Design Methodology, Measurement and Application of MMIC Transmission Line Transformers

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Abstract

A design methodology is presented for MMIC transmission line transformers (TLTs) typically having three conductor levels. Several such structures on GaAs have been processed and measured showing good broadband performance. The usefulness of TLTs is demonstrated by an MMIC amplifier design.

Introduction

In the design of RF systems, various kinds of transformers play an important role. They are used for example in impedance matching networks, balanced–to–unbalanced transformations or as splitters and combiners. However, the conventional planar version used in MMIC technology [1], [2], [3] so far shows poor electrical performance and has not been used very much, therefore. In order to describe the layout of such conventional transformers, fig. 1 shows an S-band transformer coupled balanced MMIC amplifier example, where these planar versions have been used [4].

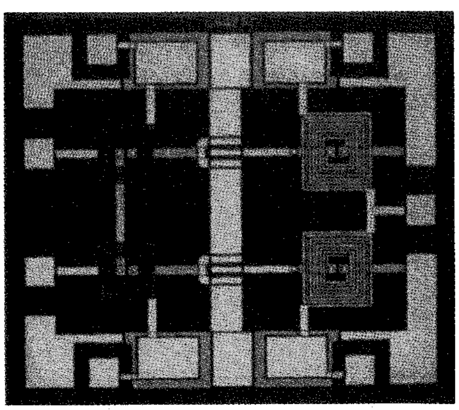

Figure 1: S-band transformer coupled balanced MMIC amplifier, ref. [4]

Their main problem is, that they consist of interwound spirals sitting in the same plane, which results in a relatively low inductive coupling. As they are on top of high permittivity material (GaAs), their capacitive effects are quite pronounced limiting their useful range of frequency with acceptable performance drastically. Also, with these transformers it is usually impracticable and even worse to have a different number of turns on the primary and the secondary side.

To avoid the problems associated with conventional MMIC transformers, TLTs, as already proposed in [5] can be used in a MMIC technology, that allows for more than two conductor levels. Our design methodology for such MMIC TLTs is presented in this paper. Since a number of these components have been processed on GaAs for different applications recently in our group, measured data is also available now for verification and for demonstration of the design success. Finally, the design of a 2–stage broadband HEMT MMIC amplifier using these novel transformers is discussed showing reduced space consumption for the matching networks.

Design methodology

In order to outline the problems associated with MMIC TLT design, fig. 2a and 2b shows the historical coaxial version of a 1:4 transformer in comparison to the respective schematic of its MMIC counterpart. In fig. 2a the characteristic impedance of

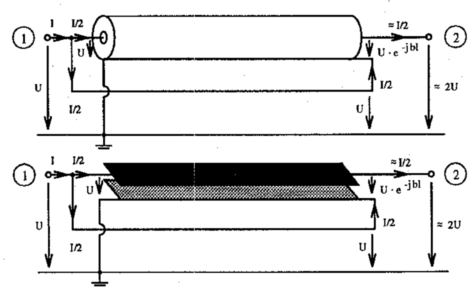

Figure 2: Schematic of a coaxial in comparison to a MMIC transmissionline transformer

the coaxial line has to be chosen as the geometric mean value of the port impedances to obtain optimum performance [6]. In the monolithically integrated TLT, the coaxial line is replaced by a two–strip transmission line consisting of a pair of stacked strip conductors. Note, that the straight line versions shown in fig. 2 are wound up in reality so that for the MMIC TLT the layout of fig. 3 results. With the hybrid mode character of the coupled strip structure in the MMIC TLT, the mode properties are different from the coaxial version. However a relatively simple design methodology has been found to successfully design such transformers for quite arbitrary port impedances and

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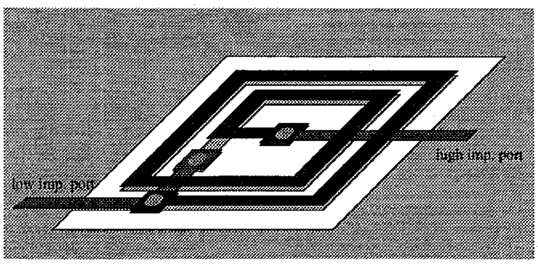

Figure 3: Layout of a MMIC transmissionline transformer

frequencies.

