Bandwidth Improvements for Loaded-Line Traveling Wave Electro-optic Modulators

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Abstract

A circuit theory approach is employed to analyze the factors and limitations affecting impedance, optical/electrical velocity matching, and bandwidth in loaded-line traveling-wave modulators. Design improvements are explored that show a potential 30- to 40-percent improvement in bandwidth over previous designs.

Introduction

To date, electro-optic phase modulators have been built demonstrating excellent sensitivity figures-of-merit, ranging up to 500 to 600°/V-mm[1],[2]. This sensitivity is orders of magnitude greater than achieved using ferro-electric crystals such as lithium niobate. Semiconductor modulators are usually fabricated in the form of a p-i-n diode structure appropriately confined so as to perform optical waveguiding. By applying an electrical signal to selected sections of this waveguide, the optical signal can be phase shifted using various band-edge or quantum-confined phenomena. This facilitates the modulation of the optical carrier by an electrical signal. Despite their great sensitivity, these modulators tend to be highly capacitive and are therefore difficult to impedance match at high frequencies.

One of the best techniques for accomplishing impedance matching to capacitive elements is by using traveling-wave techniques[3]. By integrating the capacitive electrodes with inductive or high impedance transmission lines, an artificial transmission line that can be designed for a specific impedance and propagation constant is created. For the entire length of the traveling-wave modulator to be effective, the velocity of the electrical signals must be matched to the propagating optical signal. In III-V semiconductor materials, the fields of the electrical signal propagate in both the substrate and air, and thus have a higher propagation velocity than the optical signal, which is mostly confined to the substrate. Proper design of modulator electrodes therefore requires both matching the capacitive elements of the electrodes while slowing down the electrical signal. Integration of the modulator electrodes into an artificial line accomplishes both of these tasks and greatly improving the bandwidth figure-of-merit of a modulator[4].

The most effective method to date of achieving both impedance and velocity matches in a transmission line electrode was presented by Walker[5]. His technique adjusts the impedance of the interconnecting transmission lines and capacitive electrodes to simultaneously achieve proper impedance and velocity matching. This technique was successfully used to achieve a record 36-GHz bandwidth for a semiconductor modulator, having an impressive 6.4-GHz/volt figure of merit.

In this paper, a circuit design approach rather than slow-wave analysis is used to determine which properties of the interconnecting lines have an impact on the impedance, velocity, and bandwidth of the circuit. Afterwards, several transmission line configurations are analyzed to determine what performance improvements are possible. By using raised microstrip interconnects, a ~30-percent improvement in bandwidth can be achieved. Additionally, by incorporating input impedance matching, improvements on the order of ~40 percent can be realized.

Theory

In general, a traveling-wave modulator (or any traveling-wave structure) is constructed such that each capacitive element (the active element) is interconnected with inductors to create an artificial transmission line as shown in Figure 1(a). The overall impedance, Z_a, and phase velocity, v_a, of this artificial line are determined by

$$Z_a = \sqrt{\frac{L'}{C'}} \qquad v_a = \frac{1}{\sqrt{L'C'}} \tag{1}$$

where L′ and C′ are the total inductance and capacitance per unit length of the segments along the line.

At microwave frequencies, the interconnecting inductance is generally realized by a short length of high impedance transmission line as shown in Figure 1(b). These transmission

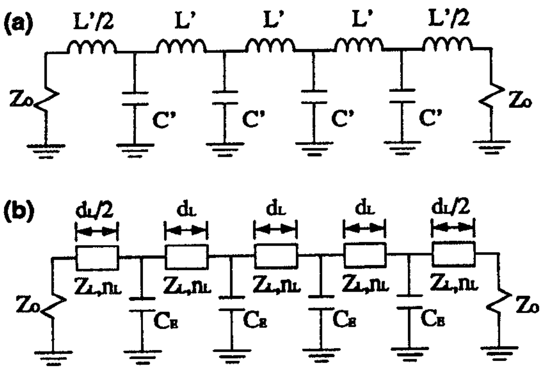

Figure 1. Transmission Lines

lines have an impedance Z_L, an effective index of n_L, and a physical length of d_L. One can model the effects of these lines with a pi-equivalent circuit as shown in Figure 2. The element values in this model are determined by [6]

$$L = \frac{Z_L}{2\pi f} Sin\left[\frac{2\pi f d_L n_L}{c}\right] \sim \frac{Z_L d_L n_L}{c} \tag{2}$$

$$C_L = \frac{1}{\pi f Z_L} Tan\left[\frac{\pi f d_L n_L}{c}\right] \sim \frac{d_L n_L}{Z_L c} \tag{3}$$

