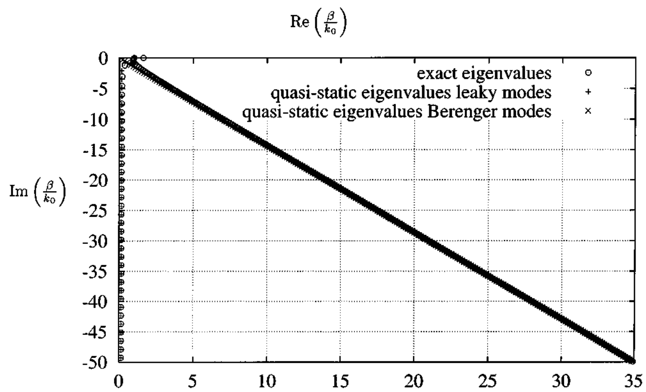

Fig. 4. Location of the eigenvalues β/k_0 of the TM modes for a microstrip–PML configuration.

V. Conclusions

Analytic expressions in the quasi-static limit are derived for two sets of zeros, which exist in a microstrip substrate terminated by a PML. For a strongly absorbing PML, it is shown that the first set corresponds to the leaky modes of the microstrip substrate, whereas the second set, termed Berenger modes, is not influenced by the microstrip substrate. As demonstrated by examples, the analytical formulas allow a quick calculation of the modal constants whenever $\omega\sqrt{\epsilon_0\mu_0} \ll \beta$. Moreover, the quasi-static expressions can be used as initial guesses in a numerical root-finding routine that can provide more accurate values for the zeros of the dispersion relation. Thereby, our results can speed up the calculation of the Green's function of the microstrip substrate terminated by a PML, as discussed in [5].

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Wide-Bandwidth Millimeter-Wave Bond-Wire Interconnects

Thomas P. Budka

***Abstract*—A new type of interconnect has been developed that significantly extends the bandwidth of fixed-length bond-wire interconnects between microwave circuits. This interconnect maximizes bond-wire length, as well as landing pad size while simultaneously extending the cutoff frequency of the interconnect. The bond-wire interconnect is treated as a five-stage low-pass filter where basic filter theory is used to develop an interconnect prototype. Microstrip interconnects are designed using electromagnetic simulators, which match a specific low-pass filter response on a 5-mil-thick (127 μm) glass substrates. The measurements indicate a return loss greater than 12 dB and an insertion loss from 0.0 to 0.3 dB from dc to 80 GHz using two 17-mil-long (432 μm) 1-mil-diameter (25 μm) ball bonds with a tolerance of $\pm$ 2 mil (50 μm). For comparison, an uncompensated interconnect with two 17-mil-long (432 μm) bond wires has 1-dB insertion loss and 10-dB return loss at 40 GHz and continues to degrade at higher frequencies.**

I. Introduction

Historically, chip-to-chip interconnects using bond wires have been analyzed as having a single electrical component: a series inductor. In order to improve the high-frequency performance of a bond-wire interconnect, efforts have usually focused on reducing the length of the bond wire and also reducing the chip-to-chip spacing. However, limitations in manufacturing require longer bond-wire lengths and wider chip-to-chip spacing to improve the yields of millimeter-wave multichip assemblies (MCAs). Therefore, the goal for millimeter-wave interconnect design is to maximize bond-wire length to improve manufacturability and maximize bond pad size so that mechanical tolerances are eased. This paper demonstrates that this is possible with a filter theory approach to interconnect design.

A single bond-wire interconnect can be treated as a single-stage low-pass filter with a fixed cutoff frequency. As more stages of the filter are added, the bandwidth of the filter increases until adding additional stages becomes inappropriate due to unacceptable filter losses. Usually low-pass filters are between three and seven stages. For this specific design example, the bond-wire interconnect is treated as a five-stage low-pass filter.

