

ANALYSIS AND DESIGN OF TAPERED FINLINE ARRAYS FOR SPATIAL POWER COMBINING

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INTRODUCTION

A broadband waveguide-based spatial power combiner was previously proposed by Alexanian and York [1]. The combiner consists of a array of tapered slotline antennas and MMIC power amplifiers, and was placed between two rectangular waveguides, as shown in Figure 1. As well as receiving and transmitting signals from and to the waveguides, the antenna sections also serve as impedance transformers, providing matched load to the source. This paper describes our effort in the analysis and design of an optimal tapered finline array, which resulted in desirable characteristics of gain flatness and return loss reduction, and therefore, improvement of power performance and efficiency of the combiner. The design work is based on the theory of small reflection and analytical results for Klopfenstein tapers [2-5], with the understanding of the relationship between the propagation constant and the geometrical parameters of finline. The spectral domain method [6] was employed to analyze the finline array structure to set up the relationship. Based on this information, a tapered finline array was designed and fabricated. The return loss measurement showed good agreement with the theoretical results, indicating promising outlook in performance improvement for the spatial power combining circuit.

FORMULATION

The design rules for optimal tapers involve an extension of the well-known theory of small reflection to a non-TEM structure, and analytical results for Klopfenstein tapers [2-5]. Using the theory of small reflection [4], it can be shown that an gradual impedance taper of length L on a non-TEM line has an input reflection coefficient given by

$$\Gamma_{in}(f) = \frac{1}{2} \int_0^{\theta} e^{-j\theta} \frac{d}{d\theta} \ln \frac{Z(\theta)}{Z_0} d\theta \quad \text{with} \quad \theta(f, z) = \int_0^z 2\beta(f, z') dz' \quad (1)$$

$\theta(f, z)$ is the round-trip phase delay to a point z along the taper, and $\theta_i = \theta(f, L)$. The function $Z(\theta)$ describes the variation in wave impedance along the taper. It has been shown by Klopfenstein [6] that, in order to maintain an input reflection coefficient to be less than the maximum allowable reflection coefficient Γ_m over the passband, $Z(\theta)$ must take the form,

$$\ln \frac{Z(\theta)}{Z_0} = \frac{1}{2} \ln \frac{Z_L}{Z_0} + \Gamma_m A^2 F(2\theta / \theta_i - 1, A) \quad \text{with} \quad F(x, A) = \int_0^x \frac{I_1(A\sqrt{1-y^2})}{A\sqrt{1-y^2}} dy \quad (2)$$

$I_1(u)$ is a modified Bessel function of the first kind and first order. Assuming waves propagating along the finline structure are quasi-TE in character, equation (2) can be rewritten as

$$\beta(f_0, z) = \sqrt{\beta_L \beta_0} e^{-\Gamma_m A^2 F(2\theta(f_0, z)/\theta_0 - 1, A)} \quad (3)$$

where $\beta(f_0, z)$ represents the propagation constant at point z along the taper line. The passband is defined as $\theta_0 \geq 2A$, from which the lowest frequency of operation f_0 can be defined, assuming the propagation constant β is a monotonically increasing function of frequency. With further manipulation of equation (3), it can be shown that $\beta(L) = \beta_L e^{\Gamma_m}$, which implies that the values of L and Γ_m will both be set if any one of them is determined.

With the lowest operating frequency f_0 and either L or Γ_m specified, the propagation constant can be computed as a function of the position along the optimal taper through equations (3) and (1), based on an iterative calculation procedure.

SIMULATION

The spectral domain method [6] was employed, due to its simplicity in terms of coding effort, to set up the relationship between the propagation constant and the geometrical parameters of a finline array structure, most importantly the slot width. A table relating the propagation constant, slot width, and operating frequency was generated so that the propagation constant for a specific slot width at a particular frequency can be interpolated. A Mathematica script was written to perform the spectral domain computations and derive an interpolating function.

Figure 3 outlines the cross section of a 2x2 finline array that was used for simulation. Physically it corresponds to two 10-mil thick aluminum nitride substrates, with a separation of $c = 5$ mm, placed in an X-band (WR-90) waveguide ($a = 0.9$ in., $b = 0.4$ in.). Symmetries along the major axes were used to minimize the size of computation domain. With the normalized gap ($2g/b$) as a variable, the corresponding effective dielectric constant, $(\beta / k_0)^2$, was computed at various frequencies, and a family of curves are shown in Fig. 3. Based on this information, an optimal tapered finline array was designed and a profile of normalized gap width vs. position along the taper line was computed and shown in Figure 4. The characteristic impedance at each end of a taper line was calculated, with approximation, based on Hoffmann's formula [7] for double-strip coplanar waveguide with infinitely wide ground conductors.