First, the strip widths have to be determined to result in input and output characteristic impedances of the transformer associated with the impedance levels to be matched. This is done using a quasi–static numerical analysis for two stacked conductors forming a a pair of coupled transmission lines in the MMIC layered dielectric medium. From the knowledge of the π-mode and the c-mode characteristics, an equivalent circuit is derived based on the representation of coupled strips as outlined in ref. [7]. This procedure is visualized in fig. 4. In fig. 4b $\varepsilon_{r,c}$

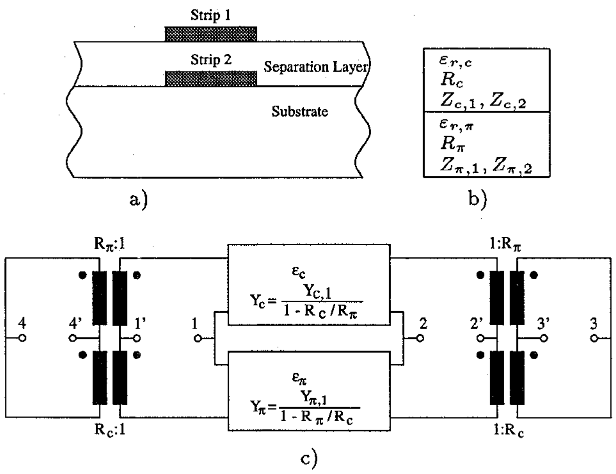

Figure 4: Derivation of equivalent circuit from the strips cross section[7]

and $\varepsilon_{r,\pi}$ are the effective dielectric constants of the c- and π-mode respectively. R_c and R_π are the associated voltage ratios of the modes and $Y_{c,\nu}$ and $Y_{\pi,\nu}$ are the mode-strip admittances of strip number ν,

Applying an equivalent circuit representation to the two coupled strips the circuit of fig. 4c results [7]. This equivalent circuit consists of two transmission lines with the characteristic impedances Z_c and Z_π which represents the two modes and ideal transformers ensuring the correct voltages on the strips. The characteristic impedances are related to the mode characteristic data as indicated in fig. 4c. Looking deeper into the mode characteristics, it can be seen, that the c-mode represents the equivalent to the coaxial transmission line mode ($U_1 \approx U, U_2 \approx 0, i_1 \approx -i_2$), whereas the π-mode is the parasitic mode ($U_1 \approx U_2, i_1 \approx 0, i_2 \approx i$). However, due to the spiralshaped configuration and the resulting increased inductance of the coupled lines, unbalanced currents are suppressed, so that the π-mode can be neglected for the first design step. It can also be seen in fig. 4c, that the presence of the transformers usually implies the characteristic impedance of the remaining transmission line to be different when referenced to port 1 and 2 as compared to reference planes at port 3 and 4, respectively. Our investigation has shown, however, that this difference disappears to a first approximation when winding up the two strips because of the suppression of unbalanced currents. So, the transformer ratio 1 : R_π in fig. 4c changes to about unity, and the characteristic impedance of the c-mode transmission line is increased by about R_π. Applying these simplifications to the equivalent circuit in fig. 4c and forming a TLT out of the two strips, the following circuit results. In

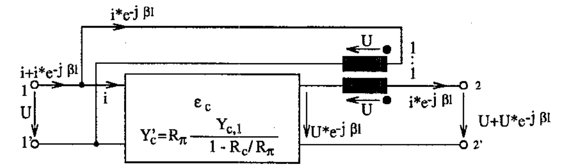

Figure 5: Simplified equivalent circuit of a MMIC TLT

this circuit, the transmission line has now to be matched for optimum broadband operation of the transformer. When neglecting the phase rotation of the transmission line $e^{-j\beta_c l}$, this condition results in the following equations.

$$Z_{ref,1} = \frac{1}{2} \cdot Z_c R_\pi$$

$$Z_{ref,2} = 2 \cdot Z_c R_\pi$$

$$Z_c = (1 - R_c/R_\pi) \cdot Z_{c,1}$$

Using these formulae, the required strip widths are determined iteratively using our previously mentioned quasi-static 2D analysis program. Since the only variables for a given MMIC process are the strip widths, the optimum values are obtained with only a few iterative steps.

The next step in the design procedure is to design for the upper bound of the useful frequency range. This can be done in complete analogy to a coaxial TLT. The frequency dependent power ratio $P_2/P_{s,avail}$ can be calculated using Ruthroffs formula [8].

$$\frac{P_2}{P_{s,avail}} = \frac{4(1+\cos(\beta_c l))^2}{(1+3\cos(\beta_c l))^2 + 4\sin^2 \beta_c l}$$

So, if a specific upper frequency of operation is required, the length l of the spiralshaped stacked conductors has to be chosen appropriately. Using the above formula, the -1 dB frequency is approximately obtained, when the length of the conductors is chosen as $\lambda_c/4$.

