BROADBAND PHASE INVARIANT ATTENUATOR

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ABSTRACT

A 6-18 GHz variable attenuator has been realized, with less then 5° phase error over 32 dB attenuation range. The approach utilizes the bi-phase nature of a reflection from a resistive load. Using quadrature couplers, it is shown that the first order errors are reduced to fourth order terms, resulting in significant phase error improvement.

INTRODUCTION

In an ideal variable attenuator phase response should remain unchanged over its attenuation range. This property is of particular importance in a growing number of broad-band phase-sensitive signal processing applications. At microwave frequencies the phase performance of conventional attenuator designs is limited by presence of inherent parasitics. A number of published works deal with the problem by using compensation techniques, [1], [2] or minimizing parasitics [3]. The improvements are limited to frequencies below X-band and/or narrow frequency ranges. The approach presented here utilizes an error cancellation scheme that is based on an earlier work on bi-phase modulators [4].

In general, a bi-phase modulator is a control device with two phase states for a given attenuation level. The phase difference of the two states is nominally 180°. It will be shown, that for a non-ideal device, the transfer functions of the two states can be expressed in the form of $(1+\mu)$ and $(-1+\mu)$ respectively, where μ is representing a complex error term to be defined below. A large reduction of the error term is achieved when a pair of identical modulators, each in opposite phase states, is cascaded, i.e $(1+\mu)(-1+\mu) = -1+\mu^2$. This mathematical relationship represents the essence of this approach.

BI-PHASE MODULATOR

As shown in Fig. 1, the bi-phase modulator is an arrangement of a pair of reflective attenuators connected by means of input and output quadrature couplers. Each reflective attenuator consists of a 3 dB quadrature coupler and a pair of voltage controlled PIN diodes at the isolated ports.

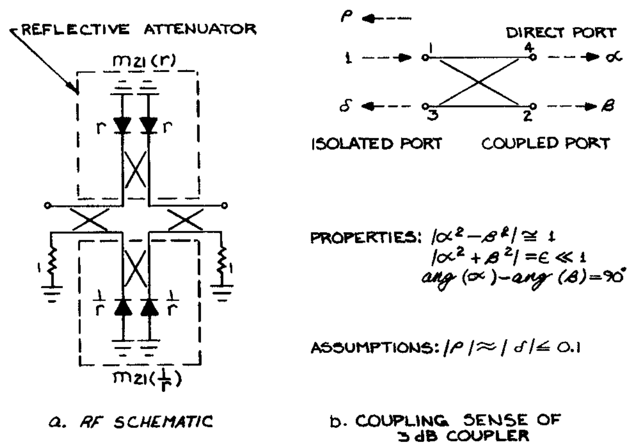

Figure 1. Balanced Modulator

The nominal response, m_o, in terms of normalized diode resistance r, is

$$m_o(r) \quad \propto \quad \frac{r-1}{r+1} \tag{1}$$

and has the property of an ideal bi-phase modulator, i.e. bi-phase response with constant phase for any value of r<1 and infinite attenuation range:

$$m_o(r) = -m_o(\frac{1}{r}) \qquad (2)$$

$$|m_o(0)| = 1 \quad \text{and} \quad m_o(1) = 0 \qquad (3)$$

In practice various parasitics are present and include (a) junction capacitance, series resistance, and inductance associated with the PIN diodes, (b) the finite directivity and non-zero reflection coefficients of the coupler, and (c) errors due to source and load mismatch. Their combined effect can be summarized by the following concept.

