AN MMIC ACTIVE PHASE SHIFTER USING A VARIABLE RESONANT CIRCUIT

Hitoshi Hayashi and Masahiro Muraguchi

NTT Wireless Systems Laboratories
1-1 Hikarinooka, Yokosuka-Shi, Kanagawa 239-0847 Japan

ABSTRACT

An MMIC active phase shifter using a variable resonant circuit is described. Phase can be changed with a constant amplitude by varying the resonant circuit's reactance. More than 100° phase shift and -4±1-dB insertion loss was obtained from 2.2 GHz to 2.8 GHz. The chip size is less than 1.0 mm^2.

INTRODUCTION

Phased array antennas are advantageous in that by changing the type or array of the antenna elements, or the way of exciting the elements, various functions can be obtained. The usage of phased array antennas has therefore expanded with the increasing demand for satellite communication, mobile communication, personal communication, and so on. Because it is necessary to excite many antenna elements respectively to set the desired amplitude and phase, the feeding circuit becomes complex and large. Thus, to make the antenna small and light-weight, the feeding circuit should be made small and light-weight. An analog phase shifter [1] is often used in an antenna of this type. This device changes the phase of the transmitted and received signal in a phased array antenna and adjusts the antenna's beam orientation. It is preferable for an analog phase shifter to be small and able to change a large amount of phase.

A conventional reflection-type analog phase shifter using varactor diodes is shown in Fig. 1. The problem with this phase shifter is that the amount of variable phase is relatively small, especially for the case in which varactor diodes are made by connecting the drain and source of a conventional FET. Furthermore, its use of a large-scale 3-dB 90° hybrid causes problems from the standpoint of circuit miniaturization.

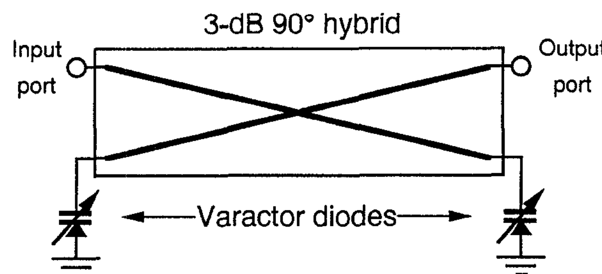


Fig. 1 Conventional reflection-type analog phase shifter using varactor diodes.

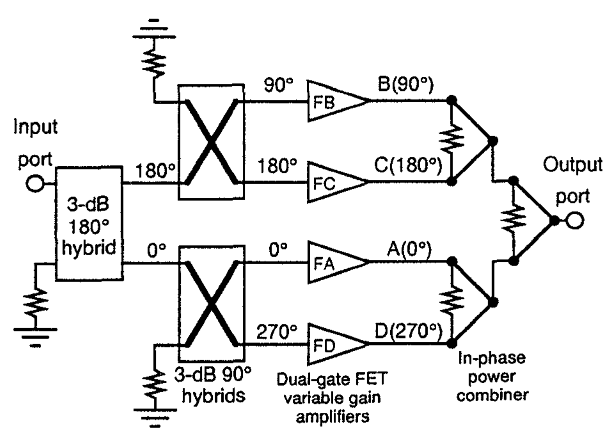


Fig. 2 360° analog phase shifter using vector combining method.

There is also a 360° analog phase shifter which uses a vector combining method [2]. This type of shifter makes four vectors, the phases of which differ by 90°, by combining a 3-dB 180° hybrid and two 90° hybrids. An arbitrary amount of phase from 0° to 360° can be obtained by controlling the amplitude with dual-gate FET variable gain amplifiers and by combining the output of these variable gain amplifiers with an in-phase power combiner. This phase shifter, however, has a relatively complex configuration, which makes miniaturizing the circuit difficult.

As an alternative to these devices, we have developed an MMIC active phase shifter with a large amount of variable phase. The phase shifter comprises a single resonant circuit, thus enabling it to be miniaturized.

