ADVANCED MILLIMETER WAVE TECHNOLOGY FOR MOBILE COMMUNICATION

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ABSTRACT

E-plane and conventional waveguide technologies are applied to the design of millimeter wave subsystems, including oscillators, mixers, filters, modulators and hybrids. Different design approches are described. Transitions between E-plane guiding structures are considered. Advantages and drawbacks regarding practical design and performances are emphasised. Experimental results at Ka-band (26.5-40 GHz) and V-band (50-75 Ghz) are presented.

INTRODUCTION

The need of low-cost millimeter wave communication systems is increasing as new frequency allocations are foreseen. As an example, new bands have been regulated at 38, 55 and 58 GHz in the UK [1]. Many applications, including mobile cellular radio, automotive radar and vehicle communication have since emerged as commercial systems requiring a big technology development effort [2,3]. Different approches are possible in designing the required subsystems. The hybrid (MIC) technology, including E-plane devices, is commonly used today, although MMIC is rapidly expanding.

This paper describes our work carried out to develop E-plane (metal insert, finline, coplanar waveguide -CPW-) and microstrip technologies, for the design of mm-wave subsystems (filters, mixers, ASK and PSK modulators, hybrids) at 30 and 60 GHz. A 0.254 and 0.127 mm thick CuClad-217 substrate has been used. Waveguide technology is also considered, but only for two specific applications at 60 GHz (Gunn diode oscillator and PSK reflection modulator). The 60 GHz subsystems described have their primarily application to indoor-communication systems.

TECHNOLOGY AND DESIGN MODELS

E-plane transmission structures are commonly used in the mm-wave range. Some advantages are low losses, rectangular-waveguide compatibility, subsystem integration capability and repeatability. Numerical methods are required to compute their propagation characteristics as well as the effects associated to discontinuities and transitions. Several methods proposed in the literature were developed to investigate the transmission properties of E-plane infinite structures: approximate Cohn's method [4], Spectral Domain (SD) [5], Generalized Transverse Resonance (GTR) [6] and the Method of Lines (MOL) [7]. As an example, the characteristic impedance and λ_g/λ_o (finline wavelength normalized to free space wavelength) of unilateral, centered finline in a WR-15 waveguide at 60 GHz, as a function of w/b (w: slot width, b: waveguide height), are compared in the following table.

	λ_g/λ_o		characteristic impedance (Ω)	
	w/b = 0.05	w/b = 0.5	w/b = 0.05	w/b = 0.5
Cohn	0.87839	1.04690	145.5	380.5
SD	0.88181	1.06088	151.1	404.0
GTR	0.88169	1.05868		
MOL	0.98222	1.11469		

As expected, approximate Cohn's method gives accurate results when w/b is small. SD and GTR give comparable results. MOL is very well suited to complex guiding structures like double-face E-plane.

Transitions from waveguide to finline, microstrip and CPW have also been the subject of intensive experimental research in our Department. Triangular and exponential finline tapers were tested at Ka-band using the finline housing proposed by Adelseck [8]. Two lengths (λ and 1.5 λ) were considered in both cases. Insertion loss (IL) for a double transition ranged from 0.5 to 2 dB and return loss (RL) from 16 to 30 dB. While the triangular taper resulted in advantage over the exponential one, particularly at the lower frequency range (26.5-32 GHz), the longer taper did not show a significant improvement. Some absortion peaks were also detected but could be removed later using the housing proposed by Meier [9]. Similar results were obtained at V-band (50 -75 GHz). Triangular finline tapers demonstrated

16 dB RL and 0.6 IL (double transition) at 60 GHz, corresponding to 0.3 dB per transition. These performances are well in advantage over transitions using $\lambda/4$ finline step-transformers (7 dB RL and 0.8 dB IL per transition) but they are similar to those measured for circular-arc finline transitions.

Waveguide-to-microstrip tapered transitions through antipodal finline (figure 1) were designed and tested at Ka-band. The taper contour is sinusoidal. A circular slot is included to avoid unwanted resonances, as recommended in the literature [10]. Optimum performance was empirically obtained for a circular slot width equal to that of the microstrip line. Experimental results showed 15 dB RL, 0.5 dB IL per transition, over a bandwidth of 30% centered at 30 GHz. Microstrip-to-CPW (quasi-TEM mode) tapered transitions were also designed at Ka-band (figure 2). The transition is double faced and it was implemented using Hermite interpolating techniques to assure a very smooth characteristic impedance variation along the taper. To avoid the propagation of the undesired CPW non-TEM mode, a reduction in the waveguide height "a" is required along the CPW length. For a centered CPW, with dimensions (in mm) a=1.6, w=0.25 , 2s=1.53 and ϵ_r=2.17, its cut-off frequency is 39.98 GHz and the characteristic impedance of the quasi-TEM mode is 55.3Ω. Both transitions were combined to construct a waveguide-microstrip-CPW transition, and empirically optimized. Best results, obtained when printing the metalic half-circle in both faces, showed a 1 dB IL and better than 15 dB RL from 28 GHz to 37 GHz for a double waveguide-microstrip-CPW transition. Each waveguide-to-microstrip transition is 21 mm long; each microstrip-to-CPW transition is 5.4 mm long; CPW length is 3.3 mm. The bandwidth is mainly restricted by the waveguide-to-microstrip transition.

