I-2. Log-Periodic Phase Difference Circuits

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A schematic diagram of the constant phase difference circuits to be discussed is given in Fig. 1. It consists of two similar two-port log-periodic circuits with their inputs connected in parallel. Ideally, the two-port circuits are designed such that their scattering coefficients are given by

$$S_{11} = S_{22} = 0, \quad S_{12} = S_{21} = \exp\left[-j\left(\frac{2\pi \ln f}{|\ln \tau|} + \varphi\right)\right], \tag{1}$$

where f is the frequency, τ is the design ratio of the log-periodic circuits, and φ is an arbitrary constant. The arbitrary constant φ may be varied by

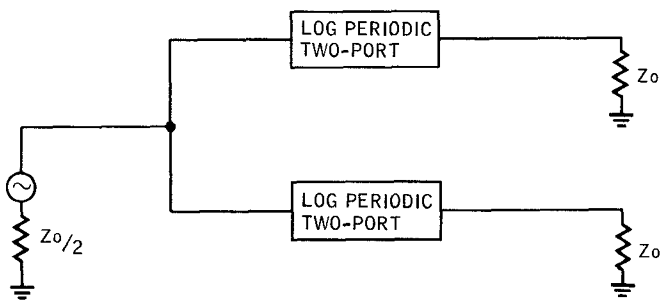

Fig. 1 Constant phase difference circuit.

scaling the log-periodic circuit as discussed in the preceding paper. This allows the two-port circuits to be designed such that the coupled outputs have a phase difference which is independent of frequency. Although the phase difference circuit performs its function in a frequency-independent manner, the circuit is actually dispersive because of the form of S_{12}.

Figure 2 illustrates several forms of two-port log-periodic circuits which may be used as components in the phase difference circuit. The lines may be considered to represent the center conductors of strip transmission line circuits. It will be noticed that the circuits of Figures 2 (b), (c), and (d) have a one-fold symmetry about their center lines. If the coupled strips of Fig. 2 (a) were placed one above the other, then that structure would be symmetric in the sense that a rotation of 180 degrees about the center line of the structure would leave it unchanged. The analysis of symmetrical circuits such as these may be greatly simplified by making use of the normal modes. The normal modes or eigenvectors for circuits with either of the above types of symmetry are the common even (+ +) and odd (+ −) modes.

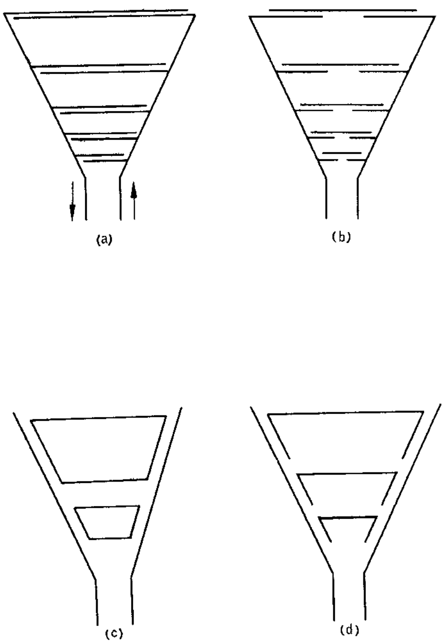

Fig. 2 Two-port log-periodic transmission line networks.

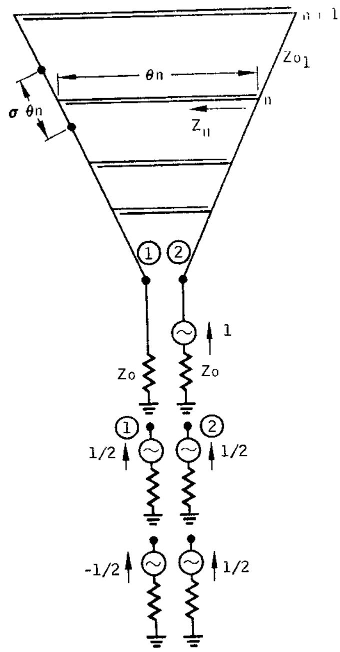

Fig. 3 Normal mode excitation of a log-periodic network.

The scattering matrix for these circuits may be written as

$$S = \frac{1}{2}\begin{bmatrix} \Gamma_e + \Gamma_o & \Gamma_e - \Gamma_o \\ \Gamma_e - \Gamma_o & \Gamma_e + \Gamma_o \end{bmatrix}, \tag{2}$$

where Γ_e and Γ_o are the even and odd mode input reflection coefficients which are normalized to Z_o.

Figure 3 illustrates the even and odd mode excitation of the network of Fig. 2 (a). Now for circuits with one-fold symmetry, magnetic or electric walls may be inserted along the center line for the even and odd modes of excitation, respectively. Thus, Γ_e or Γ_o is the input reflection coefficient of a single transmission line loaded in a log-periodic manner, as discussed in the preceding paper. Γ_e and Γ_o may be calculated for given circuit parameters by straightforward, but tedious matrix multiplication. If the circuit is designed properly (see preceding paper), then ideally the phases of Γ_e and Γ_o are linearly proportional to log f. The analysis of circuits with 180° symmetry can also be reduced to that of a single loaded transmission line, but the computation of the loading impedances is not as straightforward.

