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A DIODE CONTROLLED CONTINUOUSLY ADJUSTABLE DIGITAL PHASE CHANGER AND ITS APPLICATION AT SLAC*

This note will describe the functioning, design and utility at SLAC of a diode controlled continuously adjustable digital phase changer, CDP. The CDP is continuous in the following sense: The phase difference between maximum and minimum phase shift across the device is 360^o. If the CDP is "turned" in the direction of increasing phase and its phase shift reaches a maximum it will decrease by 360^o (reset to zero) and then the phase will continue to increase. Similarly, when the phase is decreasing and it reaches a minimum the CDP will reload and the phase shift will increase by 360^o and then will continue to decrease. If one disregards what happens during the transition, the changes of -360 or +360 cannot be detected (by an electron beam or any other means).

The name Serrodyne[1] was given to an analog and the name Digilator[2] was given to a digital continuous phase changer. In this note unless otherwise qualified CDP will stand for a diode controlled microstrip adjustable digital phase changer.

A. Functioning of the CDP

The CDP consists of several binarily weighted differential phase shift sections in series, each section controlled by the corresponding bit in a counter. An outline of a CDP is shown in Fig. 1. With a number n in the counter the phase shift is

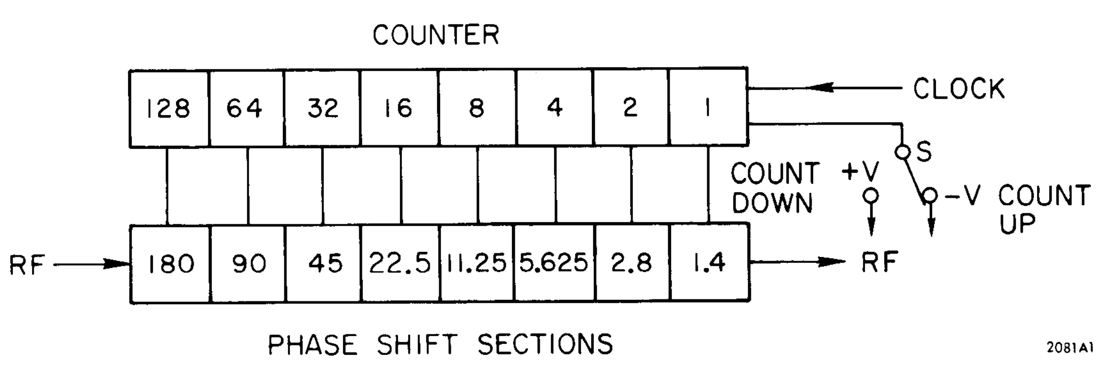

FIG. 1--Continuous digital phase shifter.

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$n(360/2)^N$ degrees where N is the number of sections used, which is also the number of bits in the counter. The resolution, the least significant bit, LSB, is $360/2^N$ degrees. With 8 sections (8 bits) the resolution is 1.4^o. With each additional section the resolution is halved. It is possible to obtain a one-to-one correspondence between degrees and the number in the counter by making the LSB 1^o and using nine sections. The total differential phase shift is 511^o but the counter is reset at 359^o. The differential phase shifts can also be in BCD in which case ten sections will yield a one-to-one correspondence between the number in the counter and degrees. In the latter case the phase shift can be inserted and read out by means of thumbwheel switches. It can also be read out by BCD to seven segment converter. The price for simplified encoding and readout is two additional phase shift sections.

It is possible to obtain higher precision by the use of several digital and a "vernier" analog phase shift section as shown in Fig. 2. The number in the first

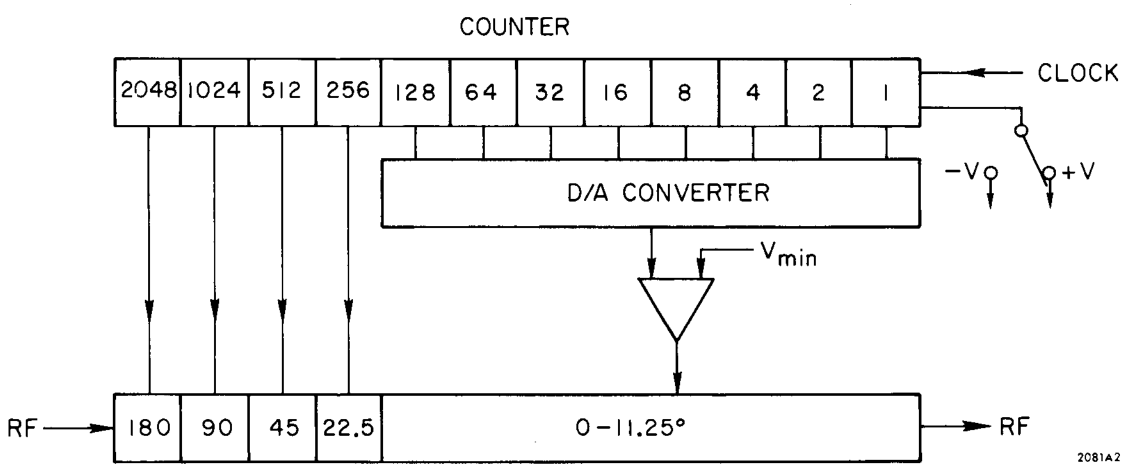

FIG. 2--Hybrid digital-analog-continuous phase shifter.

eight bits of the counter is converted to an analog voltage ΔV, which is used to control the vernier phase shift. If the vernier phase shift is linear with ΔV and is 11.25^o when ΔV is a maximum then the number in the counter is the phase shift across the CDP in units of $360/2^{12}$ ($.008^o$ for the 12 bit counter).

