A Measurement Fixture Suitable for measuring Substrate Noise in the UWB Frequency Band
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A measurement fixture suitable for measuring substrate noise in the UWB frequency band

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Abstract This paper presents a measurement fixture suitable for measuring substrate carried noise for lightly doped substrates within the UWB frequency band. Signals coupling through the substrate are usually fairly weak and special precautions are taken to avoid any distortions that may be caused by the test fixture. The proposed measurement fixture is based on a modified ground-signal-ground (GSG) pad fixture. Parasitic effects in the measurement fixture are evaluated with help of an equivalent circuit model. From measured results, the presented fixture is shown to provide a measurement band from 3 to 10 GHz. As an example of the usability of the presented fixture a test-case using a class-E PA is presented. The noise injected by the class-E PA is measured using the proposed fixture as noise detector and the results are show to be accurate to within ±1.5 dB.

Keywords Wide band measurements · Substrate noise · UWB · GSG

1 Introduction

UWB communication is one of the technology platforms currently being investigated in support of both low data rate and high data rate applications. It is considered to be one of the most promising wireless technologies for short-range systems [1]. Due to an advantage in cost and device density, CMOS technologies, especially lightly doped technologies, hold a great potential of complying with the needs for cheap and compact implementation of UWB devices. However, UWB systems demand stringent noise figure owing to its ultra wideband nature (larger than 500 MHz within the allocated frequency band from 3.1 to 10.6 GHz) and the low power spectral density of the signal [2]. The noise issue is of particular concern for UWB systems implemented on a single chip as substrate noise here will further degrade performance. In such scenarios, sensitive RF circuits such as low noise amplifiers have to be integrated together with noisy circuits, such as digital circuits, switch-mode power amplifiers, frequency dividers etc. Those noisy circuits result in the generation of substrate noise in a wide frequency band. Due to the low resistivity of the substrate (10–20 Ohm-cm for lightly doped substrates), the substrate noise can easily propagate to the RF circuit area and deteriorate the noise figure.

While it may be easy to conceptualize the coupling through the substrate, it is much more difficult to measure this noise in any real circuit. This is especially the case measurements are needed for in an ultra wide frequency band where the valid measurement frequency band usually is limited by the parasitic effects in the measurement setup. So far, a number of studies have been conducted on the measurement of substrate noise [3–11]. The reported studies attempt to address the problem of substrate noise measurement from different perspectives. Some try to
verify the coupling model of the substrate [3–5]; some focus on heavily doped CMOS or other integrated circuit processes [6–8]; some investigate the waveforms or narrow band power spectral density (PSD) of substrate noise [9]; while others propose complicated circuits or devices as substrate noise sensors [10, 11]. However, few publications have investigated an appropriate wide band measurement method for measuring substrate noise in UWB systems implemented using lightly doped CMOS processes.

This paper presents a measurement fixture that is suitable for measuring substrate carried noise for lightly doped substrates within the UWB frequency band. The aim is here to obtain a measurement of the wide band PSD of the substrate noise. A measurement fixture based on a modified GSG pad is shown to have high potential for accomplishing this. A test chip is designed using a 0.18 \( \mu m \) lightly doped CMOS process to evaluate the effects of two key parasitic factors, the distance-based substrate resistance and the capacitance between the substrate and the Metal_1 ground. Several practical design rules for reducing the parasitical effects are concluded based on measurement results. As an example of the usability of the proposed method, switching noise measurements are presented for a switch-mode class-E PA.

This paper is organized as following: Sect. 2 describes the substrate noise measurement fixture. Section 3 discusses the evaluation of the effects of different measurement distances and the capacitive coupling between the substrate and the Metal_1 ground plane of the measurement setups. Section 4 presents the experiment on measuring of switching noise generated by a class-E PA. In Sect. 5 conclusions are drawn.

2 The substrate noise measurement fixture

To experimentally investigate the substrate noise in an UWB system, the measurement setup must have a sufficiently wide measurement bandwidth. It is therefore important to avoid introducing complicated parasitic components or devices in the measurement setup, which can dramatically affect the measurement results. The presented substrate noise measurement fixture, also referred to as the substrate noise detector, is designed based on a modified GSG pad structure composed of two 85 \( \times \) 85 \( \mu m \) ground pads and one 85 \( \times \) 85 \( \mu m \) signal pad. The center-to-center distance between the signal pad and each ground pad is 150 \( \mu m \). Thus, no extra fixture apart from a GSG probe is needed for the measurement of the substrate noise. The cross section view of the fixture is shown in Fig. 1. For simplicity only one metal layer, the Metal_1 layer, is shown here. In the actual setup all metal layers are connected to the top metal layer.

