"Cul-de-sac" microstrip resonators for high speed integrated optical commutator switches

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ABSTRACT

A novel microstrip resonator structure for use with integrated Y-branch optical modulators fabricated in Ti:LiNbO₃ is proposed. The legs of the structure are intended to act as the electrodes of the modulator, with light being directed into each of the output waveguides of the Y-branch on alternate half-cycles of the standing wave excited in the resonator; forming an optical commutator switch. Such resonators having Al₂O₃ substrates were designed, fabricated, and tested. Measurements on one such resonator, operating at 7.12 GHz and having an unloaded quality factor of 123, indicate that 50 V should develop across the ends of its legs for 35 mW dissipated power; the corresponding values, from the model used to design the resonator, were 179, 50 V, and 24 mW, respectively. Using the model it is shown that a similar resonator fabricated on LiNbO₃ should be able to develop about 50 V for 100 mW dissipated power at 15 GHz.

1. INTRODUCTION

In this paper we present some of design considerations, fabrication details, and test results for a novel, half-wave, microstrip resonator, the "cul-de-sac" resonator, that is intended to be used in conjunction with a Y-branch optical modulator¹ to form an optical commutator. The name of the resonator type reflects the similarity between the resonator structure and the well known street sign; i.e., the resonator consists of two substantially parallel legs open on one end and connected on the other end by an open ring, see figure 1. While the resonators fabricated to date, and reported on here, were made on Al₂O₃ substrates, the parallel legs are intended to act as the electrodes of a Y-branch optical modulator fabricated using a Ti:LiNbO₃ technology; directing pulses of light alternately into each of the output waveguides on successive half cycles of the standing wave excited in the resonator.

Although there has been consistent research into Y-branch type optical modulators in Ti:LiNbO₃²⁻⁴, the devices studied tended to have relatively small branch angles, allowing for lower operating voltages and insertion losses due to radiation, which led to long devices. Devices having larger branch angles are shorter and require higher voltages and they have higher losses due to radiation but have lower losses due to absorption and scattering; in fact, by using large voltages on devices with larger branch angles some of the radiated light is recaptured and steered into the guiding branch improving the insertion loss, in some cases by more than 1 dB⁵.

It is the general thrust of this work to explore the use of resonators in the design of high frequency, low power consumption, electrooptic modulators. Due to the resonator's inherently narrow-band nature, such electrooptic modulators would probably find application as optical commutators in time division multiplexer/demultiplexer systems^{6,7}.

Figure 1. A cul-de-sac resonator acting as the electrodes of a Y-branch optical modulator.

2. THE RESONATOR STRUCTURE AND MODEL

The resonator is to be a half-wave (or odd multiple of a half-wave) resonator so that the potential on each of the parallel legs will be of opposite polarity, establishing a voltage between the legs across the gap. It is intended that the potential difference between the end of the resonator's legs be much higher than that on the microstrip feed line; i.e., the resonator is to act as a voltage transformer. In this way, the resonator electrode can apply a high voltage to the optical modulator while requiring relatively little power from the source. It was our goal to limit the supplied power to 100 mW while obtaining a modulating voltage of about 50 V across the gap.

Losses are a major factor when determining the structure of a resonator. There are two main kinds of loss in microstrip resonators: ohmic skin loss and radiation loss. Since, for a fixed frequency, the ohmic loss is proportional to the length of the line, a half-wave resonator (as opposed to higher order odd multiples of a half-wave) will have the least loss of this kind. Radiation loss is primarily due to discontinuities in the transmission line, i.e., it is mainly dependent on the geometrical structure.

The hairpin⁸ and the open ring⁹ structures were initially considered as candidates for the resonator electrodes. They were modelled to determine their suitability for the intended application. "Current crowding"¹⁰ near the inner edges of the two closely spaced legs of the hairpin resonator increased the ohmic loss to unacceptable levels. In the open ring structure, if the radial length of the gap was to be sufficient for the microstrip constituting the resonator to serve as the optical

modulator's electrodes, the width of the microstrip would have been too wide; resulting in the energy being stored primarily as a large current, rather than as a large voltage, i.e., reducing the desired voltage transformation effect. The best design that we have studied to date, in that it will meet our power/voltage goals, namely the cul-de-sac resonator, is a hybrid of the hairpin and the open ring structures.

As can be seen from figure 1, the cul-de-sac structure has two major discontinuities, one at either end of the parallel legs. Still, since the interelectrode gap is narrow, $\sim 10 \mu\text{m}$, the dipoles (in fact quadrupoles when the images in the ground plane are considered) formed at the ends are inefficient radiators. For our model the radiation loss was calculated using a formula for planar resonators¹¹ using the method of images (see for example Ref. 12). On the other hand, the narrow gap causes current crowding so that the legs should be kept short. Since the legs are intended to act as the electrodes of Y-branch optical modulators with relatively large branch angles, $1\text{-}2^\circ$, they can be relatively short, e.g., the legs need only account for about one fourth of the total length of a 15 GHz resonator on LiNbO_3 .

