BROADBAND SWITCHED-BIT PHASE SHIFTER USING ALL-PASS NETWORKS

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ABSTRACT

A method of broadband phase shifting, utilizing an unbalanced all-pass network topology, has been developed. By taking advantage of the intrinsically matched characteristics of these networks, it is shown that multi-octave response can be achieved by cascading of two or more similar networks. Utilizing this approach, an octave band 4-bit phase shifter has been realized, having less than 18° total phase error and better than -30 dBc carrier suppression when operated as a frequency translator.

INTRODUCTION

A number of broadband phase shifting techniques at microwave frequencies have been developed. The most widely used include resistively controlled I-Q modulators [1], [2] and various reactive networks in a switched bit arrangement [3], [4]. The former method is preferable for low power high resolution requirements, while the latter is chosen when high speed, and/or high power handling are the primary requisites.

This paper reports on the development of a new approach to broadband phase shifter design using all-pass networks. By definition, all-pass networks are intrinsically matched, thus simplifying the synthesis of its phase response over large bandwidths as compared to hi-low pass, and loaded line phase shifter designs. By cascading several all-pass networks, it becomes relatively straight-forward to synthesize single- and multi- octave responses.

THEORY

Reactive all-pass networks have been analyzed by [5], [6], in balanced configuration, and require the use of a balun to achieve common input/output ground [7]. However, for stripline and microstrip applications an unbalanced configuration with a common ground would be desirable to permit integration with other components. Such a configuration is shown in the two all-pass circuits of Fig. 1. Both the series L and the series C configurations have equivalent phase responses. In the following analysis, let us express the associated reactance $x=\omega L/Z_0$ and susceptance $b=\omega C Z_o$ in terms of a normalized frequency, Ω, and a normalized impedance, z, where

$$z = \frac{1}{Z_o}\sqrt{\frac{L}{C}}\,, \quad \Omega = \frac{\omega}{\omega_o}\,, \quad \text{and} \quad \omega_o = \frac{1}{\sqrt{LC}}\,,$$

where we define ω_o as the transition frequency. Below the transition frequency the phase response approaches that of a low-pass filter, while above it, that of a high-pass filter. These equations lead to the relationships:

$$x = \Omega z \quad \text{and} \quad b = \frac{\Omega}{z}\,.$$

By applying the principles of a symmetrical lossless network [8], it can be shown that the S-parameters s_{11} and s_{21} of both configurations, in terms of even- and odd- mode reflection coefficients, Γ_e and Γ_o, are respectively

$$s_{11} = \frac{1}{2}(\Gamma_e + \Gamma_o) \quad \text{and} \quad s_{21} = \frac{1}{2}(\Gamma_e - \Gamma_o)\,,$$

$$\Gamma_e = \frac{jWz-1}{jWz+1} \quad \text{and} \quad \Gamma_o = \frac{1-jW/z}{1+jW/z}\,,$$

where $W=\Omega-1/\Omega$.

It is evident that if $z=1$, i.e. $L/C=Z_o^2$, then $s_{11}=0$ and the network is matched at all frequencies. It follows that the transmission coefficient s_{21} has an unity magnitude with a phase response

$$\psi = \pi - 2\tan^{-1}(\Omega - \frac{1}{\Omega}) \qquad (1)$$

It is readily seen from this equation, that the rate of change $d\psi/d\Omega$ at transition frequency, $\Omega=1$, is equal to -4. In other methods, such as [3], [4] and [7], the equivalent phase response is a function of $\pi-2\tan^{-1}\Omega$, with an associated rate of change of phase, $d\psi/d\Omega$, equal to -1 at $\Omega=1$. The four fold improvement of this method leads to phase differential circuits with significantly reduced reactance ratios.

SWITCHED BIT DESIGN CONCEPT

Consider now a pair of all-pass networks, shown in Fig. 2 as networks A and B, with associated transition frequencies ω_A and ω_B. Let us specify ω_m as the geometric mean of ω_A and ω_B,

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i.e $\omega_m^2=\omega_A\omega_B$, introduce a design parameter $p^2=\omega_A/\omega_B$ and redefine Ω as ω/ω_m. Using Eq. (1), yields the insertion phase of the two networks

$$\psi_A=\pi-2\tan^{-1}(\Omega p-\frac{1}{\Omega p}) \quad \text{and} \quad \psi_B=\pi-2\tan^{-1}(\frac{\Omega}{p}-\frac{p}{\Omega}),$$

leading to a differential phase expression

$$\phi=2\{\tan^{-1}(\Omega p-\frac{1}{\Omega p})-\tan^{-1}(\frac{\Omega}{p}-\frac{p}{\Omega})\}. \quad (2)$$