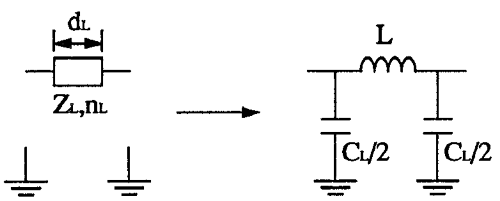

Figure 2. Pi-Equivalent Circuit

where f is the frequency of interest, c is the free space velocity of light, and d_L is the length of the transmission line. The approximations on the right are valid for values of d_L less than a quarter-wave long. Replacing the transmission lines with this model yields an equivalent circuit shown in Figure 3. The impedance of this artificial line is given by

$$Z_a = \sqrt{\frac{L}{C_E + C_L}} = \frac{Z_L}{\sqrt{1 + \frac{c Z_L C_E}{d_L n_L}}} \tag{4}$$

while the effective microwave index of the line calculates to be

$$n_a = \frac{c\sqrt{L(C_E + C_L)}}{d_L} = \left[\frac{Z_L^2}{Z_o^2} - 1\right]\frac{n_L^2 d_L}{c Z_L C_E}\sqrt{1 + \frac{c Z_L C_E}{d_L n_L}} \tag{5}$$

Meanwhile, the cutoff frequency of this constant-k filter network is given by

$$f_c = \frac{1}{\pi\sqrt{L(C_E + C_L)}} = \frac{c}{\pi d_L n_L \sqrt{1 + \frac{c Z_L C_E}{d_L n_L}}} \tag{6}$$

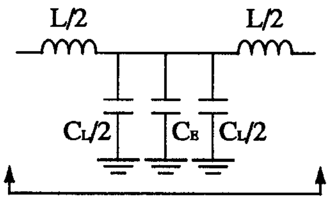

Figure 3. Single-Section of Artificial Transmission Line

Generally, the frequency response is primarily determined by the size of the active element, in this case the modulator electrode, which determines C_E. As a higher and higher frequency is desired, it is necessary to make smaller active elements. The question at hand is how can the interconnecting lines be adjusted to give the maximum frequency response for a given size of C_E.

We can constrain the length, d_L, of the inductive line section to give us a desired travelling wave impedance, Z_o, by setting

$$d_L = \frac{c Z_L C_E}{\left[\frac{Z_L^2}{Z_o^2} - 1\right] n_L} \tag{7}$$

Inserting this constraint into Equations (5) and (6) show that

$$n_a = \frac{Z_L n_L}{Z_o} \tag{8}$$

$$f_c = \frac{1}{Z_o \pi C_E}\left[1 - \frac{Z_o^2}{Z_L^2}\right] \tag{9}$$

Several observations can be made by studying these last two equations. First, the cutoff frequency of the network is independent of n_L and is maximized by making Z_L as high an impedance as possible. Second, increasing the impedance Z_L

also increases the microwave index and therefore slows down the microwave signal. Since it is necessary to have the microwave index, n_a, equal the optical index, n_o, any increase in impedance must be accompanied by a corresponding decrease in n_L. The challenge is to find layout configurations that can maximize Z_L while minimizing n_L to allow n_a to match the optical index n_o. To this end, several different configurations are considered.

