allows the real frequency transmission zeros to be extracted independently and realized by individual elements. This is important from a sensitivity view point. The remaining transmission zeros are realized in the center of the network with a cross-coupled array structure.

One of the most important results is that selective linear phase filters with finite real frequency transmission zeros may be realized with only positive couplings in the entire network. This is a necessary condition for many types of physical realizations particularly in waveguide structures when cavities only supporting a single mode of operation are used.

From this prototype, many narrow-band bandpass structures may be designed using the reactance slope parameter technique. At microwave frequencies, the phase shifters are realized by lengths of line or waveguide and the remaining elements by iris coupled cavities. To demonstrate the feasibility of the method a sixth degree pseudoelliptic bandpass filter has been constructed using low-loss TE_{011} mode cavities.

Having established the procedure for realizing combined pseudoelliptic and nonminimum phase (self-equalized) characteristics with the low-loss TE_{011} mode cavity, the other advantages of using this mode should be mentioned. The dimensions are comparatively large which mitigates multipactor effect in space, ensures high power handling and renders construction easy at millimeter wave frequencies up to about 40 GHz. The planar construction enables the device to be mounted on a flat cooling plate for efficient transfer of dissipated heat. The independent tuning for the pole cavities simplifies the tuning procedure for the filter.

Perhaps the most important application for this type of filter realization will be in the high-power low-distortion output multiplexion of contiguous or near-contiguous channels. A procedure is currently under development to match-multiplex extracted pole filters onto a common waveguide manifold.

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Short Papers

On the Design of Temperature Stabilized Delay Lines

PEITRO DE SANTIS

Abstract—**Recently published design formulas for delay lines with transmission phase temperature stabilization are shown to be approximate. Their validity range is assessed. New exact formulas of broader validity are presented.**

With the advent of low-loss microwave ceramic materials featuring negative dielectric constant temperature coefficients as well as high dielectric constants, a new class of temperature stabilized delay lines has recently become feasible. In [1] design criteria have been presented for temperature stabilized transmission line delay lines (TLDL) made up of barium tetratitanate and sapphire MIC's. These devices, to be used in satellite born regenerators for PSK data transmission, must provide a fixed group delay and a temperature stable transmission phase. To meet these requirements the lengths l_A and l_B of the $BaTi_4O_9$ and Al_2O_3 TLDL's (hereafter referred to as partial delay lines (A) and (B)) are calculated by solving the system

$$\frac{l_A}{v_{gA}} + \frac{l_B}{v_{gB}} = \tau \tag{1}$$

$$\left(\frac{\Delta\phi_A}{\Delta T} + \frac{\Delta\phi_B}{\Delta T}\right)(\phi_A+\phi_B)^{-1} = \alpha_\phi = 0 \tag{2}$$

where $V_{gA,B}$ and $\phi_{A,B}$ are the group velocities and the transmission phases associated with the two partial delay lines. Furthermore τ is the total group delay and α_ϕ the total transmission phase temperature coefficient over the temperature range ΔT.

In [1], (2) was approximated by

$$\frac{\alpha_A l_A}{v_{gA}} + \frac{\alpha_B l_B}{v_{gB}} = 0 \tag{3}$$

where $\alpha_{A,B}$ are the transmission phase temperature coefficients of the two partial delay lines.

Obviously the approximation is represented by the fact that phase velocities $v_{\mathrm{ph}\,A,B}$ have been replaced by the corresponding group velocities $v_{gA,B}$. On the basis of that, one might think that

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such an approximation is valued whenever phase and group velocities are not too different from each other. It is the purpose of the present letter to show that this may not be a correct criterion to choose. In fact, even though group and phase velocities differ by a small amount the above approximation may still produce an imperfect temperature compensation (i.e., $\alpha_\phi \neq 0$). In the following, a quantitative evaluation of this phenomenon is presented in conjunction with "unapproximated" design formulas for the TLDL.

Let us use the primed quantities $l'_{A,B}$ to indicate the solutions to the system of (1) and (3). They are found to be [1]:

$$l'_A = \frac{\alpha_B \tau C}{(\alpha_B + |\alpha_A|)(1+\Delta_A)\sqrt{\epsilon_{\text{eff}\,A}}}$$
$$l'_B = \frac{|\alpha_A| \tau C}{(\alpha_B + |\alpha_A|)(1+\Delta_B)\sqrt{\epsilon_{\text{eff}\,B}}} \tag{4}$$

where C is the velocity of light in vacuum, $\Delta_{A,B} = (\omega/2\epsilon_{\text{eff}\,A,B})(\partial\epsilon_{\text{eff}\,A,B}/\partial\omega)$, $\epsilon_{\text{eff}\,A,B}$ are the effective dielectric constants of delay lines (A) and (B), and ω is the angular operation frequency. Note that $1+\Delta$ is the ratio between phase and group velocity along a transmission line.