The value of the parameter p is related to the differential phase at the center frequency $\phi_m\equiv\phi\,|_{\Omega=1}$. Therefore, to synthesize a phase shifter with a nominal phase ϕ_m, the value of p is computed from

$$p=\frac{1}{2}\tan(\frac{\phi_m}{4})+\sqrt{1+\frac{1}{4}\tan^2(\frac{\phi_m}{4})} \quad (3)$$

which is obtained by solving Eq. (2) for p, with $\Omega=1$. Given that ω_m is the center frequency of the operating bandwidth, the elements of the phase shifter are determined as follows:

$$L_A=\frac{pZ_o}{\omega_m}, \quad L_B=\frac{Z_o}{p\omega_m}$$
$$C_A=\frac{p}{Z_o\omega_m} \text{ and } C_B=\frac{1}{pZ_o\omega_m}. \quad (4)$$

Table 1 gives values of p for selected phases ϕ_m.

Phase ϕ_m:	22.5°	45°	90°	180°
p:	1.050	1.104	1.228	1.618

Table 1: Values of a Design Parameter p for a Single Section All-pass Phase Shifter

Note should be taken of the value of p, which defines the ratio of element values in networks A and B. Although it is not directly evident from the above treatment, it can be shown that low values of p permit more design flexibility in MIC construction. The other methods of reactive phase shifting, e.g hi-low pass and balanced all-pass require equivalently larger ratios for the same phase shift.

THEORETICAL BANDWIDTH

If the operating bandwidth B is defined as equal to ω_2/ω_1, the theoretical peak to peak phase flatness $\Delta\phi$ is obtained from Eg. (2)

$$\Delta\phi=\phi(\Omega=\sqrt{B})-\phi_o.$$

Examples of the peak-to-peak flatness are given in Table 2.

Phase ϕ_m:	22.5°	45°	90°	180°
B = 1.2:1	0.6°	1.2°	2.0°	1.9°
B = 1.6:1	3.6°	6.9°	12.1°	12.5°
B = 2.0:1	6.5°	12.8°	23.0°	26.5°

Table 2: Peak-to-peak Phase Flatness of a Single Section All-pass Phase Shifter

PRACTICAL CONSIDERATIONS

The phase response given by Eq. (2) assumes that the values of L and C are chosen in accordance with Eq. (4). In reality, it is necessary to allow for some tolerance on these theoretical values. It can be shown that if the ratios $L_A/L_B = C_A/C_B$ are equal to p^2, then the variation in the phase response due to their absolute value is negligible to the first order, except at the band edges. For example, if the circuit capacitances have a tolerance $1\pm k$, then the useful bandwidth is reduced by a factor 1-k. Thus if k=0.1 and the required bandwidth is 1.5:1, then the design bandwidth should be 1.65:1. It should be noted that although the network is no longer matched under these conditions, the degree of mismatch is comparatively small (max VSWR$\approx 1+k$).

Additional analysis also show that the insertion loss due to circuit losses peaks at the transition frequencies. Unless these losses are minimized, this effect can give rise to a undesirable amplitude modulation at ω_m/p and $\omega_m p$ frequencies.

Realizing the all-pass phase shifter by means of MIC techniques, where the L's and C's are "printed" on a common substrate, addresses the design issues raised above. The inductive components are printed as short high impedance lines and the capacitors are formed by means of low impedance stubs, with the ground printed on the reverse side of the substrate. To form a series capacitor, the ground has to "float" for a proper design. Once the artwork dimensions are correctly established, production tolerances due to substrate thickness, etching factors and misregistration of top and bottom artwork, do not change the ratios L_A/L_B and C_A/C_B to a first order - a condition required for uniform performance. Further, by selecting a low-loss substrate and conductors with sufficient thickness compared to skin depth, the circuit loss and therefore the amplitude modulation problem is minimized.