Fig. 1 displays a five-stage low-pass filter prototype with series inductors and shunt capacitors. There are well-known published tables in the literature of the relative values of the inductances and capacitances for filter design [1]. For example, if a 0.5-dB equal-ripple response is desired, then a single inductor filter would have an inductance of $0.70L$. A five-stage equal-ripple filter with the same cutoff frequency would have a center inductor of $2.54L$ and two outer inductors of $1.71L$ [2], where L is the inductance that selects the cutoff frequency of the filter. This implies that for the same cutoff frequency of the single- and five-stage filter, the center inductor in the five-stage design can have a 3.6 times higher inductance than a single inductor design. This directly translates into a 3.6 times longer bond wire for the same cutoff frequency. The ability to lengthen the bond wires is critical for high-yield assembly of millimeter-wave multichip modules.

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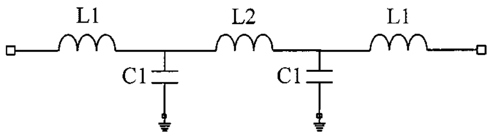

Fig. 1. Five-stage low-pass filter interconnect prototype. $L2$ is realized with the bond-wire inductance, $L1$ is realized with a short high-impedance section of transmission line, and $C2$ is realized by a combination of short low-impedance bond pad and a bond pad gap capacitance to ground pads on either side.

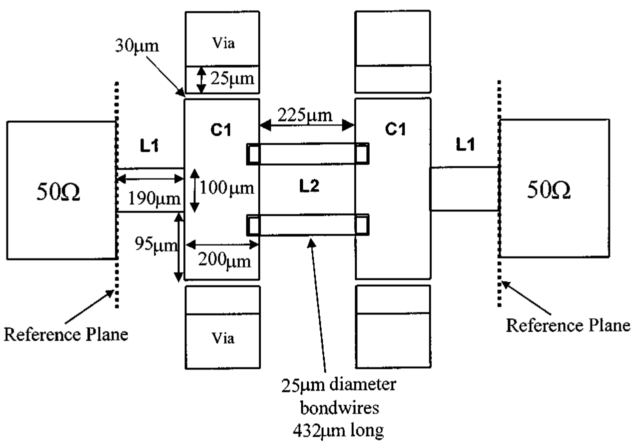

Fig. 2. Layout of a glass-to-glass interconnect that uses two 17-mil-long (432 μm) 1-mil-diameter (25 μm) bond wires.

II. Design of a Glass-to-Glass DC–80-GHz Interconnect

For automotive radar applications, there is a need for a class of bond-wire interconnects that connect 127-μm-thick glass chips to each other with a bandwidth of 85 GHz. The filter from Fig. 1 is realized with two ball bond wires for the inductance $L1$, a combination of a short low-impedance section of transmission line and a shunt gap capacitance for the capacitance $C2$, and a short high-impedance section of transmission line for inductance $L2$. During the design stage of this interconnect, the bond-wire length was first fixed to a reasonable length for manufacturing (17 mil/432 μm), but a very challenging length for 76-GHz signals.

This design was accomplished by first creating a model of the structure in Libra and optimizing the interconnect for bandwidth. Next, the structure was simulated using HP Momentum where thin-strip equivalent air bridges are used for the bond wires. An equivalent thin-strip air-bridge width is first found that has the same impedance as a round bond wire by using a two-dimensional (2-D) quasi-static simulator (PDEase2D by Macsyma Inc., Arlington, MA). For the case of a 1-mil-diameter (25.4 μm) wire bond that is 10 mil (254 mm) above a ground plane, a 2.3-mil-wide (58 μm) thin strip will have the same impedance of 213 Ω. The electromagnetic (EM) data is then fitted with a Libra physical model where the dimensions of the model match the layout dimensions. Shunt capacitances and bond-wire inductances are allowed to vary to fit the EM simulated results. Next, the Libra physical model is tweaked to find which changes in the layout will drive the performance in the right direction. The layout is changed and the process is iterated until acceptable performance is achieved.

After the best design is achieved, the final design response is verified by a second simulator (Ansoft's Maxwell Eminence). Two key design issues are addressed by this iterative procedure. First, by using HP Momentum instead of Ansoft's Maxwell Eminence, the design cycle time is reduced to less than one day for each iteration compared to three days for Ansoft's Maxwell Eminence. Also, by using Libra physical models, changes in the layout can be predicted accurately before selecting to run a CPU intensive EM simulation.