It should be noted that the signal pad is directly connected to the substrate while the ground pads are not connected to the substrate. There are two reasons for this:

\begin{equation}
C_{pad} = \frac{W \cdot L}{e_{ox} t_{ox}},
\end{equation}

where \( W, L \) are the width and the length of the pad respectively, and \( e_{ox} \) and \( t_{ox} \) are the permittivity and thickness of the oxide layer between the pads and the substrate, respectively.

For lightly doped processes, the substrate can not be treated as one single node like the case is for heavily doped substrates [8]. To account for this, \( R_1, R_2, C_{sub} \), and \( R_{sub} \) are added to form a simplified network model of the substrate.
$R_1$ and $R_2$ are used to model the spreading resistance of the substrate below the signal pad and the ground pad. $R_{\text{sub}}$ and $C_{\text{sub}}$ are used to model the resistive and capacitive coupling between the substrate noise source and the measurement fixture. The values of $R_1$, $R_2$ and $R_{\text{sub}}$ are difficult to precisely estimate from available parameters and are therefore derived from experimental data.

3 Evaluation of parasitic effects

In practical measurements of substrate noise, the substrate noise propagating from the noise source to the measurement point experiences attenuation and distortion. Distance-based substrate resistance and capacitive coupling between the substrate and the ground of the measurement setup are two key factors responsible for these effects. In order to obtain accurate measurements of the substrate noise, it is necessary to study the effects of such parasitics.

3.1 Design of the test chip

Based on the substrate noise measurement fixture in Sect. 2, a test chip has been designed to evaluate the effects of distance-based substrate resistance and the capacitive coupling between the substrate and Metal_1 ground. Figure 2 shows the topology of the test chip. Four substrate noise detectors (denoted A, B, C and D) are placed with separating distances of 230, 300 and 300 μm, respectively. For each signal pad and ground pad, all the available metal layers from Metal_1 to Metal_6 are connected. Two wide metal strip lines composed of Metal_1 connect ground pads of the substrate noise detectors to produce capacitive coupling between the substrate and the ground of the measurement setup.

Figure 3 shows an equivalent circuit model of the test structure. Here, there are four sub-circuit networks indicated by four dashed squares. Network 1 and 2 are models of the substrate noise detectors at port 1 and port 2. Network 3 is the model of the forward coupling network between the two measurement ports, including resistive coupling and capacitive coupling. Network 4, composed of $C_{\text{strip}}$, $R_3$ and $L_{\text{strip}}$, is used to model the coupling between the Metal_1 ground strip lines and the substrate. Similar to $C_{\text{pad}}$ in the model of the noise detector, $C_{\text{strip}}$ can be roughly calculated from the oxide thickness and the strip dimensions. However, for the test structure in Fig. 2 the dimension of the ground strip is so large that the electric field between it and the substrate cannot be treated as a uniform field and therefore needs more consideration. Figure 4 shows the results of an EM simulation of the electric field in the oxide layer beneath the ground strips using the finite element method. A 6 GHz signal source is placed on the signal pad of noise detector A in Fig. 2 to excite the simulation. The electric field is shown using a 1-D electric field curve taken from the oxide layer beneath one of the ground strips in Fig. 2. It can be seen that the strength of the electric field decreases quickly as the distance between the observing point and the noise detector increases. This means that $C_{\text{strip}}$ is over-estimated if its value is calculated using the actual length of 915 μm.

From Fig. 4, it can be seen that the electric field remains at a high level for distances of 50 μm and then drops quickly as the distance increases. Therefore, for calculating $C_{\text{pad}}$, the length, $L$, is decided as 50 μm and the width, $W$, is 85 μm. Then the oxide capacitance for one ground pad is calculated as 127.5 fF. Consider that two ground pads are placed in parallel, the value of $C_{\text{pad}}$ is decided as shown in Table 1. Moreover, the electric field is found concentrated within a
distance of 480 µm, which indicates the main area where the capacitive coupling happens. Thus it is reasonable to calculate $C_{\text{strip}}$ according to this distance. For example, the length, $L$, is decided as 380 µm, subtracting $2 \times 50 = 100$ µm from 480 µm, for the case of AB. Then the $C_{\text{strip}}$ for the case AB is calculated as 1.94 pF and finally decided as 1.8 pF due to the decreasing of the electric field. Similarly, $C_{\text{strip}}$ for the case of BC and BD are decided as $X_3$ pF from 480 µm, for the case of AB. Then $L_{\text{strip}}$ is the inductance of the strip line. Generally, the value of $L_{\text{strip}}$ is small and 1 pH is used in the equivalent circuit.

