High speed test interface module using MEMS technology

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ARTICLE INFO

Article history:
Received 26 August 2014
Received in revised form 12 November 2014
Accepted 12 November 2014
Available online 31 December 2014

Keywords:
Test interface module
Automatic test equipment
High speed manufacturing test
Micro-electromechanical systems (MEMS)

ABSTRACT

At frequencies above a few gigahertz, testing integrated circuits becomes a challenging task. Test signal integrity degradation due to parasitic effects of interconnects and electromagnetic coupling undermine the test results and increase the yield loss of integrated circuits at high speeds. A new test interface module based on MEMS technology is proposed in this paper. High-speed micro test-channels are designed to establish connectivity between the device under test and the tester at the die level. Experimental results indicate that the proposed architecture can be used to test integrated circuits up to 50 GHz without much loss or distortion.

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1. Introduction

With the transient frequency of available CMOS technologies exceeding 200 GHz and increasingly complex IC designs, it is now apparent that the architecture of current testers needs to be greatly improved to keep up with the formidable challenges ahead. The test requirements for modern integrated circuits are becoming more stringent, complex and costly. These requirements include an increasing number of test channels, higher test-speeds and enhanced measurement accuracy and resolution. High-speed test signals experience a broad range of nonlinearities associated with the signal paths [1–5]. When the frequency of interaction between the Device Under Test (DUT) and the Automatic Tester Equipment (ATE) approaches the gigahertz range, the effects of transmission lines, surface resistance, coupling and discontinuities become a critical issue. These undesired effects eventually undermine the test results. Current state of the art ATEs, in general, requires a Device Interface Board (DIB) to perform test on integrated circuits. The DIB provides temporary electrical connections between the DUT and measurement instruments within the tester. The DIB also provides a space for DUT-specific local circuits such as buffer amplifiers, mixers, FPGA, and load circuits [6]. It is highly desired to reduce the physical distance between the DUT and the tester to lower the transmission line effects and electromagnetic coupling. The traditional approach is to locate ATE Pin-Electronics (PE) as close as possible to the DUT. In practice, general purpose pin-electronics are designed to meet a variety of requirements to cover different test scenarios. As a result they are commonly bulky and cannot be readily integrated and placed adjacent to the DUT I/O pins. Even in the state of the art ATEs, the physical separation between DUT and pin-electronics exceeds several inches.

To reduce the length of the transmission line between tester resources and the DUT, various techniques have been introduced in the literature. Mydill proposed a technique in which most of the test head circuitry have been removed and located remotely in an attached mainframe chassis [7]. Although the distance is reduced considerably, the loss due to skin effect and dielectric material in the 8-foot cable becomes significant beyond the 100 MHz range.

In [8], Keezer and Zhou proposed a solution to replace general purpose pin-electronics with technology-specific transceivers. This architecture minimizes the transmission line effects but at the cost of extra buffer ICs and restricting the flexibility of the general purpose pin electronics.

To meet the test requirements for high-speed system-on-chip devices, leading ATE manufacturers offer additional extension instrumentation to generate high-speed signals. Verigy, for example, extends the performance of its scalable V93000 series with a Pin Scale HX extension card generating signals up to 12.8 Gbps [9]. The Pin Scale HX channel card combines the functionality of at-speed ATE channels with high-integrity loopback. A similar card is proposed by Credence that supports signal up to 6.4 Gbps with a jitter less than 30 ps peak-to-peak [10]. These cards are relatively expensive and they require a major capital investment.

In [11], a test module is designed using standalone mini-testers which consist of a digital logic core (DLC) designed with an FPGA.
This method is suffering from some timing errors due to XOR multiplexing gates which are known for data dependent jitter (DDJ) in addition to random jitter. Moreover, the ATE DC and AC parametric test instruments and functional test channels must be connected to the device under test via an array of RF switches which add some complexity to the test environment.