The characteristics of a two-port circuit will be outlined for the circuit of Fig. 3. It consists of two radial transmission lines of characteristic impedance Z_{01}, interconnected at log-periodic intervals by coupled transmission lines with a coupling coefficient c and of electrical length θ_n. The mean of the even and odd mode characteristic impedances of the coupled lines is denoted by Z_{02}. The spacing of the coupling sections is determined by σ. The shunt impedances, Z_{ne} and Z_{no}, which the coupling sections present to the radial transmission line are illustrated in Fig. 4. Effectively, the coupling between the shunt stubs makes the effective stub length different for the even and odd modes. Thus, it is to be expected that Γ_e and Γ_o will not, in general, be in phase, since the active or short circuit region for the even mode is at a greater distance from the input than that for the odd mode.

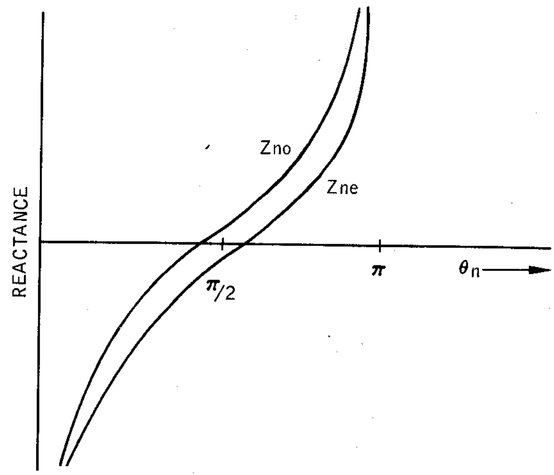

Fig. 4 Normal mode shunt loading impedance.

The characteristic impedances for the two modes are also different and are given by

$$Z_{oe} = Z_{01}\left[1 + \frac{Z_{01}k}{2Z_{02}\sigma}\right]^{-1/2},$$
$$Z_{oo} = Z_{01}\left[1 + \frac{Z_{01}}{2Z_{02}\sigma k}\right]^{-1/2}. \qquad (3)$$

where $k = \sqrt{(1-c)/(1+c)}$. The solid curves of Fig. 5 illustrate the ideal variation of Γ_e and Γ_o when normalized to Z_{oe} and Z_{oo} respectively. The coupling coefficient c may be chosen to provide a phase difference between Γ_e and Γ_o of 180°. When Γ_e and Γ_o are normalized to $Z_o = \sqrt{Z_{oe}Z_{oo}}$, the dashed curves are obtained. Calculation of the scattering coefficients gives approximately:

$$S_{11} = S_{22} \approx 0$$
$$S_{12} = S_{21} \approx \exp -j\left[\frac{2\pi \ln f}{|\ln \tau|} + \varphi + \frac{1}{2}\left(\ln \frac{Z_{oe}}{Z_{oo}}\right)\sin\left(\frac{2\pi \ln f}{|\ln \tau|} + \varphi\right)\right]. \qquad (4)$$

In order to keep the ripple or phase deviation from linear small, the ratio Z_{oe}/Z_{oo} should be near unity.

The results of extensive computer investigations of two-port circuits will be given in the form of design curves and concepts. Of those studied, the circuit of Fig. 2 (b) is best suited for the phase difference circuit applica-

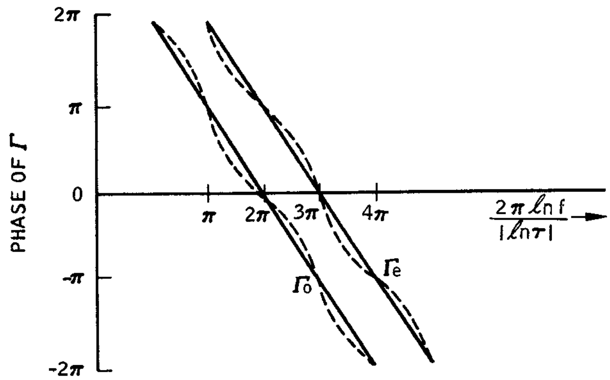

Fig. 5. Normal mode reflection coefficient phase characteristics.

tion. Theoretical results predict that it is possible to make $S_{11} < .03$, and the phase ripple of S_{12} less than 1.5 degrees.

Experimental results for a balun (i.e., a phase difference of 180°) designed for the frequency band of .5 to 5 Gc are given. The circuit is contained in a triangular-shaped low-loss dielectric loaded package approximately .5 inches thick, 10 inches wide and 24 inches long. By modifying the design and increasing the cost, a smaller package could be obtained. Printed circuit versions of the two-port structure of Fig. 2 (b) were utilized.