Layout Configurations

Coplanar waveguide/coplanar stripline - Coplanar stripline is the transmission media originally chosen by Walker for design of his traveling-wave modulators. As such, it provides a good basis for comparison. With coplanar lines, the electric field is evenly split between the air and the dielectric substrate. The impedance of transmission lines are primarily determined by the width of the lines, w, and their spacing, s, to the ground plane(s). This configuration also possesses a dielectric constant that is relatively independent of impedance, and is given by

$$n_L = \sqrt{\frac{\varepsilon_R+1}{2}} \tag{10}$$

where ε_R is the dielectric constant of the substrate. If we use a gallium arsenide substrate with ε_R= 12.9, then n_L= 2.636. Equation (8) shows that achieving a velocity match to an optical index of 3.4 (for gallium arsenide waveguides) requires an impedance of 64.5 Ω. With an electrode capacitance of 0.0905 pF (example used by Walker), the cutoff frequency shown in Equation (9) is 28.06 GHz. These results are essentially the same given by Walker, though obtained through a different derivation.

Microstrip- Microstrip is perhaps the most commonly used form of microwave transmission medium. Microstrip possesses a non-TEM mode of transmission where impedance and effective dielectric constant are complicated functions of the line width, w, substrate height, h, and conductor thickness, t. The most widely used expressions for impedance, Z_L, and effective dielectric constant, n_L^2, of microstrip lines is given by Hammerstad and Jensen[7]. Knowing that high impedances and low dielectric constants are required of the interconnecting lines, a thicker substrate (6 mils is common) and thick plated lines (t = 0.2 mil) are chosen on gallium arsenide. Using Hammerstad and Jensen's equations, calculations of impedance and index were made for various widths of transmission line and are shown in Figure 4. Shown in Figure 5 is the resulting microwave index, n_a, as a function of linewidth for the artificial transmission line as given by Equation (8). The resulting strip width yields a microwave index of 3.4 is 2.6 mils. This width has a line impedance of 60.46 Ω, an index of 2.785, and a cutoff frequency of 22.2 GHz. It is apparent that microstrip is unable to generate a high enough impedance with a low index to provide any improvement in cutoff frequency.

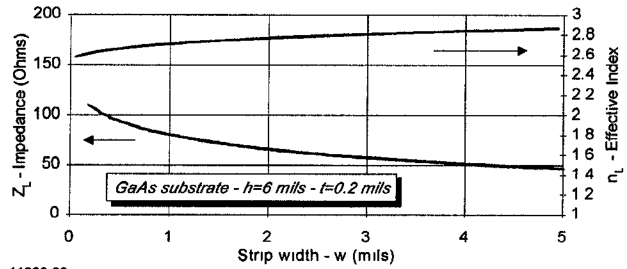

Figure 4. Microstrip Characteristics

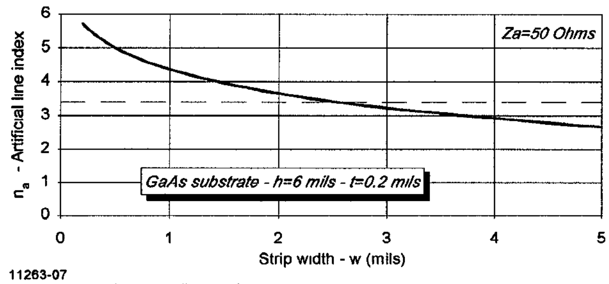

Figure 5. Microstrip Characteristics

Raised microstrip- Raised microstrip is commonly used in the layout and design of spiral inductors because it maximizes the inductance/unit length and reduces the coupling between adjacent conductors (which are usually spaced close together). This configuration is essentially the same as microstrip, except that the conductor is raised up off the substrate by a distance h′. This added height corresponds to the height of airbridges used during the fabrication of MMICs. This technique significantly increases the line impedance and lowers the dielectric constant because raising the conductor above the substrate puts more of the field in the air rather than in the substrate. Unfortunately, there are no empirical equations that describe the impedance and effective dielectric constant for this particular multilayer configuration. Calculations are generally accomplished with the aid of an electromagnetic simulator. Using typical values for microwave IC processing of h′ = 2 μm and h = 6 mils yields the results shown in Figure 6. These results were obtained