When $l'_{A,B}$ are substituted into (2), it turns out that

$$\left(\frac{l'_A}{v_{\text{ph}\,A}} + \frac{l'_B}{v_{\text{ph}\,B}}\right)^{-1} \left(\frac{-|\alpha_A| l'_A}{v_{\text{ph}\,A}} + \frac{\alpha_B l'_B}{v_{\text{ph}\,B}}\right) = \alpha^*_\phi \neq 0. \tag{5}$$

The fact that $\alpha^*_\phi \neq 0$ indicates that the approximate lengths $l'_{A,B}$ cause the transmission phase of the device to be thermally unstable. From (5), via (3) and (4) α^*_ϕ may be cast under the form

$$\alpha^*_\phi = \frac{\dfrac{1+\Delta_A}{1+\Delta_B} - 1}{\dfrac{1}{\alpha_B}\dfrac{1+\Delta_A}{1+\Delta_B} + \dfrac{1}{|\alpha_A|}}. \tag{6}$$

From (6) it is recognized that, given two substrate materials with α_A and α_B, it is sufficient that $1+\Delta_A = 1+\Delta_B$ to have $\alpha^*_\phi = 0$. However this condition is unrealistic as, in general, the two MIC's have different frequency dispersion. In a practical situation with $1+\Delta_A \neq 1+\Delta_B$, α^*_ϕ is different from zero and may be calculated by use of (6).

For the case reported in [1] of a composite TLDL with $BaTi_4O_9$ and Al_2O_3 substrates, $|\alpha_A| = 3.9\times10^{-6}/°C$, $\alpha_B = 80.1\times10^{-6}/°C$, $\Delta_A = 1.077$, $\Delta_B = 1.033$, and $\alpha^*_\phi = 0.148\times10^{-6}/°C$. Whether this value of α^*_ϕ is acceptable depends on the system's specs. Note that the measured value of the total transmission phase temperature coefficient α_ϕ reported in [1] is $0.6\pm0.3\times10^{-6}/°C$.

A different situation wherein the approximation is certainly not valid because α^*_ϕ is a considerable portion of the maximum acceptable α_ϕ, however, may be encountered in practice for materials with different physical properties than those reported in [1]. In fact, one may wish to use MIC's with higher negative dielectric constant temperature coefficients and compensate them with MIC's on substrates other than single crystal sapphire (e.g., with ceramic allumina). For instance a type of commercially available $BaTi_4O_9$ exists with $|\alpha_A| = 15\times10^{-6}/°C$ [2]. Using this material together with allumina ($\alpha_B = 50\times10^{-6}/°C$) we built a composite TLDL operating at 14.125 GHz with a group delay time of 16.66 ns, corresponding to a 1-symbol duration in a 120 Mbit/s DC-QPSK signal. In this device the characteristic impedances of the two partial delay lines at zero frequency were 30 Ω and 50 Ω, respectively.

As a consequence $1+\Delta_A/1+\Delta_B = 1.0471$ and $\alpha^*_\phi = 0.538\times10^{-6}/°C$. This value of is 34.3 percent of the spec value of $\alpha_\phi = 1.57\times10^{-6}/°C$ corresponding to a stability of ±2 degrees over a temperature interval of 30°C. Under these circumstances the approximation of [1] is not accurate and the partial delay line lengths l_A and l_B must be calculated using the following "exact" formulas obtained from the (1) and (2)

$$l_A = \frac{\alpha_B \tau C}{[\alpha_B(1+\Delta_A) + |\alpha_A|(1+\Delta_B)]\sqrt{\epsilon_{\text{eff}\,A}}}$$
$$l_B = \frac{|\alpha_A| \tau C}{[\alpha_B(1+\Delta_A) + |\alpha_A|(1+\Delta_B)]\sqrt{\epsilon_{\text{eff}\,B}}}. \tag{7}$$

On the basis of the above results we conclude that, in general, formulas (4) are a good approximation to formulas (7) whenever the substrate material with negative dielectric constant temperature coefficient is highly stable, i.e., α_A is very small. Furthermore, inspection of (6) reveals that this result is pretty much independent of the value of α_B for all practical situations wherein $1+\Delta_A/1+\Delta_B \simeq 1$.

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Periodically Loaded Transmission Lines

JOSE PERINI

***Abstract*—In this paper equations for the transmission parameters of a periodically loaded line are derived in closed form with no restriction on the size, type, or number of discontinuities. The equations also take into consideration any attenuation that may exist on the line.**

Several plots of the input reflection coefficient and transmission coefficient are presented and compared with experimental results. The agreement is very good.

I. Introduction

It is common practice to try to maximize the transfer of power to a load at the end of a long transmission line by minimizing the input reflection coefficient of the system. In this paper it is shown that if the line has periodically distributed discontinuities, which is quite common, then this procedure may lead to quite the opposite result. More recently, interest in the switching characteristics of pulses in such lines has resulted in their being approximately analyzed. [1]. Pulse switching is an important problem in the design of high-speed digital computers. In this paper a closed form solution for the periodically loaded line is obtained. The equations can handle any type of discontinuity

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