The use of a vernier analog section increases the precision but not the accuracy of the CDP. If the CDP reads 180^o and 180^o is in error by say 1^o one cannot set the CDP to 180.5^o. However, the CDP can be used as a coarse phase shifter to set the phase to near the middle range of a fine phase shifter.

Although the use of the analog phase shift section lowers the power capability of the phase shifter, the small phase shift excursion demanded from it allows the adjustment of the average phase point for maximum power handling and will have low phase sensitivity with temperature and power variations and a low, constant insertion loss over its phase range. If it were not for the above limitations the "vernier" could be 360^o thus eliminating the discrete sections. Varactor phase shifters with greater than 360^o phase shift range have been constructed.[3,4]

The phase shifter is turned by causing the counter to count up or down depending on the position of switch s. Refer to Fig. 1. When the phase shift is increasing and reaches 358.6^o the counter automatically clears and the next phase shift is zero or its equivalent 360^o, and the phase shift continues to increase. Similarly when the counter is counting down the phase shift is decreasing and when it reaches 0 the counter will reload and the next phase shift will be 358.6^o and will continue to decrease. Thus there is no discontinuity in phase shift.

The CDP can be turned using a clock with a period between pulses from nsec to seconds or with manually operated switches. The turning speed is limited by the switching speed of the diodes. The speed can be increased by using binary varactors instead of diodes. The phase can be read out as a binary number indicated by, say, LEDs, or by displaying the number that is in the counter. Simple systems exist that convert the binary numbers to decimal readout.

An automatic phase measuring system can be constructed by making the CDP part of a closed loop system. The output from the CDP, P_2, and a reference signal P_1, are fed into a phase detector whose output voltage is proportional to sin ϕ, where ϕ is the phase lag (delay) of P_2 with respect to P_1. If ϕ is between zero and 180, P_2 lags P_1, the detector output is positive and the counter will count down removing phase delay from the path of P_2, until the detector output drops to zero. If ϕ is between 0 and 180, P_2 leads P_1, the detector output is negative and the counter will count up inserting phase delay in the path of P_2 until the detector output is zero. Thus the phase of P_2 will always be driven toward the null located on the positive slope portion of voltage vs phase characteristic, and the phase between P_2 and P_1 at the detector input is maintained constant (zero). The variation of the reference phase is duplicated by the CDP. The change in phase between P_2 and P_1 is indicated by the change of the number in the counter.

If an auxiliary counter with no control of CDP but ganged to master counter is used and if at the fiducial phase between P_2 and P_1 the number in it is set to zero then the phase between P_2 and P_1 relative to the fiducial phase is given directly by the number in the auxiliary counter. This is illustrated as follows. Let us assume that the number in the counter is also the phase shift across the CDP. Let us further assume that the CDP reads 330^o and we delay the reference by 40^o. The output of the detector is negative and the counter will count up until the phase shift across the CDP is 359, when it will reset to zero, i.e., change by -359^o. This change has no effect on the detector output and the counter will continue to count up until the phase shift across the CDP is 10^o. Thus the change in phase in the CDP duplicates and the auxiliary counter indicates the change in phase of the reference. (The 359 change does not count.) Another way of looking at this is to note that the phase across the CDP changed by -320^o which is equivalent to 40^o since 360^o can always be added or subtracted.

The counter readout can be used to measure the phase shift of a pulsed signal without an oscilloscope as long as there is sufficient time for counting and changing of phase of the CDP. Counters can work at 100 MHz rates and the phase shifter can be made to switch in 10 nsec therefore 10 nsec pulses can be handled, as long as they are repetitive.

An alternative to counter readout is a D/A converter. The voltage output of the D/A is a sawtooth voltage. Operation at the extremities of the sawtooth can be avoided by using an auxiliary counter and setting the number in it in the middle of its range at the initial phase setting. The D/A output can be displayed on a dc meter even though the signal is pulsed.

B. Design of Bit Phase Shift Sections

It is possible to make a two value phase shift section by switching in a differential path, by switching in a two position short circuit at one port of a three port circulator or at two ports of a 3 dB hybrid and by loading the line with two different susceptances.[5] In this note a phase shift section consisting of two susceptances, B, separated by a transmission line of length $\beta \ell$ as shown in Fig. 3 will be considered.