Let $m_{21}(r)$ and $m_{21}(\frac{1}{r})$ be two phase states of a typical bi-phase modulator. Then it is possible to relate them to the ideal response $m_o(r)$:

$$m_{21}(r) = m_o(r) + \mu(r) \qquad (4)$$

$$m_{21}(\frac{1}{r}) = -m_o(r) + \mu(r) \qquad (5)$$

where error coefficient $\mu(r)$ has been introduced, having the following properties:

$$\mu(r)=\mu(\frac{1}{r}) \quad \text{and} \quad \mu(1)=m_{21}(1).$$

Note that $\mu(1)=m_{21}(1)$ is a "null" point at which the normalizing resistance of the diode resistance r is defined. Depending on parasitics and coupler's even and odd mode impedances and phase velocities, this normalizing resistance is not necessarily the characteristic impedance, i.e. 50 ohms. Further more if $|m21(r)| = |m_{21}(\frac{1}{r})|$ then $\mu(r)$ and $m_o(r)$ are in quadrature, as is evident from eqns. (4) and (5).

In the balanced modulator arrangement, Fig. 1, it can be shown [4], that its response is

$$m'_{21}(r) = \alpha^2 m_{21}(r) + \beta^2 m_{21}(\frac{1}{r}) \qquad (6)$$

where the coupling coefficients α and β are defined in Fig. 1b. In terms of eqns. (4), (5) and the coupling balance $\varepsilon=|\alpha^2+\beta^2|<<1$, eqn. (6) above yields the transfer functions of the two phase states

$$m'_{21}(r) = A\{ m_o(r) + \varepsilon\mu(r) \} \qquad (7)$$

and

$$m'_{21}(\frac{1}{r}) = A\{-m_o(r) + \varepsilon\mu(r) \} \qquad (8)$$

where the error term, now represented by $\varepsilon\mu(r)$, has been effectively reduced by an order of magnitude, and $A=\alpha^2-\beta^2$ is an insertion phase constant. It is desirable that both terms in eqn. (7) are in quadrature. In practice this condition is essentially achieved over wide band by proper choice of bias voltages such that as to minimize the amplitude error between $m'_{21}(r)$ and $m'_{21}(\frac{1}{r})$.

PHASE INVARIANT ATTENUATOR

Cascading two balanced modulators with opposite phase states, i.e multiplying equations (7) and (8), yields

$$M_{21}(r) = m'_{21}(r)\, m'_{21}(\frac{1}{r}) =$$

$$= A^2\{ -m_o^2(r) + \varepsilon^2\mu^2(r) \} \qquad (9)$$

further reducing the error term by two orders of magnitude. In addition, if the quadrature relationship between terms in eqn. (7) is established as discussed above, then the error term in eqn. (9) is in phase with $-m_o^2(r)$ and, in fact, does not contribute to phase error.

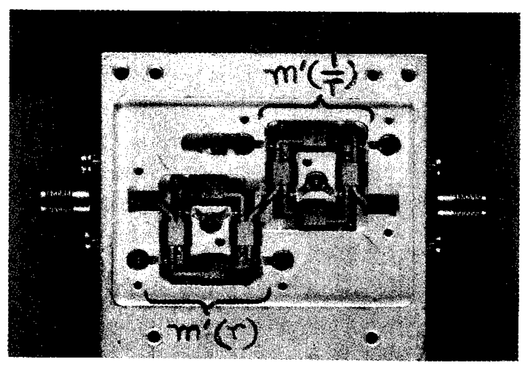

Figure 2. Rf Detail of Phase Invariant Attenuator

To illustrate the improvement factor, assume $|\mu|\leq 0.1$ and $|\varepsilon|\leq 0.1$ (typical for a 3:1 bandwidth coupler with 20 dB directivity). At 20 dB attenuation, $m_o(r)=0.1$, the comparison of the worst case phase uncertainties (i.e when error term is in quadrature) yields:

(i) from eqn.(4) $\quad$ ang{ $m_{21}(r)/m_{21}(0)$ } = 39°

and

(ii) from eqn.(9) $\quad$ ang{ $M_{21}(r)/M_{21}(0)$ } = 0.05°,

an improvement factor of almost three orders of magnitude. In practice, the improvement factor is limited by the ability to ensure electrical uniformity of the PIN diodes, by bias circuit perturbations, and non-ideal couplers.

A rf assembly is shown in Fig. 2. Its response, given in Fig. 3, shows that in the 6-18 GHz band the actual phase error at 20 dB is less than 3° and at 32 dB, less then 5°. Insertion and return losses are also shown.

