A Self-Steering Array Repeater

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A scheme is disclosed whereby an antenna array is automatically directed by a simple intermodulation of signal components. In reception, each array element feeds a pilot signal and the modulated signal to a third-order mixer wherein the phase associated with the signal in that element is automatically cancelled. This allows in-phase addition of the contributions from the many elements irrespective of the array shape or the direction of the incoming signal. For transmission, a pilot signal received from the distant receiver location provides by intermodulation a phase compensation to the signal radiated from each transmitting element so as to automatically direct the radiated signal to the distant receiver. There are no significant restrictions as to the shape of the array or the frequencies used.

The scheme lends itself to multiple-element, low-power circuitry and may be used in either space or terrestrial systems to give a high repeater directivity without requiring stabilized platforms or control of antenna orientation. An experimental verification of the basic principle is described.

I. INTRODUCTION

Antenna directivity has become widely used to provide a high effective radiated power with only modest transmitted power, particularly at microwave frequencies where antennas having apertures of many wavelengths are of reasonable size. The precise aiming of the antennas made necessary by this high directivity requires the use of sturdy towers in terrestrial systems, and the proposed use of stabilized platforms for space applications.

A number of methods of avoiding the requirement of accurate orientation and stabilization based upon the Van Atta array concept[1] have been proposed,[2,3,4,5] but each puts strict requirements upon the array shape, and none provides a common intermediate terminal where one may drop and/or add channels. Another method[6] uses a phase-locked loop or servo control to automatically phase the reception from each element for in-phase addition.

The scheme proposed herein avoids many of the limitations of the earlier ones, puts no restrictions on the shape of the array, and does not involve servo control or feedback — either electrical or mechanical. The new scheme compensates for the relative phase of each array element by an intermodulation (frequency mixing) process like that used in some radio diversity[7,8] receivers.

An experimental circuit has been constructed from available hardware to demonstrate in its simplest form the basic principle — coherent addition of microwave signals regardless of relative phase at the input. In-phase addition was obtained as predicted.

Since state-of-the-art microwave solid-state devices provide low power but at relatively high efficiency (i.e., Esaki diodes, varactor multipliers, and microwave transistors), paralleling the power output from many such units in an array provides four distinct advantages:* (*i*) efficient addition of the power from many repeaters, (*ii*) steerability of the beam, (*iii*) high directivity and antenna gain, and (*iv*) reliability—failure of individual units is of little consequence.

Thus it is expected that by the use of modern solid-state devices and micro transmission line techniques, a very simple lightweight self-steering repeater can be built requiring only a fraction of the power of more conventional repeaters with no necessity for orientation control and with the inherently increased reliability provided by a multiplicity of independent parallel circuits.

II. A SELF-STEERING ARRAY REPEATER — BASIC IDEAS

The pointing angle of a steerable array is determined by the relative phasing of the individual elements. Signals received from a distant transmitter by elements of an array differ in phase by an amount which depends upon the geometry of the array, and the relative phases are distributed in a manner exactly opposite to that required for retransmission back toward the source. Van Atta used this fact to show that by suitable interconnection of the elements of a regular linear array, the phase differences can be canceled out, resulting in a return characteristic from an array much like that of a corner reflector[1,2,3] (see Fig. 1).

An alternative which avoids many of the limitations and difficulties encountered when the basic Van Atta array is used in an actual repeater can be explained with reference to a satellite whose orientation is uncontrolled, as shown schematically in Fig. 2. This figure shows an array

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* These advantages were pointed out by R. C. Hanson in Ref. 3 for the case of the conventional active Van Atta array.

