PHASE SHIFTERS

A microwave phase shifter is a two-port device capable of producing a true delay of a microwave signal flowing through it. The produced time delay can be fixed or adjustable, and the phase shifter is accordingly called fixed phase shifter or tunable phase shifter. The geometry of the phase shifter is intimately related to the guiding structure which is used to design it and is also related to the operational frequency. Most phase shifters are realized in waveguides or in planar structures. Electrically, a phase shifter can be characterized by its scattering parameters matrix. The scattering matrix for an ideal phase shifter (see Fig. 1) takes the form

$$\boldsymbol{S} = \begin{bmatrix} 0 & e^{-j\phi_1} \\ e^{-j\phi_2} & 0 \end{bmatrix} \tag{1}$$

The signal arriving at port 1 will appear at port 2 with a phase shift ϕ_1 without being reflected at port 1 and with no attenuation, while the signal arriving at port 2 will appear at port 1 with phase shift ϕ_2 and without reflection or attenuation.

In general, $\phi_1 \neq \phi_2$ and the phase shifter is called nonreciprocal, and if $\phi_1 = \phi_2$ the phase shifter is called reciprocal. In practice it is impossible to achieve perfect matching at the two ports (no reflection) and to avoid some attenuation while the signal flows through the phase shifter. For these reasons the scattering matrix of a real phase shifter can be written in general as

$$\boldsymbol{S} = \begin{bmatrix} S_{11} & |S_{21}|e^{-j\phi_1} \\ |S_{12}|e^{-j\phi_2} & S_{22} \end{bmatrix} \tag{2}$$

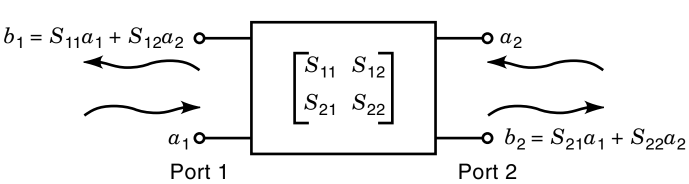


Figure 1. Phase shifter viewed as a two-port device.

The attenuation of the microwave signal due to the presence of the phase shifter can be calculated from its S parameters, and it is expressed in decibels as

$$(\text{Insertion loss})_1 = 20 \log |S_{21}| \tag{3}$$

$$(\text{Insertion loss})_2 = 20 \log |S_{12}| \tag{4}$$

The subscripts 1 and 2, respectively, refer to the phase shifter when the input signal is at port 1 or port 2. The mismatch is expressed as standing wave ratio (VSWR) at each port and is given by

$$(\text{VSWR})_1 = \frac{1 + |S_{11}|}{1 - |S_{11}|} \tag{5}$$

$$(\text{VSWR})_2 = \frac{1 + |S_{22}|}{1 - |S_{22}|} \tag{6}$$

In order to evaluate the performance of a phase shifter, it is necessary to introduce a quality factor. For a phase shifter operating at a specific frequency we can define a *figure of merit* as the ratio between the maximum phase shift (in degrees) and the corresponding attenuation in dB at that frequency. This parameter can be expressed as

$$\text{figure of merit} = \frac{\Delta(\text{phase}\, S_{21})}{|S_{21}|} \tag{7}$$

The performance of a phase shifter can be measured using a standard S parameter setup, including a network analyzer and a test-set of calibration standards suitable for the specific guiding structure (1).

PHASE SHIFTER CLASSIFICATION

A first classification of a phase shifter can be based on its phase shifting capability, according to which it can be identified as fixed or adjustable. A fixed phase shifter will provide a constant phase change between the two ports, while an adjustable phase shifter will provide a phase change between the ports which can be controlled mechanically or electrically. Further classification for the adjustable type is based on electrical performance and operational principle. Within the adjustable phase shifters we can distinguish between those where the phase change is achieved through a mechanical tuning and those where the change is obtained with an electrical signal. Furthermore, for the electrically tunable type we can distinguish between those where the phase can be changed continuously (analog) and those where the phase can only be changed by discrete steps (digital). Figure 2 shows a graphical classification of different types of phase shifters.

PHASE SHIFTER PERFORMANCE

In the evaluation of a phase shifter performance, besides the quantities derived from its S parameters such as insertion loss, quality factor, and VSWR, other quantities are important for practical design. Below we discuss such parameters and their corresponding meaning.

- *Operational Bandwidth.* This is defined as the 3 dB bandwidth (2), which is expressed as the frequency range

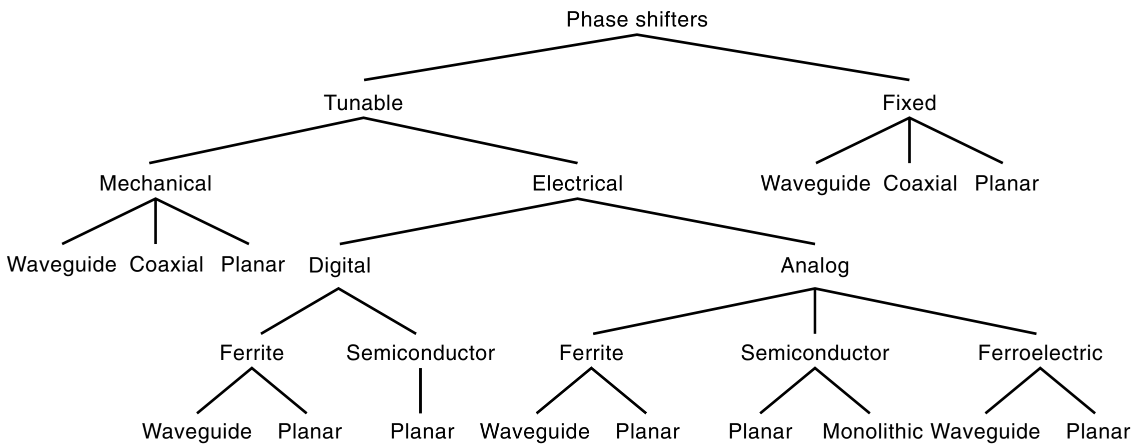


Figure 2. Phase shifter classification chart.

in which the insertion loss is contained within 3 dB change.

- *Power Handling Capabilities.* This is expressed as the maximum power which can flow in the phase shifter without overheating its components or without introducing nonlinear phenomena due to the amplitude of the microwave field. This second limitation is particularly important for phase shifters which employ discrete devices such as field-effect transistors (FETs) or diodes.
- *Switching Speed.* This is the time needed by the phase shifter to switch between two different states, usually at the two ends of the achievable phase shift (larger allowable jump).
- *Temperature Sensitivity.* This expresses the sensitivity in terms of degree of phase shift degradation per °C change. This parameter should be small to avoid the necessity to adopt thermal compensation.
- *Physical Size.* This parameter can be very important, especially when the phase shifter is employed in a radar system where thousands of units are required. Physical dimensions and weight must be minimized even at the cost of other parameters. As an example, think of a radar system which needs to be mounted on the front side of a jet fighter.