Finline-to-CPW transitions were also designed (figure 3). For signals propagating from each side (finline and CPW -quasi-TEM mode-) the transition acts as a 180º hybrid [11]. When using a pair of antiparallel Schottky diodes at the junction, the result is a balanced mixer [11]. The junction can also be used as a "switchable balun" when putting a pair of antiparallel PIN diodes, with application to PSK transmission modulators [12]. Two practical profiles were considered at Ka-band: linear (progressive) and circular. To evaluate the transition performance, the non-TEM mode IL along a double transition was measured in two cases: (a): propagation, (b): cut-off (waveguide height reduced). The linear transition showed better performances in (a): 1-3 dB in the 26.5-40 GHz range (1-5 dB, circular transition), while circular transition was better in (b): > 20 dB (26.5-33 GHz) (>20 dB, 26.5-31 Ghz - linear transition).

Alternative waveguide-to-microstrip transitions using ridged waveguide were investigated. Two designs were considered: ladder step-transformer type and linear-profile taper (figure 4). The former has better a-priori electrical performances (RL and IL), being however more influenced by mechanical tolerances. A Montecarlo analysis was performed considering a 50 μm machining tolerance at 60 Ghz. Ridge width is 1 mm. Results show that similar electrical performances can be expected in both cases (statistically), with higher taper IL due to the fact that it is usually longer. Both transitions were implemented at V-band into a WR-15 waveguide considering three possible microstrip widths: 0.5, 0.8 and 1 mm. Best experimental results were acheived for 0.5 mm, corresponding to the impedance-matching case. At 60 GHz they are: 14-18 dB RL, 1.1-1.5 dB IL for a double 14.39 mm-long linear taper (the step-transformer was 8 mm long) and a 24 mm long microstrip line, corresponding to 0.1 dB per transition. The ladder step-transformer exhibit better performances but it is more difficult to implement.

MM-WAVE SUBSYSTEMS DEVELOPED

60 GHz hybrids

The design of microstrip branch-line and rat-race hybrids at V-band is highly affected by T-junction discontinuities. Specific design techniques of microstrip V-band hybrids with any degree of coupling, based on an analytical compensation of T-junction effects, were implemented. They overcome limitations detected in similar methods [13] also simplifying the design procedure. Two 3-dB hybrids were designed at 60 Ghz, a 180º rat-race and a 90º branch-line (three branches). The masks are in figure 5. The access ports are WR-15 rectangular waveguide. The waveguide-to-microstrip transition is the linear-taper ridged waveguide described before. To attain good aspect ratios and to minimize T-junction effects, a high microstrip-line characteristic-impedance was selected. To match the ridged waveguide impedance to the microstrip-line impedance, a linear microstrip taper was used. Measurements showed a 4-4.5 dB insertion loss, in a very good agreement with simulations, as well as a previous measurement of transitions losses and microstrip-line losses. Transmission unbalance was 0.3-0.5 dB, at center frequency. Return loss was 15-20 dB, a figure probably limited by transitions. Isolation was 17 dB.

