A Comparison of Tunable Ferroelectric Π- and T-Matching Networks

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***Abstract*— A comparison of tunable Π- and T-matching networks based on ferroelectric capacitors is presented. Both topologies are promising candidates for automated antenna tuners in mobile communication systems. Assets and drawbacks of both circuits are discussed comparing linear and nonlinear simulations and measurements of two exemplary circuits at 850 MHz. The requirements to achieve sufficient linearity are determined by means of simulations. Both matching networks cover an impedance range of approximately $|\Gamma_{in}| \leq 0.6$ with a maximum transducer power loss of 4 dB and 6 dB, respectively.**

I. Introduction

Today's mobile communication devices are used in almost all imaginable environments, including near body area, in cars, on different surfaces and of course in talking position near the head. The environment of the antenna and the resulting field distribution around it has unfortunately an eminent impact on its impedance [1], [2]. The resulting mismatch between antenna and RF-frontend reduces its power efficiency, linearity and lowers the power of the input signal.

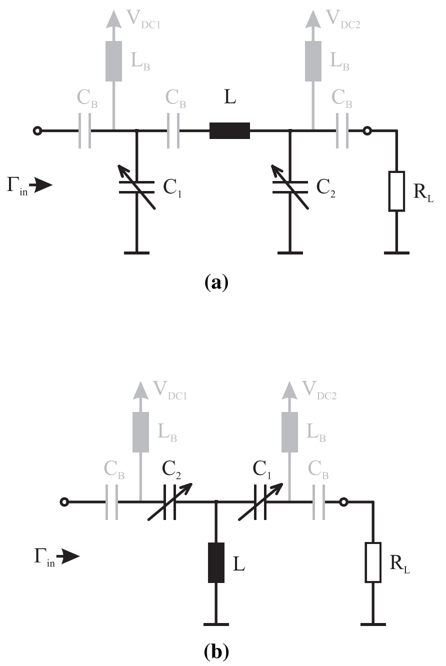


Fig. 1. Tunable matching networks with bias circuitry: **(a)** Π-MN terminated with R_L **(b)** T-MN terminated with R_L.

To overcome these problems a tunable matching network (MN) with very low insertion loss, high linearity and a wide tuning range is desirable. One possibility to introduce tunability into such networks is the use of barium-strontium-titanate (BST) varactors. Thin-film BST varactors attract attention due to their very high tunability at a comparatively low bias voltage. Furthermore, the high permittivity of BST enables large capacitance densities and accordingly small device sizes. Unfortunately, the quality factors of today's capacitors are still moderate at higher frequencies which leads to increased insertion loss. A second severe drawback of thin-film BST varactors with low tuning voltages is the inherent nonlinear distortion.

The simplest types of tunable MNs are lowpass Π- and highpass T-circuits [3], [4]. They combine a broad impedance coverage with a small component count and therefore a small size. In order to reveal which topology fits best these demands, both types were carefully investigated by means of linear and nonlinear circuit simulations and measurements.

II. Matching Area

Figures 1(a) and (b) show the circuit diagrams of Π- and T-MNs, respectively. To determine the required tunability τ_{C}, the capacitance for the capacitors C_1 and C_2 and the required inductance L for a concentric distribution of the matching area some quite lengthy calculations are necessary. Due to the limited space of this paper we will present them in near future somewhere else. One of the results of these calculations is, that there is only one solution for a minimum network Q-factor and therefore for minimum loss at a given matching area $|\Gamma|_{\max}$. Other result are, that Π- and T-networks show a mirror inverted small-signal behaviour and the capacitors C_1 and C_2 must have the same size (only in the lossless case). The results are depicted in Fig. 2(a) and (b). The parameters for the networks are given in Tab. I. We can mention that Π-MN require capacitors with comparatively high capacitances and

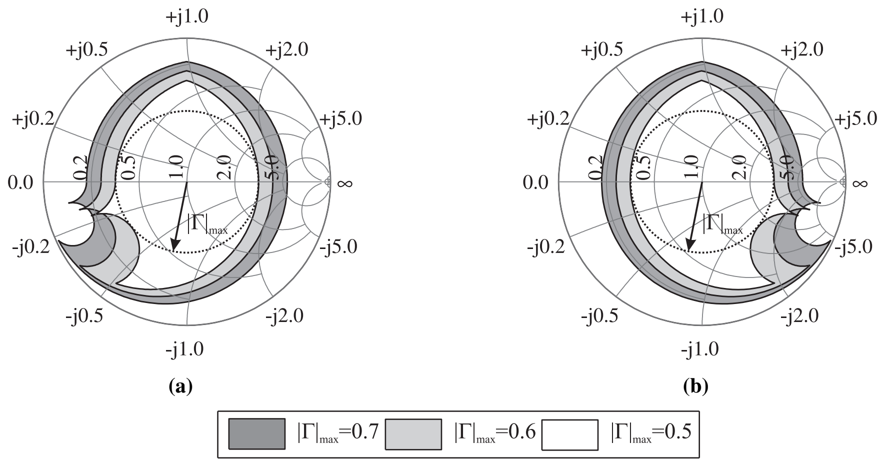


Fig. 2. Matching area of networks with minimum network Q-factor: **(a)** Π-MN. **(b)** T-MN.