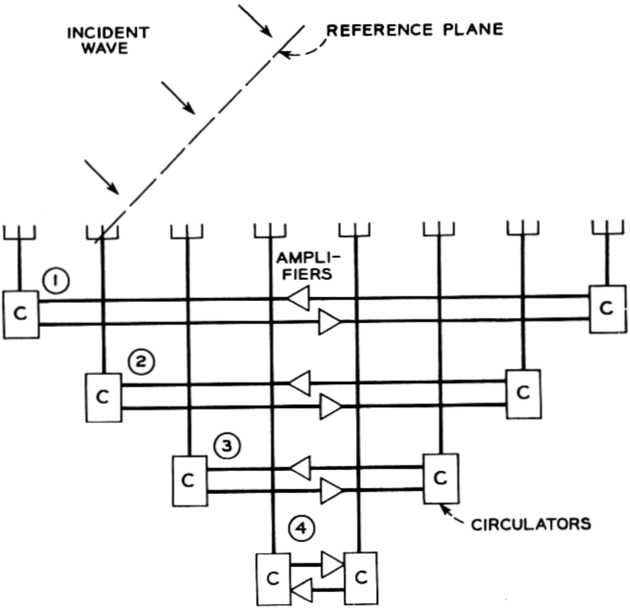

Fig. 1 — An active Van Atta array.

in which the signal received by an element is heterodyned with a locally generated beating oscillator, and the difference frequency is connected back to the same antenna element. All elements are treated alike, and all are excited from a common local oscillator supply. Let the signal received by the ith element be of the form:

$$e_i(t) = \exp j(\omega_R t + \varphi_i) \tag{1}$$

where φ_i is the phase of $e_i(t)$ relative to an arbitrary reference plane normal to the transmission path. The output of the mixer will be:

$$E_i(t) = \exp j[(\omega_B - \omega_R)t - \varphi_i]. \tag{2}$$

If the local oscillator, which is common to all elements of the array, is adjusted to a frequency larger than ω_R, then $(\omega_B - \omega_R)$ is positive, and the phase φ_i will be reversed in sign with respect to that of the received signal, as shown[9] in (2). If, further, ω_B is adjusted to be about twice ω_R

$$(\omega_B - \omega_R) \approx \omega_R \tag{3}$$

and the resultant can be diplexed onto the same array element. The excess phase on transmission just cancels that on reception. This is true

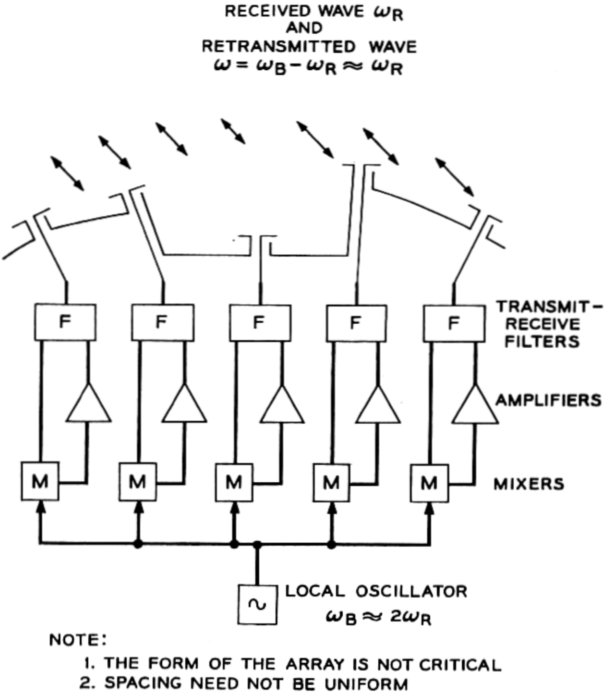

Fig. 2 — An elementary form of active converting array.

for all array elements and their associated circuits. Hence the retransmitted signals are phased just right to form a beam directed back toward the distant transmitter. Note that the foregoing is true regardless of the position of the ith element or the shape of the array, and that no interconnection of array elements is necessary except for the common local oscillator.