In this paper, a new solution based on MEMS technology is proposed to reduce the physical distance between the tester and device-under-test to a few hundred of micrometers. As a result, the undesired effects of transmission lines are significantly attenuated allowing the test channels to operate up to 50 GHz without considerable signal integrity degradation. The rest of the paper is organized as follows. Section 2 describes the architecture and the building blocks of the proposed MEMS interface module. Circuit models are derived in Section 3. Simulation and experimental results are discussed in Sections 4 and 5, respectively, followed by conclusion in Section 6.

2. MEMS DIB architecture

The architecture of the proposed MEMS Test Interface Module (TIM), which is designed using IntelliSuite CAD tool, is shown in Fig. 1a. It includes a fixed and a removable socket [15]. The fixed socket acts as an interface between the removable socket and the ATE test channels. The removable socket provides necessary means to support temporary connectivity between the device under test and the tester. It also provides a physical location for the device under test and includes contact springs for temporary connections. The removable socket can be customized based on the size of DUT and the number of I/O pins. In the test phase, a pressure mass is positioned on top of the die under test to keep it tightly pressed against the removable socket connecting the die to the tester.

2.1. Fixed socket

It includes solder pads on the bottom side to connect to ATE pin electronics. The electrical connections between the removable socket and the tester resources are established through a bed of micro pins located on the top side of the fixed socket. These micro pins are in fact the substitutes for the conventional pogo pins [16] commonly used in automatic test equipment.

2.2. Removable socket

The removable socket incorporates cantilever type contact springs to provide temporary electrical connections between the DUT I/O pins and the ATE resources. The die under test is positioned on top of the removable socket. Four circular alignment keys are integrated in the fixed and removable sockets to ensure proper positioning of the structure during the test stage. Each contact spring, shown in Fig. 1b has a square geometry of $100 \times 100 \mu m$ with a contact pad dimensions of $50 \times 50 \mu m$ and $10 \mu m$ height. It is supported by eight beams of equal length to maintain symmetrical pressure around the contact pad. The contact surface is $10 \mu m$ above the substrate plane to allow deflection during the operation. The contact spring is designed to satisfy the following objectives: (a) smooth and flat contact surface to maximize the contact area in order to achieve a lower contact resistance, (b) square area to match the die contact pads, (c) negligible shear and high elastic capability. Moreover, the design must avoid a concave deformation at the top surface of the contact pad that confines the contact area within the concavity shape.

Fig. 1c shows the cross section of a contact spring and a DUT I/O pad in the test phase after an induced pressure. The contact spring experiences elastic deformation due to the applied pressure in the test phase. This temporary deformation ensures the electrical connectivity between the die I/O pads and the fixed socket. The proposed MEMS interface can be modified to accommodate test interface circuits usually required to perform functional tests on high speed analog and RF circuits [17,18]. MEMS TIM as compared to the currently used device interface boards provides necessary means for fault detection at the die level. This will enable manufacturers to detect faults before packaging. Thus, the cost of packaging which is a major portion of the overall cost of fabrication is eliminated for faulty DUTs. Furthermore, MEMS TIM can readily meet the requirements of next generation of small pitch devices predicted in the International Technology Roadmap for Semiconductors (ITRS) [16]. The proposed MEMS based test solution includes necessary contact springs to simultaneously contact all pads of a die-under-test in one touchdown. This will allow the proposed probing method to be used for either wafer probing or die testing. To conduct wafer level probing, each die can be tested separately in one touchdown until all dies are covered. It is also possible to have multiple MEMS probes for parallel wafer-level probing.

3. Circuit models

A circuit model for the proposed MEMS TIM has been developed to predict its behavior at high frequencies. A close view of the
MEMS contact spring which connects the top and bottom of the removable MEMS socket in Fig. 1 is shown in Fig. 2a. The effects of fixed socket can be compensated through proper calibration of ATE test channels thus in practice the signal path on the removable socket dominates the behavior of the entire MEMS interface module. Fig. 2b presents a cross section of one contact spring path. To develop a circuit model for the signal path, the transmission line between ATE test channels and DUT I/O pads and the discontinuity at the contacts have been taken into consideration. The circuit model for the proposed MEMS TIM is shown in Fig. 2c. It includes three distinct parts of (a) the top contact which is formed between a contact spring and an I/O pin of DUT (b) the conductive path between the top and bottom pads through the proposed structure and (c) the bottom contact that is formed between the contact spring and the ATE test channel. Due to the geometrical symmetry of the top and bottom contact pads, an identical model for both contacts has been employed.