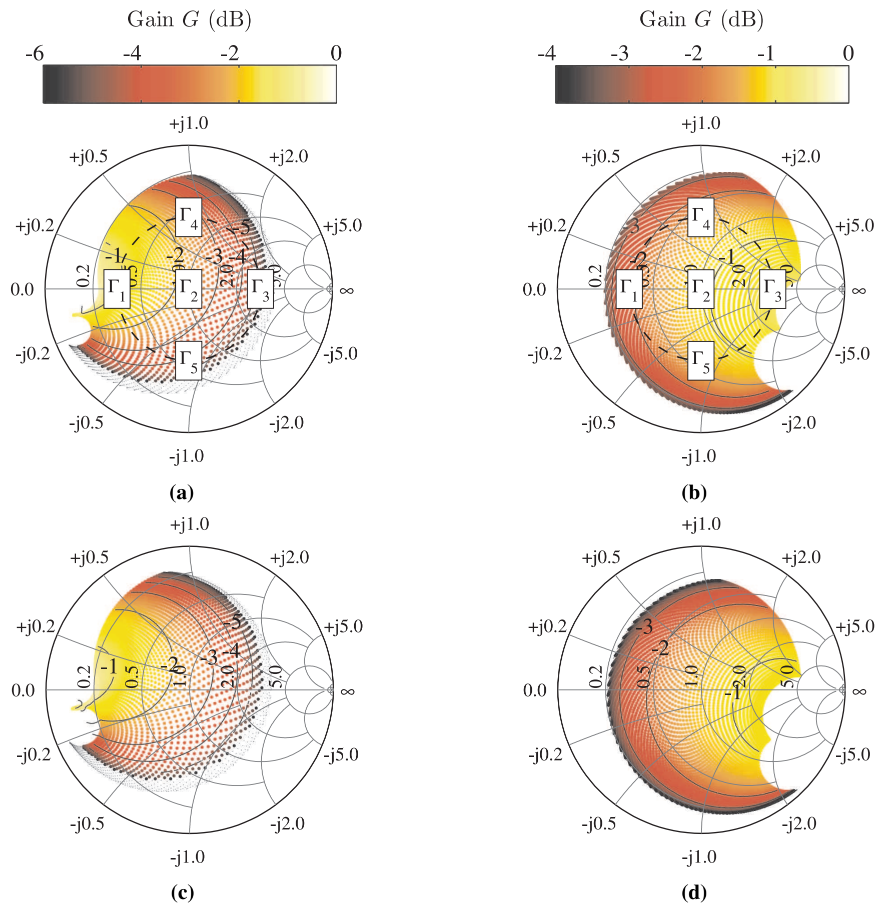


Fig. 3. Simulated and measured insertion loss of Π- and T-MNs at $f = 850\,\mathrm{MHz}$ in dependence of Γ_{in}: **(a)** Simulation of a Π-MN. **(b)** Simulation of a T-MN. **(c)** Measurement results of a Π-MN. **(d)** Measurement results of a T-MN.

coils with low inductance values. The T-MN shows a mirror-inverted picture. Their capacitance values are small and the inductance of the coil is high.

III. Insertion Loss

At the first glance it is very easy to define the requirements for the insertion loss of a matching network. The added insertion loss of the matching network should be smaller than the mismatch loss of the detuned antenna without the network. Due to the fact that a matching network does not work in a $50\,\Omega$ environment, we have to calculate the transducer power gain G_T to determine the loss. In a frontend module the second port of the MN is normally terminated with $50\,\Omega$ and the MN is adjusted so that the input impedance is conjugate complex

