

Tunable Antenna Impedance Matching for 4G Mobile Communications

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Abstract—We analyze a Π -network for tunable antenna impedance matching in 4G mobile communications. Fundamental limits on designability and practical limits on implementation of the network are presented. The performance of the network is evaluated in terms of coverage (the area on Smith chart that can be matched) and the maximum achievable tuning bandwidth. We also derive the required component values and component Q from coverage, bandwidth, as well as the efficiency of the network. We conclude that the Π -network (synthesized for the maximum achievable bandwidth) is able to cover E-UTRA Band 4 and 10 for VSWR up to 8 and all other bands for VSWR up to 10. To achieve that, in a $50\ \Omega$ RF environment, it requires a maximum capacitance within the range 13.64–2.7 pF, and a maximum inductance in the range 34.1–6.8 nH in frequency range 700–3500 MHz. If the power loss is limited to 0.5–0.6 dB, it requires Q_L of 70 and Q_C of 70–100.

Index Terms—Antenna tuning, 4G mobile communication, impedance matching, Q factor, Π -network.

I. INTRODUCTION

Mobile communication systems are evolving into the 4th generation (4G) which is targeting a downlink peak data rate of 1 Gbps (100 Mbit/s for high and 1 Gbit/s for low mobility) and an uplink peak data rate of 500 Mbps [1], [2]. To reliably achieve such high-speed communication, efficient transmission and reception of signals are required. One of the main components influencing transmission and reception is the mobile antenna. Because mobile devices operate in proximity to the human body, the antenna impedance is affected by the body and the hand that holds the device [4], [5]. A change in antenna impedance creates mismatch between the antenna and the RF front-end which significantly degrades the power efficiency of the radio link [7]. We have presented in [8] that, in receive antenna diversity systems, antenna mismatch can cause severe degradation of the system performance in circumstances of multiple diversity antennas being simultaneously mismatched.

To maintain link quality, an antenna tuning unit (ATU) is therefore used to dynamically match the antenna impedance to the RF front-end. ATUs typically use lumped Π - or L -networks, which are composed of tunable capacitors and fixed inductors, to achieve tunable antenna impedance matching. The topology, performance and tuning method have been studied in a variety of publications, e.g., [7], [9], [10]. However, none of these papers is focused on the 4G LTE application. This gives rise to the motivation of our work.

In this paper, we start from the 4G LTE application requirements. Then we analyze one of the most widely used matching networks — the low-pass Π -network, and present

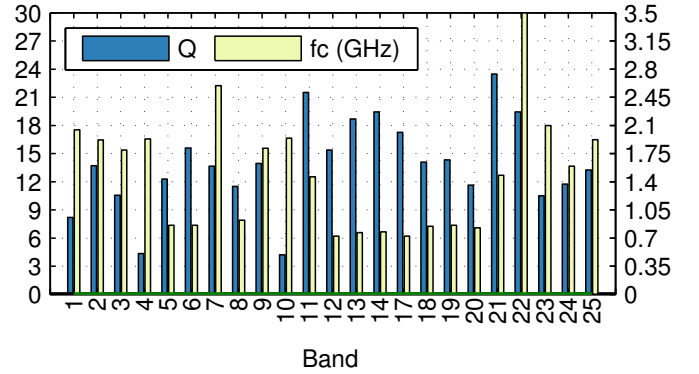


Fig. 1. Quality factor of E-UTRA FDD frequency bands. f_c represents the center frequency of each band.

its performance in terms of coverage (the area on Smith chart that can be matched), the achievable tuning bandwidth and power efficiency. From these performance indicators, the required component values and component Q (quality factor) are derived.

Different from papers which show the performance of matching networks at specific frequencies, we study the matching network in a normalized way and present results that can be de-normalized for different frequencies, bandwidths and application requirements. The results achieved will show how does the Π -network fit 4G LTE applications and also give guidelines on how to determine what tuning components are required.