Different physical taper is required for different frequency due to the frequency dependence of the propagation constant, which is obviously impossible to implement. Therefore, the center frequency at the band of interest was chosen to be the operating frequency.

MEASUREMENT

An optimal tapered finline array was fabricated with 10 mil aluminum nitride substrates. Multi-section tapers were used to realize the ideal continuous tapers. The taper length was chosen to be 17.5mm and 50 sections (350 μ m each) were used for each taper. All the tapered finlines were terminated with 100 Ω single wrap-around chip resistors by using 1 μ m gold wires, as shown in Figure 5. Measurement was performed in a waveguide environment and the results presented very good agreement with the simulation result, as shown in Figure 6. A little less than 20dB return loss covers the entire band from 7.5 to 12.5 GHz and the response is broadband in character.

It should be noted that 100 Ω chip resistors were used as matched terminations. Several measurements were performed for terminating resistors with different values, ranging from 30 Ω to 100 Ω , with 100 Ω yielding the best return loss, suggesting that the slotline dimensions at the end of the taper yielded a characteristic impedance close to 100 Ω . This in turn suggests that the formulas in [7] are not suitable for computing characteristic impedance in this specific finline structure.

CONCLUSION

In an effort to enhance the performance of a previously proposed waveguide-based spatial power combiner, an optimal tapered finline array was designed to improve the gain flatness and return loss of the power combiner. An optimal taper was designed based on the theory of small reflection and the design equations for Klopfenstein tapers. Due to its simplicity in developing computer codes, spectral domain method was used to compute the propagation constants corresponding to finline structure with various slot widths. The results of return loss measurements shown good agreement with the simulation results, presenting broadband frequency response and good return loss characteristics. Finite difference scheme is currently under investigation in an effort to model the groove effect and larger scale of finline arrays, since it is relatively simple to develop computer codes.

ACKNOWLEDGEMENT

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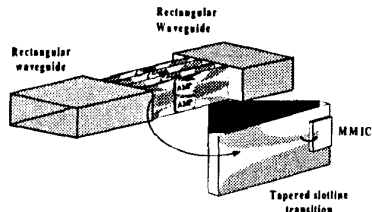


Figure 1 – Topology of spatial combiner

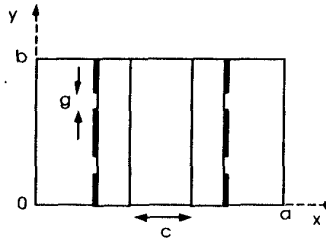


Figure 2 – Two card array in a rectangular waveguide

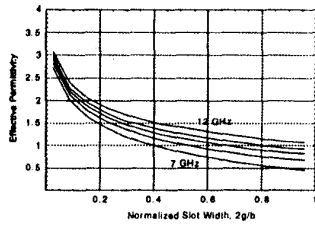


Figure 3 – Effective dielectric constant vs. normalized slot width at various frequencies

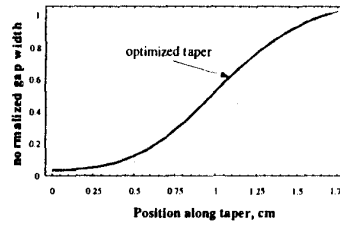


Figure 4 – Normalized gap width vs. location along the optimal tapered finline structure

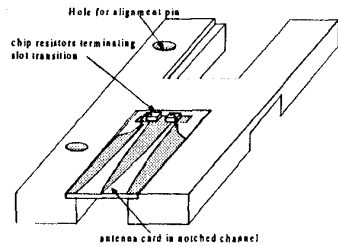


Figure 5 – Tapers terminated with chip resistors (one card only)

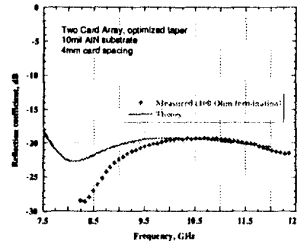


Figure 6 – Comparison between measurement and simulation results