TABLE I

ELEMENT VALUES FOR Π- AND T-NETWORKS WITH MINIMUM NETWORK Q-FACTOR

Π	$\|\Gamma\|_{\max}$	$\omega C_{1,2}$	τ_C	ωL
	0.5	91 mS	0.62	28.9 Ω
	0.6	117 mS	0.66	25.1 Ω
	0.7	159 mS	0.7	21 Ω
T				
	0.5	11.5 mS	0.62	86.8 Ω
	0.6	10 mS	0.66	100 Ω
	0.7	8.4 mS	0.7	119.1 Ω

to the antenna impedance. Both conditions inserted in the G_T definition lead to

$$G = \frac{|S_{21}|^2}{1 - |S_{11}|^2}. \tag{1}$$

To predict the overall performance of the matching networks in the simulation we used a model previously introduced in [5]. At the operating frequency of 850 MHz the used capacitors showed quality factors in the range of $Q = 35$ at a bias voltage of 0 V up to $Q = 60$ at 25 V bias voltage. The relative tunability at V_{DC} =25 V was 70 %. We used a 1 pF and a 2 pF capacitor for the T-MN and a 16 pF and a 22 pF capacitor for the Π-MN. The different size of the capacitors is necessary to compensate for the reduction of the matching area due to loss. To minimize the losses we used wire wound air coils with quality factors $Q > 120$. As a result of the careful modeling the simulation results of the T-MN depicted in Fig. 3(b) match excellently with the measurement results in Fig. 3(d). The same applies to the results of the Π-MN in Figs. 3(a) and (c). For both plots the bias voltages $V_{\mathrm{DC}1,2}$ were swept from 0 V to 25 V in 0.25 V steps represented by dots in Fig. 3. Bias states with more than 6 dB loss were plotted with small black dots in the backround to enhance the picture resolution.

Referring to these results we can conclude that Π-MNs are a good choice for matching low impedances whereas T-MN are preferable for matching high impedances. At the opposite side of the respective impedance area the losses increase significantly. Remarkable is the fact that the overall loss of the T-MN is about 2 dB lower than the loss of the Π-MN. Simulations showed that the main reason for this behaviour is the high sensitivity of the Π-network to the parasitic influence of the bond wires. So we expect that a flip-chip mounting technique would bring similar results for Π- and T-networks. Nevertheless capacitors with higher quality factors are required to reduce the loss in both topologies.

IV. Nonlinear Behavior

For modern complex modulation formats linearity is even more important than insertion loss. To determine the nonlinear behavior a 2-tone source-pull setup depicted in Fig. 4 was used. The IP3 of the measurement system was 75 dBm. The tone distance was 1 MHz at a center frequency of $f_0 = 850$ MHz. To keep a reasonable measurement time only five different impedance points Γ_{1-5} depicted in Figs. 3(a) and (b)