### 3.1. Circuit model for micro-scale contacts

The behavior of contacts can be approximated by $R_{\text{cont}}$, $C_{\text{cont}}$ representing the contact resistance, contact capacitance parameters respectively [19].

#### 3.1.1. Contact resistance $R_{\text{cont}}$

On microscale level, all solid surfaces are rough. The existence of surface roughness between two contacting materials results in electrical contact resistance. The actual contact is created at distinct spots formed by the surface roughness at spots known as a-spots [20]. Consequently, the electrical current lines cluster together to pass through microscopic contact areas. The constriction of electrical current at a-spots reduces the total electrical current passing through the bulk interface creating contact resistance. In the proposed model, a single circular a-spot is considered to derive the electrical contact resistance. Using Laplace’s equation [21,22], the constriction resistance for a circular a-spot in a cylindrical conductor with radius $R$ is given by

$$
R_{\text{cont}} = \frac{\rho}{2\pi} \left[ 1 - 1.41581 \left( \frac{a}{R} \right)^2 + 0.06321 \left( \frac{a}{R} \right)^4 \right] + 0.15261 \left( \frac{a}{R} \right)^5 + 0.19998 \left( \frac{a}{R} \right)^6
$$

where $\rho$ is the electrical resistivity, and $a$ is the constriction radius. For $a \ll R$, Eq. (1) reduces to [23,24].

$$
R_{\text{cont}} = \frac{\rho}{2\pi}
$$

It is widely accepted [25,26] that the true contact area is determined by the plastic deformation of a-spots projecting from the contact surface. Under this assumption, the area of mechanical contact can be determined from

$$
F = A_c \times H
$$

where $F$ is the applied load and $H$ represents the material hardness. From $R_{\text{cont}} = \rho/2\pi$, $x = (A_c/\pi)^{1/2}$ and (3), the contact resistance can be obtained from

$$
R_{\text{cont}} = \sqrt{\frac{\rho^2 \pi H}{4F}}
$$

As expected, the contact resistance is inversely proportional to the applied pressure. Although a low contact resistance is desired, the pressure has to remain below the critical point of plasticity region where the deformation of the material is irreversible.

#### 3.1.2. Contact capacitance $C_{\text{cont}}$

The actual contact area at the contact region is considerably lower than the apparent contact area. The multi-spot nature of contact surface creates regions of parallel capacitances. Two different types of capacitors are formed between the contacts interface. In regions filled with air voids, micro air voids capacitors $C_{\text{air}}$ are created. The actual contact points which are separated by thin insulating layers form a set of parallel micro contact capacitors $C_{\text{mc}}$. Thus, the total capacitance can be represented by Eq. (5).

$$
C_{\text{cont}} = \sum C_{\text{air}} + \sum C_{\text{mc}}
$$

As the actual contact area is much smaller than the apparent area, the capacitance due to the micro-contacts can be neglected. Hence, the effective contact capacitance can be approximated by the sum of the parallel elements of non-contacting regions filled with air voids [27]. For the proposed MEMS structure, the surface area of a contact pad is much larger than the separation between DUT pins and contact springs; therefore, the electric field fringing effect can be neglected. With this assumption, the contact capacitance due to the air voids can be approximated by the standard parallel plate capacitor model using an average gap between the two surfaces.

$$
C_{\text{cont}} = \frac{\varepsilon_0 \varepsilon_r A}{g}
$$

where $\varepsilon_0$ is the permittivity of free space $8.85 \times 10^{-12} \text{ F/m}$, $\varepsilon_r$ is the relative permittivity of the dielectric material and $g$ is the average gap at the interface. Due to the short path between the contacts where they are connected, the contact inductance has no significant effect and can be ignored from the contact model.
3.2. Contact spring’s conductive path circuit model