60 GHz BPSK modulators

Two kinds of BPSK modulators can be found in the literature: reflection-type and transmission-type. Transmission balanced modulators (TBM) are based in two switching diodes and equal path transmission lines in a balanced configuration. Switching produces two possible phase states at the output signal, with a phase shift of 180º. The balanced structure implies inherently broad PSK bandwidth and good input/output isolation. In [14], a TBM using a planar microstrip/slotline configuration at 60 Ghz is described. A 30 GHz TBM was designed (figure 6) which combines finline, CPW and microstrip lines mounted in an Adelseck housing [8]. Two antiparallel PIN diodes at the finline-to-CPW transition and a $\lambda/4$ CPW act as a switchable balun, as described before. Experimental results show 2 dB IL and 10 dB RL over a 5.7 GHz bandwidth (28.3-34 GHz) . Phase shift between states is 180º with 3º deviation over the entire Ka-band. Reflection modulators (RM) offer a simpler design and lower insertion loss but they require an external circulator. One easy way to implement a RM is to switch a signal path-length using a single PIN diode [15]. A RM was implemented at 60 GHz in finline (figure 7) obtaining 1.5 dB IL (ON state) and 6 dB IL (OFF state) with 10º phase shift deviation from 180º (nominal) over a narrow bandwidth (2%) [23]. The bandwidth limitation comes from a mathematical condition [16] that the impedance-transforming network (a shorted finline length in our case) has to met. The same approach can be implemented in an alternative design which combines a waveguide housing and a planar circuit containing the PIN diode (figure 8) [17]. The substrate is placed in the transverse plane of the waveguide. It permits the use of a beam lead diode with only slightly worse insertion losses as compared to conventional waveguide post-mounting. A RM was constructed in a V-band (WR-15) housing at 58 GHz, resulting in 1.5 IL and 8º phase shift error.

Planar Balanced Mixers

As explained above, a finline-to-CPW is a 180° hybrid which can be used for designing balanced mixers. Some prototypes of such mixers have been fabricated at 30 and 60 GHz with good performance. Figure 6 shows an example of a 30-GHz balanced mixer. The general structure follows that presented in [11], with the peculiarity that the LO port is also in waveguide; to couple the LO power with the required phase relationship to the pair of diodes mounted in the E-plane CPW section, there has been chosen a waveguide-microstrip-CPW (quasi-TEM mode) transition without via holes between the opposite substrate faces. The waveguide-to-microstrip part of the transition is that commented previously and referenced in [10], which presents relatively wideband behaviour. Although not mandatory in this design, wideband characteristics of the complete transition have been sought through a gradual microstrip-CPW (quasi-TEM mode) transition (figure 2). Such a mixer has given conversion loss less than 7.5 dB and RF-port RL greater than 6 dB (typically 8 dB) over a 28-33 GHz frequency range for a 10 dBm local oscillator pumping at 27.03 GHz.

A crossbar balanced mixer at 60 GHz has also been built and tested [21]. The general structure is shown in figure 12. The RF is coupled from a WR-15 waveguide to a microstrip 50Ω line through a stepped ridge transition. The RF is then coupled to the quasi-TEM mode of a CPW using the transition shown in figure 12. This transition can be considered a simplified version of one reported in [22]; it achieves the continuity between the microstrip ground plane, which ends abruptly at the beginning of the CPW section, and the ground planes of the CPW quasi-TEM mode trough quarter wave sectorial resonators placed on top of the microstrip ground plane with the narrow end connected to the CPW ground planes. The CPW goes into a WR-15 waveguide through a 0.5 mm wide, non current-cutting slot in the waveguide narrow wall. The Schottky beam-lead diodes are placed across the CPW slots in the middle of the WR-15 waveguide. The local oscillator is fed through the waveguide containing the CPW with the diodes, the opposite end being terminated with a sliding short circuit that gives a degree of freedom in the OL matching. The IF is extracted through the opposite WR-15 waveguide narrow wall after another CPW-to-microstrip transition. Due to the big size of the beam-lead diodes as compared to the 0.1 mm wide slots of the CPW, performances have shown to be very dependent on the exact diode position after mounting. Conversion loss achieved are typically under 12 dB between 59 and 61 GHz for a LO pumping of 15 dBm at 58.7 GHz, although much better conversion loss (around 7 dB) has been occasionally observed.

Metal insert E-plane filter

Figure 9 shows a sketch of an E-plane metal-insert (septa) band-pass filter. It consists of a thin metal plate with alternating square holes, inserted in a standard waveguide to form coupled resonant cavities [18]. A tolerance analysis was carried out and showed than a precision of 10 microns in the dimensions of the septa was necessary to achieve good results on a very narrow bandwidth (1% or less). If a precision of 100 microns is considered, then the bandwidth must be wider than about 2%. Two different plate widths (100 u and 35 u) have been tested for the design of 3 sections Chebyscheff filters centered at 28 GHz and 60 GHz. A standard photolitografic procedure was used for the fabrication of the septa, and good experimental results were obtained for filters, having 2.2 GHz bandwidth at 28 GHz (7.8%) and 800 MHz at 60 GHz (1.33%). IL were respectively 0.5 dB and 1.8 dB, being both well centered in the designed center frequency.