FIXED PHASE SHIFTERS

A fixed phase shifter must provide a constant phase change between its two ports. Theoretically any transmission line would be suitable for producing such a function as illustrated in Fig. 3. For instance, in the X band (1) a coaxial cable could be used as a fixed phase shifter, while in the Ku band a waveguide can be employed for the same purpose. In many applications, it is desirable to achieve a differential phase shift between two lines having the same length. For this purpose, lines with different time delay must be used. A possible approach to this problem is to change the propagation constant of the line, loading it with lumped or distributed elements. So if β_1 is the propagation constant of the unloaded line and β_2 is the propagation constant of the loaded one, the achieved differential phase shift will be given by (3)

$$\Delta\phi = (\beta_1 - \beta_2)x \tag{8}$$

where x is the length of the line. So the basic idea in the realization of this type of phase shifter is to change the propagation constant of the transmission line by properly loading it. As an example, consider the realization of a fixed phase shifter in circular waveguide geometry. The waveguide is loaded with metal inserts as shown in Fig. 4. The equivalent circuits for the loaded and unloaded cases, assuming that the guide is operating with the fundamental mode TE_{11} (4), are reported in Fig. 5. Both lines have the same length, and the differential phase shift between the two TE_{11} modes is related

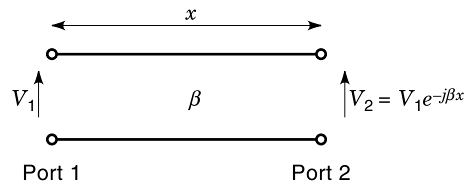


Figure 3. Transmission line acting as a phase shifter.

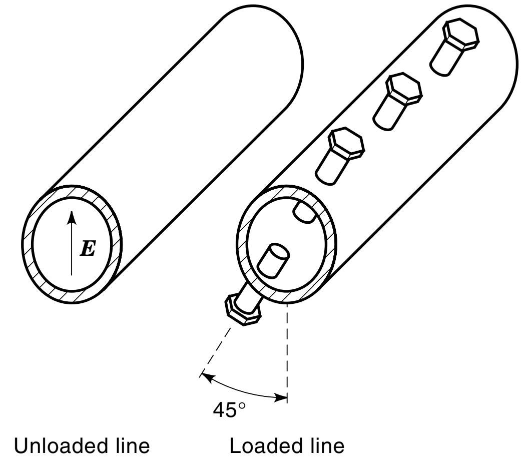


Figure 4. Loaded circular waveguide.

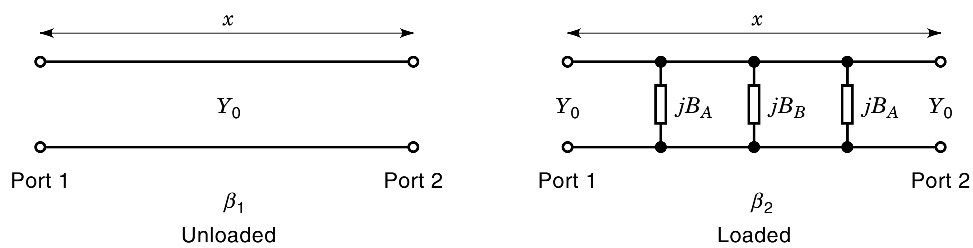


Figure 5. Equivalent circuits for loaded and unloaded circular waveguides operating with the fundamental mode.

to the normalized susceptance of the loads (5) by the following equations:

$$\overline{B}_B = \frac{B_B}{Y_0} = \frac{\sin 2\beta x - \sin(2\beta x + \Delta\phi)}{\sin^2 \beta x} \tag{9}$$

$$\overline{B}_A = \frac{B_A}{Y_0} = \frac{\sin \Delta\phi \cos \beta x - (1 - \cos \Delta\phi) \sin \beta x}{\sin \beta x \sin(2\beta x + \Delta\phi)} \tag{10}$$

The use of Eqs. (9) and (10) allows one to design the loads necessary to achieve a desired phase shift. Using a similar concept, depending on the transmission line geometry, different type of loads can be devised as shown in Fig. 6. A quarter-wave transformer is used to avoid reflection at the load interface. Figures 6(a) and 6(b) show realization in circular waveguide geometry using dielectric or metallic loads, while Figs. 6(c) and Fig. 6(d) are rectangular waveguide geometry using dielectric loads.

MECHANICALLY TUNED PHASE SHIFTERS

Mechanically tunable phase shifters are capable of varying the signal delay in a transmission line using some moving parts. The specific geometry depends on the operational frequency and on the guiding structure. As an example, three classical implementations—a coaxial cable, a waveguide, and a microstrip line, respectively—are outlined below.

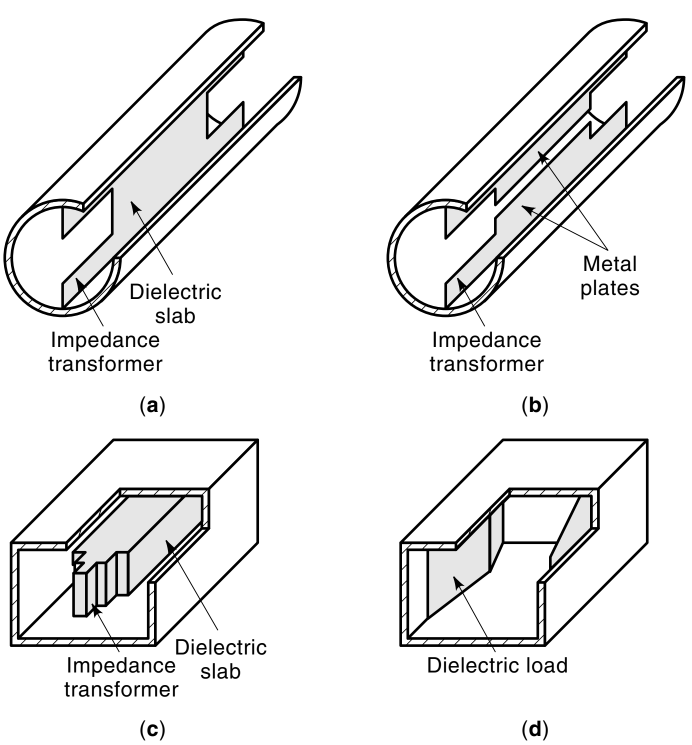


Figure 6. Different types of loaded transmission line.

Coaxial Cable Phase Shifter

In a coaxial cable the dominant mode is TEM (see ELECTROMAGNETIC FIELD MEASUREMENT) (6) so the phase of the signal propagating over a length between two cable ends points is given by

$$\phi = \frac{\omega\sqrt{\epsilon_r}}{c} x \tag{11}$$

where x is the cable length, ω is the operating frequency, ϵ_r is the dielectric constant of the inner core of the cable, and c is the speed of light in free space. A Δx change in its length will produce a change in phase ($\Delta\phi$) between the two cable points end expressed by

$$\Delta\phi = \frac{\omega\sqrt{\epsilon_r}}{c} \Delta x \tag{12}$$

Figure 7 illustrates a section view of this type of phase shifter. To allow for the stretch, the coaxial cable has concentric air lines which can slide one into another, maintaining the characteristic impedance of the cable constant while changing length.

Waveguide Phase Shifter

In waveguide geometry, one way of obtaining a tunable phase shift without changing its length is to change the effective dielectric constant in some region of the guide, inserting a movable dielectric slab. Figure 8 illustrates one version of this mechanical tunable phase shifter. The insertion of the flap in

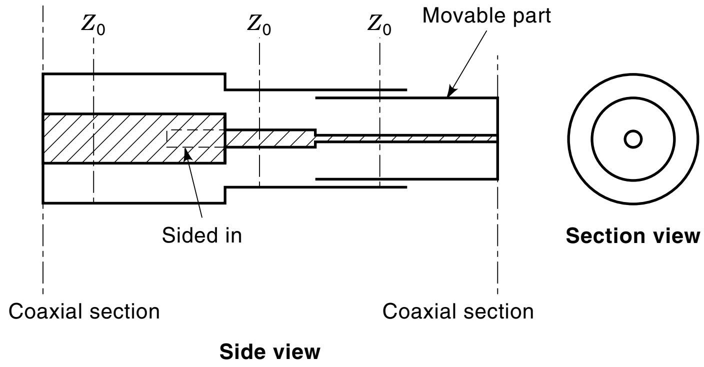


Figure 7. Mechanically tuned coaxial phase shifter.