### 3.2 Measurement of the DC resistances

To determine the distance dependency of the substrate resistance, the DC resistance between the different signal pads of noise detector A, B, C and D are measured and the results are shown in Fig. 5. The DC resistances with distance of 230 µm (AB), 300 µm (BC), 530 µm (AC) and 600 µm (BD) are measured as 360 Ω, 380 Ω, 405 Ω and 410 Ω, respectively. It is seen that the resistances are on the order of hundreds of Ohms and the variation of the resistances is insignificant. This indicates that, for measurement distances from 230 µm to 600 µm, the attenuations caused by substrate resistances are large but fairly constant. Furthermore, the values of $R_1$ and $R_{\text{sub}}$ in the equivalent circuit of Fig. 3 are derived based on these measurement results. For instance, in the case of BC, the value of $2 \times R_1 + R_{\text{sub}}$ is decided as 380 Ω, which equals the measured DC resistance. Since the ground pads are not connected to the substrate, it is difficult to measure the spreading resistance between the signal pad and the ground pad. But its value can be approximated based on the measured DC resistances in Fig. 5. The ground pads have the same dimensions as the signal pads and the central to central distance is 150 µm. So the spreading resistance of the substrate below the signal pad and the ground pad can be approximately decided as 360 Ω. Further, considering that two ground pads are connected in parallel, the spreading resistance values of $R_2$ and $R_3$ are approximated as shown in Table 1.

### 3.3 Measurement of S-parameters

To determine the frequency dependency of the propagation effects, the test chip is studied using S-parameter measurements over 45 MHz to 10 GHz. The measured S21s over distances of 230 µm (AB), 300 µm (BC) and 600 µm (BD) are illustrated in Fig. 6. The simulation results of the equivalent circuit with parameters in Table 1 are also shown in Fig. 6. From these results the equivalent circuit model is seen to match measured data accurately, especially in case AB and BC. Relatively larger deviations are seen in case BD. For this case, the magnitude of S21 is around $-45$ to $-50$ dB at high frequencies, which makes accurate measurement a non-trivial task. And thus a reduced accuracy is expected over high frequencies. It is also seen in Fig. 6(a) that the attenuations at 45 MHz with different distances are almost identical at $-14$ dB. For such a low frequency, all capacitive couplings can be approximately treated as open circuit and the attenuation is mainly caused by the substrate resistance. Since the values of the substrate resistances of different distances are similar the attenuations are of course similar too.

Moreover, high attenuations are observed at high frequencies. For instance, attenuations of $-32$, $-38$ and $-47$ dB are observed at 8 GHz for the three different cases. From Table 1, it is noticed that $R_{\text{sub}}, C_{\text{sub}}, R_3$ and $C_{\text{strip}}$ are the only four changing variables that might be responsible for this. $C_{\text{sub}}$ is insignificant due to its small value. $R_{\text{sub}}$ is also excluded by reason of its near constant value in the three cases. Thus the different attenuations over high frequencies of three cases must be resulted from different values of $R_3$ and the intentionally built capacitances between the Metal_1 ground strip lines and the substrate, i.e. $C_{\text{strip}}$. For higher frequencies, coupling through $R_3$ and $C_{\text{strip}}$ to ground is more

<table>
<thead>
<tr>
<th>Items</th>
<th>AB</th>
<th>BC</th>
<th>BD</th>
</tr>
</thead>
<tbody>
<tr>
<td>$R_{\text{sub}}$ (Ω)</td>
<td>80</td>
<td>100</td>
<td>130</td>
</tr>
<tr>
<td>$C_{\text{sub}}$ (fF)</td>
<td>30</td>
<td>20</td>
<td>10</td>
</tr>
<tr>
<td>$R_{\text{tip}}$ (Ω)</td>
<td>3</td>
<td>3</td>
<td>3</td>
</tr>
<tr>
<td>$R_1$ (Ω)</td>
<td>140</td>
<td>140</td>
<td>140</td>
</tr>
<tr>
<td>$R_2$ (Ω)</td>
<td>40</td>
<td>40</td>
<td>40</td>
</tr>
<tr>
<td>$R_3$ (Ω)</td>
<td>60</td>
<td>30</td>
<td>10</td>
</tr>
<tr>
<td>$C_{\text{pad}}$ (fF)</td>
<td>250</td>
<td>250</td>
<td>250</td>
</tr>
<tr>
<td>$C_{\text{strip}}$ (fF)</td>
<td>1800</td>
<td>2000</td>
<td>2400</td>
</tr>
<tr>
<td>$L_{\text{strip}}$ (pH)</td>
<td>1</td>
<td>1</td>
<td>1</td>
</tr>
</tbody>
</table>
pronounced than that of low frequencies and thus the magnitude of S21 decrease. However, all S21 magnitude responses are flat over the frequency band of 3 to 10 GHz (with fluctuation less than 3 dB), which means the measurement setup can be used to measure the PSD of substrate noise even under conditions of strong parasitic effects.