If the characteristic admittance of the intervening line is unity then that the constants of the prototype section are related by:

$$B^2 - 2B \cot \beta\ell + (G^2 - 1) = 0 \qquad (1)$$

where G is the iterative admittance of the section. The admittances of the prototype section are normalized to the intervening line section. If the line is to be

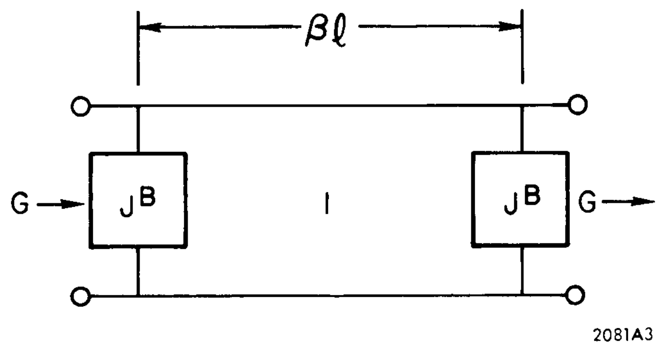

FIG. 3--Prototype differential phase shift section.

terminated by G_0 instead of G then all admittances should be multiplied by G_0/G. This type of admittance leveling is widely used in filter synthesis.

The phase shift, ϕ, of the section is:

$$\phi = \cos^{-1}(\cos\beta\ell - B\sin\beta\ell) \qquad (2)$$

Equation (1) is a quadratic in B and therefore in general there are two values of B, B_1 and B_2 for a given G and a given $\beta\ell$. The change in phase shift as B changes from B_1 to B_2 is

$$\Delta\phi = \cos^{-1}(\cos\beta\ell - B_2\sin\beta\ell) - \cos^{-1}(\cos\beta\ell - B_1\sin\beta\ell) \qquad (3)$$

Five special solutions of (1) are possible. They can be delineated with the aid of the Smith chart. Inspection of the chart will show that the conductances are symmetric and the susceptances antisymmetric about the real axis. As a consequence an admittance G + jB on a line $\phi_\Gamma = \beta\ell$, or $\phi_\Gamma = -\beta\ell$ transformed by a line length $\beta\ell$ or $180 - \beta\ell$, (the angle on the Smith chart is twice the line length) will become its complex conjugate G - jB. If jB is added at this point the section input admittance becomes G which is, therefore, the iterative admittance of the section. Thus if there are two susceptances for the same G, on the line $\phi_\Gamma = \beta\ell$ or one $\phi_\Gamma = \beta\ell$ and the other on $\phi_\Gamma = 180 - \beta\ell$ we have a perfect match for both. The two values for B that yield a match are plotted on a Smith chart shown in Fig. 4. The Smith chart plot does not give information about the phase differential due to the change in B. This is obtained from Eq. (3).