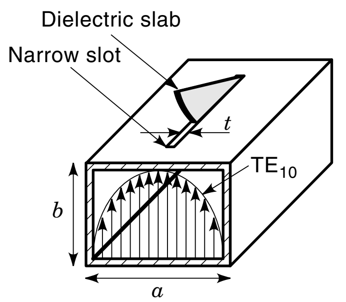


Figure 8. Mechanically tuned waveguide phase shifter.

the center of the waveguide, where the electric field is maximum assuming that the fundamental mode is propagating, will delay the signal, producing a phase shift. This type of device is only usable with some restrictions, since the thickness of the flap and its dielectric constant must be calculated to avoid the propagation of higher-order modes (i.e., TE_{30}). A simple equation for the design of this type of phase shifter, which avoids higher modes, was proposed by Gardiol (7) and leads to the following relation:

$$\sqrt{\epsilon_r}\tan[\pi(a-d)/\lambda_c] = \cot(\pi\sqrt{\epsilon_r}d/\lambda_c) \tag{13}$$

Based on the same concept, it is possible to have a movable dielectric inside a rectangular waveguide operating with the fundamental mode TE_{10} as shown in Fig. 9, where the interaction of the field with the dielectric will be maximum when the plate is in the center of the guide (maximum delay corresponding to maximum phase shift) and minimum when it is on the side walls. Proper design of the $\lambda/4$ transformer (8) is necessary to avoid reflections at the phase shifter interface.

Microstrip Phase Shifter

For the microstrip geometry a mechanically tuned phase shifter was proposed by Joines (9). In this geometry it is possible to achieve a phase shift by changing the dielectric constant of the substrate above and below the strip as depicted in Fig. 10.

The change in the dielectric constant will induce a change in the propagation constant, and consequently different phase shifts will be achieved. This structure is attractive because it yields a continuous phase shift while maintaining the characteristic impedance constant. If we observe its section view depicted in Fig. 10, we notice that by a proper design of the thicknesses t_1 and t_2, accordingly with the dielectric constant (9), it is possible to keep the characteristic impedance of the strip constant while changing its propagation constant.

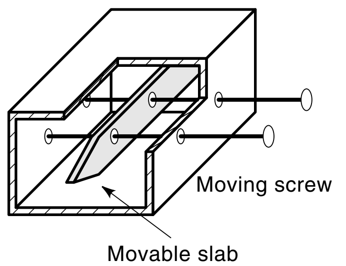


Figure 9. Dielectric loaded rectangular waveguide phase shifter.

These are just a few examples of mechanically tunable phase shifters; of course, many others are possible, but the basic concept on which they operate is the same and can be summarized as follows. In order to obtain a phase shift, it is necessary to delay the electric signal independently from the type of guiding structure used. This can be achieved in two ways: One way is to change the *physical length* of the transmission line Δx which produces a delay of the signal at the output port, inducing a phase shift change given by $\Delta\phi = \beta\Delta x$. In the second case, a change of the *wave propagation velocity* obtained by changing the propagation constant of the line $\Delta\beta$ will produce a phase shift at the output port given by $\Delta\phi = \Delta\beta x$.

ELECTRICALLY TUNED PHASE SHIFTERS

In electrically tuned phase shifters the phase change is controlled by an electric signal (driving signal) such as a voltage or a current. Since no moving parts are involved in the phase control process, electrically controlled phase shifters can achieve faster phase shift compared to mechanical ones. They can be subdivided into two major categories—*digital* and *analog*—depending on the type of control on the phase shift they provide. One of the most important application of electrically tuned phase shifters is the so-called phased array system. A phased array system is an array of antennas of which by electronically controlling the phase of the electromagnetic signal at each antenna element one can change its pointing direction. As an example, let us consider a linear array of antenna as illustrated in Fig. 11. If all the elements are excited with the same phase signal, the radiated signal adds coherently and forms a wave front parallel to the array direction (line joining all the elements). The beam pointing direction is perpendicular to the wave front, so the radiated beam will point in a broadside direction. In a phased array, this direction is adjustable by acting on the phase of the electromagnetic signal at the aperture of each radiating element. In a linear array with equispaced elements the beam can be steered by introducing a progressive phase shift between successive elements. If θ_0 is the scan angle with respect to the broadside direction, then the phase delay to be introduced between adjacent antenna apertures can be calculated from

$$\Delta\phi = \frac{\omega}{c}d\sin\theta_0 \tag{14}$$

where d is the element spacing, and c is the free space light speed. For scanning the beam continuously, $\Delta\phi$ is varied by analog phase shifters; and for switching the beam from one scan angle to another, $\Delta\phi$ is varied in discrete steps by digital phase shifters. The same principle applies to planar array for achieving three-dimensional scanning and switching.

Digital Phase Shifters

Digital phase shifters use electronic devices such as pin diodes or FETs as switching elements; this allows the digital phase shifter to direct the microwave signal through paths of different length, obtaining in this way the phase shift. The use of a pin diode as a switching circuit allows biasing of the diode forward (to obtain a trough) or reverse (to obtain a open circuit) by means of a dedicated bias circuitry. In a similar

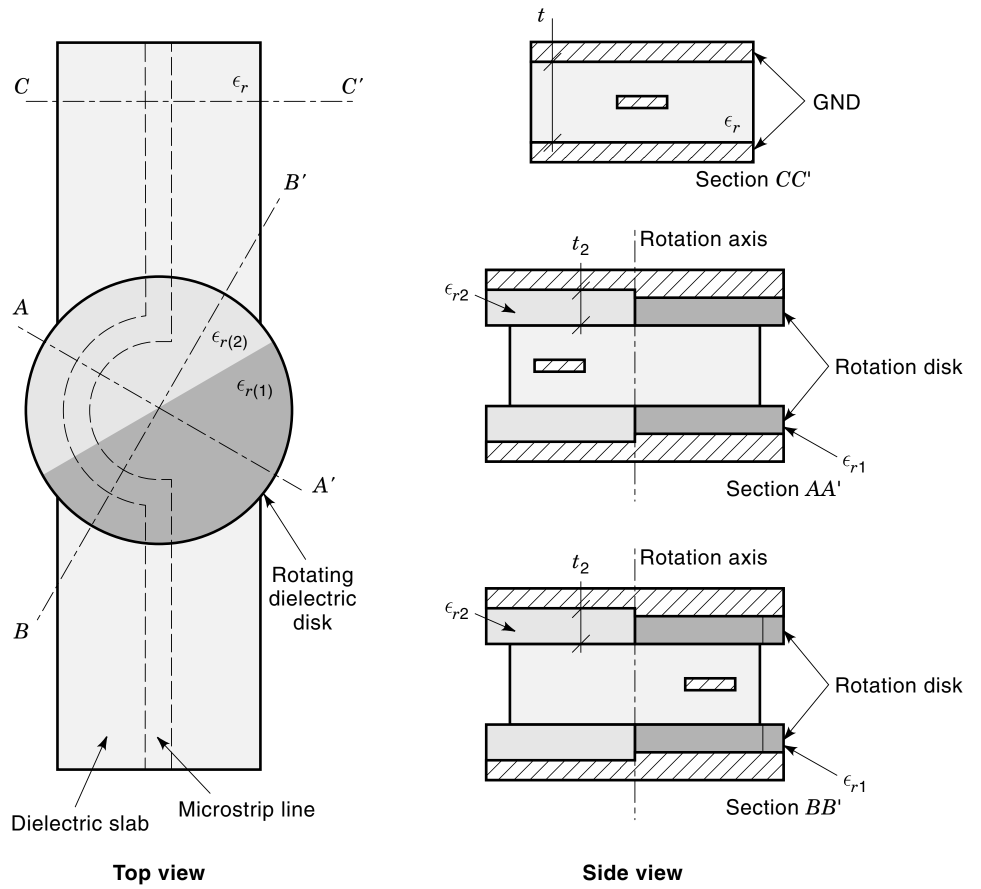


Figure 10. Microstrip mechanical tuned phase shifter.