This cancellation of phase by mixing is basic to all of the systems described herein. In case it is desired to receive independently, the phase mixing can be accomplished as shown in Fig. 3 (see Ref. 8). The received signal is first divided into two parts in a branching filter, and the carrier or a separate pilot frequency is amplified and further separated from the modulation products in a narrow-band amplifier. The three frequencies — i.e., carrier or pilot, modulation, and local oscillator — are then mixed in a third-order mixer (or two separate more conventional mixers) and the third-order product is selected for subsequent demodulation. The phase of the incoming waves (φ_i) is relative to a plane perpendicular to the incoming wave normal, and will be different for each element.

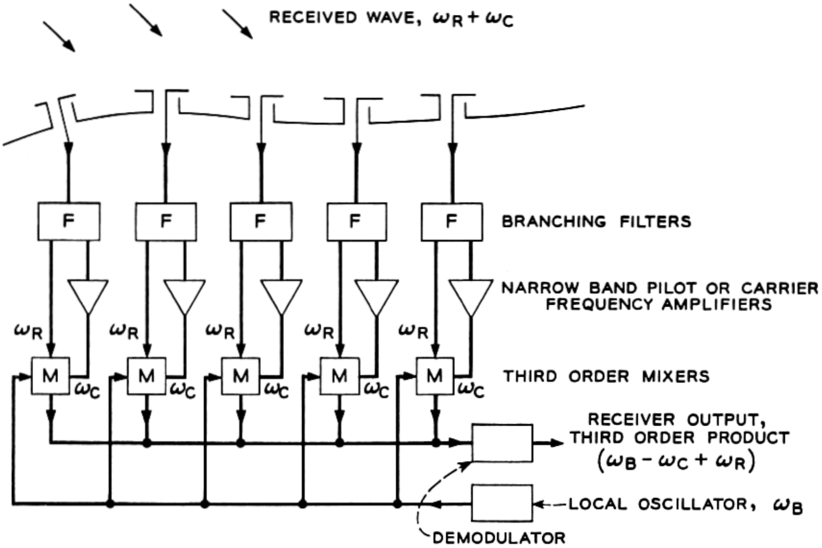

Fig. 3 — An active converting array receiver.

The third-order product is:

$$E_i t = \exp j[(\omega_B - \omega_C + \omega_R)t - \varphi_i + \varphi_i] = \exp j(\omega_B - \omega_C + \omega_R) \quad (4)$$

where ω_B is the local oscillator (radian) frequency, ω_C the received carrier, or a separate pilot frequency, and ω_R the rest of the received signal. Since (4) contains no phase term, evidently the voltages from several such channels can be added.

For transmission of a locally generated modulation, it is evident that modulation applied to the local oscillator of Fig. 2 will be contained in the retransmitted signal. In such a case, the incoming signal acts as a pilot to direct the transmission.

III. ARRAY STEERING BY PILOT FREQUENCY CONTROL

In many applications it is not desired to retransmit in the direction of the received signal. In such a case, a separate pilot signal sent from the distant receiving terminal can serve to define the direction for retransmission, and by the described frequency mixing operation, this can be accomplished automatically. Since the antenna beam is directed or steered toward the distant receiver regardless of its location, the antenna gain can be as large as desired, independent of the changing satellite orientation in space systems or of movement of towers in terrestrial systems.

The advantages of a satellite repeater which does not require orientation control are quite apparent; the advantages to be gained in the application of this scheme to a terrestrial system are also worth noting. Since changes in pointing angles in terrestrial relay systems will be small, the elemental antennas of the array can be relatively high gain, and thus relatively few elements are required to implement a steerable array (STAR) repeater. Hence the advantages of reliability and self-steering can be obtained in terrestrial systems with only small increase in the amount of repeater electronics.

IV. SELF-STEERING SATELLITE REPEATER

Before going into detail, we will describe a prototype repeater embodying the principles described. Since we are interested in partially oriented terrestrial as well as nonoriented satellite repeaters, we will attempt to generalize the discussion to cover both situations. In the terrestrial case, the array elements can profitably use area directivity, and it is desirable to interconnect two separate arrays with elementary repeaters, as shown in Fig. 4. In the satellite case it is more desirable to combine the functions in a single array, or to separate transmitting and receiving functions. The satellite repeater is visualized as spherical and entirely covered with elemental antennas, as shown in Fig. 5.

