Status Report on NsG-615

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STATUS REPORT ON NsG-615

NATIONAL AERONAUTICS AND SPACE ADMINISTRATION

This is a status report covering the period October 1, 1965 to March 31, 1966 for Grant NsG-615/21-02-029. The funding of this grant was not accomplished until March 9, 1966. Therefore, little progress can be reported during this period. However, since the time when the funds became available, the work has been actively pursued and the array which was discussed in the proposal is now under construction. The engineer in charge of the construction and operation of the array is moving to the Clark Lake Radio Observatory this month. Most of the required equipment has been received, is under construction, or is on order. The foundations for the antennas are complete, and the antennas are being erected.

In lieu of other items to report we are submitting a paper on the design of large arrays which has been written recently by Mr. Komesaroff. Mr. Komesaroff has now completed his two year visit to the United States, and is returning to Australia.

THE DESIGN OF LARGE STEERABLE ARRAYS USING VOLTAGE-VARIABLE CAPACITORS

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Summary

A method is proposed for building a rapidly steerable high resolution antenna array, suitable for radio astronomy. The method employs voltage variable capacitors spaced at equal intervals along the transmission line of a "travelling wave" array. Steering is accomplished by varying a DC voltage applied between the transmission line conductors.

The technique is most readily applicable at decameter wavelengths, since the necessarily low ohmic efficiency of the system is not a drawback at these wavelengths. It could, however, also be used at shorter wavelengths since integrated circuit amplifiers make it possible at relatively low cost, to amplify the output of each antenna element before coupling it into the transmission line.

I. INTRODUCTION

There are a number of astronomical problems relating to the Solar System, the Galaxy, and external galaxies which could be investigated by means of a decameter wavelength radio telescope with angular resolution considerably better than one degree. The present communication describes a simple and relatively inexpensive method by which such an array could be made fully and rapidly steerable.

It is obviously advantageous to make any radio telescope rapidly steerable, but in the decameter range the advantages are even greater than at shorter wavelengths, because of ionospheric effects. Due to refraction one cannot make accurate absolute position measurements of sources, but if the aerial beam could be swept rapidly from source to source, good relative measurements of neighboring sources would be possible. The same consideration applies to flux density measurements when absorption becomes significant. Also, because of scintillations, accurate flux densities may require long observing periods, and these may be achieved in one night with an instrument capable of following. The distortion of extended brightness distributions which results from time dependent refraction changing as the region drifts through the antenna beam, may be overcome by rapid scanning, especially with a multi-beam instrument. Finally, a steerable multi-beam instrument would be ideally suited to a study of these ionospheric effects.

To achieve full steering it would be necessary to place phase adjusting devices between each adjacent pair of antenna elements. If the elements were simple dipoles, there would be some hundreds or thousands of phase adjusters, distributed over several miles. For the steering to be rapid, all of them would need to be controllable from a central point. The following Sections show how steering may be achieved by shunting the transmission line of a "traveling wave" array with voltage-variable capacitors. The effect of the capacitors is to change the phase velocity and hence the phase gradient along the array; thus steering may be accomplished by merely varying a DC potential applied between the transmission line conductors.

It is not possible to build up a linear array of unlimited length in this way, since losses in the capacitors will impart an undesirable taper to the illumination pattern which in turn will produce unwanted sidelobes. However, using commercially available voltage variable capacitors it is practicable to build up a block of about twenty dipoles, which at decameter wavelengths is equal in collecting area to a reflector about one hundred feet in diameter. Having thus reduced the phasing problem by more than an order of magnitude, one can then use other phasing techniques in order to connect the blocks together.

An array of this kind necessarily has a low ohmic efficiency, due to losses in the capacitors, and also because of the need to use lightly coupled dipoles in order to avoid severe tapering of the illumination pattern. At decameter wavelengths the low efficiency is not a disadvantage, since the sensitivity limit is set by the general sky brightness temperature, which is one or two orders of magnitude greater than easily achievable receiver noise temperatures. The technique could, however, also be applied at shorter wavelengths, since the current development of inexpensive semi-conductor "integrated circuits" makes it economically feasible to amplify the signal from each individual radiating element.

In Section II of this paper, expressions are derived for the effective phase velocity and characteristic impedance, as well as the frequency dispersion and attenuation, along a transmission line loaded with identical equally spaced capacitors. In Section III these results are applied to the design of a "voltage steerable" array, and in Section IV some preliminary experimental results, obtained with a very small array, are described.

II. THE CAPACITY LOADED LINE

A. Closely-Spaced Capacitors. Consider a transmission line shunted with identical capacitors C, at intervals L along its length. The total added capacity per unit length is then $\mathcal{C}$ where $\mathcal{C} = \frac{C}{L}$, and it can be shown that for any wavelength λ such that $\lambda \gg L$ the effective phase velocity v and characteristic impedance Z_c of the loaded transmission line are given by

$$v = \frac{1}{\sqrt{\mathcal{L}_o(\mathcal{C}_o + \mathcal{C})}} \qquad (1)$$

$$Z_c = \sqrt{\frac{\mathcal{L}_o}{\mathcal{C}_o + \mathcal{C}}}$$

Here $\mathcal{L}_o$ and $\mathcal{C}_o$ are the distributed inductance and capacity per unit length of the line, all quantities being measured in M.K.S. units. For the unloaded line, v is equal to c, the free space velocity, i.e.

$$c = \frac{1}{\sqrt{\mathcal{L}_o \mathcal{C}_o}} \qquad (2)$$

Thus, the effect of adding the capacitors is to reduce both the phase velocity and characteristic impedance in the same ratio:

$$\frac{v}{c} = \frac{Z_c}{Z_o} = \sqrt{\frac{\mathcal{C}_o}{\mathcal{C}_o + \mathcal{C}}} \qquad (3)$$

If we lightly couple a number of equally spaced dipoles along the length of the line, then by changing the magnitude of the shunting capacitors, we can change the phase gradient along the array and hence the direction of its main response. An obvious disadvantage is the large change in Z_c which would accompany any change in pointing; this would lead to formidable impedance matching problems.

