three different devices. Four straight lines have been fitted to the four sets of data points by least squares and the average loss obtained from their slopes is 10dB/cm. This is 7dB below what we previously best result, itself the lowest reported propagation loss for extended cavity laser fabrication by IFVD [8]. This reduction is thought to be due to the better confinement provided by the rib waveguide structure employed in the present lasers, together with the use of high quality epitaxial material. The measured transparency current density (defined as the current density needed to overcome all losses except mirror losses) and internal quantum efficiency for the material used in these experiments are 330A/cm² and 90%, respectively.

![Graph](image)

**Fig. 3** Loss as function of extended cavity length for devices annealed at 900°C for 90s

Slopes of straight line equations give average loss in dB/mm

- 60mA, a(L) = 1.097L + 0.555
- 80mA, a(L) = 1.095L + 0.515
- 100mA, a(L) = 1.07L + 0.541
- 120mA, a(L) = 0.925L + 0.590

The fabrication process was repeated while performing IFVD at the higher temperature of 950°C. By widening the bandgap further, the residual absorption in the passive waveguide sections, which causes the propagation losses, should decrease. Fabrication problems were encountered in the removal of the SrF₂ after RTA at such a high temperature, resulting in many non-working devices. From those regions where the SrF₂ layer could be removed completely, very low loss devices were obtained. The best result was obtained from a laser in which the extended cavity length L was 400µm and for which the loss was 3.6dB/cm. More experiments are under way to produce extended cavity lasers on DQW structures annealed at temperatures between 900 and 950°C.

In conclusion, the fabrication of very low loss extended cavity lasers by IFVD has been demonstrated. Rib waveguide lasers exhibited an average loss of 10dB/cm in the passive sections and the lowest loss of 3.6dB/cm was obtained from a laser with an extended cavity 400µm long, rapid thermal annealed at 950°C for 30s. To the best of our knowledge, both the average and the best results above are the lowest losses obtained from GaAs/AlGaAS extended cavity lasers fabricated by either IFVD or IID.

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MODES OF A SHIELDED CONDUCTOR-BACKED COPLANAR WAVEGUIDE

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**Abstract**

The dispersion characteristics of conductor backed coplanar waveguide (CBCPW) in a metal enclosure, including higher order modes, are obtained by means of an efficient numerical technique, namely the method of lines (MOL). Knowledge of higher order modes is essential for estimating 'singlemode' bandwidth and for characterising discontinuities.

Introduction: CPW structures have attracted considerable attention because of their features such as ease of incorporation of series and shunt elements and suitability for monolithic microwave integrated circuits. To obtain certain desirable features, the original CPW [1] has been modified to include conductor backing [2]; multilayers [3], conducting top cover [4], and shielding [5] etc. Some recent studies have investigated the important phenomenon of leakage of power from the dominant mode in conventional CPW to the dielectric slab, conductor-backed dielectric slab, or parallel-plate waveguide modes [6-8]. These phenomena place an upper limit on the useful frequency range of the conventional structure. Lateral confinement of the CPW has been suggested to control such leakages and has been the subject of many recent studies [9]. Thus, to avoid leakage from the dominant mode and to provide shielding and mechanical integrity, CBCPW in a metal enclosure appears to be a very useful structure. As pointed out in [10], the dispersion in such a structure can be controlled by adopting the suspended substrate configuration.

For the CBCPW in a metal enclosure, only limited results based on full-wave analysis have been reported [5-10]. Information on

References


Additional contributions from various authors have been acknowledged.
the various modes supported by such a structure is very important in the design of various components from the points of view of determining 'singlemode' bandwidth, characterising discontinuities and transitions, and to avoid mode conversion. This Letter reports the modal characteristics of a CBCPW in a metal enclosure. The analysis is carried out using the method of lines, with non-equidistant discretisation for numerical efficiency [11].

![Plane of symmetry](electric/magnetic wall)

Fig. 1 Conductor backed coplanar waveguide (CBCPW) in a metal enclosure

Also shown are the non-equidistant discretisation lines for $T'$ and $I'$.

Analysis: The structure analysed is shown in Fig. 1. By virtue of symmetry, either an electric wall or a magnetic wall can be placed at $x = a/2$. The electric and magnetic fields of this structure are given by

\[
E = V \times X (i(t') z (jw)) - V \times (i(t') \mu z) \\
H = V \times (i(t') z) + V \times (i(t') \mu z) w)
\]

where the scalar potentials $T'$ and $I'$ satisfy the Helmholtz equation. By discretising $T'$ and $I'$ as shown in Fig. 1, and using the notation of [11], we obtain the following characteristic equation:

\[
\begin{bmatrix}
[i\mu]
\end{bmatrix}

\begin{bmatrix}
\mu_I
\end{bmatrix}

\begin{bmatrix}
E_x
\end{bmatrix}

\begin{bmatrix}
\chi
\end{bmatrix}

\begin{bmatrix}
-j\alpha
\end{bmatrix}

= 0
\]

(2)

where $E_x$ and $E_z$ are the $x$ and $z$ electric field components in the slot region. Finally, the solutions are obtained by

\[
\det \{y(f, e, z, a, i) \} = 0
\]

(3)

Results: The eigenvalues of eqn. 3 for a GaAs substrate are plotted in Figs. 2 and 3. The results in Fig. 2, $V(e, r, f, a)$ against frequency, are computed up to 1000 GHz for the same dimensions as in [10]. The fine structure of the first few higher order modes has been brought out more clearly here. Curves 1 and 2 of this Figure are obtained by placing a magnetic wall at $x = a/2$ whereas curve 3 corresponds to an electric wall. Curve B is obtained by removing the centre strip and placing a magnetic wall at $x = a/2$. Curves 1 and B are identical to those reported in [10]. Curve 1, starting from DC, corresponds to the CPW mode; a study of field configuration at different frequencies shows that at high frequencies, the fields for this mode strongly deviate from the quasi-TEM field configuration and a strong $j$-directed electric field develops in the dielectric layer. This feature, and to some extent the field distribution in the slot, are in common with mode 3, explaining the marked dispersion of curve 1 and the merging of curves 1 and 3 at high frequencies.