Gunn Oscillators

Both CW and Varactor-controlled Gunn oscillators have been designed. The diodes are mounted on posts inside a cavity formed by a short circuited waveguide of reduced height. The short circuit is movable in order to have a mechanical frequency adjustment and, in some designs, a capacitive diaphragm has been inserted between the cavity and the output waveguide. Extensive simulation has been performed to compute the impedance seen by the Gunn diode using the methodology of [19] and [20]. Good agreement has been found in 18 GHz and 26 GHz CW oscillators, but some discrepancies were encountered in voltage controlled oscillators in these bands, especially with respect to the tuning bandwidth, and in 60 GHz CW. Some practical results achieved are: 10 dBm output power at 27 GHz with electronic tuning bandwidth of 300 MHz, and 13 dBm output power at 60 GHz in CW oscillation. Figure 10 shows a view of a 60 GHz CW oscillator.

60 GHz ASK Modulator

Several finline PIN-diode switches (or amplitude modulators) at different frequencies have been designed. The configuration is simple and it consists of two PIN beam-lead diodes connected between the two conductors of a finline and separated a distance L. The spacing between the two conductors of the finline has been chosen to be 0.1 mm, and one of them is DC isolated in order to allow biasing the diodes. Two tapered finline-to-waveguide transitions are used at both ends to interface with standard flanges. The most critical parameter of the design is the distance between the two diodes, which has to be optimized for having high isolation and minimum IL simultaneously. Figure 11 shows the total attenuation as a function of the diode bias voltage and a sketch of the circuit for a modulator at 60 GHz. An isolation of 14 dB and insertion loss of 2 dB have been achieved with 1.2 V excursion.

CONCLUSIONS

Millimeter wave technology is mature today and it is succesfully applied to mobile-communication systems. The allocation of new frequency bands will help to define technological requirements more accurately. To demonstrate the attained performances as well as to compare different approaches, a number of subsystems have been presented using E-plane and conventional waveguide techniques at Ka-band (26.5-40 GHz) and V-band (50-75 GHz). Practical implementation requires sophisticated technical means. Statistical analysis is often required to account for fabrication tolerances. The need of advanced numerical methods to compute the electrical characteristics of complex guiding structures, including discontinuities and transitions, has also been outlined. They should be integrated in specific CAD programs to ease the design-implementation cycle. Commercial software can be used in the design of millimeter wave microstrip circuits, although some user pre-processing is ussually needed. This leads to specific design procedures and software not available at the moment.

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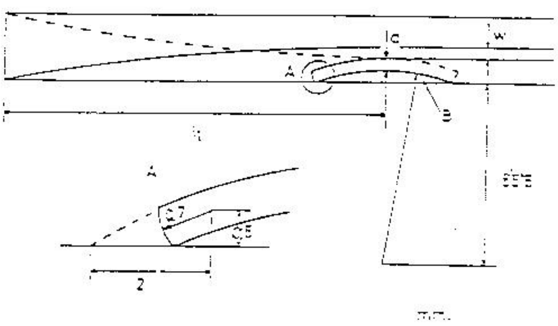