OCTAVE BAND DESIGN

It is evident from Table 1 that a single section design for a 4-bit phase shifter has a phase flatness varying from about $\pm3°$ to about $\pm13°$ over an octave band. Taking advantage of the inherently matched performance of the all-pass network, one can readily synthesize a broader band phase shifter by cascading two or more such networks with offset center frequencies. In particular, consider the two stage design shown in Fig. 3, where networks A2 and B2 are cascaded with networks A1 and B1 respectively. It can be shown that the phase response, is

$$\frac{\phi(\Omega)}{2} = \tan^{-1}\left(\frac{\Omega}{pq} - \frac{pq}{\Omega}\right) - \tan^{-1}\left(\frac{\Omega p}{q} - \frac{q}{\Omega p}\right) + \tan^{-1}\left(\frac{q\Omega}{p} - \frac{p}{q\Omega}\right) - \tan^{-1}\left(pq\Omega - \frac{1}{pq\Omega}\right) \quad (5)$$

where q is expressed in terms of transition frequencies

$$q = \sqrt{\frac{\omega_{A2}}{\omega_{A1}}} = \sqrt{\frac{\omega_{B2}}{\omega_{B1}}} .$$

In Table 3, values of p and q are given for a 3:1 band design. The theoretical peak-to-peak phase flatness, computed according to Eq. (5), is listed in Table 4 for several values of phase and bandwidth.

Phase:	22.5°	45°	90°	180°
p:	1.040	1.082	1.174	1.413
q:	1.543	1.549	1.569	1.667

Table 3: Values of p and q for a 3:1 band design.

Phase:	22.5°	45°	90°	180°
B = 3:1	2°	4°	7°	7°
B = 4:1	4°	7°	12°	13°
B = 6:1	7°	13°	22°	25°

Table 4: Theoretical Peak-to-peak Phase Flatness of a Two Section All-pass Phase Shifter

MEASURED DATA

A prototype model of a 4-bit phase shifter, operating in the 4-8 GHz band has been assembled and is shown in Fig. 4. A 2.5:1 design bandwidth was actually used, for reasons discussed above. The all-pass networks were printed on a 0.005" thick alumina substrate. A series C all-pass configuration has been used. Measured room temperature data are summarized in Figs. 5-8. The relatively good carrier and sideband suppression data in Fig. 8 is evidence that the phase error of each bit is relatively small, in most cases less than ±6°. (A phase error of ±6° in any bit with phase shift $360°/2^n$, where n=1,2,3 or 4, results in about a -30 dBc side level at the 2^n harmonic of the translation frequency.) This device, which uses a high speed switch driver, has a translation bandwidth of several megahertz. The phase-to-phase switching time is less than 25 nsec. Over a temperature range of -55°C to 95°C, the performance degradation is small (within ±3°), as is expected from the balanced nature of a switched bit configuration.

CONCLUSION

It has been demonstrated that all-pass networks are versatile tools for a broadband phase shift design. In theory, one can realize decade-wide bandwidth responses, with only three or four stages. In practice, however, one has to deal with the problem of maintaining the non-distributive nature of the components over such wide frequency ranges. Although, in the current effort, MIC design techniques were employed, monolithic MMIC technology, with the capability of realizing wideband "lumped" reactive components, should permit realization of the full potential of this broad-band method.

ACKNOWLEDGEMENT

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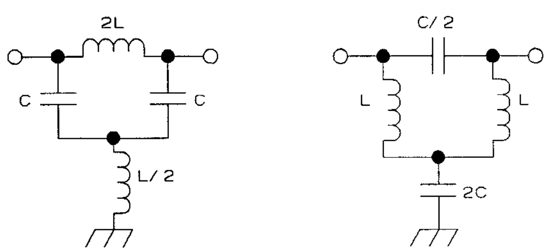