DESIGN APPROACH

The operating principle of the proposed phase shifter is easy to explain. Assuming that input voltage is V_1, output voltage is V_2, central angular frequency is ω_0, the Q factor is Q, and $s=j\omega$ (Here, ω is an angular frequency), the voltage transfer function H(s) of a general second-order all-pass network is expressed as follows:

$$H(s) = \frac{V_2}{V_1} = \frac{s^2 - \frac{\omega_0}{Q}s + \omega_0^2}{s^2 + \frac{\omega_0}{Q}s + \omega_0^2}. \tag{1}$$

Thus, If Y_{21} in the admittance matrix Y of a certain phase shifter using a resonant circuit is expressed by the next Equation (2), a frequency characteristic is obtained for which only the phase

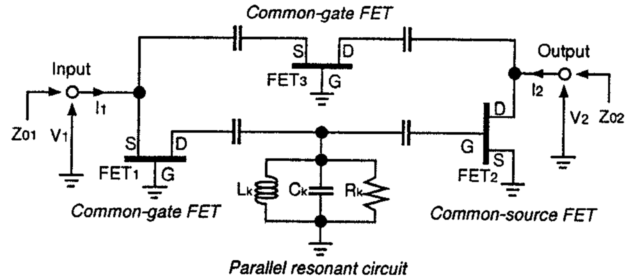


Fig. 3 Circuit configuration of active phase shifter.

changes, with the amplitude remaining constant even if the central angular frequency changes.

$$Y_{21}(s) = \frac{I_2}{V_1} = G \cdot \frac{s^2 - \frac{\omega_0}{Q}s + \omega_0^2}{s^2 + \frac{\omega_0}{Q}s + \omega_0^2} \tag{2}$$

In this Equation, G is the gain of this circuit.

A circuit diagram of the newly proposed phase shifter is shown in Fig. 3. As the figure shows, a commnon-gate FET (FET1) and a common-source FET (FET2) are cascade-connected and a parallel circuit comprising an inductor L_k, a capacitor C_k and a resistor R_k is inserted between these FETs. In addition, another common-gate FET (FET3) is connected to this circuit in parallel. To operate the phase shifter, the first step is to obtain its admittance matrix Y. To simplify the analysis, the FET equivalent circuit is assumed to be a combination of the transconductance g_m and the gate-to-source capacitance C_{gs} only, and the FETs are assumed to have the same cut-off frequency f_T ($=g_m/2\pi C_{gs}$). Under these assumptions, the admittance matrix Y is expressed as follows:

$$\begin{aligned}
Y_{11}(s) &= \frac{I_1}{V_1} = s(C_{gs1} + C_{gs3}) + (g_{m1} + g_{m3}) \\
Y_{12}(s) &= \frac{I_1}{V_2} = 0 \\
Y_{21}(s) &= \frac{I_2}{V_1} = -g_{m3} \cdot \frac{s^2 - \left(\frac{g_{m1}g_{m2}}{g_{m3}} - \frac{1}{R_k}\right)\frac{1}{(C_k + C_{gs2})}s + \frac{1}{L_k(C_k + C_{gs2})}}{s^2 + \frac{1}{R_k(C_k + C_{gs2})}s + \frac{1}{L_k(C_k + C_{gs2})}}. \\
Y_{22}(s) &= \frac{I_2}{V_2} = 0
\end{aligned} \tag{3}$$

From Equations (2) and (3), the condition that only the phase may change while the amplitude remains constant even if central angular frequency changes is expressed by the next Equation (4).

$$\frac{g_{m1}g_{m2}}{g_{m3}} - \frac{1}{R_k} = \frac{1}{R_k} \qquad \therefore R_k = \frac{2 \cdot g_{m3}}{g_{m1}g_{m2}} \tag{4}$$