II. THE E-UTRA FREQUENCY BANDS

The spectrum allocation for 4G LTE (also known as E-UTRA, the Evolved Universal Terrestrial Radio Access) is defined in the 3GPP technical specification [3]. The E-UTRA band definition contains FDD (frequency division duplexing) and TDD (time division duplexing) bands.

From the matching network design point of view, we are concerned with the Q of the bands, which is defined as

$$Q = \frac{\text{Center frequency of the band}}{\text{Bandwidth}}. \quad (1)$$

The band Q is a suitable measure to assess the tunability requirements of a matching network.

The Q of the FDD bands (Band 1 to 25) are shown in Fig. 1. We see that Q varies from 4 on Band 4 and 10 both at around 1.9 GHz to 23 on Band 21 at around 1.5 GHz. It is

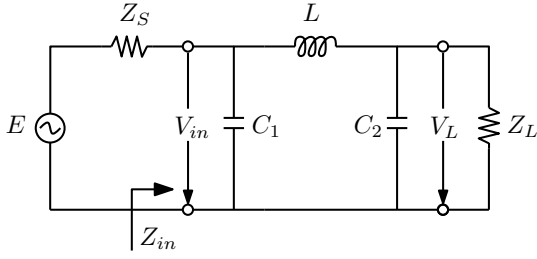


Fig. 2. Π -network

also interesting to notice that, except for Band 4 and 10, all the other FDD bands have a Q no less than 8.

The Q for TDD bands varies from 13 on Band 41 at around 2.6 GHz to 134 on Band 34, which has a very small bandwidth of 15 MHz, at around 2 GHz.

Despite the fact that some bands have an extremely large Q , we do not have to design extremely narrow band matching networks accordingly, because the primary goal of designing a matching network is to cover the band with sufficiently low return loss, rather than to suppress out-of-band transmission. On the contrary, we will see it is the small Q , which means large bandwidth, that makes designing of the matching network a challenging task. Therefore, we put more emphasis on small Q and will show how Q , together with the antenna impedance, is related to the designability of the matching network.

III. THE Π IMPEDANCE MATCHING NETWORK

A. Topology

The objective of antenna matching is to match the antenna impedance to the RF front-end, which typically has an impedance of $50\ \Omega$, to minimize reflection, i.e., to minimize VSWR (voltage standing wave ratio) which can go up to $10 : 1$ [6].

The low-pass Π -network, shown in Fig. 2, is one of the most simple and widely used networks for antenna impedance matching [7], [10] due to its harmonic rejection capability and wide coverage [12]. Unlike L-networks, whose Q is uniquely determined by the load, the Π -network has one more degree of freedom for the user to design the bandwidth of matching.

In theory, the Π -network can provide complete Smith chart coverage [10]. In practice, its coverage is limited by available component values, or tunability of the components. Furthermore, the achievable bandwidth of the matching network is determined by the load impedance. Now questions arise: For a given area, can the network match all the load impedances falling in it? Does it meet the bandwidth requirements? What components are needed? What efficiency can we expect? The following sections will answer these.

B. Fundamental Limitations

It is customary to analyze a matching network by terminating it with a real load. In this subsection we assume that in Fig. 2 $Z_S = R_S$ and $Z_L = R_L$.

The loaded Q of the Π -network can be found to be

$$Q_L = \frac{1}{2} (\omega C_1 R_S + \omega C_2 R_L). \quad (2)$$

The necessary and sufficient conditions for designability of the Π -network are given in [12] as

$$Q_L \geq \begin{cases} \frac{1}{2} \sqrt{\frac{R_L}{R_S} - 1} & \text{for } R_L \geq R_S \\ \frac{1}{2} \sqrt{\frac{R_S}{R_L} - 1} & \text{for } R_S > R_L \end{cases}. \quad (3)$$

The fundamental limitations on the designability given in (3) indicate that the degree of mismatch, measured by R_L/R_S , determines the minimum achievable Q , or equivalently, the maximum achievable bandwidth. Qualitatively, the larger the degree of mismatch, the lower is the achievable bandwidth. From another viewpoint, once Q is given, the maximum allowed mismatch is determined and thus the coverage of the network is determined. For bands with a small Q , the coverage realizable by the Π -network is correspondingly small.