Fig. 2 displays the top-view layout of the glass-to-glass interconnect using two 1-mil-diameter (25.4 μm) bond wires designed for a fixed length of 17 mil ($\sim$432 μm). The final fitted Libra schematic based on measured data for this interconnect is presented in Fig. 3. The measured results represent a first-pass design success for production capable bond-wire lengths of 17 mil (432 μm)$\pm$2 mil (50 μm). The combination of EM optimization using HP's Momentum two-and-one-half dimensional (2.5-D) EM simulator, as well as a second verification by Ansoft's Maxwell three-dimensional (3-D) EM simulator helped produce these first results. To fit the model to the measured data, only the bond-wire inductance and the gap capacitance were varied for the 17-mil-long bond-wire case. Next, the inductance step was determined that fit the 15- (381 μm) and 19-mil-long (483 μm) case. The inductance of these two bond wires is simulated to be between 0.115–0.155 nH if the bond-wire lengths vary between 15- (381 μm) and 19-mil long (483 μm). If these bond wires were uncompensated, then they would only pass frequencies up to 40 GHz with 1-dB insertion loss with a return loss of 7 dB into 50-Ω ports. At 76 GHz, the uncompensated interconnects have an insertion loss of 3.4 dB and a return loss of 2.7 dB. Figs. 4 and 5 display the measured and modeled insertion and return loss of the compensated bond wires and uncompensated designs.

The compensated interconnects are clearly superior to the uncompensated ones, which uniformly degrade with higher frequencies. The results display that the design goal of $>$10 dB return loss and $<$0.3 dB insertion loss is easily met for automotive applications at 76 GHz for 1-mil diameter (25 μm) bond-wire lengths from 15 (381 μm) to 19 mil (483 μm).

III. Conclusions

This paper has presented a new type of interconnect topology that uses filter design techniques to create a dc–80-GHz five-stage low-pass filter bond-wire interconnect on 5-mil-thick (127 μm) glass. The design techniques may be applied at any frequency and on any other substrate (Si, GaAs, Duroid, etc, ...) and enable the use of significantly longer bond wires, as long as each chip has filter-like compensation on the RF bonding pads. The interconnects tested represented the closest possible structure that would be manufactured with automated ball-bonding equipment. Additional testing should be performed to gauge the effects of bond-wire shapes or spacings. These variations may change the equivalent series inductance of the bond wires and change the bond-wire length tolerances predicted by these measurements. This design technique may also be applied to plastic packages with long wire leads or to ball grid array packages, as long as a small shunt capacitive element can be realized at either end of the lead.

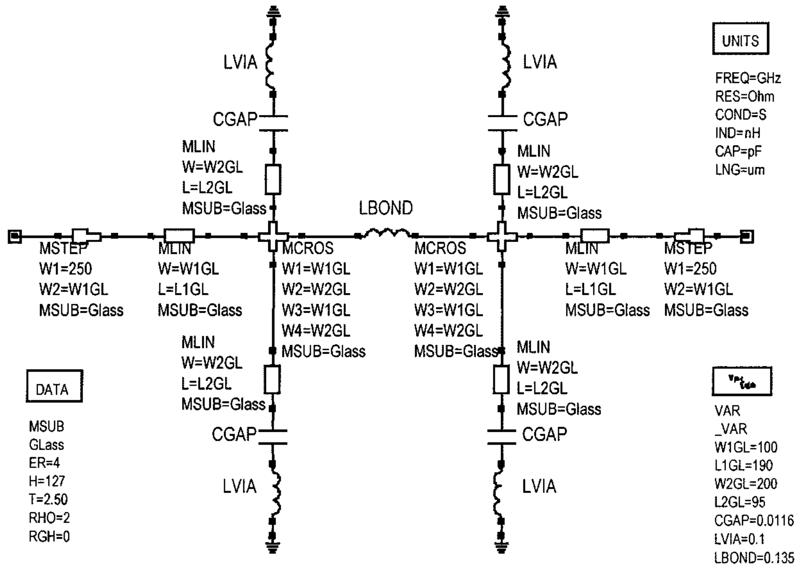

Fig. 3. Equivalent schematic of the glass-chip-to-glass-chip interconnect using 1-mil-diameter (25.4 μm) 17-mil-long (432 μm) ball bonds.