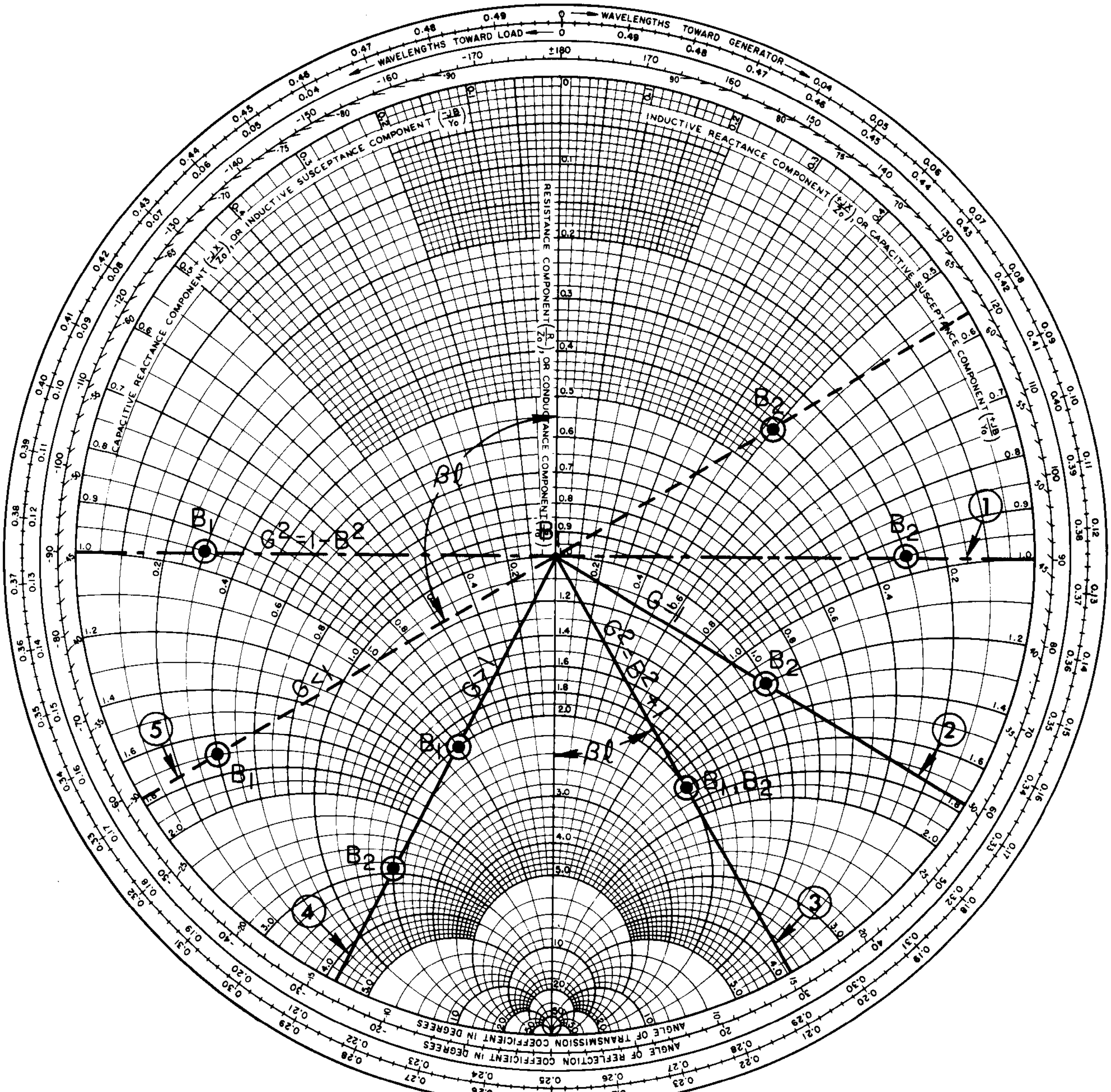

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FIG. 4--Two susceptances yielding identical characteristic admittance.

The five special solutions of Eq. (1) are:

CASE 1. One root is zero. Since the product of the roots of a quadratic is c/a, we obtain from (1)

$$G = 1 \qquad B_2 = 2 \cot \beta \ell$$

On the Smith chart B_1 is at the origin and B_2 is at the intersection of the line $\phi_\Gamma = \beta\ell$ or $\phi_\Gamma = -\beta\ell$ and the G = 1 constant conductance circle.

Substitution of zero for B_1 and 2 cot $\beta\ell$ for B_2 in (3) yields:

$$\Delta\phi = \cos^{-1} (\cos \beta\ell - 2 \cos \beta\ell) - \cos^{-1} \cos \beta\ell = (180 - \beta\ell) - \beta\ell = 180 - 2\beta\ell$$

As a function of $\Delta\phi$: $\beta\ell = 90 - \Delta\phi/2$; $B_2 = 2 \tan \Delta\phi/2$. G = 1. In the neighborhood of $\Delta\phi = 0$, $B_2 = \Delta\phi$. Table 1 lists the values of B_2 and $\beta\ell$ required for a given phase shift change $\Delta\phi$, as the shunt susceptance change from zero to B_2.

TABLE 1

Shunt Susceptances and Line Length for a Phase Differential Section Design Based on Case 1. G = 1, $B_1 = 0$

$\Delta\phi$	$\beta\ell$	B_2
1.0	89.5	.0175
2.0	89.5	.0349
4.0	88.0	.0689
8.0	86.0	.149
22.5	78.75	.398
45.0	67.5	.828
90.0	45.0	2.0
120.0	30.0	3.464
150.0	15.0	7.464
175.0	2.5	45.8
180.0	0	∞

As $\Delta\phi$ varies, from 0 to 180, B_2 varies from zero to ∞ and $\beta\ell$ from, 90° to zero. The table is valid under the following transformations $\Delta\phi \longrightarrow -\Delta\phi$, $B_2 \longrightarrow -B_2$ and $\beta\ell \longrightarrow 180 - \beta\ell$. In this case as $\Delta\phi$ varies from 0 to -180 B_2 varies from zero to $-\infty$ and $\beta\ell$ from 90 to 180.

CASE 2. The roots are equal in magnitude and opposite in sign $B_2 = -B_1$. Since the sum of the roots of a quadratic is -b/a we obtain from (1)

$$\beta \ell = 90 \qquad G^2 = 1 - B^2$$

On the Smith chart the values of B are at the intersection of the line $\phi_\Gamma = \pm 90$ and a $(0 < G < 1)$ constant conductance circle.

Substitution of 90° for $\beta\ell$, -B for B_1, and B for B_2 in (3) yields:

$$\Delta\phi = \cos^{-1}(-B) - \cos^{-1}B = 90 + \sin^{-1} B - (90 - \sin^{-1}B) = 2 \sin^{-1} B$$

As a function of $\Delta\phi$

$$B = \sin \Delta\phi/2; \qquad G = \cos \Delta\phi/2;$$

In the neighborhood of zero $\Delta\phi = \Delta B$. Table 2 lists the magnitude of B and the iterative admittance of the prototype section for a phase shift change $\Delta\phi$ as B changes from negative to positive. Of course, $\Delta\phi$ is negative as B changes from positive to negative.

TABLE 2

Shunt Susceptance and Line Characteristic Admittance for a Matched, $\Delta\phi$ Differential Phase Section, for a Design Based on Case 2. B Changes from a Negative to a Positive Value. $\beta\ell = 90^{\circ}$.

$\Delta\phi$	$\lvert B \rvert$	G
1	.0087	1.000
2	.017	1.000
4	.035	.999
8	.070	.998
22.5	.195	.981
45	.383	.924
90	.707	.707
120	.866	.500
150	.996	.259
175	.999	.044
180	1.000	0.000

It can be shown that a network consisting of two susceptances of normalized amplitude ϵ, where $\epsilon \ll 1$, separated by a 90° line section, has an input VSWR of $1 + \epsilon^2/2$. Hence for small $\Delta\phi$ VSWR $= 1 + B^2/2 = 1 + (\Delta\phi)^2/4$. For $\Delta\phi$ less than 25° the VSWR is less than 1.05 and there is no need to change the characteristic impedance of the intervening line.