Further down we will show that, in order to obtain a large voltage between the legs of our resonator, we will want a large unloaded quality factor, Q_u . It is well known that $Q_u = (2\pi f_0 W_s)/P_l$; where f_0 is the resonant frequency of the resonator, W_s is the stored energy, and P_l is the average power loss. Also, since the energy in a resonator is stored in the electric and magnetic fields, i.e., in the voltage and current distributions, if we wish to obtain a large voltage between the ends of the resonator's legs then the microstrip forming the resonator should have a high characteristic impedance.

To determine structure parameters, such as the width of the strips, the radius of the ring, and the length of the parallel legs, it is important to know the voltage (or current) and impedance distributions along the microstrip line forming the resonator. To do this it is convenient to model the resonator as a straight line resonator. Figure 2a is the straight microstrip line equivalent circuit where: section AB corresponds to the ring, CA and BD correspond to the parallel legs, EC and DF are the excess lengths caused by the open ends, and O is the midpoint of the resonator. The parallel legs, being virtually identical to coupled microstrips in their structure, have a characteristic impedance Z_l that is smaller than that of the ring Z_r , therefore, the corresponding equivalent microstrip is wider. The steps in the characteristic impedances have the effect of making the voltage distribution along the microstrip piecewise sinusoidal; also, they cause the propagation constant for the parallel legs k_l to be different from that for the ring section k_r . Knowing Z_l , Z_r , k_l , and k_r the voltage distribution can be obtained. Furthermore, the attenuation constants for the two parts, α_l and α_r , are different as well. Using α_l and α_r the total ohmic loss can be calculated. For our model α_l was assumed to be the same as that for two coupled microstrip lines (see for example Ref. 13) and α_r was taken to be that for the microstrip constituting the ring.

Although the characteristic impedance of the microstrip composing the resonator should be large, the input impedance of the resonator is a function of the coupling position which can, theoretically, be any value between zero and infinity. However, in order to reduce the reflection loss, the power should be coupled into the resonator at a point where the input impedance is matched to that of the power source; this may require the use of a quarter-wave transformer. Here

(a)

(b)

Figure 2. (a) The straight microstrip line equivalent circuit for the cul-de-sac resonator. (b) The equivalent circuit used to calculate the input impedance and the coupling.

we have assumed that the entire source power may be delivered to our resonator and is dissipated therein.

For the purpose of modelling the input impedance and the coupling between the source and the resonator, the equivalent circuit in figure 2a can be further simplified, as shown in figure 2b, as two shunt, open-circuited, uniform transmission lines. Here θ is the electrical length from the coupling point P to the centre of the resonator and Z_s is the source impedance. In this model the characteristic impedance of the transmission line forming the resonator is the same as that of the microstrip constituting the ring. This circuit is equivalent to the cul-de-sac resonator if we are concerned only with the input impedance looking into a point on the ring. Using basic transmission line theory with this model it can be shown that the input impedance is

$$Z_i = \frac{2Z_r Q_u}{\pi} \sin^2 \theta . \quad (1)$$

The relationship between the voltage across the gap at the open ends of the parallel legs V_o and the source voltage V_s is

$$\frac{V_o^2}{Z_r} \sim \frac{4\beta}{(1+\beta)^2} \frac{8Q_u}{\pi} \frac{V_s^2}{Z_s} \quad (2)$$

where β is the coupling coefficient. In other words the source voltage can be transformed by a factor of $\{2\beta^{1/2}/(1+\beta)\} \times (8Q_u Z_r / \pi Z_s)^{1/2}$. Clearly, $\beta = 1$ maximizes the factor $2\beta^{1/2}/(1+\beta)$. Also, Z_s is usually a fixed value. While a large Z_r is desirable, the losses associated with its increase put limits on its possible value. It is, therefore, desirable to design a resonator having a relatively large Q_u .