However, if $\frac{L}{\lambda}$ is not negligibly small, it is found that as $\mathcal{C}$ increases, Z_c decreases less rapidly than $\frac{v}{c}$. In particular, for $\frac{L}{\lambda} \simeq 1/10$, a variation of $\frac{v}{c}$ sufficient to steer the beam through more than ±45° may be achieved with only a small change in Z_c. In Section IIB, general expressions are derived for $\frac{v}{c}$ and Z_c for arbitrary values of L.

B. The General Ladder Network. Fig. 1 represents a uniform transmission line shunted with identical admittances Y spaced at intervals L along its length. The admittances at the ends are Y/2. We can analyze the response of such a network to a voltage impressed on one end if we know its iterative admittance and its propagation constant.

The iterative admittance has a value Y_c such that if Y_c is placed across one end of the network, the same admittance is seen looking into the other end. Thus, Y_c is clearly analogous to the reciprocal of the characteristic impedance of an ordinary transmission line. If the network is terminated at one end in Y_c, the propagation constant P per section is defined in terms of the complex ratio of the voltages across adjacent admittances by the relation

$$\exp(P) = \frac{V_k}{V_{k+1}} \tag{4}$$

where the k's are integers which increase from the "sending end".

To calculate Y_c and P we represent the ladder network as a cascaded arrangement of "half sections" as shown in Fig. 2. For one such half-section, there are two "image admittances", Y_{I1} and Y_{I2}, such that if Y_{I1} is placed across the terminals AB, and Y_{I2} across the terminals CD (Fig. 3), then the half-section is correctly matched at both ends. Clearly $Y_{I1} = Y_c$. If Y_{oc} is the admittance seen looking into terminals AB when no connections are made to C and D, and if Y_{sc} is the admittance when C and D are short circuited, it may be shown (cf. LePage and Seeley, p. 152)

$$Y_c = Y_{I1} = \sqrt{Y_{oc} \cdot Y_{sc}} \tag{5}$$

and

$$\tan h\left(\frac{P}{2}\right) = \sqrt{\frac{Y_{oc}}{Y_{sc}}} \tag{6}$$

(Y_{I2} could be calculated by carrying out the same procedure with the half section reversed.)

From the usual transmission line equations it follows that

$$Y_{oc} = (Y/2 + Y_o \tanh \frac{\gamma}{2}) \qquad (7)$$

$$Y_{sc} = (Y/2 + Y_o \coth \frac{\gamma}{2}) \qquad (8)$$

where Y_o is the reciprocal of the characteristic impedance Z_o of the transmission line, and γ is the propagation constant of a length L of the transmission line.

If we normalize all admittances with respect to Y_o and represent the normalized values with lower case letters, we may write, from (5, 6, 7, and 8)

$$y_c = [(\frac{y}{2} + \tanh \frac{\gamma}{2}) (\frac{y}{2} + \coth \frac{\gamma}{2})]^{\frac{1}{2}} \qquad (9)$$

$$\tanh \frac{P}{2} = \left[\frac{\frac{y}{2} + \tanh \frac{\gamma}{2}}{\frac{y}{2} + \coth \frac{\gamma}{2}}\right]^{\frac{1}{2}}$$

From the identity

$$\sinh P = \frac{2 \tanh \frac{P}{2}}{1 - \tanh^2 \frac{P}{2}}$$

it then follows that

$$\sinh P = y_c \sinh (\gamma) \qquad (10)$$

Which may be written

$$Z_c \sinh P = Z_o \sinh \gamma$$

C. Capacitors at Arbitrary Spacing. (1) Lossless Case. For the special case of perfect capacitors of capacity C spaced at equal intervals L along a lossless transmission line, as shown in Fig. 4,

$$y = Y/Y_o = i \frac{\omega C}{Y_o} = i\ b \qquad (11)$$

and $\gamma = i\theta$,

where $$\theta = \frac{2\pi L}{\lambda} \qquad (13)$$

Equations (9) and (10) then become

$$y_c = [(\tan \frac{\theta}{2} + \frac{b}{2}) (\cot \frac{\theta}{2} - \frac{b}{2})]^{\frac{1}{2}} \quad (13)$$

$$\sinh P = i\, y_c \sin \theta$$

We see that y_c is real (i.e. Z_c is resistive) provided $b < 2 \cot \frac{\theta}{2}$. Within the same range of b, sinh P is a pure imaginary, and therefore P is a pure imaginary. Writing $P = i\varphi$ it follows that

$$\sin \varphi = y_c \sin \theta \quad (14)$$

and

$$\cos \varphi = \cos \theta - \frac{b}{2} \sin \theta \quad (15)$$

Thus, if the line is terminated at one end in Y_c and has a voltage impressed on the other end, provided only that $b < 2 \cot \frac{\theta}{2}$, voltages of the same magnitude will appear across all capacitors. This follows from Eqn. 4, and the fact that P is a pure imaginary. The relative phases of the voltages are given by equation 14. If we lightly couple a dipole across every m-th capacitor, then except for radiation losses, each dipole will receive an excitation of the same amplitude, but the phase of the excitation will decrease by $m\varphi$ from dipole to dipole.

It is convenient to present the foregoing results in a slightly different form. For a uniform transmission line (without capacity loading) the characteristic impedance Z_o is given by

$$Z_o = \sqrt{\frac{\mathcal{L}_o}{\mathcal{C}_o}}$$

From this result and equations (2, 11, and 12), it follows that

$$b = \frac{2\pi L}{\lambda} \frac{\mathcal{C}}{\mathcal{C}_o}$$

Hence

$$\frac{Z_o}{Z_c} = y_c = [(\tan \frac{\pi L}{\lambda} + \frac{\mathcal{C}}{\mathcal{C}_o} \frac{\pi L}{\lambda}) (\cot \frac{\pi L}{\lambda} - \frac{\mathcal{C}}{\mathcal{C}_o} \frac{\pi L}{\lambda})]^{\frac{1}{2}} \quad (16)$$

$$\sin \varphi = \frac{Z_o}{Z_c} \sin \theta \quad (14a)$$

Writing $\frac{c}{v} = \frac{\varphi}{\theta}$, it is easily seen that for vanishingly small values of $\frac{L}{\lambda}$, equations 16 and 14a are equivalent to (3).

