MONOLITHIC KA-BAND 180-DEGREE ANALOG PHASE SHIFTER EMPLOYING HEMT-BASED VARACTOR DIODES

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Abstract
The design and performance of an MMIC 180 degree analogue phase shifter, with flat relative phase-shift, over a wide bandwidth, is presented. The phase shifter, which measures only 1.3 x 1.3mm, employs two non-identical cascaded stages to achieve 180° tuning range over a wide bandwidth. Reflection-type topology using Lange couplers is employed with two sets of back-to-back varactor diodes connected to the coupled and direct ports. The varactor diodes are realised from the standard pHEMT device geometry to maintain compatibility with standard foundry processing. It will be seen that the design can also be used to achieve direct carrier BPSK modulation, used in order to reduce hardware complexity and cost. The modulator operation will be demonstrated at 38 GHz with a 2 Mbit/s data rate.

Introduction
Most of today’s microwave applications require variable control devices, such as phase shifters and attenuators, with good dynamic range and precision at reasonable prices. When compared to purely digital implementation, analog control devices can provide many more significant advantages for signal processing in communications applications [1]. For example, they require only one control wire per device, they require almost no control power with a passive reflection topology incorporating varactor diodes or cold-FETs, they do not suffer from quantization errors and they can make much more use of expensive chip space with a reflection topology that uses active circulators. Hence analog control devices appear ideal for large adaptive phased arrays and other high performance applications. The synthesis and realisation of a wideband analog phase shifter, which employs non-identical stages, and can be employed as a direct BPSK modulator, is presented in this paper.

BPSK
Binary phase-shift keying (BPSK) is frequently used for millimetre-wave digital communications since it is robust and spectral efficiency is not critical. A BPSK modulator alters the phase of the input RF carrier signal from a reference phase of 0° to an opposite phase of 180° in correspondence to the modulating control signal. For an ideal BPSK modulator, the phase shift between the two states (ON and OFF) must be exactly 180°. In practice however, this is not achievable and hence leads to phase error characteristics. Modulating directly at carrier frequency has been shown before to be an attractive means of reducing the hardware complexity and cost for wireless applications [2]. If a conventional microwave mixer is used to upconvert the modulated signal from a low frequency to the transmission frequency, a complex chain of mixers, filters and amplifiers must be employed. However, for millimeter-wave applications, where cost remains a major factor, direct modulation is an attractive means of realising low cost VSAT transmitters. A simple direct BPSK modulator can be realised using a 180 degree phase shifter.

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Phase shifter

In the case of analogue control most phase shifters are based on the continuous voltage-control of a varactor diode's capacitance. Dual-gate FETs have also been used as analogue phase shifters, but their performance is somewhat limited. The reflection type phase shifter (RTPS) was first introduced by Hardin [3], and almost all analog phase shifters employ this topology [4,5]. With reference to Fig. 1, it can be seen that identical reflection terminations are connected to the direct and coupled ports of the 50Ω Lange couplers, which is employed to achieve constant insertion loss and good port matches. The other two ports form the circuit's input and output. When the reflection terminations employ voltage dependent reactance, such as an active inductor or voltage-dependent varactor devices, the relative phase difference between the output voltage wave vector and the input voltage wave vector can be varied electronically. This will therefore result in a shift in the phase angle of the total voltage transmission coefficient.

With varactor diodes connected to the coupled and direct ports of the Lange coupler, the signal from the output port is that reflected from the two varactors. If the varactors represent ideal variable capacitances, then the magnitude of their reflection coefficient is fixed at unity, but the phase varies from zero (no capacitance) to -180° (maximum capacitance-virtually short circuit). Thus the phase of the output signal can be varied continuously using the bias voltage on the varactor diodes. For wideband applications, small size MMIC 180 degree phase shifters have been reported [6], but their phase error characteristics were shown to deviate by as much as 25° over 6 to 18 GHz, and that is considered to be too large. The single-stage phase shifter illustrated in Fig. 1, with the reflection terminations employing voltage-dependent varactor diodes, is not an ideal phase shifter since the phase error performance is also rather poor across a wide bandwidth. The phase shift range is limited in practice, and is determined by the capacitance tuning ratio of the diodes. If the terminations are more than simply diodes (eg. some inductance is introduced) then the control range can be improved, but the bandwidth of the phase shifter is then rather limited. In recent years, a number of variations to the basic analogue reflection type topology have been reported in monolithic form without significant improvements in their narrow bandwidth performance [7]. The phase shifter presented in this paper employs a cascade of two carefully matched non-identical phase shifters in order to overcome the narrow bandwidth problem and achieve the 180° tuning range required, a topology referred to as the Cascaded Match Reflection Type Phase Shifter (CMRTPS) [8].