way, an FET channel can be switched on or off by proper bias (10). The use of FETs and diodes allows four basic designs. The simplest one is shown in Fig. 12, where the shift is obtained by switching the signal between two different length transmission lines; Fig. 12 also shows a microstrip implementation of this type. The phase shift is proportional to the difference between the length of the two lines and is given by $\Delta\phi = \Delta x\beta$. In a similar way, as shown in Fig. 13, the use of different loads on the transmission lines, which are switched on and off by the diodes, allows a phase change between the biased and unbiased condition ($\Delta\phi$) related to the impedance of the line (Z_0) and to the susceptance of the loads (B) according to the relation (5)

$$\Delta\phi = 2\arctan\left(\frac{B_n}{1 - B_n^2/2}\right) \tag{15}$$

where $B_n = Z_0B$. A microstrip implementation of this circuit is also shown in Fig. 13; the reactive and inductive loads are obtained using open stubs of different length, and their values can be calculated for a given susceptance using (11)

$$B = Z_S \cot\left(\frac{2\pi f\sqrt{\epsilon_r}}{c}\, l_S\right) \tag{16}$$

where Z_s is the stub impedance, f is the operation frequency, ϵ_r is the relative dielectric constant, l_s is the stub length, and c is the speed of light in free space. Another design using a pin diode in combination with a 90° hybrid circuit is illustrated in Fig. 14. In this case the differential phase shift obtained between the biased and unbiased condition is given by

$$\Delta\phi = \beta\frac{x}{2} \tag{17}$$

The bias circuit must be designed carefully here, in order to avoid degradation of performance and direct-current (dc) leaks. The bias circuit must allow biasing of all the active devices independently, while insulating the dc biasing signal from the radio-frequency (RF) signal. A simple design for a microstrip topology is illustrated in Fig. 15. The two dc blocks are acting as series capacitors for the RF signal, allowing the RF to go through the diode while stopping the dc component from the rest of the circuit. The two $\lambda/2$ high-impedance microstrip lines are operating as an open stub (11). At the microstrip junction they will result in an open circuit transparent to the microwave signal while allowing the dc signal to provide the necessary bias for the diode. The high impedance of the open stub makes it look like an open circuit for a larger bandwidth (11).

The use of FETs as a switching element is similar to that of the pin diode: The source and drain are grounded (only for the dc signal). In the off state, the gate-source and the gate-drain capacitances are equal. Because of this, the drain is not isolated from the gate terminal. In real circuits, the bias network is configured so as to provide high impedance for the RF at the gate terminal. This is achieved by using a low-pass filter such that it presents an effective RF open at the gate. This arrangement is shown in Fig. 16.

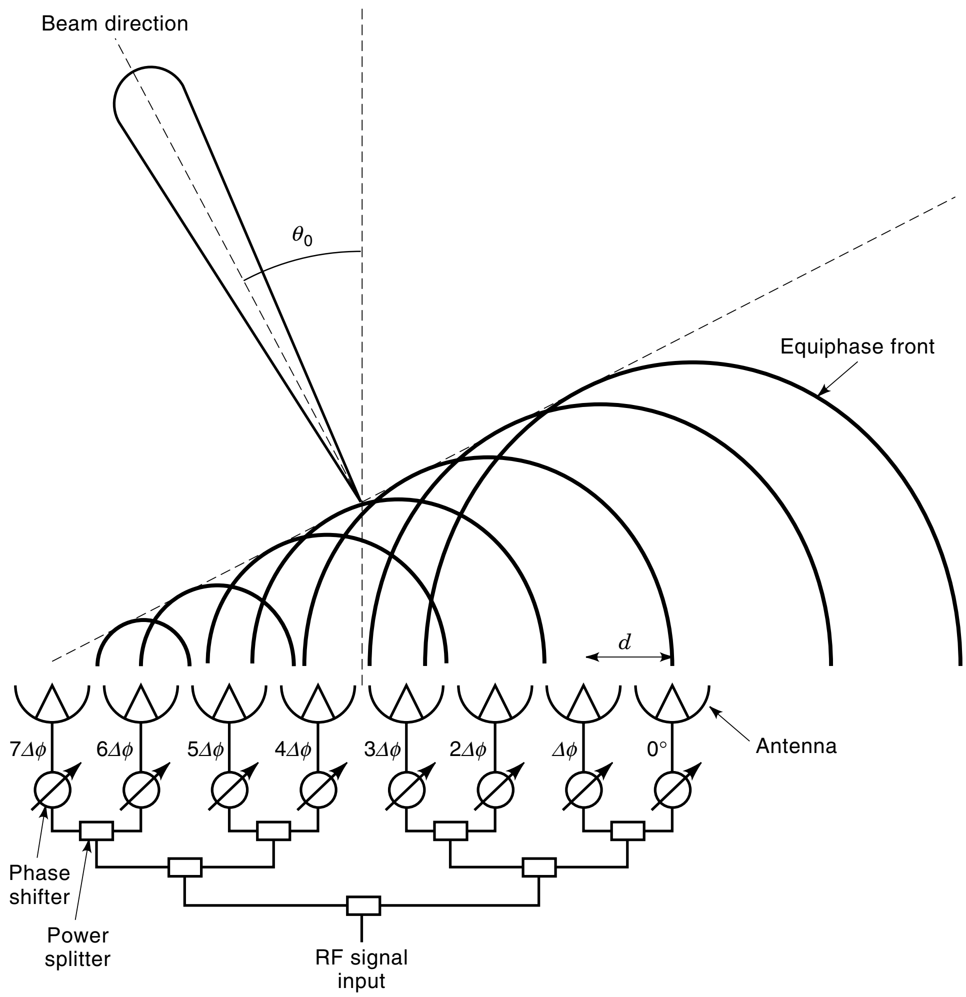


Figure 11. Beam-steering concept using a phase shifter at each radiating element.