C. Practical Limitations

In addition to the fundamental limitations, there are also practical limitations on the feasibility of the matching network. We highlight these aspects: (1) feasibility of the component values, (2) tunability of the tuning components, and (3) availability of the component Q which are derived from the power efficiency of the network.

For tunable capacitors, the tunability is defined as $(C_{\max} : C_{\min})$, and for tunable inductors as $(L_{\max} : L_{\min})$. The maximum and minimum component values are determined by the goal of matching as well as the architecture and technology of tuning devices. [10] introduces some of the architectures for equivalently tunable inductance and capacitance.

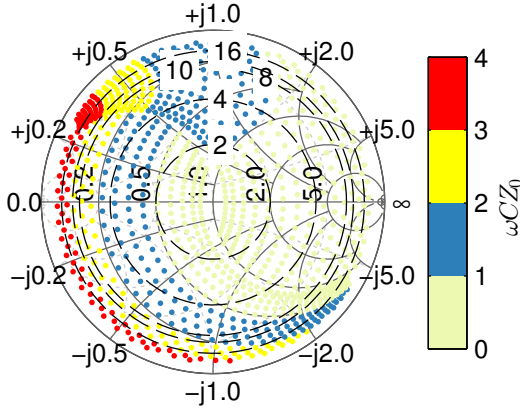
The coverage of the Π -network (synthesized for the maximum achievable bandwidth) with respect to $b_C = \omega C Z_0$ ($C = \max(C_1, C_2)$) and $x_L = \omega L / Z_0$ is illustrated in Fig. 3. It is shown in Fig. 3a that to match the impedances located in the outer left part of the Smith chart larger b_C , thus larger C , is required and in Fig. 3b that to match the impedances located in the outer right part larger x_L , thus larger L , is needed. To cover the area defined by $\text{VSWR} \leq 10$, the required $b_{C,\max}$ and $x_{L,\max}$ are both 3, which correspond to C_{\max} of 13.64–2.7 pF and L_{\max} of 34.1–6.8 nH in the frequency range 700–3500 MHz accordingly.

D. Bandwidth

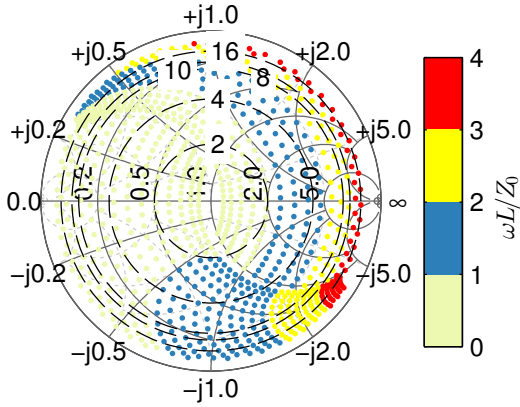
We measure the 6 dB bandwidth $\text{BW}_{6\text{dB}}$ of the complex-loaded matching network as defined by $|S_{11}| \leq -6\ \text{dB}$. To allow direct comparison between the maximum achievable bandwidths and the E-UTRA bandwidth requirements, we compute the corresponding quality factor from

$$Q_{L,6\text{dB}} = \frac{f_C}{\text{BW}_{6\text{dB}}}, \quad (4)$$

where f_C is center frequency of the 6 dB band.



(a) Coverage vs. susceptance of capacitor



(b) Coverage vs. reactance of inductor

Fig. 3. Coverage of the II-network (synthesized for maximum bandwidth) versus component values. The characteristic impedance Z_0 is used for normalization. The labeled concentric circles indicate different VSWR values.