Active Device Characterisation

The varactor diodes can simply be implemented by connecting together the drain and source terminations of a standard pHEMT to form the cathode, and the gate alone forming the anode. The bias potential is applied across the cathode and anode. Fig. 2 illustrates the equivalent circuit model of an Interdigitated Planar Schottky Varactor diode (IPSVD). The active layer under the gate, with length L and width W can be regarded as a distributed R-C network, where R is the conducting channel resistance and C is the depletion region capacitance.

The capacitance of the Schottky junction varies approximately to the standard expression for an abrupt junction:

\[ C(V) = C_0 / [1 + (V/V_0)^{\gamma}] \]  \hspace{1cm} (1)

Where \( \gamma \geq 1 \),

- \( C_0 \) is the zero bias capacitance of the device,
- \( V \) is the applied bias voltage to the device, and
- \( V_0 \) is the built-in potential of the Schottky junction.

Applying reverse bias voltage to the device will cause the depletion layer capacitance to decrease since \( \gamma \geq 1 \). The depletion layer charge capacity should be increased through a high density of doping profiles in order to achieve a large tuning ratio. Assuming the gate length to be quite short (less than 0.5μm), the varactor diode can be characterised as a variable capacitor according to its applied bias voltage, with associated parasitic elements.

With reference to Fig. 2 it can be seen that the equivalent circuit model has three bias dependent elements. These are \( C(V) \), the junction capacitance, \( R_j(V) \), the junction leakage resistance, and
The series resistance, when a high reverse bias voltage is applied to the device, the depletion layer extends into the semi-insulating substrate. Hence C(V) is at its minimum value, and \( R_s(V) \) is an open circuit. However, when the reverse bias voltage is lowered, the capacitance of the diode is the total sum of the depletion layer capacitance and the extrinsic capacitance, while \( R_s(V) \) has a low resistance value and \( R_j(V) \) is still an open circuit. With forward bias voltage applied to the device, \( C_j(V) \) is at its maximum value, while both \( R_s(V) \) and \( R_j(V) \) are at a minimum.

**Circuit Synthesis**

The basic principle of operation of the phase shifter illustrated in Fig. 1 is that the \( S_{11} \) of the reflection termination is transformed into the \( S_{21} \) of the circuit. The \( S_{11} \) magnitude will be fixed at unity when employing a pure variable capacitance, while the phase varies continuously through a range determined by the capacitance tuning ratio of the varactor. The incident signal at the input port of the Lange coupler is split equally to the coupled and direct ports, with a 90 degree phase difference between them. A further 3dB split and 90 degree phase difference is introduced when each of the reflected waves passes back through the coupler. When the two reflected waves are superimposed at the input port they are anti-phase and thus cancel. At the output port of the coupler the two reflected signals are in-phase and form the output signal. For identical loads and an ideal coupler, the S-parameter matrix of the circuit is given by

\[
\begin{bmatrix}
0 & -R_f \gamma \\
R_f \gamma & 0
\end{bmatrix}
\]

where \( \gamma \) = voltage reflection coefficient

The resultant phase shift of the device is

\[
\angle S_{21} = \angle \Gamma_f - 90^\circ
\]

(3)

It can be seen from (3) that the phase shift level is dependent on the phase shift of the reflection coefficient of the reflection termination. The voltage reflection coefficient is simply expressed as

\[
\Gamma_f = \frac{Z_r - Z_0}{Z_r + Z_0}
\]

where \( Z_0 = 50 \, \Omega \) in most cases

\[
Z_r = R_f + jX_f
\]

\[
X_f = 1/\left(2\pi f \times \text{Capacitance}\right)
\]