3. RESULTS AND DISCUSSION

To verify our model, cul-de-sac resonators designed to operate at 7 GHz were fabricated on Al_2O_3 substrates and tested. The structure parameters were optimized, i.e., to give a maximum V_o for the least amount of power dissipated, assuming $Z_s = 50 \Omega$ and $\beta = 1$. The thickness of the substrate was 0.89 mm. The microstrip forming the resonators was made of 6 μm thick, electrolytically deposited Au; in order to obtain this thickness a double masking technique was used in the patterning¹¹. The gap between the parallel legs was 12 μm . Each of the legs was 1.3 mm long and 0.14 mm wide. The mean radius of the ring was 0.86 mm and the width was 0.14 mm. The calculated Z_r was 95 Ω and the predicted Q_u was 179. This gave an expected V_o of 50 V for 24 mW input power. The devices were tested using a scalar network analyzer. A plot of the normalized reflected power vs. frequency is shown in figure 3. The measured Q_u for this device was 123 and f_o was 7.12 GHz (since we are using a scalar network analyzer to take a reflection measurement, the value for Q_u differs slightly from that obtained by simply using the ratio $f_o/\Delta f_{3\text{dB}}$ ¹¹). Using the model above and assuming that the resonator impedances would be nearly equal to the design values, a V_o of 50 V would require 35 mW input to this resonator.

Figure 3. A plot of the normalized reflected power vs. frequency for a cul-de-sac resonator.

Using our model a resonator was designed to operate at 15 GHz on a 0.3 mm thick z-cut plate of LiNbO_3 . The structure parameters were optimized assuming $Z_s = 50 \Omega$, $\beta = 1$, a gap width of $4 \mu\text{m}$, and Au microstrip $6 \mu\text{m}$ thick. The length of the each parallel leg would be 0.35 mm and the width would be 0.03 mm. The mean radius of the ring would be 0.79 mm and the width would be 0.02 mm. The calculated Z_r was 62Ω and the predicted Q_u was 71. This gave an expected V_o of 59 V for 100 mW input power. Assuming a similar reduction of 30% between the Q_u 's of the fabricated resonator and the model, we have calculated that a V_o of 49 V would result for 100 mW input power. In other words such a resonator should meet our power/voltage requirements.

4. SUMMARY

A novel microstrip resonator, the cul-de-sac resonator, that is intended to be use with Y-branch optical modulators has been proposed. Some of the considerations taken into account when modelling and designing these devices were described. Resonators of this type were fabricated on Al_2O_3 substrates and were tested to verify the model. Measurements indicate that resonators of this type fabricated on LiNbO_3 should develop about 50 V between the ends of their legs while dissipating 100 mW of power at 15 GHz.

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6. REFERENCES

1. N.A.F. Jaeger and W.C. Lai, "Y-Branch Optical Modulator," *SPIE Vol. 1583 Integrated Optical Circuits*, pp. 202-209, 1991.
2. W.K. Burns, A.B. Lee, and A.F. Milton, "Active Branching Waveguide Modulator," *Appl. Phys. Lett.*, vol. 29, no. 12, pp. 790-792, 1976.
3. Y. Silberberg, P. Perlmutter, and J.E. Baran, "Digital Optical Switch," *Appl. Phys. Lett.*, vol. 51, no. 16, pp. 1230-1232, 1987.
4. P. Granstrand, B. Lagerström, P. Svensson, L. Thylén, B. Stoltz, K. Bergvall, J.E. Falk, and H. Olofsson, "Integrated Optics 4×4 Switch Matrix with Digital Optical Switches," *Electron. Lett.*, vol. 26, no. 1, pp. 4-5, 1990.
5. W.C. Lai, "LiNbO₃ Y-Branch Optical Modulator," *M.A.Sc. Thesis, Univ. of British Columbia*, 1991.
6. A. Djupsjöbacka, "Time Division Multiplexing Using Optical Switches," *IEEE J. Select. Areas Commun.*, vol. 6, no. 7, pp. 1227-1231, 1988.
7. N.A.F. Jaeger, "High Speed Integrated Optical Modulators in Lithium Niobate and Compound Semiconductors," *CCTA Technical Papers*, pp. 117-122, 1992.
8. R.J. Roberts and B. Easter, "Microstrip Resonators having Reduced Radiation Loss," *Electron. Lett.*, vol. 7, no. 8, pp. 365-368, 1971.
9. I. Wolff and V.K. Tripathi, "The Microstrip Open-Ring Resonator," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-32, no. 1, pp. 102-107, 1984.
10. S.E. Schwarz, M.D. Prouty, and K.K. Mei, "Radiation from Planar Resonators," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-39, no. 3, pp. 521-525, 1991.
11. M. Chen, "Microstrip Resonators for High Speed Optical Commutator Switches," *M.A.Sc. Thesis, Univ. of British Columbia*, 1992.
12. R.F. Harrington, Time-Harmonic Electromagnetic Fields, chapter 3, McGraw-Hill, New York, 1961.
13. K.C. Gupta, R. Garg, and I.J. Bahl, Microstrip lines and Slotlines, pp. 153-197, Artech House, Norwood, 1979.