Fig. 5 is a graphical representation of equations (16) and (14a). It shows the dependence of $\frac{Z_o}{Z_c}$ and $\frac{c}{v}$ on $\frac{\mathcal{C}}{\mathcal{C}_o}$ and $\frac{L}{\lambda}$. Each solid line is drawn for a constant value of $\frac{L}{\lambda}$, and shows the variation of the first two quantities with $\frac{\mathcal{C}}{\mathcal{C}_o}$. The dashed lines refer to constant values of $\frac{\mathcal{C}}{\mathcal{C}_o}$ and show the variation with $\frac{L}{\lambda}$. Clearly, the dashed lines indicate the effect of keeping the physical configuration fixed while varying the operating wavelength.

It can be seen from the solid lines, that wheras $\frac{c}{v}$ is a montonically increasing function of $\frac{\mathcal{C}}{\mathcal{C}_o}$, for any fixed (non-zero) value of $\frac{L}{\lambda}$, $\frac{Z_o}{Z_c}$ increases to a maximum as $\frac{\mathcal{C}}{\mathcal{C}_o}$ increases, and then decreases. Thus, by varying $\frac{\mathcal{C}}{\mathcal{C}_o}$ it is possible to obtain a fairly larger change in $\frac{c}{v}$ with only a small change in $\frac{Z_o}{Z_c}$. For example, if $\frac{L}{\lambda} = \frac{1}{10}$, $(\frac{Z_o}{Z_c})_{max} = 1.7$ and the corresponding value of $\frac{c}{v}$ is 2.5.

(2) Effect of Losses.

In practice the main sources of loss in a line loaded with voltage-variable capacitors will be the capacitors themselves. If these are of high quality, the effect on $\frac{c}{v}$ and y_c will be small, but there will of course be some power attenuation.

Each capacitor may be represented as a pure capacitance shunted with a conductance G_p. If P is the power flowing along the loaded transmission line and ΔP is the power dissipated in the capacitor,

$$\frac{\Delta P}{P} = \frac{G_P}{Y_c} = \frac{g_P}{y_c} ,$$

provided

$$G_p \ll Y_c$$

If Q is the ratio of susceptance to conductance for the capacitor,

$$\frac{\Delta P}{P} = \frac{b}{Qy_c} = \frac{2\pi L \, \mathcal{C}}{y_c Q \lambda \mathcal{C}_o} \qquad (17)$$

(3) Frequency Dispersion

As $\frac{\ell}{\ell_o}$ increases the line becomes increasingly dispersive. If we write

$$\varphi = \frac{c}{v}\,\theta = \frac{c}{v} \cdot \frac{2\pi L}{\lambda} \ ,$$

then it follows that

$$\frac{d\varphi}{d\lambda} = -\frac{c}{v}\frac{\theta}{\lambda}\left\{1 - \frac{\lambda v}{c}\frac{d}{d\lambda}\left(\frac{c}{v}\right)\right\} \ ,$$

and hence

$$\frac{d}{d\lambda}(\cos\theta) = -\frac{c}{v}\frac{\theta}{\lambda}\left\{1 - \frac{\lambda v}{c}\frac{d}{\lambda}\left(\frac{c}{v}\right)\right\}\sin\theta$$

Differentiating the right hand side of equation (15) and equating the result with the expression just derived we find that

$$\frac{\lambda v}{c}\frac{d}{d\lambda}\left(\frac{c}{v}\right) = -\left\{\frac{v}{c} \cdot \frac{Z_c}{Z_o}\left[1 + \frac{\ell}{\ell_o}\left(1 + \frac{\theta}{\tan\theta}\right)\right] - 1\right\} \qquad (18)$$

For $\frac{L}{\lambda} = \frac{1}{10}$ and a value of $\frac{\ell}{\ell_o}$ such that $\frac{Z_o}{Z_c}$ has its maximum value,

$$\lambda\frac{v}{c}\frac{d}{d\lambda}\left(\frac{c}{v}\right) \simeq -0.20$$

III. DESIGN OF A VOLTAGE-STEERABLE ARRAY

A. General Considerations. To illustrate the design procedure, the results just derived will be applied to the case of an array operating at $\lambda = 11$ m. Various factors (which are discussed below) limit the number of radiating elements which can usefully be connected to a single voltage variable transmission line. Therefore, in order to synthesize a very large array it would be necessary to combine the outputs of many such blocks of elements. The present discussion will be restricted to the design of the individual blocks, since the problem of combining their outputs is analogous to building a large array from a number of reflectors, and this has been discussed extensively elsewhere. It will be assumed that the composite array is to have a pencil beam response representing the correlated outputs of two large orthogonal linear arrays. We will therefore be interested in the voltage rather than the power patterns of the linear arrays.

B. Range of Steering. In order that the array have no secondary response the spacing d between dipoles must not exceed approximately one-half wavelength, and if the direction of maximum response is to be approximately in the broadside direction at minimum capacity, the transmission line must be transposed between dipoles. Referring to Fig. 6, the phase difference ψ between the signals from adjacent dipoles due to a wave front making angle χ with the array axis is given by

$$\psi = \frac{2\pi d}{\lambda}\left(\frac{c}{v} - \sin\chi\right) - \pi \qquad (19)$$

If the array has its maximum response in the direction χ, then $\psi = 0$ and,

$$\sin\chi = \frac{c}{v} - \frac{\lambda}{2d} \qquad (20)$$

As $\frac{c}{v}$ increases χ increases. We will make $\left(\frac{c}{v}\right)_{min}$ only slightly less than $\frac{\lambda}{2d}$ so that χ has only a small negative range. In order to give χ larger negative values the "feed" and "termination" connections of the array will need to be interchanged. Voltage variable capacitors are available commercially, having capacitance which is variable through the range $10\ \mu\mu F \le C \le 33\ \mu\mu F$ and with Q's of 400 or better at this wavelength. The assumed characteristic impedance of the transmission line is 500Ω, corresponding to $\mathcal{C}_o = 6.67\ \mu\mu F/\text{meter}$. If the capacitors are spaced at intervals of $\frac{\lambda}{10}$, then