3.2.1. AC resistance

One of the critical factors that affect the signal integrity is the AC resistance. At high-frequencies, current flows near the surface reducing the effective cross section of the conductor. The current density reduces exponentially from the surface towards the center. This phenomenon, called skin effect, increases the resistance of wires at higher frequencies [28,29]. As the frequency of interaction increases, the effective area of the cross section used by the current becomes smaller increasing the effective resistance of the conductor. For the proposed model, with a conductive path of length \( l \) and a square cross section as shown in Fig. 2b, the AC resistance \( R_{AC} \) is equal to

\[
R_{AC} = \frac{\rho l}{4\delta(a - d)} = \frac{\rho l}{A}
\]

where \( a \) is the width, \( l \) is the length, \( \rho \) is the resistivity of conductor, \( A \) is the effective cross section area, and \( \delta \) is the skin depth. For the contact spring module with two identical parallel paths, the total AC resistance is equal to \( RAC/2 \).

3.2.2. Self and mutual inductance \( L_{self} \), \( L_{m} \)

Partial inductance represents the component of inductance that results only from the part of the current loop that is explicitly being modeled. To calculate the partial inductances of rectangular cross-sectional wires, closed form equations presented in [30,31] are used. The equations for self and mutual inductances of a rectangular wire with \( l \gg w + t \) are given in (8) and (9) respectively.

\[
L_{self} = \frac{\mu_0 l}{2\pi} \left[ \ln \left( \frac{2l}{w + t} \right) + 0.2335 \left( \frac{w + t}{l} \right) + 1 \right]
\]

(8)

\[
L_{m} = \frac{\mu_0 l}{2\pi} \left[ \ln \left( \frac{l}{d} + \sqrt{\frac{1}{d^2} + \frac{1}{l^2}} \right) - \sqrt{1 + \frac{d^2}{l^2} - \frac{d}{l}} \right]
\]

(9)

where \( d \) is the distance between wires, \( \mu_0 \) is the magnetic permeability of free space, \( w \) is the width, \( t \) is the thickness, and \( l \) is the length of wire. The self-inductance is not affected by the presence of a neighboring contact springs. The total partial inductance of each conductive path segment is

\[
L_{total} = L_{self} + \sum_{i} L_{mi}
\]

(10)

where \( L_{mi} \) is the mutual inductance of the \( i_{th} \) neighboring contact spring.

3.2.3. Mutual capacitance \( C_m \)

Mutual coupling exists between any two adjacent conductors. The coupling is a function of area, length and the separation between them, as well as dielectric barrier thickness and permittivity. The mutual capacitance \( C_m \) per unit length for two identical cylindrical conductors [32,33]

\[
C_m = \frac{\pi \varepsilon_{0} \varepsilon_r}{\cos^{-1}(2a/d) - a/d} = \frac{\pi \varepsilon_{0} \varepsilon_r}{\ln \frac{d + W}{W} + \sqrt{d^2 + W^2} - 1}
\]

(11)

where \( \varepsilon_0 \) is the permittivity of free space, \( \varepsilon_r \) is the relative permittivity of dielectric material, \( d \) is the distance between the two conductors, and \( a \) is the radius of the conductor. If the distance \( d \) between two rectangular conductors is equal or larger than the width \( W \), then mutual capacitance \( C_m \) can be approximated as two cylindrical wires with similar cross-sectional area with center to center distance of \( (d + W) \) is given by [34,35]

\[
C_m = \frac{12.1 \times 10^{-12} \varepsilon_r}{\log \left[ \frac{2a}{W} + \sqrt{\left(\frac{2a}{W}ight)^2 - 1} \right]}
\]

(12)

4. Simulation results

Mechanical reliability of MEMS microsystems is critical for proper operation of the device. Thus MEMS contact springs must remain in the elasticity region during the test stage. The elasticity depends on the yield point where the material starts to deform plastically in irreversible process. Prior to the yield point, the material will deform elastically and returns to its original form when the applied stress is removed. Once the yield point is passed, the deformation will be permanent and non-reversible until the stress reaches the material’s Ultimate Tensile Strength (UTS); which is the maximum stress that a material can withstand before breaking [36].