Figure 1 - Waveguide-to-microstrip transition

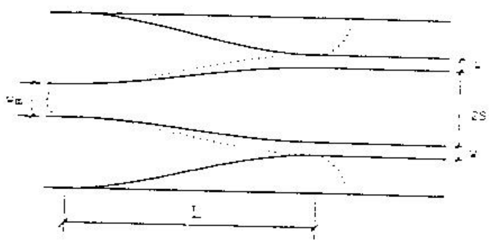

Figure 2 - Microstrip-to-CPW (quasi-TEM mode) transition

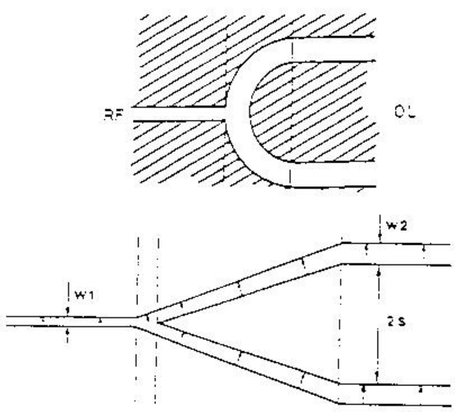

Figure 3 - Finline-to-CPW transition

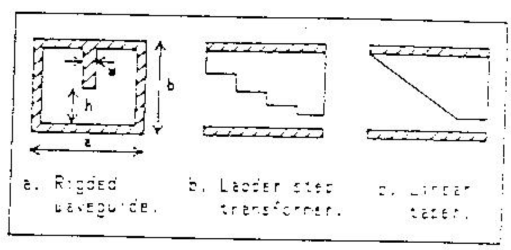

Figure 4 - Ridged-waveguide transitions

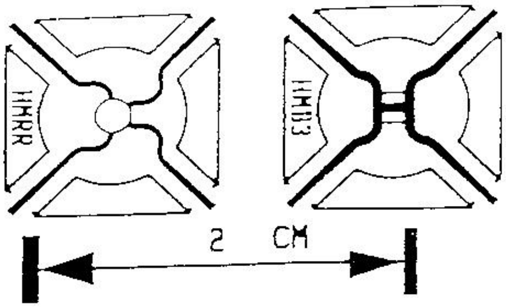

Figure 5 - Masks of microstrip 3-dB Hybrids designed (a) 180º (b) 90º

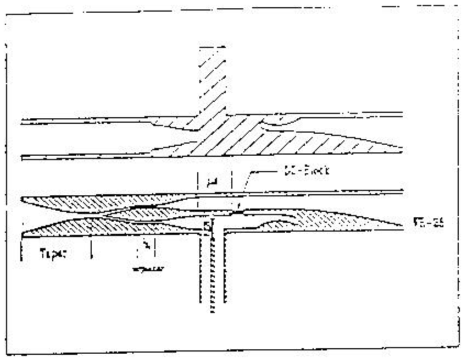

Figure 6 - Mask of E-plane Mixer / BPSK transmission balanced modulator.

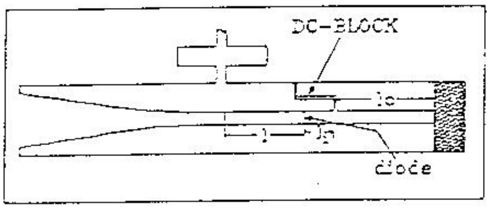

Figure 7 - Mask of finline reflection modulator

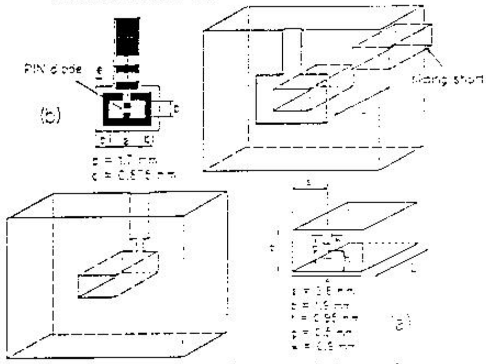

Figure 8 - Reflection modulator in a waveguide/planar configuration

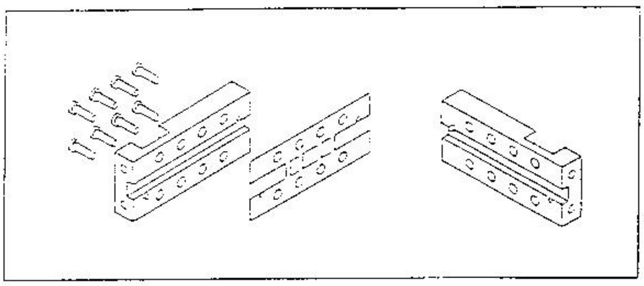

Figure 9 - Metal-insert E-plane filter

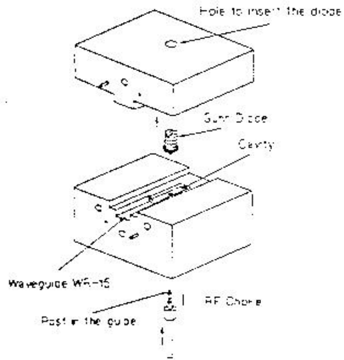

Figure 10 - 60 GHz CW oscillator

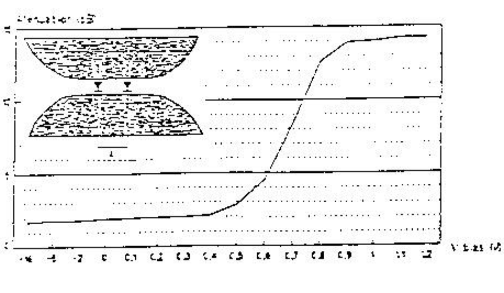

Figure 11 - ASK finline modulator

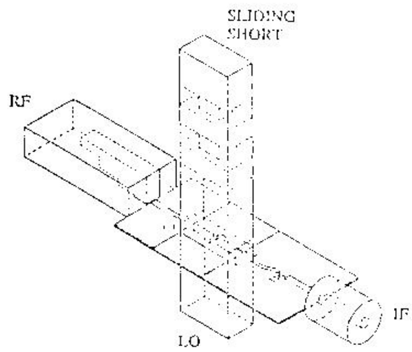

Figure 12 - 60 GHz crossbar mixer