$$1.36 \le \frac{\mathcal{C}}{\mathcal{C}_o} \le 4.50$$

Correspondingly

$$1.56 \le \frac{c}{v} \le 2.53 \qquad (21)$$

If a dipole is shunted across every third capacitor, so that the spacing d between dipoles is 0.3λ, then it follows from (20) and (21) that $\chi_{min} < 0$, which is one condition we intended to satisfy. On the other hand

$$\sin \chi_{max} = 0.86$$

and therefore

$$\chi_{max} = 59°$$

Thus by interchanging the end connections, we may sweep the beam through $\pm 59°$. Over this range $y_{c_{min}} = 1.40$ and $y_{c_{max}} = 1.70$. Correspondingly $Z_{c_{max}} \simeq 360\ \Omega$ and $Z_{c_{min}} \simeq 290\ \Omega$. If each end is terminated in $\sqrt{Z_{c_{max}} \times Z_{c_{min}}}$ the V.S.W.R. will not exceed 1.10. For many applications this would be quite acceptable. If it is not, an asymmetrical π section consisting of a length of line and two voltage variable capacitors, could be placed at each end of the array to provide a varying impedance transformation.

C. Length of Each Block. Dissipative losses in the capacitors and radiation from the dipoles lead to a tapering of the illumination pattern which limits the number of dipoles that can usefully be connected together in this way. This is so because the tapered illumination gives rise to unwanted sidelobes whose amplitudes increase as the number of dipoles increases.

This may be shown in the following way. The voltage at the receiver due to the k-th dipole may be written as

$$V_o \exp(-kz)$$

where $z = \alpha + i\psi$, α and ψ being respectively the attenuation in nepers, and phase shift in radians from dipole to dipole. Thus for a block of n dipoles, the total voltage $V_n(\psi)$ is given by

$$V_n(\psi) = V_o \sum_{k=0}^{n-1} \exp(-kz)$$

$$= V_o \cdot \frac{1 - \exp(-nz)}{1 - \exp(-z)}$$

$$= V_o \cdot \exp \frac{-(n-1)z}{2} \cdot \frac{\sinh \frac{nz}{2}}{\sinh \frac{z}{2}}$$

Therefore

$$|V_n(\psi)| = V_o \exp \frac{-(n-1)}{2}\alpha \cdot \left| \frac{\sinh \frac{nz}{2}}{\sinh \frac{z}{2}} \right|$$

$$= V_o \exp \frac{-(n-1)}{2}\alpha \cdot \left\{ \frac{\sinh^2 \frac{n\alpha}{2} + \sin^2 \frac{n\psi}{2}}{\sinh^2 \frac{\alpha}{2} + \sin^2 \frac{\psi}{2}} \right\}^{\frac{1}{2}}$$

When N such blocks are combined into a large linear array, with all blocks equally illuminated, the response becomes

$$|V_{nN}| = V_o \exp \frac{-(n-1)\alpha}{2} \left\{ \frac{\sinh^2 \frac{n\alpha}{2} + \sin^2 \frac{n\psi}{2}}{\sinh^2 \frac{\alpha}{2} + \sin^2 \frac{\psi}{2}} \right\}^{\frac{1}{2}} \frac{\sin Nn \frac{\psi}{2}}{\sin n \frac{\psi}{2}} \qquad (22)$$

If $\alpha = 0$, the response of course reduces to

$$V_o \frac{\sin Nn \frac{\psi}{2}}{\sin \frac{\psi}{2}}$$

This is perfectly acceptable, and a brightness distribution observed with such a uniformily illuminated aperture can be converted, by a convolution process, into the distribution which would be observed with an array of the same total aperture, but with any desired illumination grading.

For the uniformly illuminated aperture, nulls occur in the response pattern for

$$n\psi = k\pi \quad ,$$

where k is any integer greater than zero. The main effect of making α non zero is to replace the nulls by additional side-lobes. It is easily shown from (22) that provided $\alpha \ll 1$, the ratio of the k-th additional side lobe to the main response is

$$\frac{1}{n} \frac{\sinh \frac{n\alpha}{2}}{\sin \frac{k\pi}{n}}$$

$$\simeq \frac{n\alpha}{2k\pi}$$

Hence if the first additional sidelobe is not to exceed 4% of the main response, $n\alpha$ must not exceed .25.

Substituting $\frac{L}{\lambda} = 0.1$, $\frac{\mathcal{C}}{\mathcal{C}_o} = 4.5$, $y_c = 1.7$ and $Q = 400$ in equation (17) we find that in the maximum capacity case, the fraction of power dissipated by each capacitor is approximately .004. Since there are three capacitors per dipole, the fraction of power dissipated in the capacitors is .012 per dipole. If the dipole coupling is adjusted so that each dipole radiates this same amount of power, the total power loss per dipole will be .024, and hence α will be .012. With 20 dipoles per block $n\alpha = 0.24$, which will keep unwanted sidelobes below 4% even in the most extreme beam position.

The ohmic efficiency of the array will be only about 19%, but since the sky brightness temperature at this wavelength is about 20,000°K, this will not lead to a significant loss in sensitivity.

D. Wavelength Coverage. Two effects limit the range of wavelengths over which this kind of array may be used. The first is the inherent bandwidth of the radiating elements. Obviously by using inherently broadband elements such as rhombic/or log periodic antennas, this limit may be largely overcome. For the present purpose, however, even a thin linear dipole will function as a fairly broadband antenna. Provided the length of the dipole is less than any of the observing wavelengths, its directivity pattern will vary only slowly with wavelength. Since the dipole is assumed to be lightly coupled to the array, any change in its input impedance with wavelength will lead to only a small change in the array efficiency and in the illumination taper along each block. The latter effect will merely change the side lobe level.

The more important effect is due to the traveling wave feed system. As the wavelength changes, the direction of the main response changes. Due to misalignment of the "beams" corresponding to the various wavelengths within the pass band, there will be a reduction in the response to a point source. This effect occurs with any traveling wave feed system, but is slightly greater with the voltage steerable array because of dispersion effects along the capacity-loaded lines.