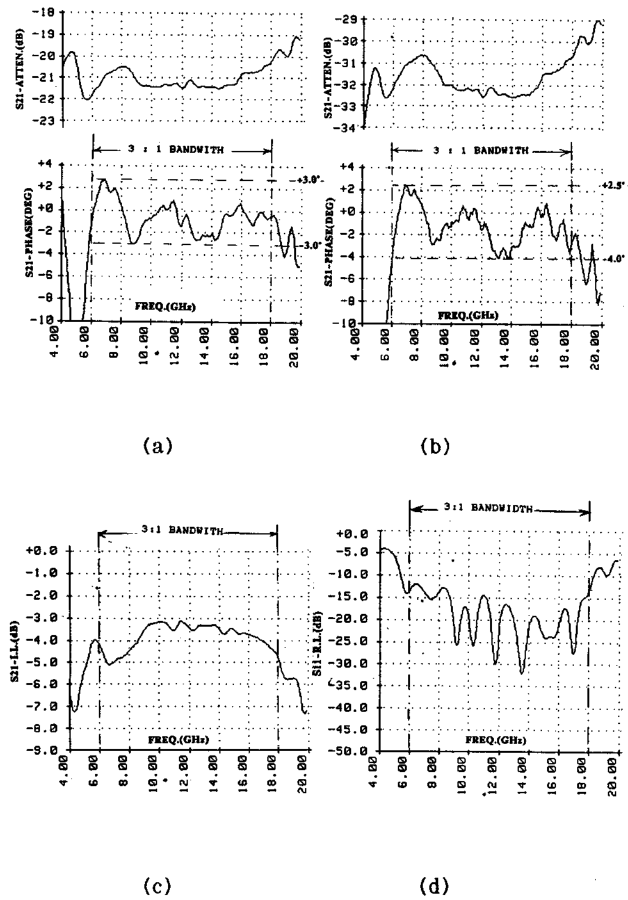

Figure 3. Rf Response of Phase Invariant Attenuator
(a) Phase and Amplitude at 20 dB, (b) at 32 dB
(c) Insertion Loss, (d) Return Loss.

BIASING REQUIREMENTS

Note that in eqn. (6), $m_{21}(r)$ and $m_{21}(\frac{1}{r})$ represent two separate, but electrically identical, reflective attenuators. Therefore two separate bias voltage functions, $V(r)$ and $V(\frac{1}{r})$, are required, which, though related, each of them have different functions of attenuation and temperature. If these voltages are generated independently, the attenuator may be prone to switching anomalies and temperature instabilities. However, a single temperature compensated, biasing control can be realized as shown by the following analysis.

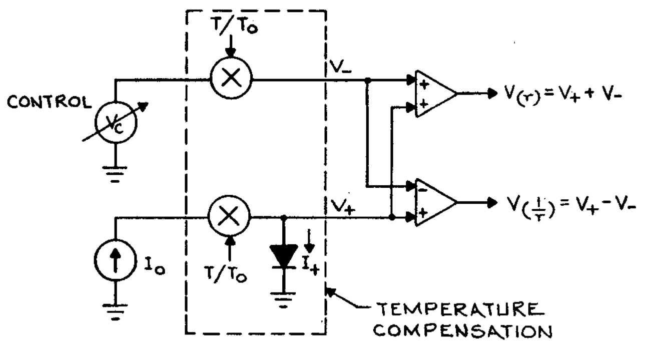

Figure 4. Biasing Scheme

Using the PN junction approximation, $I = I_s(e^{V/V_T} - 1)$, where I_s and V_T are temperature dependent constants, then, in terms of normalized diode resistance r,

$$V(r) = V_T \ln\left(\frac{T I_o}{T_o I_s} \frac{1}{r} \right) \tag{10}$$

where $I_o = I(T=T_o, r=1) >> I_s$ and $T_o = 293°K$. From that it follows

$$V_- = \frac{1}{2}\{V(r) - V(\tfrac{1}{r})\} = \frac{T}{T_o} V_{T_o} \ln(\tfrac{1}{r}) \tag{11}$$

