.5.

In order to evaluate the proposed technique, we compare the proposed HGA method with the CGA with the same evolution operations to tune the wideband impedance in the paper. We have tested the program for each example for 500 times. In order to compare them, all the genetic operations are the same. The results are shown in Table I.
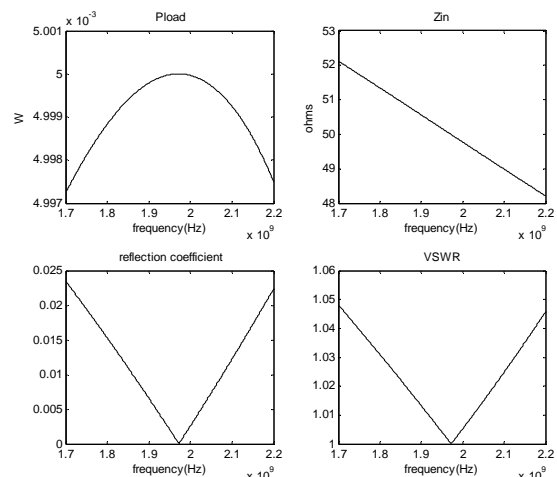

Fig.5.   Load power, input impedance, reflection coefficient, and VSWR for dual standards GSM-1800 and UMTS

Table I shows that the performance of the proposed technique is better than the CGA for GSM, UMTS and

dual standards from1.7 to 2.2GHz. The accuracies over the frequency range of the obtained wideband impedance are 90.58%, 94.43% and 97.05%, respectively. However, the corresponding accuracies of the wideband impedance are 30.00%, 41.20% and 46.40% if CGA method is adopted. The proposed method show the power transferred to load is very high with an error of at most 1.09e-4W. And the reflection coefficient and VSWR are also very close to their ideal values with an MSE error of at most 0.132. So the proposed adaptive tuning technique can achieve good accuracy of impedance matching and load power for different antenna impedances. Moreover, the proposed method can be useful for software defined radio systems using a single antenna for multiple mobile and wireless bands.

TABLE I. THE PERFORMANCE COMPARISON OF THE PROPOSED HGA TECHNIQUE AND CONVENTIONAL CGA TECHNIQUE

		HGA			CGA		
		1710–1880MHz	1900-2025MHz	1700–2200MHz	1710–1880MHz	1900-2025MHz	1700–2200MHz
Epochs		474	435	108	934	820	1795
Time(s)		13.02	12.55	9.12	17	18.87	45.36
Zin error (%)	max	1.62E-01	2.48E-01	3.70E-02	1.00E+00	1.00E+00	1.00E+00
	min	8.14E-02	3.02E-02	2.33E-02	8.51E-02	3.17E-02	2.62E-02
	MSE	9.42E-02	5.57E-02	2.96E-02	7.00E-01	5.88E-01	5.36E-01
Pload error (w)	max	4.51E-05	1.09E-04	2.49E-06	5.00E-03	5.00E-03	5.00E-03
	min	1.12E-05	1.53E-06	8.92E-07	1.18E-05	1.66E-06	1.07E-06
	MSE	1.55E-05	1.11E-05	1.48E-06	3.31E-03	2.45E-03	2.60E-03
Reflection coefficient error	max	8.43E-02	1.32E-01	1.85E-02	1.00E+00	1.00E+00	1.00E+00
	min	4.16E-02	1.52E-02	1.16E-02	4.24E-02	1.59E-02	1.30E-02
	MSE	4.84E-02	2.83E-02	1.48E-02	6.81E-01	5.37E-01	5.28E-01
VSWR error	max	1.84E-01	3.03E-01	3.77E-02	1.09E+09	5.57E+08	1.44E+09
	min	8.69E-02	3.08E-02	2.35E-02	8.85E-02	3.23E-02	2.63E-02
	MSE	1.02E-01	5.82E-02	3.00E-02	4.27E-01	2.32E-01	2.23E+00

V. CONCLUSIONS

This paper presents an adaptive wideband impedance tuning technique for multi-standard mobile communications. HGA is chosen to improve impedance matching. Examples are given for GSM, UMTS and dual standards with three different antenna impedances. The frequency range for the dual standards is from1.7 to 2.2GHz. Simulation results show good accuracy of input impedance matching and power transfer. The reflection coefficient and VSWR are also very close to their ideal values. The proposed tuning method may also be suitable for even wider range of frequencies to cover other wireless standards.

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