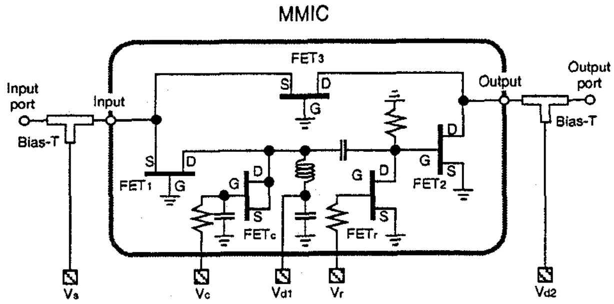
Thus, only the phase can be changed, while the amplitude is kept constant by setting the resistance value R_k determined from Equation (4) and by varying the capacitance value C_k and/or the inductance value L_k.

In addition, input matching can be achieved with the combination of the common-gate FETs (FET_1 and FET_3) by setting the transconductance value ($g_{m1}+g_{m3}$) to be a reciprocal of input impedance Z_{01}. As a result, so-called "input active matching" becomes possible, eliminating the need for passive matching elements and enabling a small-size phase shifter to be obtained. It is known from Equation (3) that the amplitude of Y_{21} is proportional to transconductance g_{m3}. Thus, the gain can be changed by changing both g_{m3} and R_k while satisfying Equation (4).

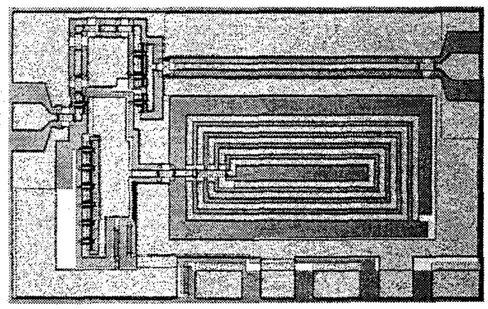
EXPERIMENTAL RESULTS

An experimental MMIC active phase shifter was fabricated using GaAs MESFETs [3]. The circuit configuration is shown in Fig. 4(a). The gate widths of FET_1, FET_2 and FET_3 are 50 µm, 50 µm and 100 µm, respectively. A photograph of the chip is shown in Fig. 4(b). The chip size is 1.28 mm x 0.78 mm, which corresponds to less than 1.0 mm^2. This MMIC structure uses 'uniplanar' techniques based on the coplanar waveguide [4].

Measured results of amplitude variation of transmission, relative phase variation of transmission, and amplitude variation of input and output reflection are shown in Fig. 5. The standard bias conditions are V_s=0.0 V, V_{d1}=4.0 V, V_{d2}=4.0 V, V_c=-6.0 V, V_r=-0.97 V, I_{d1}=8 mA, and I_{d2}=30 mA. V_c was varied from -6.0 V to +4.0 V, and V_r was varied between -0.97 V to -0.94 V to minimize the amplitude variation of transmission as much as possible in all frequency ranges. -4±1-dB insertion loss is obtained from 2.2 GHz to 2.8 GHz with more than 100° phase shift. Furthermore, measured S_{11} across this range is less than -18 dB. This good input match is attributed to the active matching configuration of this phase shifter.



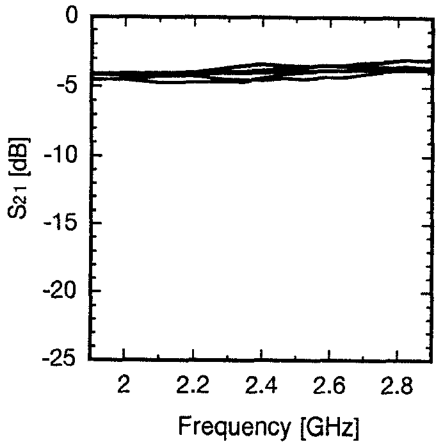
(a) circuit configuration.