As shown in Fig. 4, there is $Q_{L,6dB} < 4$ within the area defined by $VSWR \leq 8$ (except for a reasonably small area beyond this threshold) and $Q_{L,6dB} < 8$ within the $VSWR \leq 10$ area. Recall that Band 4 and 10 have $Q = 4$ and all the other bands have $Q \geq 8$.

E. Power Efficiency

All practical components exhibit a certain amount of loss that is usually modeled with an equivalent series resistance R . The degree of loss is measured by the components Q , which is defined for a capacitor as $Q_C = 1/(\omega RC)$, and for an inductor as $Q_L = \omega L/R$.

For a given component Q , we can express the equivalent component value in complex form which takes into account the loss of a capacitor as $C_e \approx C(1 - j/Q_C)$ and the loss of an inductor as $L_e = L(1 - j/Q_L)$.

To evaluate the power efficiency of the matching network, we design the network neglecting loss first and then simply replace the ideal components with practical components by substituting their equivalent complex values. Although this

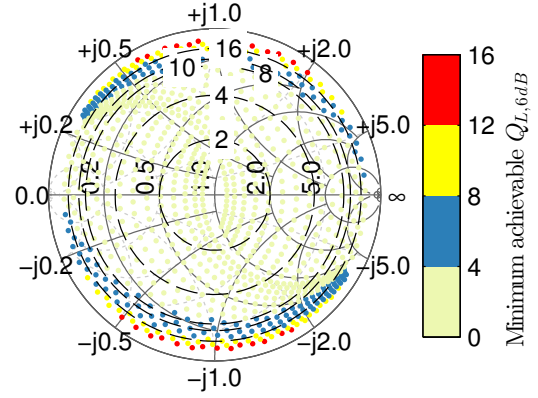


Fig. 4. Minimum achievable $Q_{L,6dB}$ (corresponding to the maximum achievable bandwidth) of the II-network. The bandwidth is defined by $|S_{11}| \leq -6$ dB. The labeled concentric circles indicate different VSWR values.

creates some mismatch, it won't affect the evaluation of the power efficiency of the matching network itself. The power efficiency of the II-network is defined as

$$\eta = \frac{P_L}{P_{in}} = \frac{|V_L|^2 G_L}{|V_{in}|^2 G_{in}}, \quad (5)$$

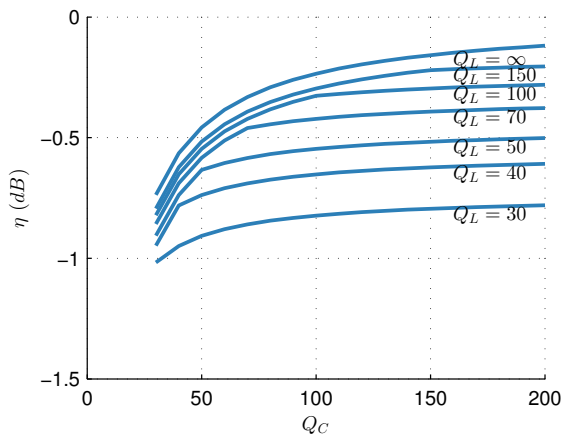
where P_L and P_{in} denote the power delivered to the load (the antenna) and the input power which is the power supplied by the source to the matching network. V_L and V_{in} are the load and input voltages. G_L and G_{in} are the load and input admittances, respectively.

The power efficiency of the II-network with respect to component Q within the area $VSWR \leq 10$ is shown in Fig. 5b. If the power loss is limited to 0.5–0.6 dB, it requires Q_L of 70 and Q_C of 70–100. Our result shows good agreement with [7] which gives an efficiency of 86.3% (−0.64 dB), 91.7% (−0.38 dB) and 95.8% (−0.19 dB) for Q_L of 60, 100 and 200 respectively.