and

$$V_+ = \frac{1}{2}\{V(r) + V(\tfrac{1}{r})\} = V_T \ln\{\frac{T I_o}{T_o I_s}\} \tag{12}$$

where voltages V_- and V_+ represent the average difference and sum of the bias voltages $V(r)$ and $V(\frac{1}{r})$ respectively. Note that $V_+ = V(r=1)$ is a condition for the maximum attenuation of the attenuator. This voltage can be obtained from a reference diode, biased at a current I_+ (substitute eqn. (12) into I-V diode law):

$$I_+ = \frac{T}{T_o} I_o \tag{13}$$

The bias requirements are thus defined by eqns. (11) and (13), and are a simple function of temperature, independent of diode parameter I_s. This biasing scheme is schematically shown in Fig. 4, where the attenuation control voltage is defined as $V_c = V_{-}(T_o)$. In terms of the magnitude of S_{21}, where $S_{21} = M_{21}(r)/M_{21}(r_1)$ and r_1 is diode resistance at insertion loss (cannot be zero for practical reasons) V_c is

$$V_c = 2V_{T_o} \tanh^{-1}\{|S_{21}|^{1/2} \tanh(\frac{V_{c1}}{2V_{T_o}})\} \quad (14)$$

where V_{c1} is the control voltage at insertion loss and $|M_{21}| \approx |m_o|^2$ has been assumed.

The control function is plotted in Fig. 5 together with actual measured data and shows a good fit over wide a temperature range. The phase error over temperature is given in Fig. 6.

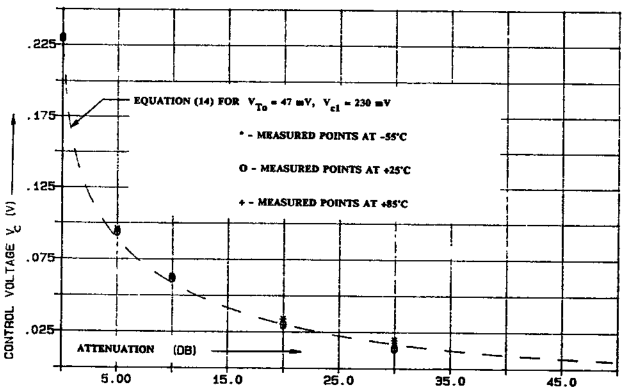

Figure 5. Control Voltage vs. Attenuation

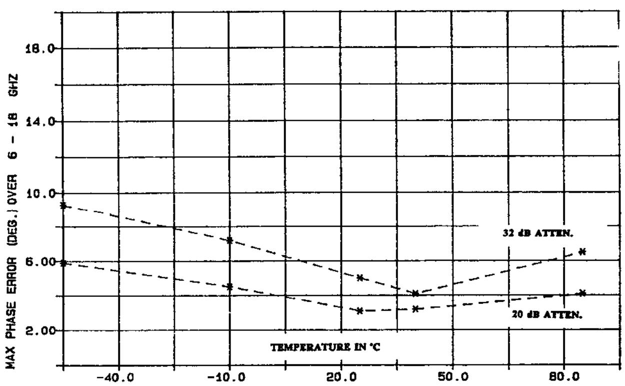

Figure 6. Max. Phase Error over Temperature

CONCLUSION

It has been shown that a Ku-band, temperature independent, low phase-shift variable attenuator can be realized by utilizing the bi-phase nature of a reflection from a resistive load and a quadrature property of symmetrical couplers. The attenuator is not limited to 3:1 bandwidth, as evident from Fig. 2. Likewise, phase error gracefully degrades at higher attenuation (15° at 45 dB), although attenuation becomes very sensitive to control voltage V_c , as can be deduced from eqn. (14). The speed of the attenuator is limited essentially by driver circuitry, and better then 150 nsec within 0.5 dB of insertion loss has been achieved.

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