Considering the measured S21 in Fig. 6 and the corresponding values of $R_3$ and $C_{strip}$ in Table 1, it is seen that a larger $R_3$ and a smaller $C_{strip}$ helps to decrease the loss. Then several design rules of the presented measurement fixture in practical applications can be drawn as following. The value of $R_3$ should be increased by reducing the dimensions of the ground pads on Metal_1 layer but keep the size of $85 \times 85 \mu m$ on the top metal layer for measurement purpose; $C_{strip}$ should be reduced by deleting the Metal_1 ground around the noise detector or using the top metal layer to connect the ground pads. Moreover, Since the distance-based substrate resistances for measurement distances from 230 to 600 $\mu m$ are similar, distance is not a key factor to avoid significant attenuation. A long distance between the substrate noise source and the noise detector like 600 $\mu m$ is acceptable if needed for other purposes such as convenient probing.

### 4 Experiment on measuring substrate noise generated by a class-E PA

As a practical example, the presented substrate noise measurement fixture is used to measure the switching noise generated by a switch-mode class-E PA. The class-E PA is designed in the same 0.18 $\mu m$ technology as the test chip in Sect. 3 and hence all conclusions and design rules drawn above can be applied. In this experiment, the PA acts as a switching noise source and the proposed measurement fixture as noise detector. When the PA is driven by high frequency signals, the generated switching noise occupies a wide frequency band, which makes it a suitable signal source to validate the wide band characteristic of the measurement fixture. Moreover, when the PA is switching on and off, the switching current is injected into the substrate mainly through the junction capacitor at the drain of the biggest NMOS transistor in the PAs output stage as shown in Fig. 7. Since the junction capacitance of the NMOS transistor is big, it is here treated as short circuit for high frequencies and the signal at the drain node is coupled into the substrate without significant distortion. Thus the switching noise measured at the substrate noise detector can be compared with the signal measured at the drain node to find out the distortion caused by the substrate noise detector over a very wide frequency band.

While designing the substrate noise detector on the chip, several factors are taken into account. Firstly, to have a clean reference ground, the substrate noise detector’s ground is closely connected to the ground pad of the DC supply. Secondly, for convenient probing, the substrate noise detector is placed at the edge of the chip with a distance of approximately 600 $\mu m$ to the output stage of the class-E PA. Finally, following the design rules drawn in Sect. 3, to avoid large attenuations, the Metal_1 ground plane around the substrate noise detector is deleted to reduce the capacitance between the Metal_1 ground plane and the substrate. The chip photo is shown in Fig. 8.

Firstly, the frequency spectrum of the signal at the drain node of the PAs output stage is measured when the class-E
PA is operating at maximal output power. Three discrete spectral components are measured at 2.44, 4.88 and 7.32 GHz, i.e. the fundamental and harmonics of the signal, with magnitudes of $-12.5$, $-49.5$ and $-35$ dBm, respectively. The fourth harmonic at 9.76 GHz is too weak to be measured. But the three spectral contents already cover a wide frequency band, which makes it usable for verifying the wide measurement bandwidth of the measurement fixture.

Based on the assumption that the signal at the drain node is coupled into the substrate without significant distortion, the substrate noise measured at the noise detector is expected to have a similar frequency profile except suffering from certain resistive attenuations. To verify this, the substrate noise is measured at the noise detector and the result is presented in Fig. 9. It shows that the spectral contents of the measured substrate noise appear at 2.44, 4.88 and 7.32 GHz with magnitudes of $-36$, $-72$ and $-60$ dBm, respectively. Then the attenuations for each component from the drain node to the noise detector are found to be $-23.5$, $-22.5$ and $-25$ dB, respectively. As shown in Fig. 10, it can be seen that the attenuation profile is quite flat in the band of 2.44–7.32 GHz with an average value of $-23.75$ dB and variation of ±$1.25$ dB. This validates the wide band characteristic of the presented measurement method.

5 Conclusion

A measurement fixture is presented for the measurement of substrate noise in lightly doped substrates within the UWB frequency band. The effects of the distance-based substrate resistance and the capacitive coupling between the substrate and the ground are evaluated by measurement of a test chip. An equivalent circuit model of the measurement fixture is given and shows accurate fit. The simulated and measured results show that this measurement fixture is able to provide reliable measurements over the UWB frequency band from 3 to 10 GHz. Furthermore, different measurement distances from 230 to 600 µm are found to have similar attenuations at low frequencies, while obvious effects of capacitive coupling are observed over high frequencies. To avoid such effects on measured substrate noise the capacitance between the substrate and the ground of the measurement setup should be minimized. Finally, the presented measurement fixture is validated by measuring the switching noise generated by a practical class-E PA.

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References


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