The direction of the main response is given by

$$\sin \chi = \frac{c}{v} - \frac{\lambda}{2d} \tag{20}$$

For a small increment, $\Delta\lambda$, in λ, χ increases by $\Delta\chi$ where

$$\Delta\chi \cdot \cos \chi = \frac{\Delta\lambda}{\lambda} \left(\lambda \frac{d(\frac{c}{v})}{d\lambda} - \frac{\lambda}{2d}\right)$$

In the most extreme beam position $\lambda \frac{d}{d\lambda}(\frac{c}{v})$ has its largest value. From equation (18) this is -0.53. Since $\frac{\lambda}{2d} = 1.67$,

$$\Delta\chi = \frac{-\Delta\lambda}{\lambda} \cdot \frac{2.2}{\cos \chi}$$

As the individual blocks have a beamwidth (to half voltage points) of about 11°, the reduction in response corresponding to $\Delta\chi = \frac{2^\circ}{\cos \chi}$ is less than 10%. Thus if the total range of wavelength covered is $\lambda_o \pm \Delta\lambda$, and $\frac{2\Delta\lambda}{\lambda_o} < .03$, the reduction in the overall response will be very much less than 10% and almost independent of beam position. The above fractional bandwidth corresponds to a frequency range of more than 800 Kc/s.

It is of course possible to make a large change in the operating wavelength, but in that case it would be necessary to modify the device providing bias voltage to the capacitors, since the relationship between voltage and direction of pointing is wavelength-dependent. One could build a swept frequency instrument on this principle, but for this it would be necessary to modulate the controlling voltage in synchronism with the frequency sweep, in order that the beam direction should remain constant.

E. Thermal Effects. According to published data, typical temperature coefficients for voltage variable capacitors are less than about 300 parts per million per degree C. Hence a 50°C change in ambient temperature changes the diode capacity by less than 1.5%. Referring to Fig. 5 and our assumed design parameters, this corresponds to a change of .02 in $\frac{c}{v}$ for the extreme beam position. From equation (20), the corresponding change in χ is less than 2°. Where many blocks of antennas are combined into a large array, the main effect would be to increase the side lobe level. In such a case it might be necessary to provide a small adjustable component of bias voltage to compensate for ambient temperature changes if these are very large.

IV. AN EXPERIMENTAL ARRAY

To make a rough test of the general principle, a small 11.4 m East-West array was built at the University of Maryland's Clark Lake Radio Observatory. The radiating elements were full wave dipoles soldered across a 470Ω line at $\frac{5}{4}\lambda$ intervals. The capacitors were spaced at intervals of $\frac{\lambda}{4}$. The transmission line was not transposed between dipoles. Since it was a preliminary test the best quality capacitors were not used, the Q's being between 25 and 50. The array was operated in conjunction with a single dipole as a phase switched interferometer having a baseline of about 110 m.

The results of four nights' observations are shown in Fig. (7). The two observed sources are Cygnus A and Cassiopeia A. It can be seen that even with the limited range of capacity available, it was possible to steer the beam through hour angles of $1\frac{1}{2}^h$ to 2^h. The arrows indicate the calculated times for the sources to pass through the beam of the array, based on measurements of the capacitance vs. bias voltage characteristic for a few sample capacitors. Although the general trend is as calculated, it is clear that the effective capacity is consistently greater than estimated, especially at low bias values. There are two possible reasons for the discrepancy.

(1) In order to obtain the widest possible capacitance range with the available components it was necessary to go to very low values of bias for which the slope of the capacitor characteristic is very large, making accurate measurement difficult.

(2) Although the dipoles were cut to a length close to antiresonance, no effort was made to tune out residual reactance. The effect of any reactance would be most marked near the broadside position of the beam (corresponding to the lowest bias). This effect could be almost entirely eliminated if appropriate simple networks were used to provide very light coupling between the dipoles and transmission line.

A much more carefully engineered array is at present in construction which uses half wave dipoles and lumped constant π sections (simulating quarter wavelengths of high impedance transmission line) as the coupling elements. With this a much more precise test of the technique should be possible.

A final point worth noting in connection with Fig. (7), is that despite fairly intense scintillations in some cases, the interferometer patterns are fairly well defined. With the long observation times which continuous steering would make available, it should be possible to make flux density measurements in a single night comparable in accuracy with those made at shorter wavelengths.

V. CONCLUSIONS

It has been shown that using currently available voltage variable capacitors, small steerable decametric arrays consisting of about 20 elements may be built up, which are capable of being steered through more than 90°, and which have bandwidths of about 0.5 - 1 MHz. If the individual elements are half wave dipoles above a ground screen the collecting area of each is about $\frac{\lambda^2}{4}$, so that the total collecting area of a 20 element array operating at λ = 15 meter is about 1000 m^2, roughly equivalent to the area of a 40 meter reflector. The cost of the array would of course be several orders of magnitude less than the equivalent reflector and steering would be far more rapid. There are many ways in which the small arrays can be connected together to produce a large array. In particular this technique seems to be suitable for simultaneous synthesis of a number of beams where all the beams lie close to the maximum of the response pattern of each of the small arrays.

VI. ACKNOWLEDGEMENTS

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VII. REFERENCES

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LEGENDS TO FIGURES

Fig. 1 - Transmission line shunted with equally-spaced admittances.

Fig. 2 - Representation of the circuit of Fig. 1 as a cascaded arrangement of "half sections".

Fig. 3 - A single "half section".

Fig. 4 - "Capacity loaded" transmission line.

Fig. 5 - Relation between effective phase velocity v and characteristic impedance Z_c for the capacity loaded line. $\frac{C}{C_0}$ is the ratio of added capacity to distributed line capacity. $\frac{L}{\lambda}$ is the capacitor spacing in wavelengths.

Fig. 6 - A "traveling wave" array. Each alternative dipole has its connection to the transmission line reversed.

Fig. 7 - Four records of the sources Cygnus A and Cassiopeia A, taken using a voltage steerable array as one element of a phase switched interferometer. The markers indicate sidereal times, and arrows indicate calculated times of passage of the sources through the beam..