Knowledge of the yield point is vital when designing a microelectromechanical structure since it represents an upper limit to the load that can be applied. In designing the removable MEMS socket, Beryllium Copper (Be Cu) was selected to implement the contact springs due to its high yield strength, anti-oxidation, and non-magnetic qualities. It is the hardest and highest strength copper alloy with a yield strength of 585–675 MPa and an ultimate strength of 620–795 MPa [37,38].

The MEMS based contact spring was designed using IntelliSuite CAD tools. Electro-mechanical analysis was performed to determine the maximum pressure “contact force” that can be applied to remain within the elasticity region to avoid any deformation of the proposed structure during the tests stage. The analysis shows a linear relationship between the deflection and the applied pressure as shown in Fig. 3a.

The stress distribution of the proposed structure was analyzed using the Finite Element Analysis (FEA) tool from IntelliSuite software. The maximum stress under various deflection points was evaluated as shown in Fig. 3b. It holds a linear relationship, which is expected as it is below the yield point. The deflection points extracted from Fig. 3a are shown on the graph to emphasize the

Fig. 3. (a) Contact Spring Deflection versus pressure load, and (b) maximum Von-Mises stress versus applied pressure, the points shows the defection at these pressures.
stress values at these points. A MEMS contact spring has to deflect 1–2 \( \mu \text{m} \) in order to establish a good electrical contact with the DUT I/O pins. For a 2 \( \mu \text{m} \) deflection, the maximum stress is 200 MPa as shown in Figs. 4b and 3b. This will ensure the operation in the elasticity region during the test stage.

Since the MEMS probe has moving parts, fatigue can be a life limiting factor where structural damage occurs after a material is subjected to cyclic loading and unloading. It is defined as the number of stress cycles sustained before structure breakdown occurs. A material performance (fatigue) is commonly characterized by an S–N curve, also known as a Wohler curve where a regular sinusoidal stress is applied by a testing machine which also calculates the number of cycles to failure \[37\]. The S–N curve shows number of cycles versus maximum stress, which is an indication for materials fatigue life cycle. For copper beryllium (Be–Cu), at 200 MPa the cycles to failure is above 10 fatigue life cycle. For gold pads has been reported in \[40\].

The electrical performance of the proposed structure was analyzed if the design cannot tolerate any expansion or contraction of the structure. In the proposed design, the configuration of the spring arms can withstand the stress caused by temperature variations. It is known that thermal expansion can cause significant stress if the design cannot tolerate any expansion or contraction of the structure. In the proposed design, the configuration of the spring arms can withstand the stress caused by temperature variations. Fig. 4b shows the stress analysis due to temperature change from normal operating room temperature. The stress exhibits a linear relationship with temperature and the MEMS probe remains within the limits of reliability and fatigue life cycle with the temperature variation from –20 °C to 50 °C.

To validate the model, the extracted characteristic parameters were used to determine the bandwidth of the transmission lines. The electrical performance of the proposed structure was analyzed using Ansoft’s 3D full-wave electromagnetic field software HFSS and Q3DExtractor. This model, which is a lumped model, can be used to predict the behavior of the implemented MEMS structure up to 60 GHz due to its micro scale dimensions.

It can be observed that the derived model Eq. (7) can predict the AC resistance up to 100 GHz with a good degree of accuracy. The contact resistance of the proposed model determined from (4) for 2 \( \mu \text{m} \) deflection at 1 MPa is 11.3 mΩ. Both extracted and mathematical values are within acceptable range and are inversely proportional to the distance. The amount of mutual capacitance between a contact spring and the nearest neighbors, Fig. 5a, is the dominant term to characterize the overall mutual capacitance. This is due to the fact that the diagonal neighbors are partly shielded by the cross conductors and the non-adjacent lines are almost completely isolated by the adjacent conductors. As discussed in Section 3 Eq. (6), the contact capacitance for an average gap ranging from 0.1 \( \mu \text{m} \) to 0.5 \( \mu \text{m} \), varies from 340 pF/m to 70 pF/m.