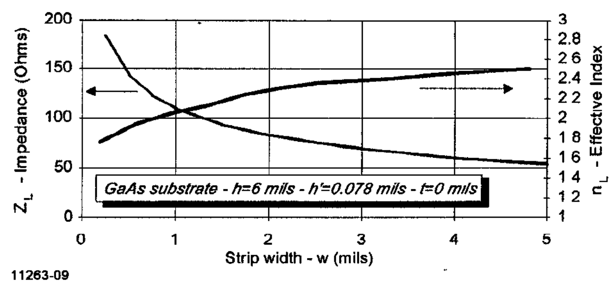

Figure 6. Raised Microstrip Characteristics

using Sonnet's *em*™ full-wave electromagnetic field simulator[8]. Figure 7 shows the resulting microwave index, n_a, as a function of linewidth for the artificial transmission line as given by Equation (8). For a width of 2.7 mils the microwave index is $n_a = n_o = 3.4$ with a line impedance of 72.0 Ω and an index of 2.374. This corresponds to a cutoff frequency of 36.4 GHz, or a 30-percent improvement in bandwidth over the coplanar configuration used by Walker.

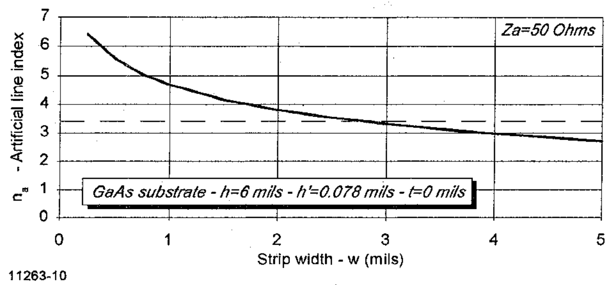

Figure 7. Raised Microstrip Characteristics

Bondwire interconnects - Though bondwire interconnects are not a producible means of constructing specific inductances at microwave frequencies, they provide an interesting test case. A general rule of thumb in microwave design is that 0.7-mil bond wires on high dielectric substrates act as transmission lines with a 150- to 200-Ω impedance. Since the fields are generally a good distance above the top of the substrate, most of the field is in the air yielding an effective index approaching unity. Equation (8) shows that using bondwire interconnects with a $n_L = 1$ means that a velocity match to $n_o = 3.4$ occurs with an impedance of 170 Ω. Consequently, bondwires with a length of 17.2 mils [from Equation (7)] would interact with $C_E = 0.0905$ pF to yield an artificial line cutoff frequency of 64.3 GHz. It is not practical to build a 60-GHz circuit that depends on specific lengths of bond wires. However, this case presents an idealistic limit to the upper frequency obtainable with a loaded-line traveling-wave modulator under the conditions given ($C_E = 0.0905$ pF).

Impedance Matching

Beyond picking transmission structures that optimize the product of $Z_L n_L$, another alternative is to choose an impedance level different than 50 Ω for the artificial line impedance. Since most microwave systems use a 50-Ω impedance, using a non-50-Ω modulator impedance would require either accepting the resultant VSWR mismatch (and accompanying ripple) or adding an auxiliary impedance matching network. This practice is common with lithium niobate modulators, which usually have an input impedance around 25 Ω (because of the high dielectric constant of lithium niobate and line width limitations). Typically, a tapered transmission line transformer is used, exhibiting a high-pass frequency response. A drawback to this scheme is that the transformer reduces the voltage at the electrodes, requiring a higher voltage to switch the modulator from full on to full off. However, many times upper end frequency response is of prime concern and this tradeoff is acceptable. Using the raised microstrip results of Figure 6 with a 40-Ω modulator impedance yields a line impedance of 54.2 Ω and line index of 2.500 for an overall microwave index of 3.4. This yields a maximum frequency of 40.0 GHz, a 40-percent increase over coplanar lines in a 50-Ω system. Obviously lower impedances could be used, but problems start to arise with properly realizing the wide line widths (low impedances) necessary.

Conclusions

Factors affecting the impedance, bandwidth, and velocity matching of electro-optic traveling-wave modulators have been explored. Optimization of modulator bandwidth requires utilizing high impedance, low dielectric constant transmission lines to interconnect modulator electrodes. Understanding this, many possibilities exist in picking materials, geometries, and tradeoffs that can improve operational bandwidth. A few factors have been explored that demonstrate that a 30- to 40-percent improvement over previous work is possible. Further research in this area will lead to continued improvements in the future.

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