CASE 3. Roots are equal. $B_1 = B_2 = B$. Since the sum of the roots of a quadratic equation equals $-b/a$ and the product of the roots equals c/a we obtain from (1)

$$B = \cot \beta\ell \qquad G^2 = B^2 + 1$$

This result is also a specific case of a general design by Dawris.[6]

On the Smith chart B is at the point of tangency of the line $\phi_\Gamma = \beta\ell$ or $\phi_\Gamma = -\beta\ell$ and a $G = (B^2 + 1)^{1/2}$ constant conductance circle.

The differential network based on design 3 is perfectly matched at a single value of B only. However it is nearly matched over a continuous range of B's for phase excursions less than $\pm 45^\circ$. Substituting $\cot \beta\ell$ for B in (2) we obtain $\phi = \cos^{-1}(\cos \beta\ell - \cos \beta\ell) = 90^\circ$. B has the same magnitude but changes from positive to negative as $\beta\ell$ changes to $180 - \beta\ell$.

$$\phi' = \left[1 - (\cos \beta\ell - B \sin \beta\ell)^2\right]^{-1/2} \sin \beta\ell = \sin \beta\ell = 1/G$$

Hence $\Delta\phi = \Delta B/G$.

CASE 4. Roots are of unequal magnitude but of the same sign, i.e., both roots are positive or both negative. On the Smith chart the susceptances are at the intersection of the line $\phi_\Gamma = \beta\ell$ (B's are both positive) or $\phi_\Gamma = -\beta\ell$ (B's are both negative) and $G > 1$ constant conductance circle. $|\beta\ell|$ is less than 90°.

CASE 5. Roots are of unequal amplitude and of opposite sign. On the Smith chart the positive susceptance is at the intersection of line $\phi_\Gamma = \beta\ell$ and the negative susceptance is at the intersection of the line $\phi_\Gamma = 180 - \beta\ell$ and a $0 < G < 1$ constant conductance circle.

Specific requirements determine on which case to base the design of a given phase shift section. All designs can be used to yield a phase excursion by varying B about its perfect match value. An idea of the resulting mismatch can be obtained by observing the variation in G as B is varied about its design value along a radius. It is obvious that G varies least with B at a point of tangency as in Case 3. For large phase variations Case 5 might yield a smaller mismatch than Case 3 over a given range. Several sections can be cascaded to yield a small VSWR over a given range.

Best suited for two value phase shift design are designs based on Case 1 and 2. Design 1 would seem simpler because it is easier to make the physical length of the intervening transmission line so as to obtain the required electrical length than to make its width so as to obtain the required characteristic impedance.

C. Design of Shunt Networks

The two susceptance values for both design 1 and design 2 are obtained by changing the termination of a section of transmission line from open to short circuit. This is accomplished by changing the bias of a PIN diode that is incorporated into a network whose input admittance is a high resistance when the diode is reverse biased and a low resistance when the diode is forward biased. For minimum loss the diode has to be a part of a network whose input is a pure resistance in both forward and reverse bias condition.[7] Such a network was developed at SLAC in microstrip for the HP82-3030 diode at 2856 MHz. The length of the open circuit line section in series with the diode was varied until the null as indicated on a slotted line shifted 90° as the diode bias reversed. The position of the purely resistive plane nearest to the diode was noted. At the input to the network the measured forward conductance, g_f, was 250 (50) mhos, and the reverse conductance, g_r, was (1/500) 50 mhos. Because the diode resistance at both open and short circuit conditions is finite there is a minimum attenuation even with optimum design. Fortunately the attenuation decreases as $\sin \Delta\phi/2$.[5,7]

A shunt network for a type 1 design whose input susceptance varies between zero and a finite value is shown in Fig. 5. It consists of a 90° line of characteristic

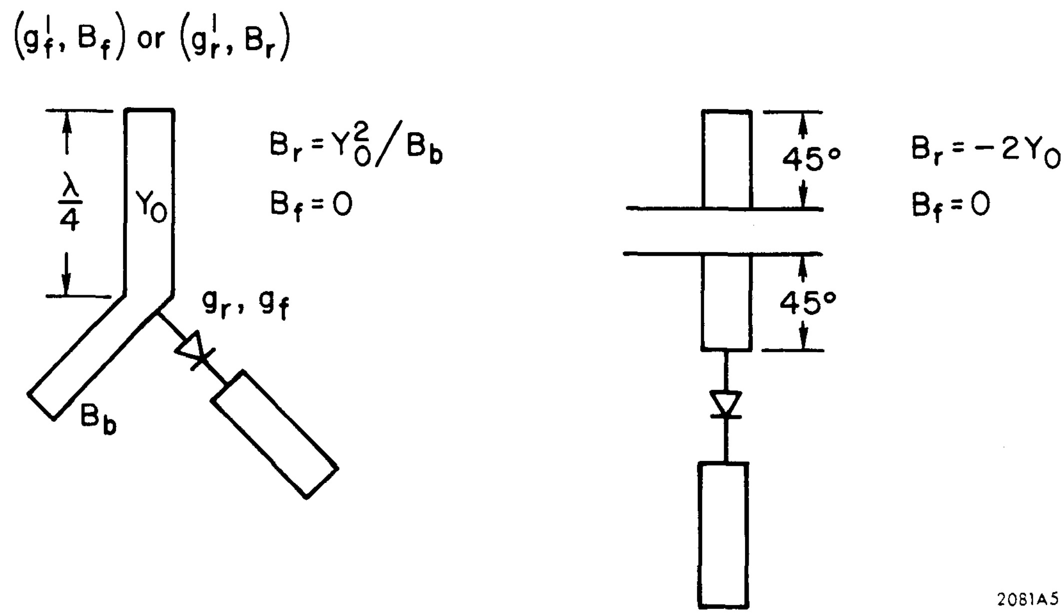

FIG. 5--Sections that yield approximately zero susceptance when diode is in forward conducting state and the proper susceptance to yield the required phase shift when in reversed state.