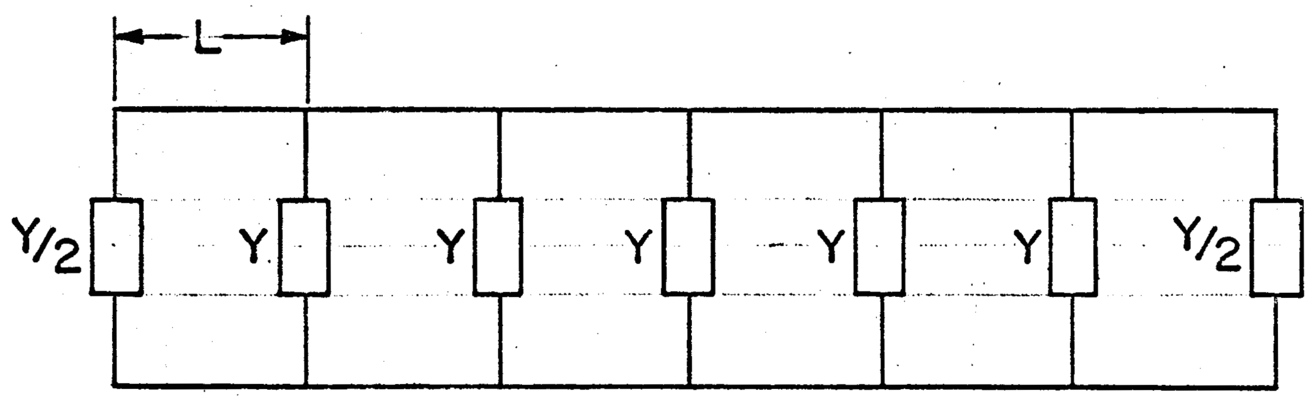


Fig. 1

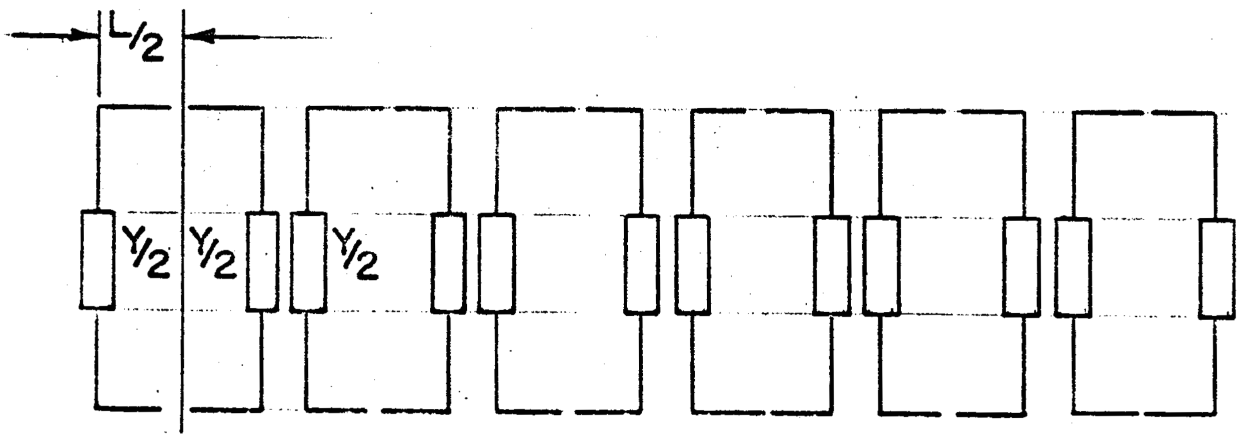


Fig. 2

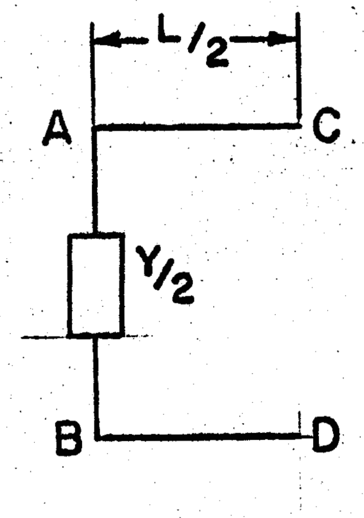


Fig. 3

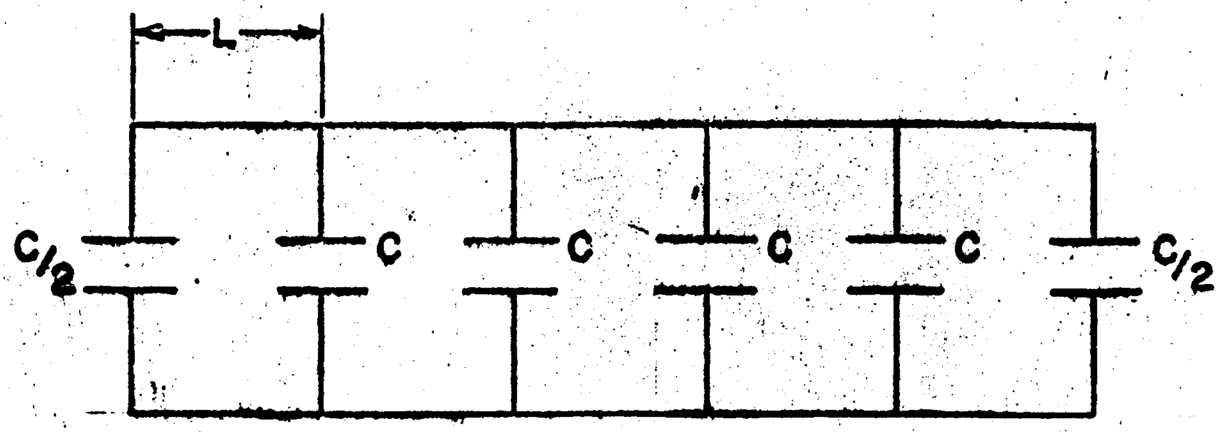
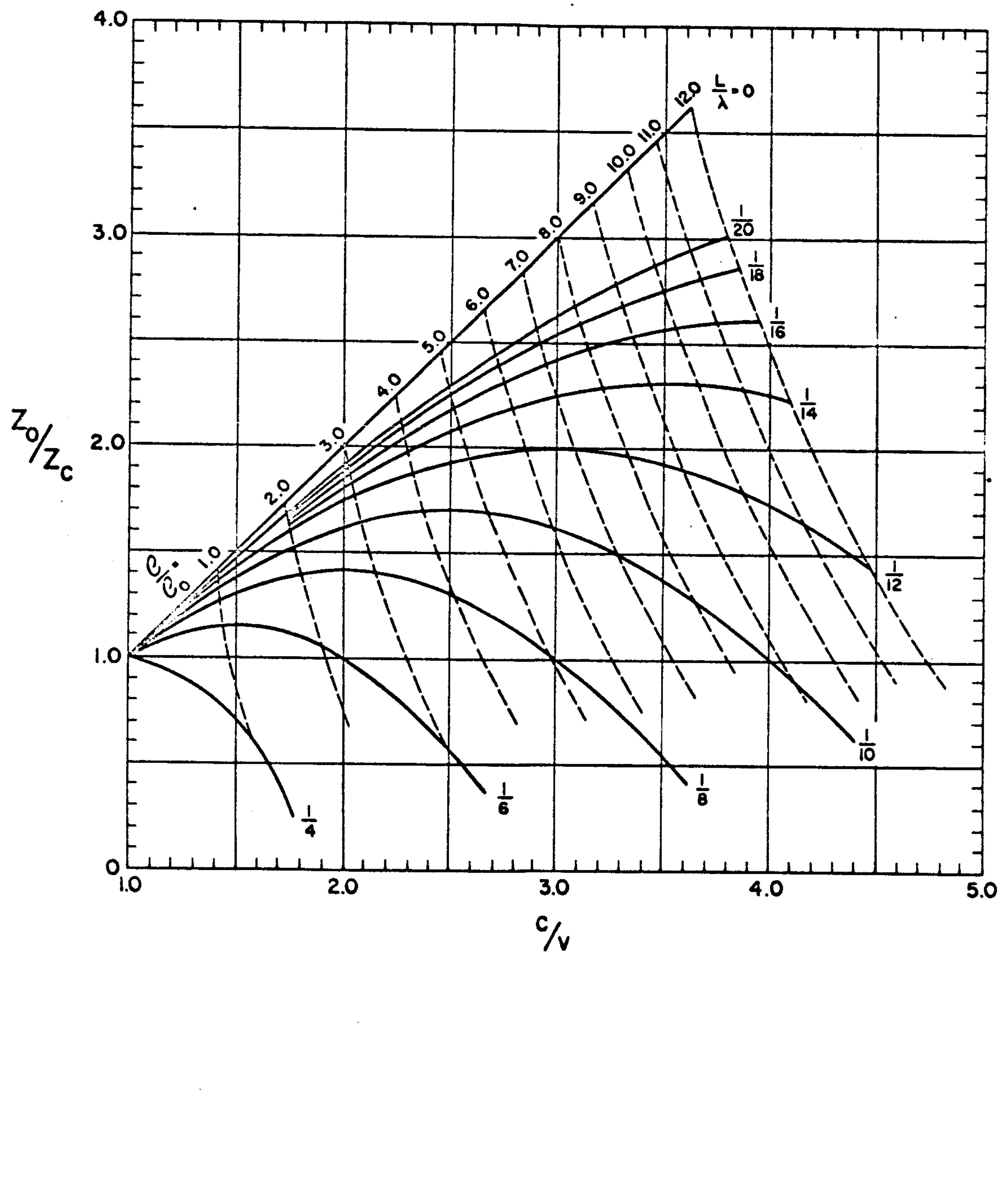


Fig. 4


4.0
3.0
2.0
1.0
0
Z_0/Z_c
1.0
2.0
3.0
4.0
5.0
c/v
$\frac{L}{\lambda}=0$
$\frac{1}{20}$
$\frac{1}{18}$
$\frac{1}{16}$
$\frac{1}{14}$
$\frac{1}{12}$
$\frac{1}{10}$
$\frac{1}{8}$
$\frac{1}{6}$
$\frac{1}{4}$
$\mathcal{C}/\mathcal{C}_0$ =
1.0
2.0
3.0
4.0
5.0
6.0
7.0
8.0
9.0
10.0
11.0
12.0