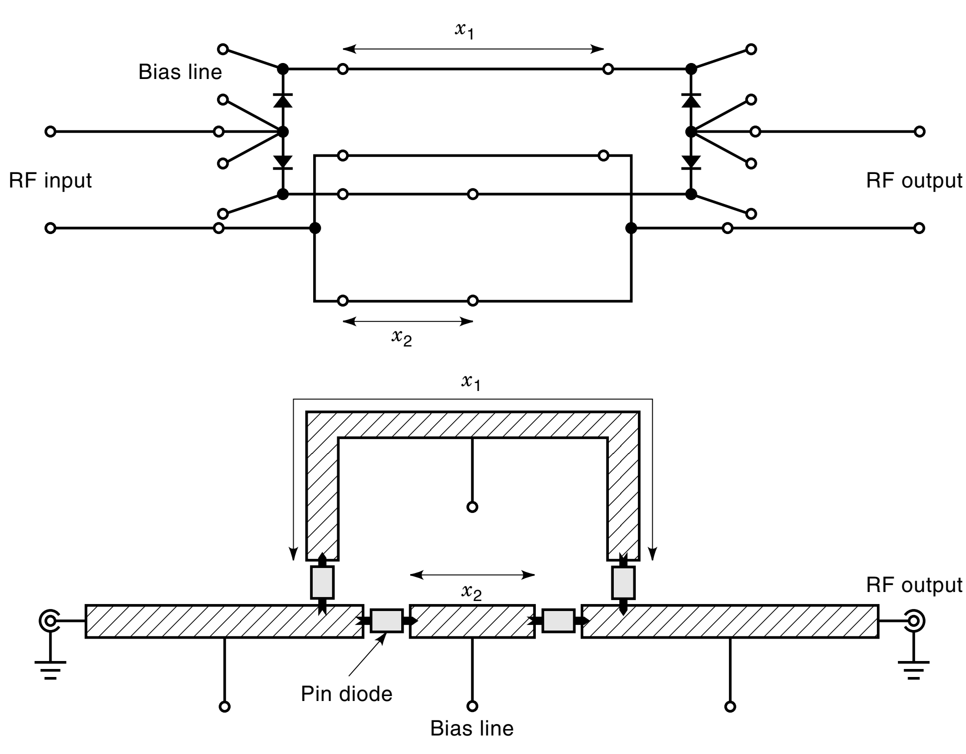


Figure 12. Pin diode type of electrically controlled digital phase shifter.

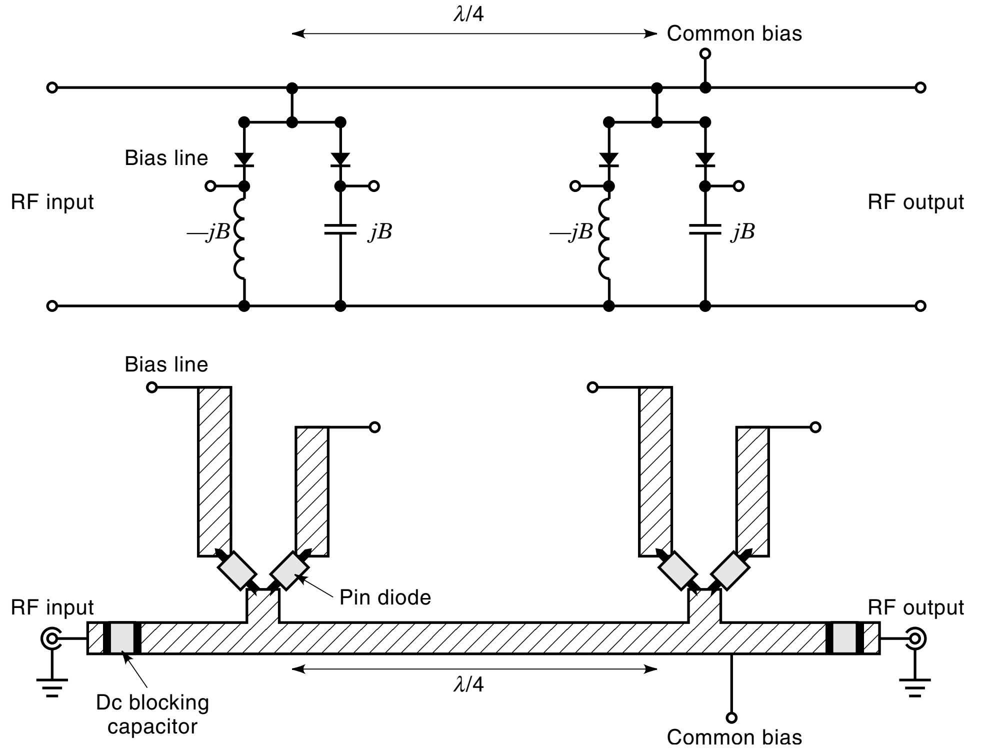


Figure 13. Loaded pin diode electrically tuned digital phase shifter.

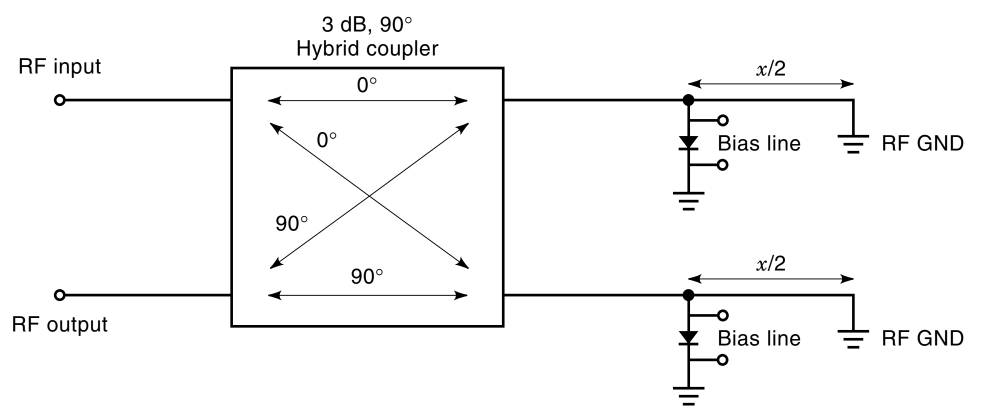


Figure 14. Pin diode and 90° hybrid circuit type of electrically controlled digital phase shifter.

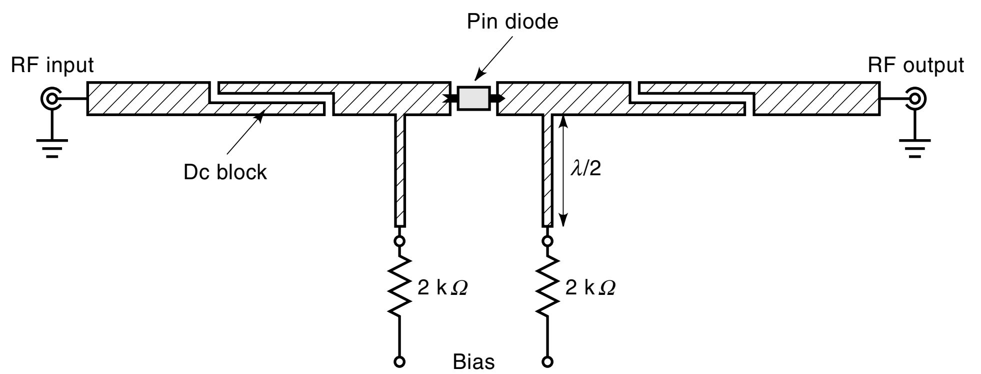


Figure 15. Bias circuit for *pin* diode type of phase shifter.