To insure that all of the antenna elements act in concert as a phased array, and hence as an antenna having an aperture nearly equal to the projected area of the array, it is necessary to combine in-phase the re-

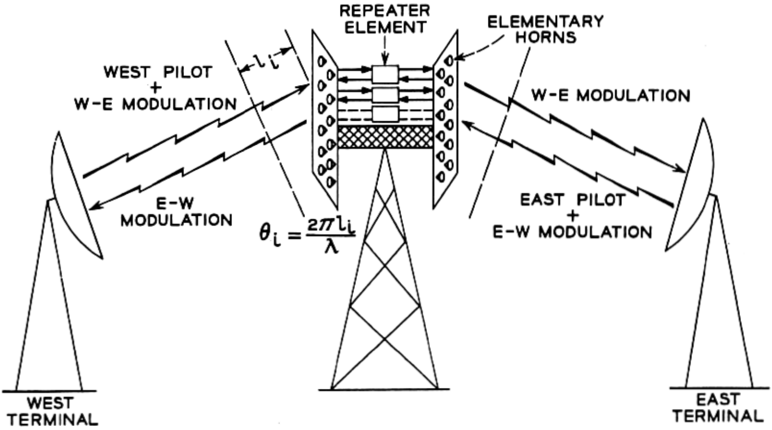

Fig. 4 — A possible two-way terrestrial repeater configuration.

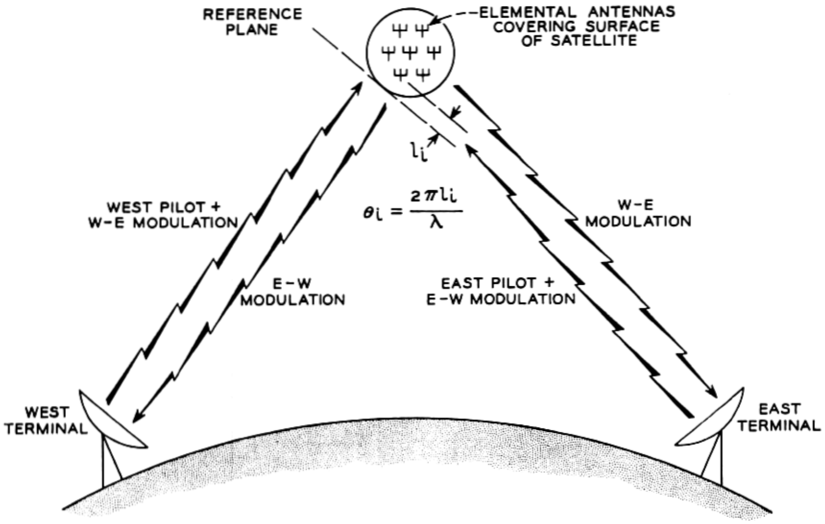

Fig. 5 — A possible two-way satellite repeater configuration.

ceived signals from all the elements. As received from the west terminal, the signals have relative phase shifts

$$\theta_i, \theta_2, \theta_3 \cdots \theta_i, \qquad \text{where } \theta_i = \frac{2\pi l_i}{\lambda},$$

and l_i is the variable distance between the ith element and a reference plane normal to the radius vector to the western terminal, as shown in Figs. 4 and 5. This distance, and hence the phase shift, depends upon the orientation of the array and changes when the array moves relative to the fixed terminals. In-phase addition of the received signals can be accomplished by the use of the pilot beam as follows: Consider first the left-hand part of the two-way repeater shown schematically in Figs. 6 or 7. Signals received from the west terminal by the ith elemental antenna are: (a) the pilot, $\exp j[\omega_P t + \theta_{i,p}]$, and (b) the modulation,

$$\exp j[\omega_{M(\text{W-E})} t + \varphi_{(\text{W-E})} t + \theta_{i,M}].$$

These are passed by the transmit-receive filter and are converted in an intermediate frequency circuit by mixing with a local oscillator in a square-law mixer. The results are

$$A(t) = \exp j[\omega_{LO_1} t - \omega_{P(\text{W})} t - \theta_{i,P}] \tag{5}$$