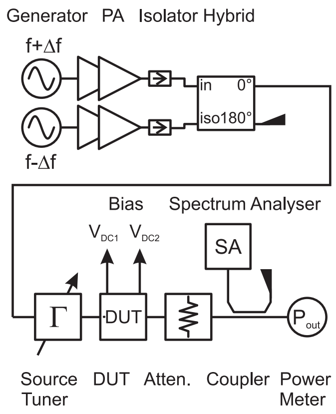


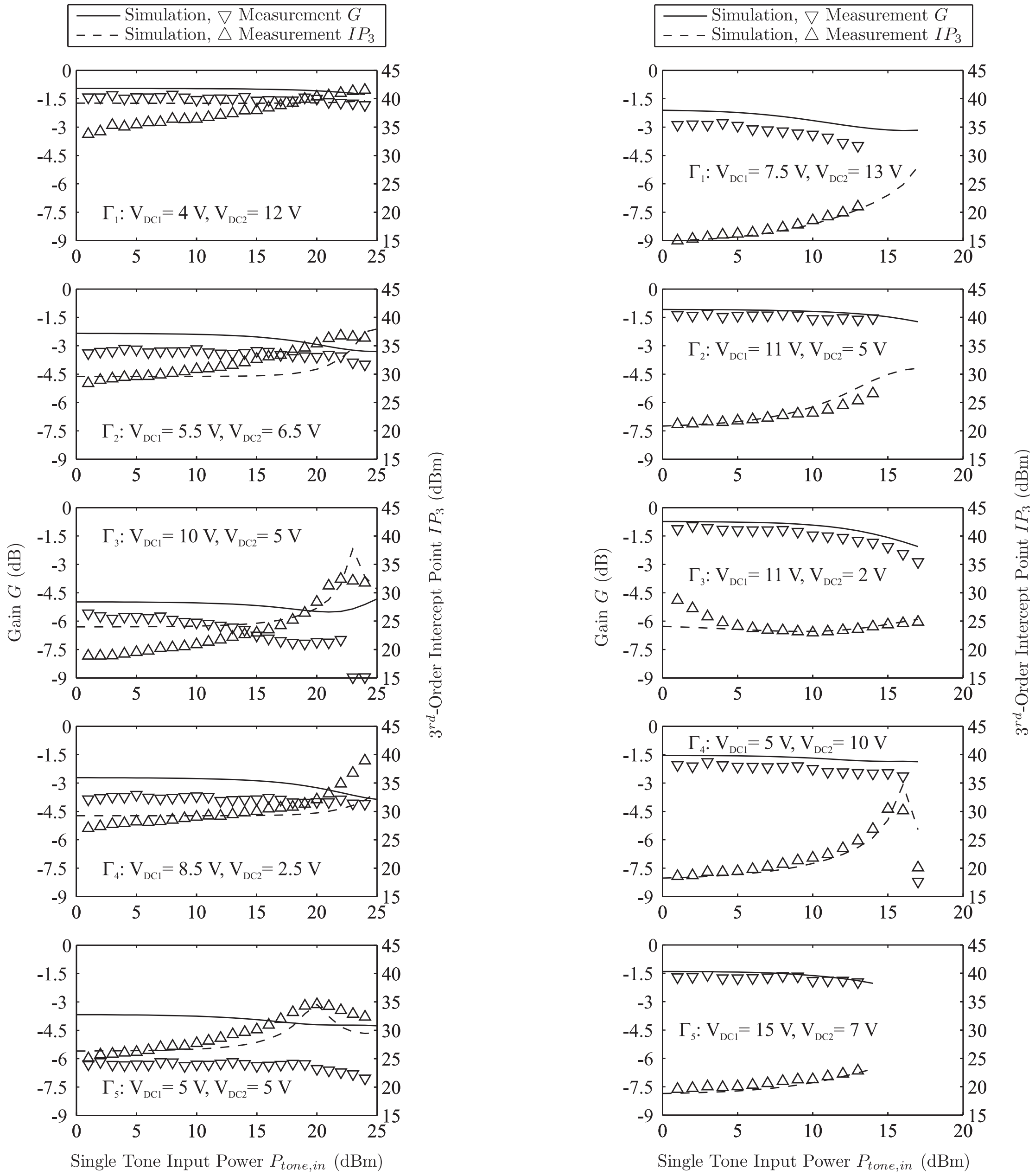
Fig. 4. Source-pull setup for nonlinear 2-tone measurements.

were selected. Due to the high tunability of the capacitors and the resultant low break-down voltage of approximately 30 V the tone power was limited to a maximum level of 25 dBm for the Π-MN and 17 dBm for the T-MN. Figure 5(a) shows the measurement results for the Π-MN and Fig. 5(b) for the T-MN. The comparison of simulation and measurement results shows clearly that it is possible to predict the nonlinear behavior very precisely. Both MNs reached a moderate IP3 of 30 dBm and 20 dBm, respectively. Some impedance states exhibited even lower linearity. Due to the superposition of higher order intermodulation products it's possible to observe 'sweet spots' with comparatively high linearity at some impedance states and power levels.

It's obvious that both networks do not fulfill the requirements for modern communication standards where an IP3 of more than 45 dBm is required. One possibility to enhance the linearity would be to use capacitors with less tunability. Simulations showed that reducing the tunability by a factor of three would lead to an IP3 of more than 45 dBm at the Γ_3-state of a Π-MN with the worst intermodulation behavior. For T-MN even a reduction by a factor of four would not be sufficient. Moreover, a further reduction of tunability would lead to even high tuning voltages for mobile devices. So other techniques must be used to enhance the linearity of a T-MN. One of this techniques could be a series connection of n capacitors with n-times the capacitance of a single capacitor like presented in [6], [7]. Simulations showed that a series connection of at least fifteen capacitors is necessary to realize an IP_3 of more than 45 dBm. Thanks to the comparatively small size of the capacitors in the T-MN the space requirements of the series connection should be acceptable.

V. Conclusions

Tunable Π- and T-MNs are able to cover a broad impedance range. Both topologies show mirror-inverted characteristics. Whereas Π- and T-MNs have the same mirror inverted linear behaviour, Π-MNs have a better linearity at the cost of higher required capacitances. Of course both topologies would benefit from higher quality factors of the capacitors. To investigate the nonlinear behavior source-pull measurements are necessary.



(a) **(b)**

Fig. 5. Nonlinear behavior for the different input reflection coefficients Γ_n plotted in Figs. 3(a) and (b). **(a)** Π-matching network **(b)** T-matching network.

Π-MNs show the worst linearity at high impedances whereas T-MNs show the highest intermodulation distortion at small impedances.

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