admittance Y_0 terminated by the parallel combination of diode conductance, g_f when the diode is forward biased or g_r, when reverse biased and an open circuited stub having a susceptance jB_b. For the conditions $g_f^2 \gg B_b^2$ and $g_r^2 \gg B_b^2$ the following approximations of the input admittances are valid

$$g_f' = \frac{Y_0^2}{g_f}, \qquad B_f = \frac{-Y_0^2 B_b}{g_f^2} \approx 0$$

$$g_r' = \frac{Y_0^2 g_r}{B_b^2}, \qquad B_r = \frac{-Y_0^2}{B_b} = B(G_0/G) = G_0 \tan \Delta\phi/2$$

hence $B_b = (-Y_0^2/G_0) \cot \Delta\phi/2$.

The attenuation due to a conductance if shunted across a line of characteristic admittance G_0, where $g/G_0 \ll 1$, is $A = 4.343\ g/G_0$ hence the attenuation due to the two diodes of a section is $A = 8.686\ g/G_0$. If $Y_0 = G_0$

$$A_f = 8.686\ G_0/g_f \qquad A_r = 8.686\ (g_r/G_0) \tan^2 \Delta\phi/2$$

It is possible to equalize the loss for the two phase shifts by making $g_f' = g_r'$. This yields

$$B_b^2 = g_f g_r$$

$$Y_0^2 = B_b B_r = B_r (g_f g_r)^{1/2} = G_0 \tan \frac{\Delta\phi}{2} (g_f g_r)^{1/2}$$

$$g_f' = g_r' = (g_r/g_f)^{1/2}\ B_r = (g_r/g_f)^{1/2}\ G_0 \tan \frac{\Delta\phi}{2}$$

and

$$A = 8.686\ (g_r/g_f)^{1/2} \tan \frac{\Delta\phi}{2}$$

Another type of shunt network for type 1 design is also shown in Fig. 5. It consists of two 45^o lines one on each side of the through line, one is terminated by the diode and the other by an open circuit. When the diode is forward biased we have two shunt susceptances of opposite sign and of amplitude, Y_0, in parallel with the line, so that the net shunt susceptance is zero. If the diode is reverse biased the net shunt susceptance is $-2Y_0$. If one of the shunt arms is terminated by a short circuit then when the diodes is reverse biased the net susceptance is zero and when the diodes are forward biased the net susceptance is $2Y_0$. The required values of Y_0 for a 90^o, 45^o and 22.5^o section respectively are $(G_0/2)^{1/2}$, $.767\ G_0/2$, $.530\ G_0/2$.

For type 2 design the shunt network whose input susceptance changes from B to -B consists of a diode placed at the end of line section 45^{o} long and having a characteristic admittance Y_0. Then the input admittances Y_f the diode is forward biased and Y_r when the diode is reverse biased are:[8]

$$Y_f = \frac{2Y_0}{g_f} - jY_0 \qquad Y_r = 2Y_0 g_r + jY_0$$

$Y_0 = B\ G_0/G = (\sin \Delta\phi/2)(G_0/G) = G_0 \tan \Delta\phi/2$. The input conductances g_f', g_r' for $\Delta\phi = 90^{o}$ are: $g_f' = .0057$, $g_r' = .0028$. The respective attenuation per section is $A_f = .05$ dB, $A_r = .02$ dB.

Another means of obtaining a $\pm$ B is to place the diode network in shunt at $90-\phi$ point of a line section and a short at $90+\phi$ point of the same line section. Then the input susceptance of the section is $Y_0 \tan \phi$ when the diode is reverse biased and $-Y_0 \tan \phi$ when the diode is forward biased.[5]

In both designs a 180^{o} bit phase shift section is impractical, and although it could be made by cascading two 90^{o} sections, nevertheless a 180^{o} section of the reflection type, will be considered.[5] The return loss of the reflection type shifter, over and above the loss of the circulator or hybrid that converts the one port into a two port, in the high VSWR limit is: $A = 17.4/\rho$, where ρ is the VSWR of the diode network. Thus the loss for the HP82-3039 diode network in the forward and reverse conditions respectively is .069 and .035 dB. The loss can be equalized by putting the diode network at the end of a 90^{o} line section of characteristic admittance chosen to yield the same VSWR in each of the two states. This yields $Y_0 = (g_f g_r)^{1/4} = (\rho_r/\rho_f)^{1/4}$, $\rho = (g_f/g_r)^{1/2}$ and specifically for the HP82-3039 diodes $Y_0 = 1.2$, $\rho = 355$, and $A = .05$ dB.