Having calculated the length of the unwound TLT, the layout can be defined by choosing an appropriate number of turns. Since the low frequency behaviour of a TLT is given by the inductance of the spiralshaped structure, the lower frequency bound can be determined approximately by the calculation of a standard MMIC single layer spiral inductor having the same floorplan as the TLT. As can be seen in fig. 2, this inductance is connected in shunt configuration to ground, so that the low frequency equivalent circuit of fig. 6 results [6]. Following the outlined step by step procedure an accurate quasi–static simulation approach [9] is finally applied to check the transformer performance before processing and use optimisation for fine tuning.

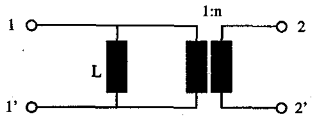

Figure 6: Low frequency equivalent circuit of a MMIC TLT, $n = 2$ for the outlined transformer type

Since our design approach usually leads to near optimal results, the final correction steps are performed very efficiently.

The described design procedure works in analogy for other kinds of transmission line transformers [6] and for other transformer ratios like $n = 3$ or $n = 4$ in the extreme. The only difference in then in the prediction of the useful frequency range.

Verification example

To demonstrate and verify the design procedure, the MMIC TLT shown in fig. 7 is used.

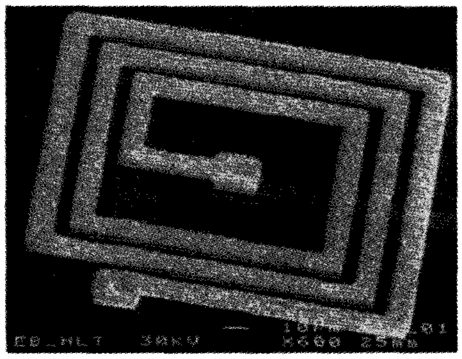

Figure 7: 500 μm GaAs substrate, $2 \times 1\,\mu$m polyimide separation, 10 μm strip widths, 5 μm lateral spacing, square 160 μm outer dimension, 1290 μm total length of winding, (lowest conductor level not visible)

Applying our 2D analysis program to the two-strip transmission line cross section the following mode characteristics result:

$R_c = 0.125$	$R_\pi = 1.12$
$Z_{c,1} = 38.41\,\Omega$	$\varepsilon_{r,c} = 2.8$

The described design procedure as applied to this TLT results in the port impedances of $19\,\Omega$ at port one and $76\,\Omega$ at port two. From the length of the unwound transmission line the approximate upper -1 dB frequency is calculated as 34.7 GHz. The value of the equivalent low frequency inductance is 1.4 nH, so that the low end -1 dB frequency is about 2.0 GHz. The measured (RFOW) S-parameters of the designed transformer referenced to the mentioned impedances are shown in fig. 8. As can be seen, be the impedance transformation is indeed achieved over a broad frequency range with low reflection at both ports. The minimum insertion loss is 1.5 dB, however, due to the use of thin metal (0.4 μm) for the lower strip and due to the relatively thin dielectric separation layers. This can be reduced further just by the use of thicker dielectric separation and metal. Looking at the transmission coefficient, it can be seen that the -1 dB frequency range is smaller than desired, which is also a result of the prevailing metal loss as our simulations [9] with different conductance values have shown.

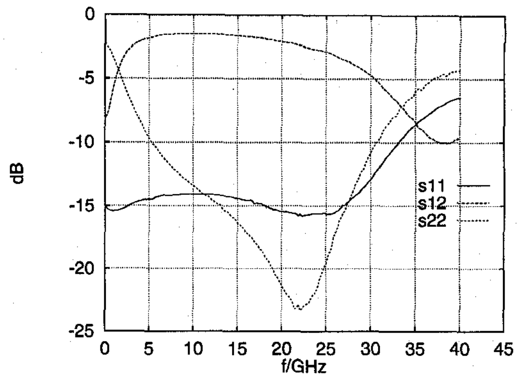

Figure 8: Measured S-parameters of the MMIC transmission line transformer of fig. 7, $Z_{ref,1} = 19\ \Omega$, $Z_{ref,2} = 76\ \Omega$.

As an example for a component on silicon, which is suitable for mobile communication applications, fig. 9 shows the measured S-parameters [10] of a MCM TLT designed for the frequency range of 460 MHz to 4.4 GHZ and for the port impedances of $50\,\Omega$ and $200\,\Omega$. Here, the minimum insertion loss of the transformer is only 0.5 dB due to the lower frequency range and also due to the use of relatively thick dielectric separation layers ($2 \times 6\,\mu$m). The measured -1 dB frequency range is also close to the values predicted using the method described here.