Q3DExtractor calculates partial inductance which represents the component of inductance that results only from the part of the current loop that is explicitly being modeled. The mutual inductive effect becomes negligible as the separation distance increases. Both extracted and mathematical models presented in Eq. (9) are within a good degree of accuracy. The self-inductance determined from (8) is 440 pH. For two coupled parallel contact springs, the total partial inductance is given by \( L_{\text{total}} = L_{\text{self}} + L_{m} \) \[41\]. Fig. 5b, shows the partial inductance as a function of distance. As expected, the partial inductance decreases as the effect of mutual inductive effect decreases with distance.

Table 1 presents the extracted lumped parameters at 50 GHz. It can be seen that the RLC parameters of the proposed contact spring are very small due to the micro-scale dimension. This will allow a test channel using the proposed MEMS module to operate at much

![Stress-Free](Image)

Fig. 4. (a) Stress analysis of the MEMS spring-probe under applied pressure of 2 MPa, and (b) stress analysis versus change in temperature.

![Mutual Capacitance IF](Image)

![Inductance pH](Image)

Fig. 5. (a) Scatter chart of capacitance contribution by neighboring contacts indicating the contribution of adjacent, diagonal and non-adjacent contact in blue, red and green colors respectively, and (b) total inductance of a contact spring versus distance. (For interpretation of the references to colour in this figure legend, the reader is referred to the web version of this article.)
higher frequency compared to conventional test channels. To demonstrate the advantages of the proposed MEMS TIM over traditional structures, a typical strip-line and coax cable were designed using Agilent ADS software tool. A strip-line of 8-cm long with cross section of $1.12 \times 0.05\text{ mm}$ and a 10-cm coax cable with 1 mm inner core diameter and 3.35 mm outer diameter were implemented. The scattered parameters of the implemented transmission lines and the MEMS contact spring were determined through simulation. Both the strip-line and the coax cable at 10 GHz exhibit a moderate insertion loss as shown in Table 2 while the MEMS structure presents a negligible insertion loss of $0.002\text{ dB}$.

3-D full wave simulations using Ansoft’s HFSS simulation tool were also carried out to determine the insertion loss. The simulation results indicate that the MEMS contact spring can operate up to 200 GHz with less than $0.3\text{ dB}$ insertion loss.

The extracted parameters in Table 1 were used to build the circuit model for the MEMS structure to perform the transient and AC analysis using Agilent’s ADS tools. A sinusoidal waveform of one volt peak-to-peak was applied to the strip-line, coax cable, and the MEMS structure as an input. Fig. 6 shows the transient response at 50 GHz for the three extracted models. The applied input signal propagates through the contact spring without much attenuation while it is attenuated for the strip-line and the coax cable.

To compare the S-parameters extracted using 3-D full wave Ansoft’s HFSS, the S-parameter analysis has been carried out using Agilent (ADS) tools to determine the insertion loss and the return loss over a frequency range spanning from DC to 300 GHz. Fig. 7a and b shows the insertion and the return loss of the three transmission lines examined in the study. It can be seen that the MEMS structure can operate up to 50 GHz without a significant insertion loss while the insertion loss becomes much higher for the coax cable and the strip line. This is not an unexpected result since the physical dimensions of the MEMS Spring are much lower than its conventional counterparts. Such a small structure can only present minute RLGC values allowing operation at high frequencies without a considerable signal loss.

In general, test probes require proper maintenance to ensure maximum life expectancy and satisfactory performance. This requirement becomes more critical for MEMS probes due to their micrometer sizes. The proposed MEMS probe requires regular base cleaning to remove debris and contamination. Manual probe cleaning and abrasive cleaning techniques are not appropriate for the MEMS probe as they may affect its structural integrity. Non-destructive probe cleaning techniques such as ultrasonic immersion cleaning and laser ablation techniques are better suited for cleaning the MEMS probe. Electromechanical analysis using Intelisuite CAD tools shows that the pressure on the MEMS probe has to be less than 65 MPa for the probe to remain in the elasticity region. The graph of the magnitude of a cyclic stress against the logarithmic scale of cycles to failure provide necessary information to analyze the effect of fatigue on the MEMS probe. Simulation results indicate that the number of cycles to failure for the proposed probe is about 500,000.