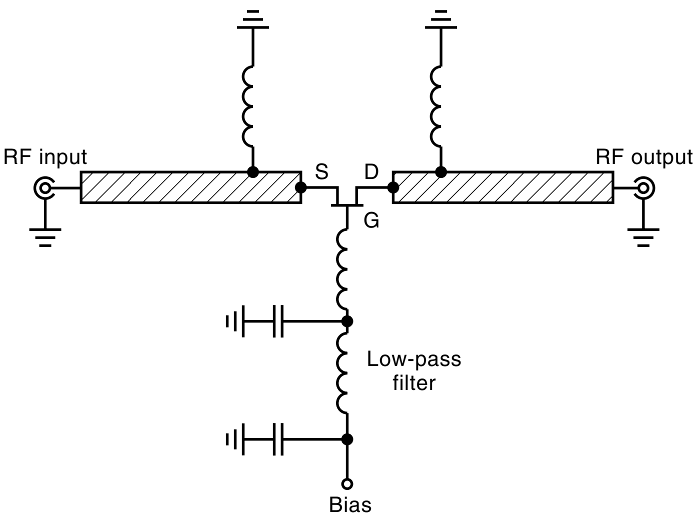


Figure 16. Bias circuit for an FET type of phase shifter.

Because digital phase shifters only allow discrete phase jumps, a cascade of them must be used when high resolution in the phase change is desired. Figure 17 shows a typical arrangement and the corresponding phase shift states for a 4-bit phase shifter, capable of giving phase jumps with an increment of 22.5° and a maximum phase shift of 360°. By increasing the number of phase shifters, higher resolution is achievable. The maximum achievable resolution will obviously depend on the number of phase shifters and is expressed by the relation

$$\Delta\phi_{\min} = \frac{2\pi}{2^n} \tag{18}$$

where n is the number of discrete phase shifters.

Analog Phase Shifters

Analog phase shifters allow time delay control of a microwave signal by using an electric driving signal. They differ from digital phase shifters due to their capability to provide continuous phase delay control. This characteristic is very attractive when a fixed phase resolution is impractical to use. Consider as an example the received signal coming from a broadcast TV, as illustrated in Fig. 18. Because a multiple reflection path exist, a double image is received. One possible solution to overcome the problem is to use two receiving antennas and by proper adjustment of the phase difference between them eliminate the reflected signal. Because of the random nature of the delay, only a continuous adjustable phase shifter can be employed.

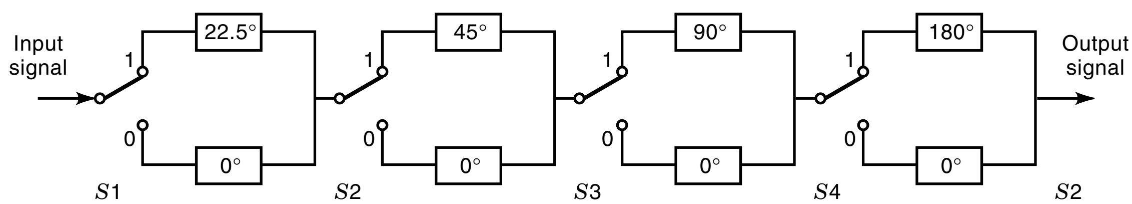
Electrically analog tunable phase shifters can be subdivided in three major subcategories as illustrated in Fig. 2. A description of the operational principle for each of them is provided below.

Ferrite Phase Shifters

Ferrite phase shifters are employed in a waveguide or a planar structure. They operate using the ferrite property of nonlinear dependence between magnetization (B) and magnetic field (H). Ferrites are nonlinear and nonreciprocal magnetic materials composed of a mixture of divalent metal and iron oxide having the general chemical structure

$$MO + Fe_2O_3 \tag{19}$$

where M is a divalent metal such as manganese, magnesium, nickel, or iron. They exhibit a hysteresis B–H dependence as reported in Fig. 19. To explain how ferrites are used in phase shifters, it is not necessary to describe in detail the material properties which are well documented in Refs. 12 and 13. The



Switch state S1 S2 S3 S4	Phase shift
0 0 0 0	0°
1 0 0 0	22.5°
0 1 0 0	45°
1 1 0 0	67.5°
0 0 1 0	90°
1 0 1 0	112.5°
0 1 1 0	135°
1 1 1 0	157.5°
0 0 0 1	180°
1 0 0 1	202.5°
0 1 0 1	225°
1 1 0 1	247.5°
0 0 1 1	270°
1 0 1 1	292.5°
0 1 1 1	315°
1 1 1 1	337.5°

Figure 17. Diagram of 4-bit phase shifter and corresponding switching scheme.

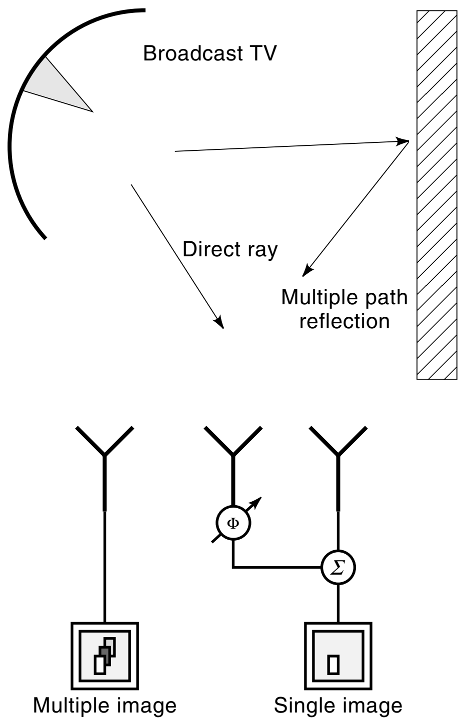


Figure 18. Phase shift recovering for a multiple path reflected signal.

nonlinear H–μ dependence will be exhibited as shown in Fig. 19. The permeability at specific magnetization value (H^*) can therefore be calculated as

$$\mu = \mu_0 + \left.\frac{\partial \boldsymbol{B}}{\partial \boldsymbol{H}}\right|_{H=H^*} \tag{20}$$

In general the ferrite permeability takes the form of a tensor because of the nonreciprocal behavior. The elements of this tensor are a function of the applied magnetic field. When the magnitude or direction of the magnetic field is changed, the permeability of the ferrite changes, thereby changing the propagation constant of the electromagnetic wave. Phase shift is a consequence of the change in the propagation constant brought about by electronically controlling the applied magnetic field. For a more extensive and complete treatment of ferrite properties at microwave frequencies, see Refs. 14 and 15. As direct application of this concept, a waveguide ferrite loaded phase shifter is described. The geometry of the device is shown in Fig. 20, the magnetization of the ferrite is

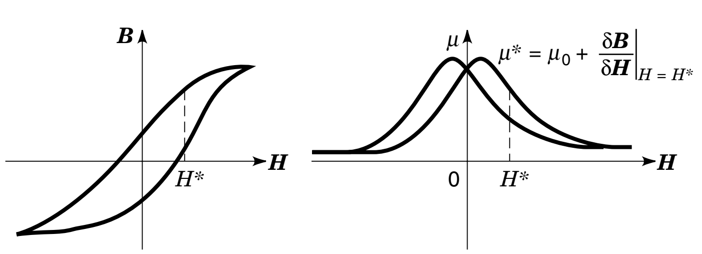


Figure 19. Nonlinear $\boldsymbol{B}$–$\boldsymbol{H}$ dependence and correspondent μ–$\boldsymbol{H}$ dependence.