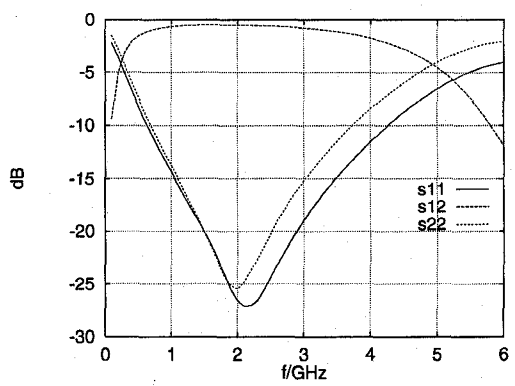

Figure 9: Measured S-parameters of a MCM transmission line transformer on thick silicon, ($Z_{ref,1} = 50\ \Omega$, $Z_{ref,2} = 200\ \Omega$) [10].

Device fabrication

The above introduced MMIC TLT is processed on a 500 μm thick GaAs-substrate. In fig. 10 the cross–section of the whole device is depicted. After processing the first metalization layer (Ti, Au) by lift off, the surface must be preconditioned before the deposition of the first polyimide layer. The wafer must be cleaned by solvents and a dehydration process. Prebake at 200°C produces a rough surface which promotes a good adhesion of the polyimide on the metal layer. The deposition of the polyimid itself is realized by spin coating. Optimized acceleration time, spinning time and spinning speed are used to produce a total layer thickness of $h_p = 1\,\mu$m and a high

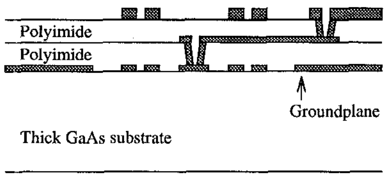

Figure 10: Cross–section of a MMIC TLT

planarization, which is most important in application of polyimide dielectric layers in MMICs. Usually the patterning of thick polyimide layer ($h > 1\,\mu$m) is done by wet or dry etching processes [11], which require an additional lithography step. In contrast to this conventional method we use, after a prebake (15 min at 95°C), a photosensitive polyimide for direct patterning. A special proximity printing technique ($\lambda = 400\,$nm) enables via hole structures with the desired profile as small as $d = 10\,\mu$m. On top of the structured polyimide a further metalization layer is deposited by lift off; again (fig. 10) the polyimide/metalization sequence is repeated, thus offering a three conducting layer stack. A conventional galvanic layer on top of the device, which may includes air bridge constructions, complete the whole structure of the TLT.

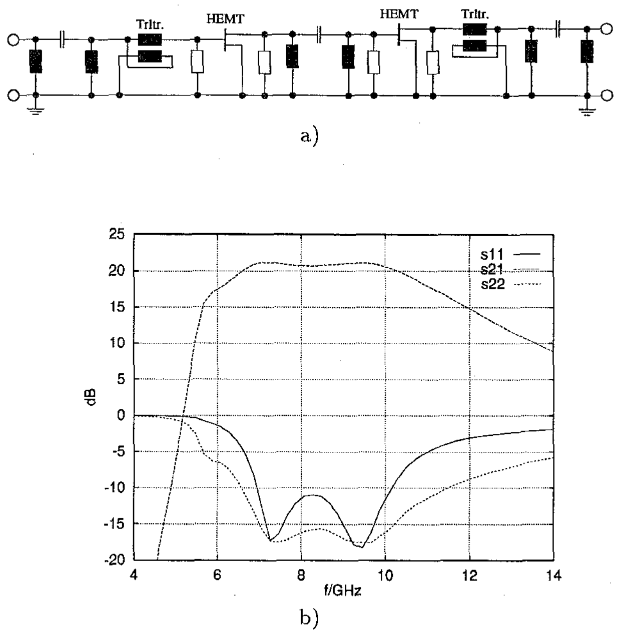

Figure 11: 2-stage HEMT amplifier using transmission-line transformers for impedance matching and simulated S-parameters

Application of MMIC TLTs

In order to demonstrate the advantages obtained when using TLTs like those described in the matching networks of MMIC amplifiers, a 2–stage 7–10 GHz amplifier using InP HEMTs has been selected as a test case. In this frequency range, broadband InP HEMT matching is not easily accomplished using conventional MMIC matching structures due to the high input impedance of these HEMTs. Connecting TLTs directly to the gate of the first HEMT and to the drain of the second one, reduces the problem of broadband matching considerably, so that straightforward further matching with realizable MMIC element values can be applied subsequently. As it turns out, the described TLTs are even well suited to match complex loads. The result of this exercise is indicated in fig. 11a and 11b. Excellent broadband matching and flat gain of more than 20 dB in obtained over the range of 7–10 GHz. The input and output matching of the two–stage InP MMIC amplifier was obtained using a commercial broadband matching CAD program [12]. Performing the same exercise of matching without the described transformers did not lead to any usefull result. Thus the new vertically integrated TLTs provide a considerable design advantage. Due to their small size, precious substrate space can be saved resulting in MMIC chip size and cost reduction.

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