5. Experimental results

To validate the simulation results, a prototype of the MEMS contact spring shown in Fig. 2 has been fabricated using a p-type silicon-on-insulator (SOI) wafer is used as the starting substrate:

- Silicon thickness: $10 \pm 1\text{ µm}$.
- Oxide thickness: $2\text{ µm}$.
- Handle wafer (substrate) thickness: $500\text{ µm}$.

![Image](image_url)
To simplify the testability of the structure, slight modifications were required to ensure that these modifications do not affect the device performance as shown in Fig. 7c. Ground pads were added to allow accurate measurement of $S$-parameter at high frequencies.

The following points were taken into consideration:

- Two ground and one signal contact pads have been incorporated at 75 $\mu$m pitch to test the device with Microtech GS high frequency probes.
- Spring type contact pads with back penetration for full deflection. These contact pads are in fact substitute for the top and bottom contacts of the removable socket.
- 500 $\mu$m separation between the contact springs resembles the two beams incorporated in the removable socket.
- Thin oxide layer to isolate the metal from the wafer.

5.1. Fabrication process

1. **Back penetration**: the Substrate is patterned and etched from the “bottom” side to the Oxide layer. This creates through-hole structures around the area of the contact springs to open free space for deflection.
2. **Physical structure**: the MEMS contact spring prototype is patterned and etched from the top side down to the Oxide layer.
3. **Oxide removal**: this will open the through holes around the physical structure for deflection.
4. **Metallization**: to establish the electrical conductivity between the two contact spring and to provide common ground for the RF probes.

As Beryllium Copper (Be Cu) was not available, the following three layers of Chromium (Cr) for adhesion (100 nm), Copper (Cu) as the conducting metal (1.3 $\mu$m) and Gold (Au) thin layer for anti-oxidation (50 nm) have been deposited to establish the electrical conductivity.

Scanning Electron Microscope (SEM) images for the fabricated prototypes are presented below. Fig. 7d shows the cross section of the through holes from the backside penetration. The side walls roughness (scalloping) is minimized.

5.2. Measurement results

The insertion ($S_{21}$) and return loss ($S_{11}$) were measured using Anritsu ME7808B Vector Network Analyser (VNA) and a Micrtech GS-probe with 40 GHz bandwidth. The scattering parameters measurements were conducted over a frequency range of 1 MHz to 40 GHz.

Fig. 8a and b shows the measured insertion loss ($S_{21}$) and return loss ($S_{11}$) respectively. The results indicate an insertion loss of 0.5 dB at 40 GHz and a return loss of –19 dB at 40 GHz. The measured $S$-Parameters results are in a good agreement with the values predicted by the simulations. The slight difference between the simulation and measurement results is primarily due to the un-modeled contact interface between the MEMS structure and the GS-probes tips.

To evaluate the maximum DC power that can be delivered by each contact probe, static DC tests have also been performed using Agilent 34401A Digital Multi-meter. The current was increased to determine the amount current that the metal layer can carry.
Fig. 8. (a) Measured return loss, (b) measured insertion loss, (c) a contact spring before and after the DC current exceed the maximum limit, and (d) SEM top view of the fabricated device.

Table 3

<table>
<thead>
<tr>
<th>Minimum pad size</th>
<th>Insertion loss @ 40 GHz</th>
<th>Return loss @ 40 GHz</th>
<th>Typical lifetime</th>
<th>Contact resistance</th>
</tr>
</thead>
<tbody>
<tr>
<td>Cascade Microtech infinity probes</td>
<td>25 × 35</td>
<td>0.9 dB</td>
<td>17 dB @ 40 GHz</td>
<td>&gt;500,000</td>
</tr>
<tr>
<td>Cascade Microtech AirCoplanar probes</td>
<td>50 × 50</td>
<td>1.25 dB</td>
<td>18 dB @ 40 GHz</td>
<td>&gt;500,000</td>
</tr>
<tr>
<td>Cascade Microtech [Z] probes GSG</td>
<td>80 × 80</td>
<td>0.8 dB</td>
<td>18 dB @ 40 GHz</td>
<td>&gt;1,000,000</td>
</tr>
<tr>
<td>Cascade Microtech unity probes</td>
<td>70 × 70</td>
<td>0.8 dB</td>
<td>15 dB @ 20 GHz</td>
<td>&gt;1,000,000</td>
</tr>
<tr>
<td>MEMS probe</td>
<td>Can be as small as 10 × 10 μm</td>
<td>&lt;0.5 dB</td>
<td>−19 dB @ 40 GHz</td>
<td>&gt;500,000 Based Copper Beryllium</td>
</tr>
</tbody>
</table>