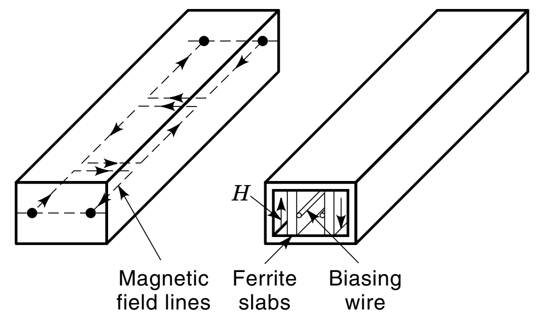


Figure 20. Nonreciprocal waveguide ferrite phase shifter.

achieved using a current loop around the ferrite slab (with the aid of a biasing wire). The ferrite is placed in the waveguide in such a way to maximize the interaction with the existing magnetic field in the guide. For waveguide operating with the fundamental mode (TE_{10}), the magnetic field will be maximum at $\frac{1}{4}$ and $\frac{3}{4}$ of the longitudinal section of the guide (6) as illustrated in Fig. 20. Because the magnetic field has opposite direction at those sections, an asymmetrical bias (see Fig. 20) will be necessary in order to obtain a phase shift. This is done using a current flowing in the two wires in opposite direction. Another concept is to use different geometries as illustrated in Fig. 21. The cross section depicted in Fig. 21(a) is an extension of the one shown in Fig. 20, with the difference being that the ferrite is placed where the maximum magnetic field exists at the bottom and top of the guide. This allows a reduction of the mismatch with the empty waveguide, making easier the design of the matching circuit. Also in this case as illustrated in the figure, a differential bias of the ferrite must be used. Figure 21(b) is also based on a simi-

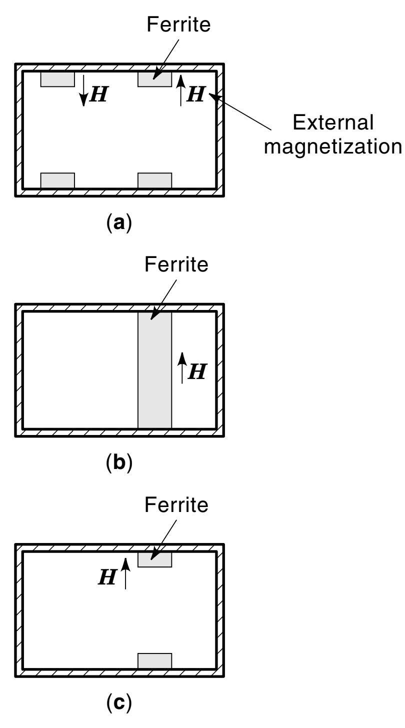


Figure 21. Different types of nonreciprocal waveguide ferrite phase shifter.

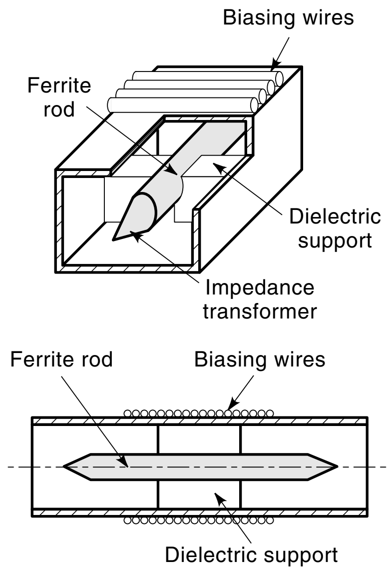


Figure 22. Reggia–Spencer reciprocal ferrite phase shifter.

lar concept; but the asymmetrical bias is replaced by an asymmetric geometry, so the ferrite is only placed on one side of the guide. Further improvement toward the matching for this structure is obtained in the case of Fig. 21(c). Several versions of a planar structure which employ ferrites as tunable elements have been proposed (16–18). Even though in principle those structures work, most ferrite phase shifters currently constructed use waveguide geometry.

Another operational principle for the ferrite reciprocal phase shifters is to use Faraday rotation to produce time delay of microwave signal. This type of phase shifter was at first proposed by Reggia and Spencer (19) and is illustrated in Fig. 22. Several modifications of the original form were proposed later, but the basic operative principle remains the same. In this phase shifter a longitudinal ferrite toroid is placed in the longitudinal section of a rectangular waveguide (see Fig. 22). The magnetic biasing field is produced by an external magnetization circuit. It is well known that when a linearly polarized wave propagates in a ferromagnetic rod, the plane of polarization of the wave in the rod rotates. Now if the rod is placed inside a rectangular waveguide with one of its dimension at cut-off, then the rotational effect is suppressed (for small size rod). Reggia and Spencer have demonstrated large changes in insertion phase with external magnetic field bias for the transmitted power. They also demonstrated that the phase variations are independent of the propagation direction. Many other authors investigated the theory beyond this effect (20,21). Practical design of the Reggia–Spencer phase shifter is mostly based on approximate equations (22) which consider the phase shift as a consequence of a small perturbation in the effective permeability.

Ferroelectric Phase Shifters

In ferroelectric phase shifters the phase-shift capability of ferroelectric materials results from the fact that if we are below their Curie temperature (23,24) (see FERROELECTRICS), the dielectric constant of such a material can be modulated under the effect of an electric bias field. Particularly, if the electric field is applied perpendicularly to the direction of propagation of the electromagnetic signal, the propagation constant ($\beta = 2\pi/\lambda$) of the signal will depend upon the bias field since $\beta = 2\pi\sqrt{\epsilon_r}/\lambda_0$ and $\epsilon_r = \epsilon_r(V_{\text{bias}})$. The total wave delay will become a function of the bias field, and therefore this will produce a phase shift $\Delta\phi = \Delta\beta l$, where l is the length of the line. Two major implementations of a ferroelectric phase shifter have been used: waveguide geometry and planar structures. In waveguide geometry the ferroelectric material is placed inside a waveguide as illustrated in Fig. 23. A voltage is applied to the center conductor, creating a vertical electric field to the grounded flange. The matching layer must be placed on either side of the sample to couple the RF energy in and out of the material. These rectangular layers of dielectric are needed in the design of the phase shifter because of the impedance mismatch between air and the high permittivity ferroelectric. One problem in the use of this type of setup is the high bias required (typically 1 kV to 2 kV) due to the thickness of the material. Ferroelectrics require a bias voltage of the order 2 V/μm to 4 V/μm in order to change significantly their dielectric constant (25,26).

Use of the planar type of ferroelectric material in microstrip geometry avoids this problem as demonstrated in Ref. 27. The geometry is illustrated in Fig. 24. The active part of the device consists of a microstrip line printed on a ferroelectric substrate whose dielectric constant is changed by bias. The length of the strip determines the maximum phase shift achievable for a fixed change of the propagation constant ($\Delta\beta$), associated with the maximum bias voltage. The complete design of this type of phase shifter is reported in Ref. 28. To reduce the required bias the ferroelectric material has a thickness of the order of 0.1 mm to 0.2 mm, allowing bias voltage of a few hundred volts. As in the case of the *pin* diode phase shifter, attention must be dedicated to the design of the biasing circuit, to avoid leakage of the dc voltage in the RF circuit.