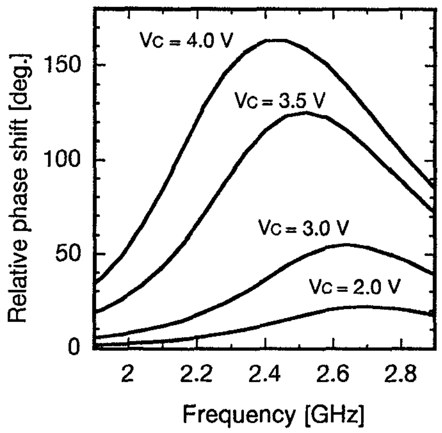
(b) photograph of chip.

Fig. 4 Experimental MMIC active phase shifter.

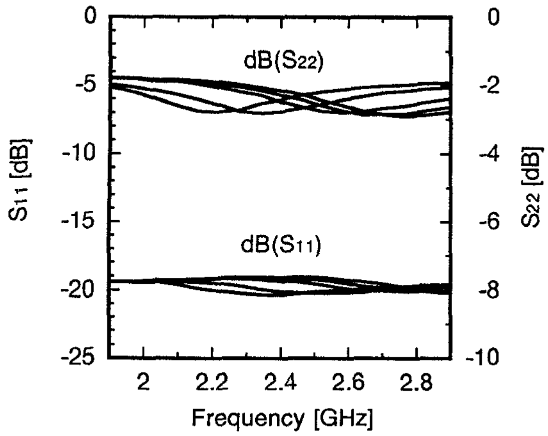
The resonant circuit in this experiment comprised a capacitor and an inductor, but only the capacitance value of the capacitor changed. If the inductance value of the inductor can also be made to change (e.g. by utilizing an active inductor [5]), it will be possible to achieve a phase shifter with an even larger amount of variable phase.



(a) amplitude variation of transmission.



(b) relative phase variation of transmission.



(c) amplitude variation of input and output reflection.

Fig. 5 Measured results.

CONCLUSION

We have developed an MMIC active phase shifter using a variable resonant circuit with second-order all-pass network characterisitcs. With the new device, phase can be changed while maintaining a constant amplitude by changing the reactance of the resonant circuit. The chip size of the experimental MMIC phase shifter was less than 1.0 mm^2, and the insertion loss was -4±1 dB from 2.2 GHz to 2.8 GHz with more than 100° phase shift.

This analog phase shifter is a first step toward achieving the so-called "active integrated antenna", in which both the microwave circuits and the antenna are integrated compactly to facilitate the miniaturizing of wireless communication devices.

ACKNOWLEDGMENT

The authors would like to thank Dr. Takehiro Murase, Mr. Masashi Nakatsugawa and Dr. Tsuneo Tokumitsu for their constant and helpful information and encouragement.

REFERENCES

[1] S. K. Koul and B. Bhat, "Microwave and Millimeter Wave Phase Shifters Volume II," MA:Artech House, 1991.

[2] M. Kumar, R. J. Menna and H. C. Huang, "Broad-band active phase shifter using dual-gate MESFET," *IEEE Trans. Microwave Theory Tech.*, vol. 29, pp. 1098-1102, Oct. 1981.

[3] T. Enoki, K. Yamasaki, K. Osafune, and K. Ohwada, "0.3-µm advanced SAINT FET's having asymmetric n+-layers for ultra-high-frequency GaAs MMIC's," *IEEE Trans. Electron Devices*, vol. 35, pp. 18-24, Jan. 1988.

[4] M. Muraguchi, T. Hirota, A. Minakawa, K. Ohwada, and T. Sugeta, "Uniplanar MMIC's and their applications," *IEEE Trans. Microwave Theory Tech.*, vol. 36, pp. 1896-1901, Dec. 1988.

[5] H. Hayashi, M. Muraguchi, Y. Umeda and T. Enoki, "A novel loss compensation technique for high-Q broad-band active inductors," *IEEE MMWMC Symp. Dig.*, pp. 103-106, Jun. 1996.