Table 4

<table>
<thead>
<tr>
<th>Pitch μm</th>
<th>Insertion loss S21 at −1 dB</th>
<th>Return loss S11 at −10 dB</th>
<th>Capacitance</th>
<th>Inductance</th>
<th>Contact resistance</th>
</tr>
</thead>
<tbody>
<tr>
<td>Ardent concepts RC10–12 test socket</td>
<td>1000</td>
<td>18.1 GHz</td>
<td>15.7 GHz</td>
<td>0.463 pF</td>
<td>1.25 nH</td>
</tr>
<tr>
<td>Johnstech PAD ROL™ 100A</td>
<td>&gt;400</td>
<td>40.0 GHz</td>
<td>14.5 GHz @ 20 dB</td>
<td>Mutual 0.050 pF</td>
<td>Mutual 0.15 nH</td>
</tr>
<tr>
<td>Cascade Microtech G40</td>
<td>400–800</td>
<td>22.0 GHz</td>
<td>12.4 GHz</td>
<td>0.129 pF</td>
<td>0.81 nH</td>
</tr>
<tr>
<td>Cascade Microtech G80</td>
<td>1000</td>
<td>20.0 GHz @ 0.6 dB</td>
<td>15.0 GHz</td>
<td>0.22 pF</td>
<td>1.0 nH</td>
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<tr>
<td>Verticon™ 100 BGA</td>
<td>500–1200</td>
<td>33.7 GHz</td>
<td>25.1 GHz</td>
<td>Mutual 0.012 pF</td>
<td>Mutual 0.068 nH</td>
</tr>
<tr>
<td>MEMS contact spring</td>
<td>100 μm</td>
<td>40 GHz @ 0.5 dB</td>
<td>40 GHz @ 19 dB</td>
<td>Mutual 0.0043 pF At 100 μm pitch</td>
<td>Mutual 0.140 nH At 100 μm pitch</td>
</tr>
</tbody>
</table>
without melting. The probes can handle maximum DC current of 500 mA. Fig. 8c shows a contact spring before and after the DC current exceed the maximum limit. The proposed MEMS contact spring can be used as a building block to design a high-density socket. Table 3 shows the performance comparison with the existing high-density test sockets [42–44].

The proposed MEMS module supports a smaller pitch which is highly desirable to support the fine-pitch requirements of next generation of integrated circuits predicted in the recent International Technology Roadmap [16]. Table 4 shows the performance comparison with the existing high-density RF probes [45]. The MEMS probe can operate at higher frequencies without a significant loss.

6. Conclusion

A MEMS test interface module has been designed to establish connections between an automatic tester and the device under test. The proposed MEMS interface can reduce the physical separation between the device under test and the tester resources by orders of magnitude due to its micro-scale dimensions. The proposed MEMS module provides a solution to the problem of undersized transmission line effects that limits the bandwidth of ATE test channels. Experimental results indicate that the designed MEMS test interface can operate up to 50 GHz while maintaining a high level of signal integrity. The MEMS interface module can also provide a place for interface circuits to facilitate accurate measurements during critical tests. Furthermore, MEMS contact springs can be used to establish necessary test channels between a die under test and the ATE measurement instruments.

Acknowledgments

The authors would like to thank CMC Microsystems and the National Science and Engineering Research Council of Canada (NSERC) for their support. The authors also thank the LMF staff at Ecole De Polytechnique for their support during the fabrication process.

References