Varactor Diode Phase Shifter

In varactor diode phase shifters a varactor diode is used as a variable-capacitance element. This variable capacitance is obtained through a voltage-tuned capacitance of the diode under a reverse-bias condition (24). The varactor diode is used in combination with a hybrid coupled circuit as illustrated in Fig. 25(a). The 3 dB 90° hybrid circuit is symmetrically terminated with the diodes. If X is the reactance of the diode, the reflection coefficient can be calculated as (11)

$$\Gamma = \frac{jX/Z_0 - 1}{jX/Z_0 + 1} \tag{21}$$

and the corresponding phase of the reflection coefficient is given by

$$\phi = \pi - 2\arctan(X/Z) \tag{22}$$

where Z_0 is the characteristic impedance of the transmission line (50 Ω typically). We notice that in order to obtain a phase variation in the range going from 0 to 2π, the reactance of the diode must go from $-\infty$ to 0 to $+\infty$ and the maximum change of phase is obtained when $X = 0$. Hence, in order to obtain a maximum phase shift, the diode must be connected in series with an inductive load to allow resonance ($X = 0$); this can be

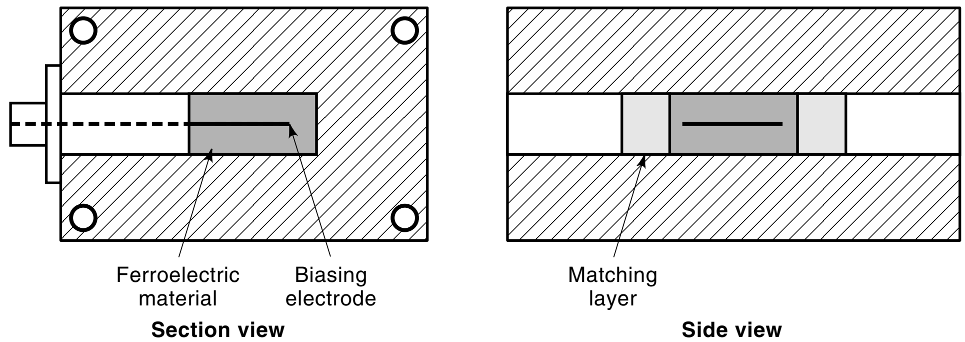


Figure 23. Waveguide ferroelectric type of analog phase shifter.

achieved with a stub as illustrated in Fig. 25(b) for a microstrip realization. The impedance of the reflecting termination (diode and stub) is given by

$$Z = R_d + j\left(Z_S \tan \beta l_S - \frac{1}{\omega C_d}\right) \tag{23}$$

where R_d and C_d are the equivalent parameters of the diode, and Z_s and l_s are the stub characteristic impedance and length, respectively. The associated reflection coefficient is calculated from

$$\Gamma = \frac{R_d - Z_0 + jX}{R_d + Z_0 + jX} \tag{24}$$

As the bias voltage changes from 0 to a negative value, C_d goes from $C_{d\max}$ to $C_{d\min}$, giving a change of X expressed by

$$\Delta X = \frac{1}{\omega C_{d\min}} - \frac{1}{\omega C_{d\max}} \tag{25}$$

For such change of X the correspondent phase change can be obtained as

$$|\Delta\phi| = 4 \arctan\left(\frac{\Delta X}{2Z_0}\right) \tag{26}$$

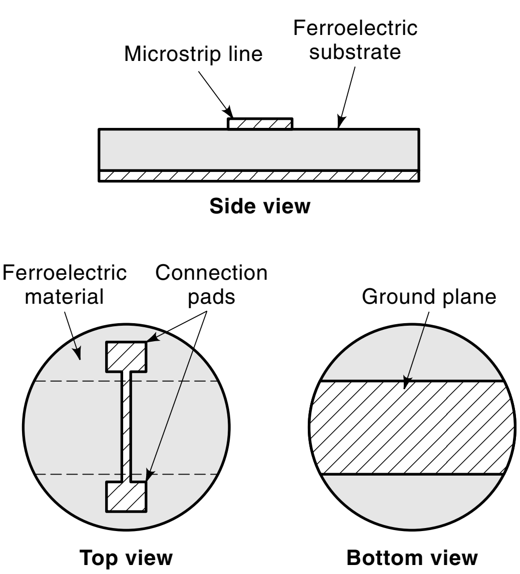


Figure 24. Microstrip ferroelectric type of analog phase shifter.

Also at $X = 0$ (resonance condition) a maximum insertion loss due to R_d will occur. The corresponding attenuation in this circumstance is

$$\alpha_{\text{dB}} = 20 \log_{10}\left(\frac{1 + R_d/Z_0}{1 - R_d/Z_0}\right) \tag{27}$$

The *figure of merit* (F) for the analyzed structure is calculated as

$$F = \frac{|\Delta\phi|_{\text{deg}}}{\alpha_{\text{dB}}} \tag{28}$$

A possible improvement of the presented structure can be obtained using two series varactor diodes as presented in Ref. 29. The operational principle remains the same, but a larger change in the capacitance is obtained.

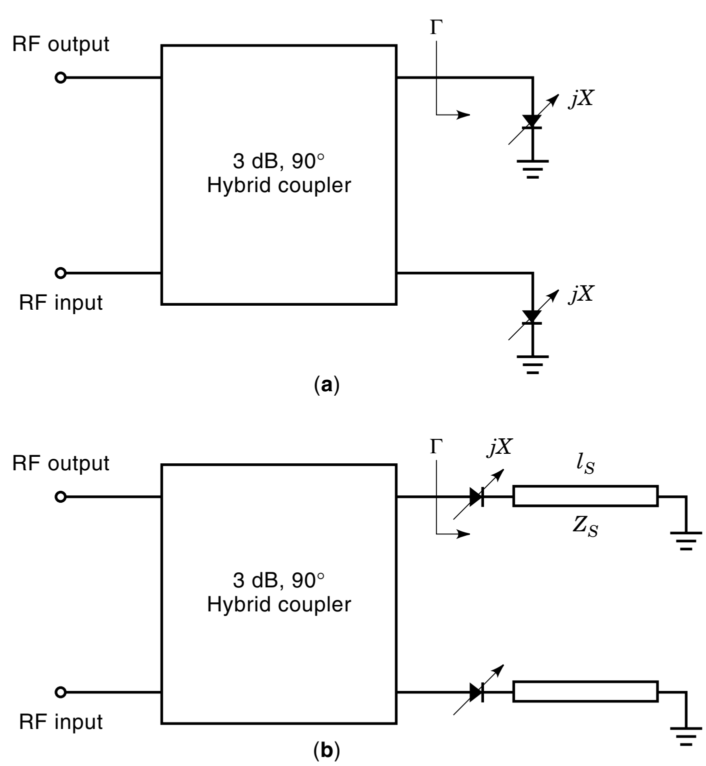


Figure 25. Varactor-based tunable phase shifter.

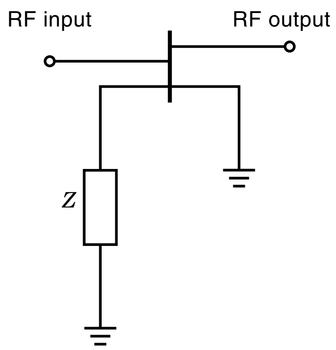


Figure 26. Active FET type of phase shifter.

Active Phase Shifter

Use of FET in an analog phase shifter (30) allows one to take advantage of the gain of the FET at microwave frequencies, while producing the time delay at the same time. Figure 26 shows the topology of this kind of phase shifter. The phase variation in the transmission coefficient (S_{21}) is achieved by controlling the bias voltage at the gate of the FET. The bias voltage is applied on the second gate of the FET, while a fixed inductive load is connected to it. The bias voltage will change the capacitance between the first gate (G_1) and the source, and this will change the amplitude and phase of the S_{21}. One limitation of this topology is the narrow bandwidth which is achieved. Use of more complicated topologies as reported in Ref. 31 will allow larger bandwidth and larger phase shifting capabilities.

ACKNOWLEDGMENTS

The author wishes to thank Professor N. G. Alexopoulos of the Electrical and Computer Engineering Department at the University of California, Irvine, for useful suggestions.

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See also FERRITE PHASE SHIFTERS; MICROWAVE PHASE SHIFTERS.