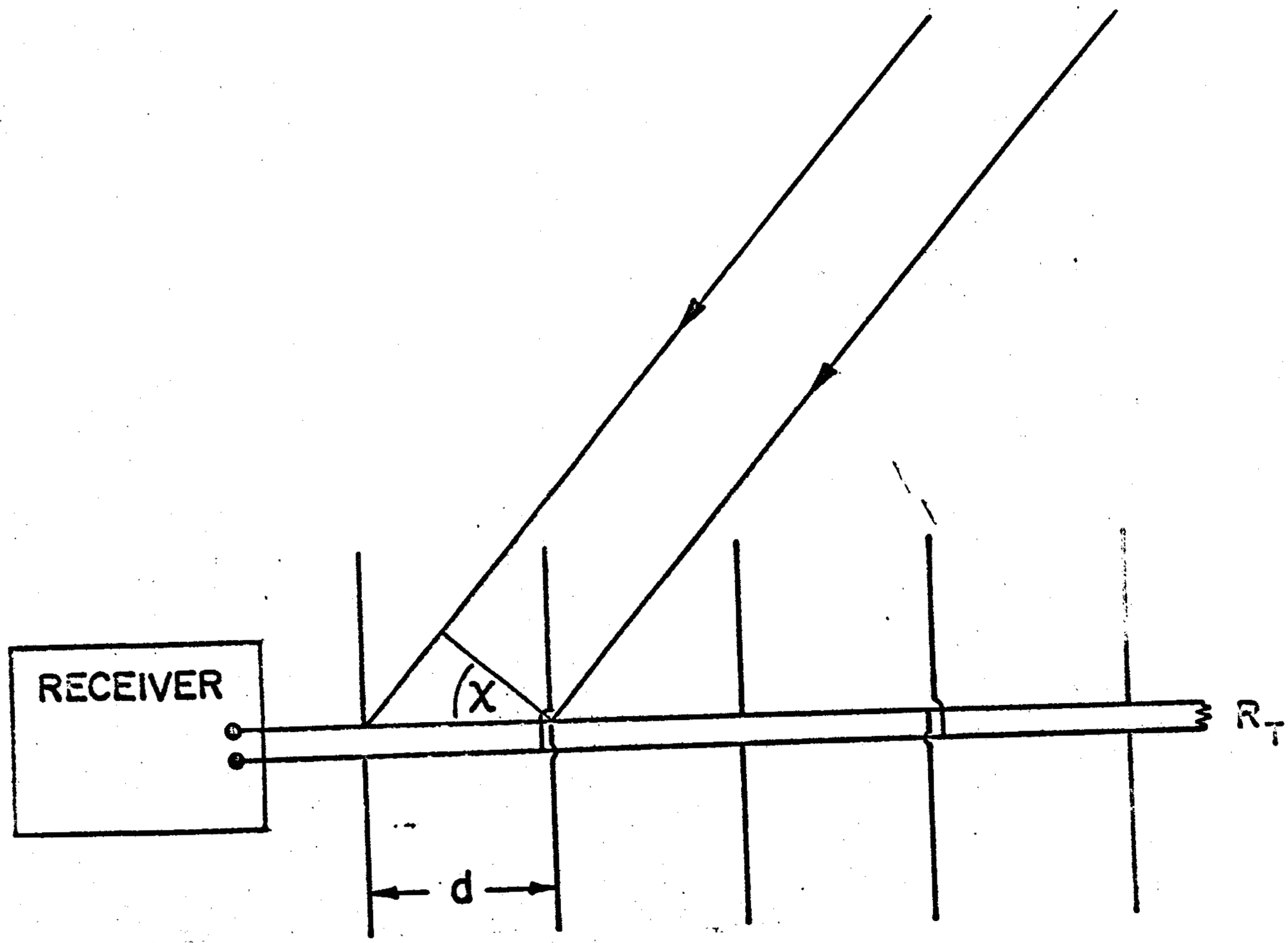
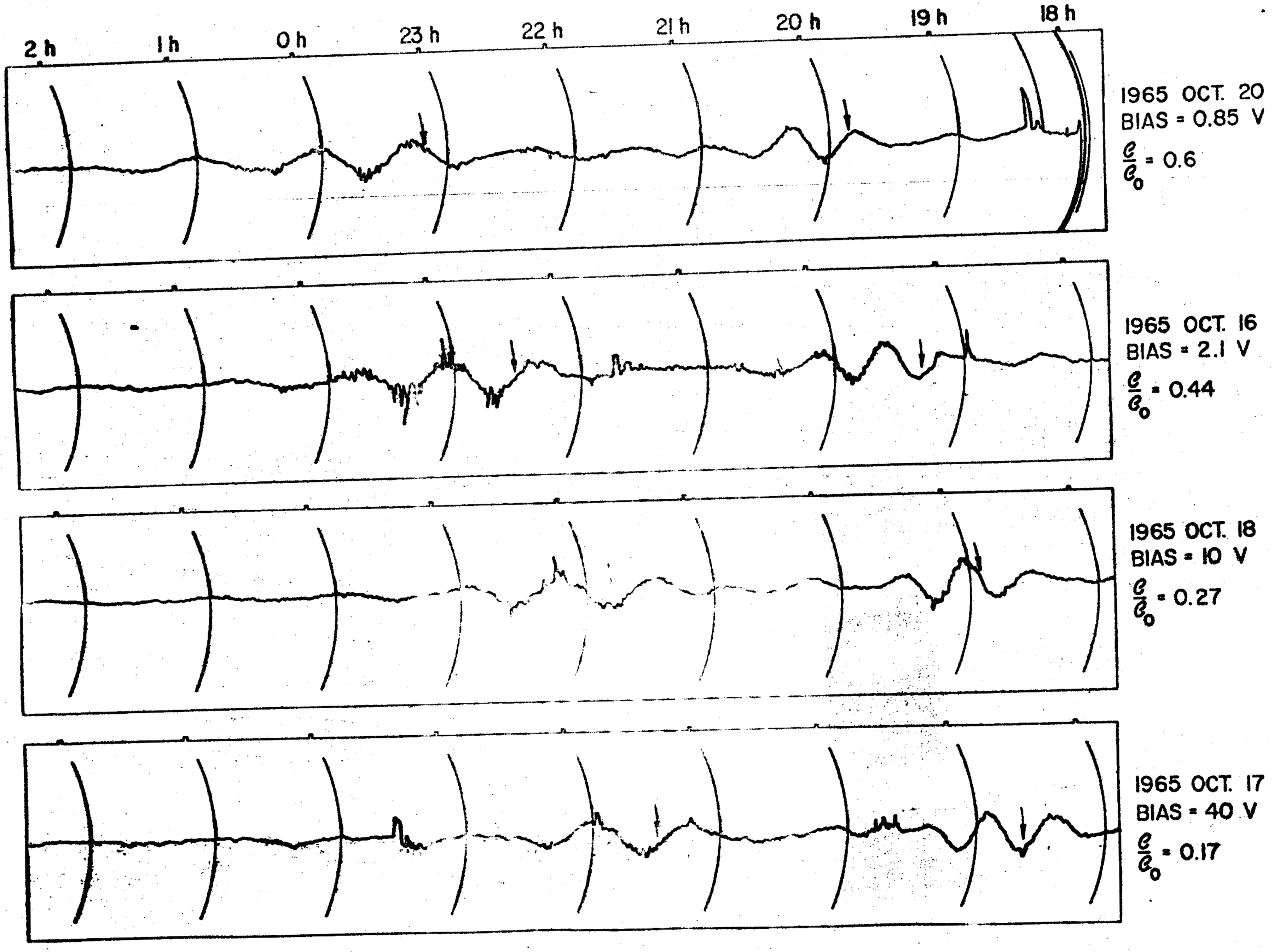


Fig. 6


2 h
1 h
0 h
23 h
22 h
21 h
20 h
19 h
18 h
1965 OCT. 20
BIAS = 0.85 V
$\frac{\mathcal{E}}{\mathcal{E}_0} = 0.6$
1965 OCT. 16
BIAS = 2.1 V
$\frac{\mathcal{E}}{\mathcal{E}_0} = 0.44$
1965 OCT. 18
BIAS = 10 V
$\frac{\mathcal{E}}{\mathcal{E}_0} = 0.27$
1965 OCT. 17
BIAS = 40 V
$\frac{\mathcal{E}}{\mathcal{E}_0} = 0.17$