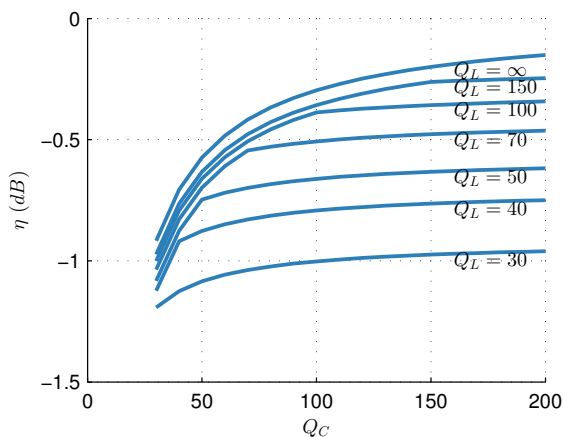
By contrast, Fig. 5a and Fig. 5c, which show the power efficiency within the areas $VSWR \leq 8$ and $VSWR \leq 16$ respectively, give very different results. Assume $Q_C \gg Q_L$, to achieve insertion loss below 0.5 dB, covering the area $VSWR \leq 8$ requires a Q_L of 50, but covering the area $VSWR \leq 16$ requires a Q_L of 100. This is due to the fact, as shown in Fig. 3, that at the edge of the Smith chart the contour lines of VSWR, Q_L and Q_C are getting increasingly dense.

Apart from the numbers, we see that when $Q_C < Q_L$, the loss of the capacitors is dominant, and when $Q_L < Q_C$, the loss of the inductor dominates. Another interesting point is that, due to the flatness of the power efficiency curves at high Q area, the larger the component Q the more difficult it is to further improve the efficiency of the matching network by increasing it.

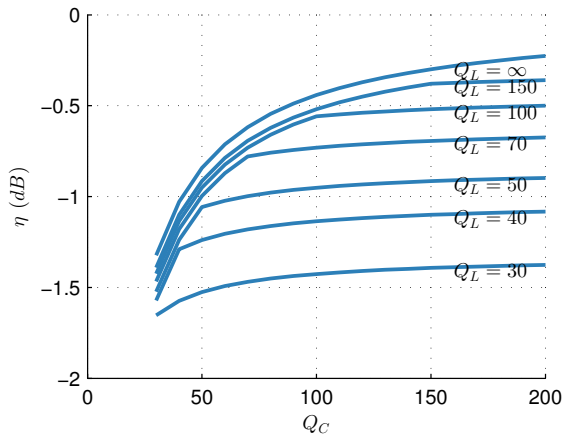
These give us practical guidelines on selecting component Q : (1) the component Q of the capacitors and the inductor should not differ too largely such that the poor one covers up



(a) $VSWR \leq 8$



(b) $VSWR \leq 10$



(c) $VSWR \leq 16$

Fig. 5. Power efficiency η of the Π -network versus component Q . η represents the worst case power efficiency in Smith chart within the area defined by VSWR.

the performance of the better one, (2) it is less efficient and more costly to further improve a high component Q to further improve the power efficiency of the network, and (3) trade-offs must be taken between the coverage of the matching network and the component Q , which is very sensitive to large VSWR.

IV. CONCLUSION

A Π impedance matching network has been studied from the 4G mobile antenna impedance matching perspective, and analyzed in terms of coverage, achievable bandwidth, the required component values and power efficiency.

We conclude that the Π -network (synthesized for the maximum achievable bandwidth) is able to cover Band 4 and 10 for VSWR up to 8 and all other bands for VSWR up to 10. To achieve that, in a typical $50\ \Omega$ RF environment, it requires C_{\max} of 13.64–2.7 pF, and L_{\max} of 34.1–6.8 nH in the frequency range 700–3500 MHz. If the power loss is limited to 0.5–0.6 dB, it requires Q_L of 70 and Q_C of 70–100.

ACKNOWLEDGMENT

This work has been performed in the ENIAC project ARTEMOS (Project No. 270683-2), in which the partner JKU is funded by the Austrian Research Promotion Agency (FFG) and the ENIAC Joint Undertaking.

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