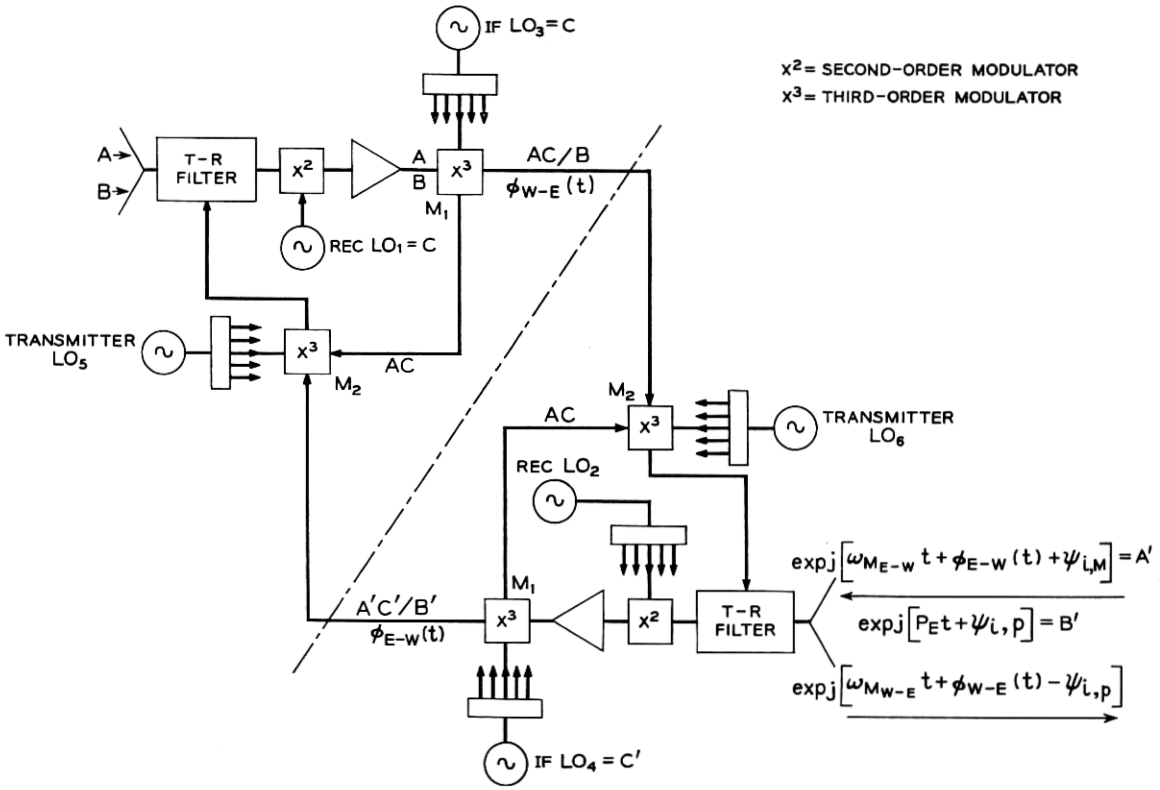

Fig. 6 — Basic elements of a two-way repeater [one of several sections joined with a common local oscillator and attached to separate input and output antenna arrays suitable for terrestrial (fixed) service].