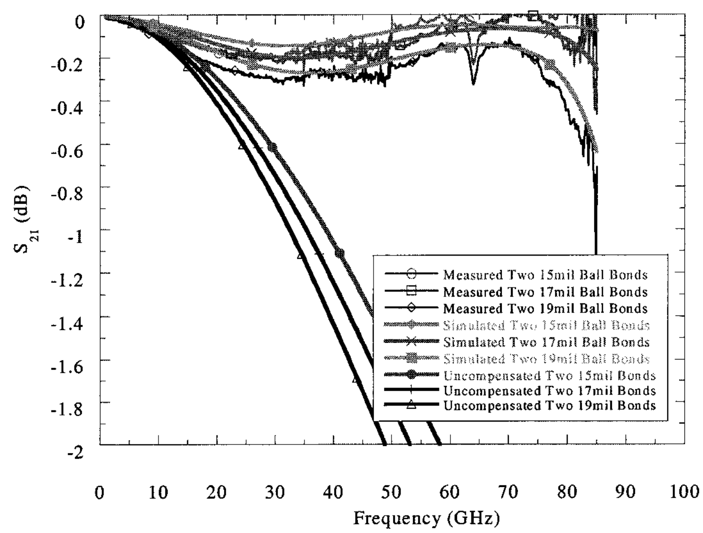

Fig. 4. Measured and modeled insertion —loss response of a compensated glass-to-glass interconnect that uses two 1-mil-diameter (25.4 μm) ball bonds of various lengths (15–19 mil/38–483-μm long). The simulated uncompensated interconnect results are presented as well.

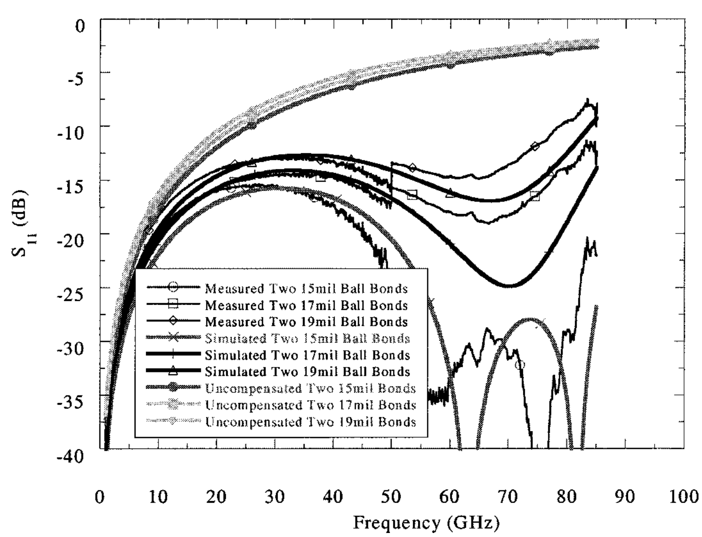

Fig. 5. Measured and modeled return-loss response of a compensated glass-to-glass interconnect that uses two 1-mil-diameter ball bonds of various lengths (15–19-mil long). The simulated uncompensated interconnect results are presented as well.