Figure 1. Series L and Series C configured All-pass Networks

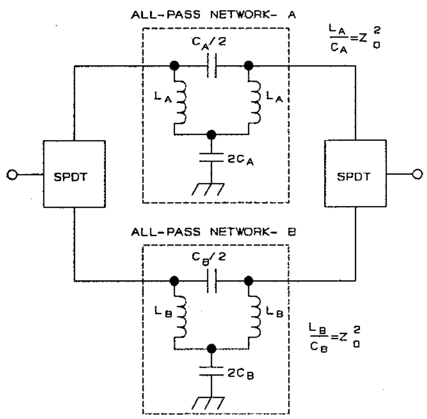

Figure 2. Switched Bit Phase Shifter Using Two All-pass Networks

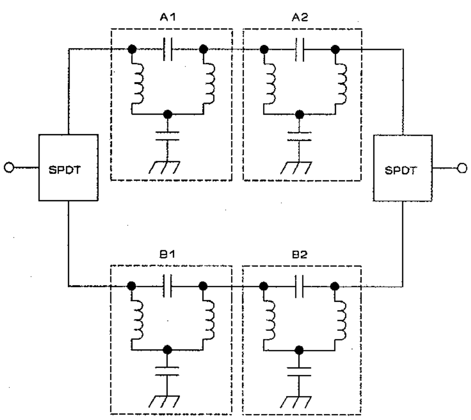

Figure 3. Two Stage Phase Shifter

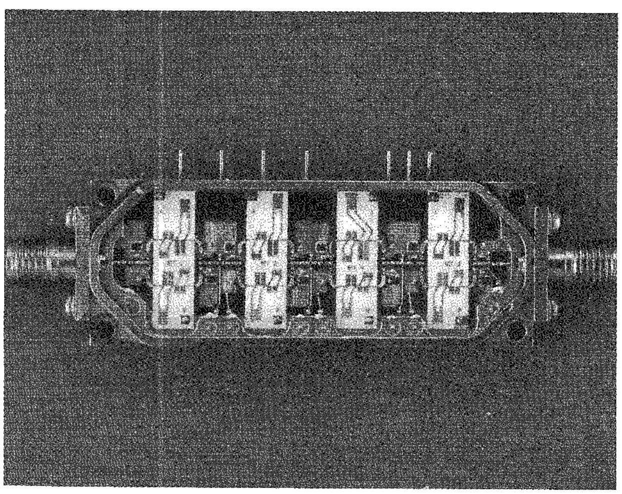

Figure 4. Rf Detail of a 4-bit Phase Shifter

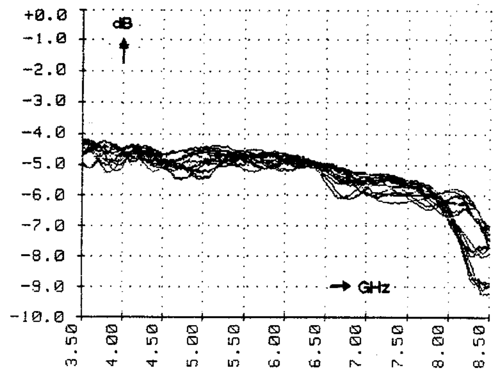

Figure 5. Measured Ins. Loss Variation at All Phase States

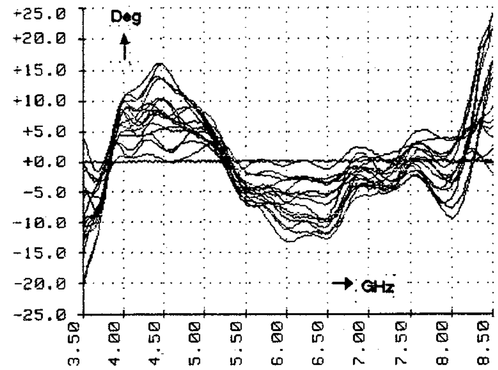

Figure 6. Measured Phase Error at All Phase States

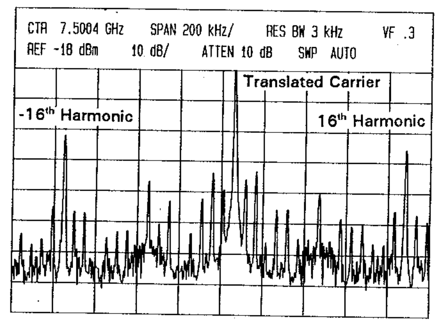

Figure 7. Typical Translation Spectrum

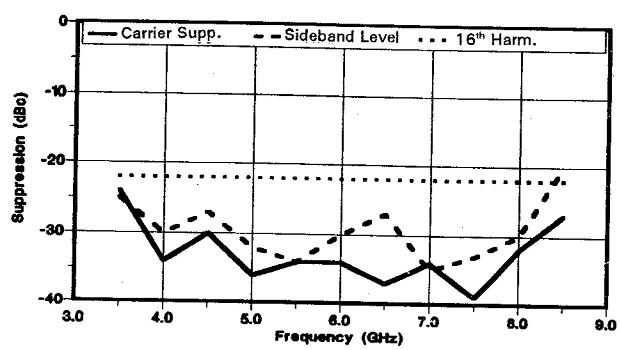

Figure 8. Carrier and Sideband Data