(4)

For an ideal phase shifter where \( R_f = 0 \, \Omega \), the reflection coefficient can be expressed by

\[
\Gamma_f = \frac{X_r^2 - Z_0^2}{X_r^2 + Z_0^2} + \left( \frac{2X_rZ_0}{X_r^2 + Z_0^2} \right)
\]

(5)

However, in reality the phase shifter is not ideal and \( R_f \neq 0 \, \Omega \). Hence the reflection coefficient is

\[
\Gamma_f = \frac{R_f^2 - Z_0^2 + X_f^2}{(R_f + Z_0)^2 + X_f^2} + \left( \frac{2X_rZ_0}{(R_f + Z_0)^2 + X_f^2} \right)
\]

(6)

The attenuation and phase-shift of the circuit can then easily be obtained from either (5) or (6).

**Modelling and Design**

The phase shifter / BPSK modulator designed consists of a cascade of two non-identical reflection-type stages. As explained earlier, a compromise is made between using two identical phase shifters (which would result in a maximum phase shift range) and two optimally-matched stages (CMFB/TPS) designed for maximum bandwidth. The result is a design which can achieve 180 degree maximum phase shift range, which is essential for modulator application, whilst still covering a wide bandwidth. The first stage uses 4x20 micron HEMTs with a micaedged ribbon inductor in series; the second
The stage uses 4x20 micron HEMTs directly connected to the Lange coupler. Both sets of reflection terminations are back-to-back varactor pairs. This is adopted in order to (i) increase the maximum RF power level of the device, (ii) reduce the errors in the resulting phase shift levels when operating at higher RF power levels, and (iii) improve the linearity. High value bias resistors are employed to prevent any RF leakage and to act as forward bias current limiters. The circuit schematic diagram of the designed circuit is illustrated in Fig. 3.

**Measured results**

The phase shifter operation of the MMIC chip, which measures $3 \times 1.3 \text{ mm}^2$, was measured using a Cascade probe and an HP8510B network analyzer. An LR calibration was performed using CPW standards. The measured phase response is shown in Fig. 4(a) for bias voltages between 1V and -3V in 0.5V steps (with the 0° reference phase taken at +1V). The design frequency was 38 GHz and the response is flat in the 30-40 GHz range. According to simulations the circuit is usable over 40 GHz. The insertion loss and return loss at each phase setting are plotted in Fig. 4(b). It can be seen that the return loss is greater than 10dB at all times and the insertion loss variation (PM/AM conversion) is satisfactory.

Modulator operation was tested using a 2 Mbit/s pseudorandom generator with TTL output feeding a pulse generator. The pulse generator allowed the careful setting of signal amplitude and DC offset, both critical for the analogue control of the phase shifter. No baseband pre-filtering was employed. The output spectrum was captured on an HP8562 spectrum analyzer and is illustrated in Fig. 5, showing a near perfect BPSSK spectrum with negligible breakthroughs of the carrier and baseband signals.

**Conclusions**

The design and performance of a Ka-band phase shifter has been described. Using the cascaded match reflection type topology it has been possible to yield a broadband phase shifter with 180° degree coverage. The analog phase shifter can also be employed as a direct BPSSK modulator, used in order to reduce hardware complexity and cost. The pHEMT-based varactor diodes use standard foundry processing and prove to give good performance for this application. Compared with earlier results for designs that use MESFET-based varactor diodes it can be seen that very similar performance levels can be achieved, but now at frequencies well into the mm-wave band.

Apart from being employed as a direct BPSSK modulator in communication systems, the designed phase shifter can also be employed as a terayne modulator in ECM systems and used for electronic beam steering in phased array radar systems.

**References**


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Fig. 1 Single-stage reflection-type phase shifter (RTPS)

Fig. 2 Equivalent circuit model of the IPSVD

Fig. 3 Schematic diagram of the designed phase shifter / modulator circuit
Fig. 4

Fig. 5 Spectrum of the BPSK output at 38 GHz for 2 Mb/s data rate