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Branch-Line Couplers Using Unequal Line Lengths

Canan Toker, Mustafa Saglam, Mustafa Ozme, and Nilgun Gunalp

***Abstract*—General solutions for branch-line couplers using different line lengths are provided in this paper. Explicit expressions are derived with which a hybrid with any given power division ratio can be designed. In the design, one of the characteristic impedances or lengths of the branches can be chosen arbitrarily to suit a given design specification. This approach brings flexibility in choosing the characteristic impedances or the lengths of the branches, and is helpful especially in monolithic-microwave integrated-circuit applications where restrictions imposed on microstrip transmission lines do not allow the use of conventional branch-line couplers employing quarter-wave-long lines. However, the resultant bandwidth is narrower compared to that of the conventional coupler.**

***Index Terms*—Branch-line couplers, hybrids.**

I. INTRODUCTION

Ever since the introduction of the branch-line couplers, it has been a common practice to choose all the branches a quarter-wave-long at the center of the frequency band of interest. With fixed lengths of the lines, attention is then concentrated on the characteristic impedances of the various branches to satisfy the given specifications. This procedure is adopted for all the work on branch line or branch-guide couplers, including multibranch couplers, which are developed to widen the inherently narrow bandwidths [1], [2]. The problem of realizability concerning the characteristic impedances of the multi-branch microstrip couplers is improved by optimization algorithms; however, in such studies, the lengths are again kept fixed and only the characteristic impedances are optimized [3]. Broad-band uniplanar branch-line couplers, which are more suitable for microwave-integrated-circuit (MIC) and monolithic-microwave integrated-circuit (MMIC) applications also use quarter-wave-long lines [4].

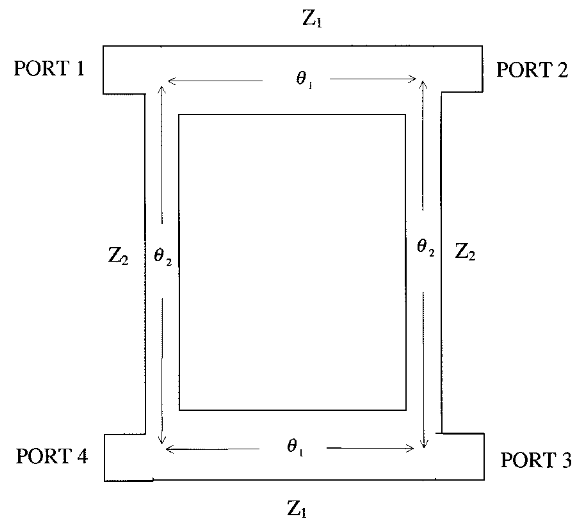

Fig. 1. Unequal line-length branch-line coupler.

Branch-line couplers with quarter-wave-long lines being satisfactory for most applications, similar constructions, but having different line lengths, are the interest of the work presented in this paper, which has not been considered previously. It is not only interesting from the academic point-of-view, but it is also useful, for some applications, to use this alternative design because the hybrids using quarter-wave-long lines do not always lead to realizable characteristic impedances. This fact is obvious in some MIC and MMIC applications where the values of the characteristic impedances of the microstrip transmission lines are outside the realizable range. Varying the lengths of the branches introduces additional freedom in choosing the characteristic impedances to overcome this realizability problem, as will be discussed below.

In the alternative design presented in this paper (see Fig. 1), the characteristic impedances and lengths of the through and branch lines are different. This type of construction enables one to deal with four variables, namely, two characteristic impedances and two lengths compared to the conventional coupler where only two characteristic impedances are used as variables. The analysis of the new branch-line coupler leads to three design equations with which a hybrid with any power division ratio can be designed readily. In this design, one of the parameters can be chosen arbitrarily, which may be very useful under certain circumstances, and the phase angle difference between the transmitted signals to the output branches is 90°. However, bandwidth reduction is unavoidable.

II. ANALYSIS

The branch-line coupler under investigation is shown in Fig. 1. This coupler has different line lengths and characteristic impedances for the adjacent arms. The lines are assumed to be lossless and reciprocal, and the reference impedance at the ports is taken as Z_0. Z_1 is the characteristic impedance of the through lines and Z_2 is the characteristic impedance of the branch lines, θ_1 and θ_2 being their respective electrical lengths. Port 1 is the input port and port 4 is the isolated port. Ports 2 and 3 are the output ports. Applying even- and odd-mode analysis [5], the chain parameters are obtained as follows:

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