The inner conductor, more or less to scale, of a microstrip (strip line) 4 bit section plus a vernier section phase shifter is shown in Fig. 6.

D. Applications

1. A remote controlled variable phase shifter with digital readout that can be used for general phase setting and phase measurement.

2. Replacement for the present SB phase shifters. The two polarity signal that turns the present Fox phase shifter can be used to tell the counter whether to count up or down. The clock rate can be set to give for the operator the most comfortable degrees/sec speed. Presently the change in phase shift indication is a dc

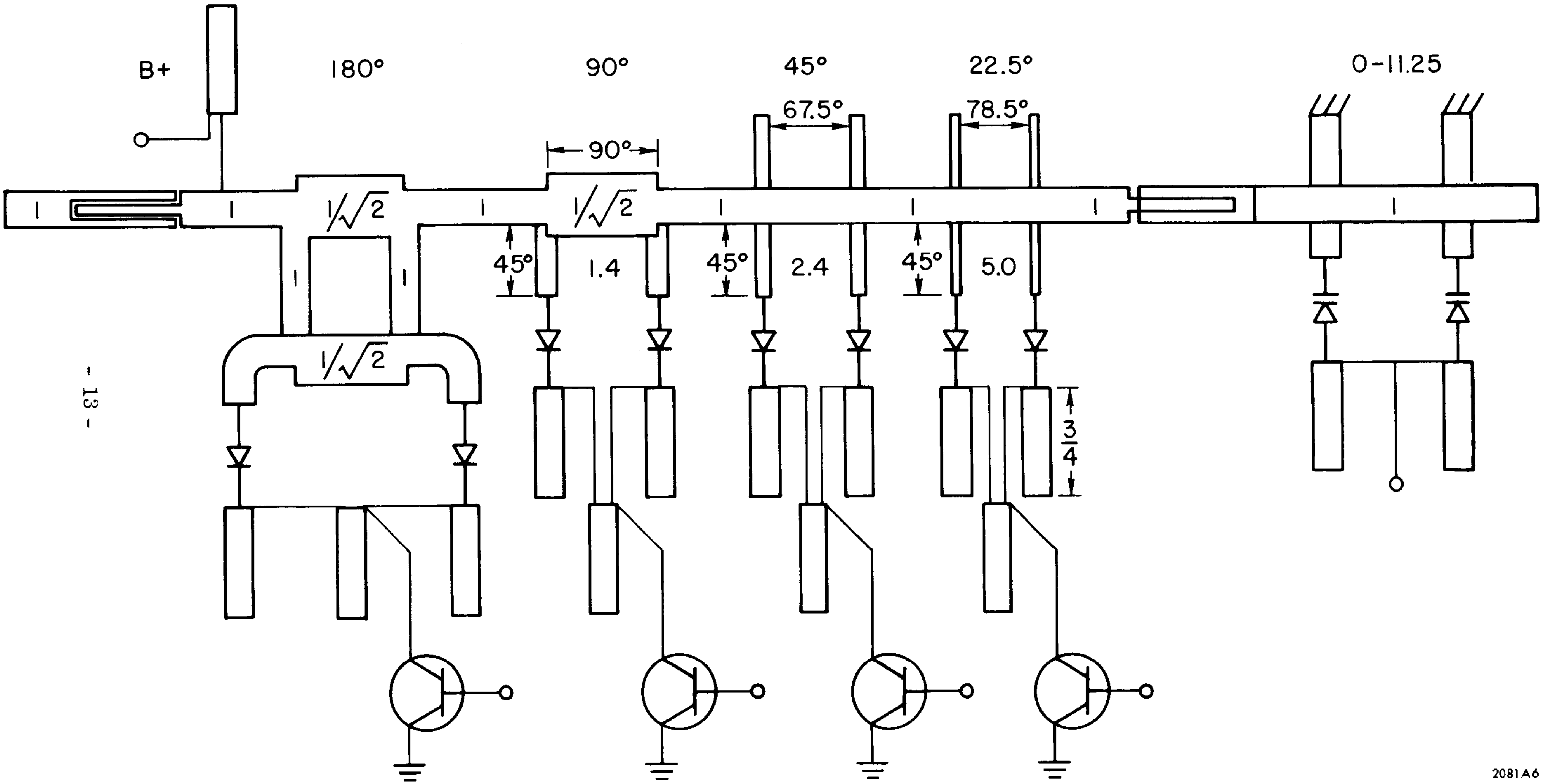

FIG. 6--Microstrip or stripline inner conductor outline approximately to scale of a 4 discrete and analog section CDP.

signal generated by means of a slide back potentiometer. Such a signal can also be obtained with the CDP by means of a D/A converter. However, a better way of indicating the change of phase shift is by keeping track of the number of pulses.

With the CDP it is not necessary to use an additional 180^{o} phase wobbler for positrons acceleration or back phasing. Once the phase is set, 180^{o} or any other number of degrees can be wobbled by a signal to the appropriate diodes. The CDP can also be used for energy transient compensation if driven so as to yield a programmed phase shift across the pulse. The rf pulse can be divided into n time intervals and the phase during each time interval can be controlled independently. This can be accomplished as follows: The CDP is driven by n counters. The outputs of the first counter, i.e., the one that controls the phase during the first time interval go to the first inputs of m, n bit counters (m is the number of phase shift bits) the outputs of the second counter go to the second input of the n bit counters and so on. The output of each n bit counter goes to a parallel to serial converter whose output in turn controls the phase shift of one CDP bit. The first controls the first bit, etc. At the beginning of each pulse the n bit counters are triggered to spill out in serial for m their stored input. In this manner each of the n counters controls the phase during one of the n time intervals.