Fig. 3, $E_x$ against frequency, gives results for all the modes supported by the structure up to 100 GHz for the dimensions mentioned. Curves 2, 3, 4, 5 and 7 are the various modes of perturbed dielectric filled waveguide ($a \times h_1$) or finline with dielectric substrate touching the waveguide wall ($h x a/2$), $h = A_1 + h_2$. The mode corresponding to curve 2 has the lowest cutoff frequency and starts as a perturbed $TE_{m}$ mode of the dielectric-filled guide. In general, the behaviour of modes corresponding to the above mentioned curves is dominated by the presence of high dielectric constant material between conducting planes. The effect of the dielectric material can be reduced by adopting the 'suspended' geometry for the CPW. The remaining curve 6 corresponds to the perturbed air filled guide ($a \times h_2$). In this frequency range, no complex modes are found.

Conclusion: Dispersion characteristics for the dominant and higher order modes of conductor-backed CPW in a metal enclosure are presented. Characteristics for the first few higher order modes have been discussed in some detail. The dispersion in the quasi-TEM CPW mode can be reduced, and the cutoff frequency of the higher order modes can be increased, by using the suspended-CPW configuration.

![Various modes up to 100 GHz for structure shown in Fig. 1, with dimensions w = 0.2mm, a = 2mm, A_1 = 0.2mm, h_2 = 0.6mm, s = 0.1mm, e = 12.9](magnetic wall, electric wave)
Novel excitation schemes for the microstrip ring resonator with lower insertion loss

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Indexing terms: Microwave resonators, Microstrip

Perturbations not affecting the intrinsic resonance of a microstrip ring are introduced to increase the coupling periphery between the feedline and the resonator. Compared to conventional excitation, the authors demonstrate an average improvement of 6 dB in the insertion loss.

The microstrip ring resonator has emerged from being just a characterisation tool for evaluating microstrip parameters [1], to the realm of circuit applications. Ring resonators based tunable/switchable filters [2,3] and microwave-optoelectronic mixers [4] have been demonstrated in the past. Recently, uniplanar resonators, analogous to the microstrip ring, in the coplanar waveguide and slotline configurations have also been reported [5]. For the microwave and optoelectronic circuit applications [4] of ring resonators, a low insertion loss is desirable. This loss is composed of conductor, dielectric, radiative (owing to the curvature of the ring) and coupling losses. Whereas at a given frequency, conductor, dielectric, and radiative losses are intrinsic to the design of the ring, the coupling loss, which is typically the dominant contributor, is dependent on the coupling efficiency. Previously, by employing dielectric overlays across the coupling gap, improved coupling has been achieved [2,5]. Alternately, active devices have also been integrated with the resonator to compensate for losses [6]. However, dielectric overlays are unsuitable for optoelectronic applications [4] as they extend out to the section of the resonator into which light is coupled. In this Letter we propose and demonstrate minimally perturbing, passive coupling schemes with lower insertion loss, for coupling into the ring resonator.

Coupling between the feedline and the resonator may be increased by decreasing the size of the coupling gap (thereby increasing the gap capacitance), by increasing the coupling periphery and by a combination of the previous two. Whereas the first approach is limited by the resolution of the lithographic equipment, the other two approaches are attractive only if they minimally perturb resonance in the ring. Wolff [7] first demonstrated that perturbations could cause splitting of the two degenerate modes of the microstrip ring. In later work [8] we found mode splitting to be strongly dependent on the position of the perturbation. When the ring was perturbed with a notch at an azimuthal angle $\phi = 0$ or $180^\circ$ there was no evidence of mode splitting. However a notch at $\phi = 45$ or $135^\circ$ caused odd modes to split. In Fig. 1 we compare the resonance of the fundamental mode of a ring perturbed with gaps at $\theta = 0$ and $180^\circ$ with an unperturbed ring. The positions of maximum electric field intensity are identical in both these structures indicating that physical perturbations at $\phi = 0$ and $180^\circ$ do not functionally alter ring [4]. The same is also true if notches are employed in place of gaps. The excitation schemes proposed in this Letter rely on the approach of perturbing the ring at $\theta = 0$ and $180^\circ$ to increase the coupling periphery.

Illustrated in Fig. 2 are the four excitations schemes considered here. Scheme A corresponds to conventional excitation, and schemes B - D are newly proposed. In scheme B, to double the coupling periphery compared to scheme A, two gaps were introduced in the ring with the feedlines extending in between these gaps. Here it is important to ensure that the physical lengths of the two halves of the resonator are identical to avoid mode splitting. This scheme is particularly attractive for optoelectronic applications that require splitting of the resonator to facilitate DC biasing [4]. Another approach to increasing the coupling periphery is by employing scheme C. Here the feedlines were extended into notches cut into the resonator to increase the coupling periphery. Compared to scheme B, scheme C does not require the introduction of gaps in the resonator, and hence there is less radiation from open ends. The third approach we considered was that of scheme D. Here the feedlines were tapered and led into "V-shaped" grooves etched into the ring. The tapers provided the necessary high concentration of fields at the tip of the feedlines to couple effectively into the resonator.