It is also possible to have independent phase adjustment for n beams. Prior to each pulse the number in each of n counters is converted to n pulse trains. Each pulse train goes to one of n gates. The output of the n gates go into the n inputs of an "or" circuit whose output is inserted into the CDP counter (which must clear after each pulse). The gates are controlled by a pattern signal with only one gate open at a time.

3. Replacement of the reference phase shifters of the phasing system. The actuating drive to the reference phase shifters, which is a 60 cps square wave will have to be converted into a double polarity dc signal with a 60 cps synchronous detector.

Table 3 compares the digital microstrip-diode shifter with the Sperry-Fox phase shifter. Table 4 compares the microstrip-diode 180^{o} phase wobbler with a commercial circulator-short wobbler. The tables were prepared in order to delineate the advantage of the microstrip-diode CDP. They show that the cost of a CDP phase shifter nearly 1/10 that of the Sperry-Fox phase shifter and has no significant disadvantage but does have the advantage of speed and versatility. If the SB and rf phase shifter were to be procurred today a savings of $100, 000 could be realized by choosing the CDP over the Sperry-Fox phase shifters.

TABLE 3

Comparison of Microstrip-Diode Continuous Digital Phase Shifter with Sperry-Fox Phase Shifter

	Microstrip-Diode	Sperry-Fox Phase Shifter
Power	sufficient	sufficient
Attenuation	.5 ± .1 dB	.75 ± .5 dB
Resolution	± .7^{o}	± 1^{o}
Readout	number in counter	mechanical angle
Driver	I. C. counter and transistor switches	mechanical gears and dc motor
DC current	250 mA	required by motor
Speed	10 nanosec/degree Also 10 nanosec to go from any fixed angle to any other fixed angle	100 millisec/degree
Cost	Diodes 100 Mat. 50 Labor 50 Total 200	1500
Volume	small	large

TABLE 4

	Microstrip-Diode	Circulator-Short
Power	sufficient	sufficient
Attenuation	.5 ± .1 dB	.75 ± .5 dB
Speed	10 nsec	2 millisec
DC driving current	25 mA on; 0 mA off	50 mA on; 50 mA off
Maximum difference in att. of the two states	.05 dB	.15 dB
Volume	smaller	larger
Cost	Diodes 10 Mat. 20 Labor 20 Total 50	Circulator 310 Short 40 Total 350

4. Replacement of klystron phase shifters. There are diodes with ratings that enable the design of 360° digital phase shifters that could be used in lieu of the 24 MW klystron Fox phase shifters. An alternate means of constructing a high power CDP is to use ferrite torroids in rectangular waveguide for phase shift bits.[2]

5. Remote controlled phase shiftless variable attenuator (in conjunction with two hybrids).

6. A normalized phase detector in a closed loop system. The digital phase error in the CDP converted to analog voltage can be the error signal necessary say to tune an oscillator in order to lock it to a reference signal or a cavity. Since the zero phase indicator output does not effect the counting rate the phase error to corrective voltage transfer term, $dV/d\phi$, is constant and hence the loop gain is not effected by oscillator or reference output amplitude variation. Another advantage of the CDP in a closed loop system is its memory.

7. Digital phase and hence digital position indicator for TWM.

8. A phase shifter and indicator in a closed loop phase measuring system.

I would have like to report the performance of a completed CDP, but this is not possible because of other work. The developments toward a CDP completed so far are:

1. Transition to microstrip
2. Width for 50 and 71 ohm and effective dielectric constant at 2856 MHz 1/16" thick Texolite microstrip line.
3. Diode network that yields a pure resistance at a given plane for both forward reverse bias conditions.
4. Rat race hybrid.

The work that remains is:

1. Development of a T-junction.[9] This involves experimentally finding the reference planes of the T by placing a short circuit in turn in each of the three arms at a distance that will block transmission between the other two arms of the junction.

2. Development of branch-line coupler 180° section. Although the rat race hybrid can be used for this purpose, it is geometrically awkward.

3. Experimental determination of characteristic impedance (for type two design) with TDR methods.

4. Development of the two susceptance sections.
5. Putting everything together and providing bias connection to the diodes.
6. Bias drive system.

There are potential applications at SLAC to warrant the further development and construction of a CDP. For the expected quantities microstrip is most practical.

E. Conclusion

It was shown how to design and construct a diode controlled continuously adjustable phase changer which is much cheaper and more versatile than the Fox phase shifter and which has many potential applications for accelerator improvements and for RLA. Although the CDP has been extensively described in the literature none were designed for a center frequency of 2856 and none meet the specifications that this note indicates can be achieved. I believe that it can be inferred from this note that the completion of the development, and construction of a CDP is warranted.

Comments by, and information from H. Hogg, R. Koontz and A. Wilmunder, R. McConnell and the work of H. Martin are greatly appreciated.

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