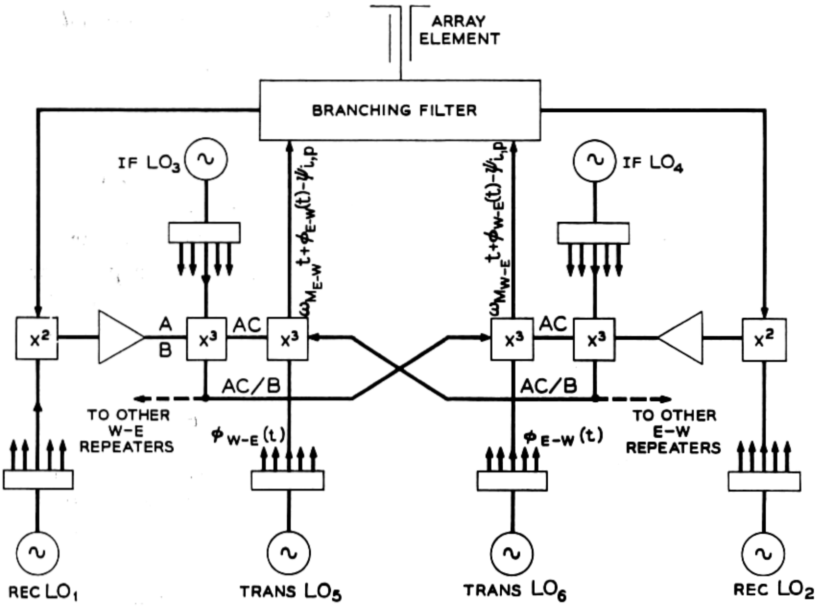

Fig. 7 — Basic elements of a two-way repeater [one of several sections joined by a common local oscillator and diplexed into a common receiving-transmitting array suitable for unstabilized satellite repeater use].

and

$$B(t) = \exp j[\omega_{LO_1}t - \omega_{M(\text{W-E})}t - \varphi_{(\text{W-E})}(t) - \theta_{i,M}] \tag{6}$$

where

ω_{LO_1} = radian frequency of common receiving local oscillator,
$\omega_{P(\text{W})}$ = radian frequency of pilot received from west terminal.

$\theta_{i,M}$ and $\theta_{i,P}$ are the phases relative to the common reference plane, which is equal to $2\pi l_i/\lambda$, where λ is the appropriate wavelength and l_i is the distance between ith antenna element and the reference plane. $\omega_{M(\text{WE})}$ = the radian frequency of the west-east modulation channel carrier and $\varphi_{(\text{W-E})}(t)$ = the angle modulation of the west-east carrier.* These are amplified and put into a third-order mixer along with IF local oscillator signal $\exp j(\omega_{LO_3}t) \equiv C(t)$. From the many modulation products generated in this mixer, two are selected by filtering. The first is

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* While the repeater is described in terms of the commonly used frequency modulation, the basic scheme can be used with any modulation technique.

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\begin{aligned}
AC/B &= \exp j[\omega_{LO_3}t + \omega_{LO_1}t - \omega_{P(\mathrm{W})}t - \theta_{i,P} - \omega_{LO_1}t + \omega_{M(\mathrm{W\text{-}E})}t \\
&\qquad + \varphi_{(\mathrm{W\text{-}E})}(t) + \theta_{i,M}] \qquad (7)\\
&= \exp j[\omega_{LO_3} - \omega_{P(\mathrm{W})} + \omega_{M(\mathrm{W\text{-}E})})t + \varphi_{(\mathrm{W\text{-}E})}(t) + (\theta_{i,M} - \theta_{i,P})].
\end{aligned}
$$

This is a carrier angle-modulated by $\varphi_{(\mathrm{W\text{-}E})}(t)$ and having the residual relative phase angle $(\theta_{i,M} - \theta_{i,P})$. If the pilot frequency is chosen nearly equal to the modulation frequency, $(\theta_{i,M} - \theta_{i,P})$* will be very small, and the W-E modulation received by the ith antenna will be in phase with that received by all of the other antennas. This completes the receiving functions; the remaining problem is to derive steering information for the outgoing beam. To this end we select the modulation product

$$AC = \exp j[(\omega_{LO_3} + \omega_{LO_1} - \omega_{P(\mathrm{W})})t - \theta_{i,P}] \qquad (8)$$

where the symbols are defined above. Now the relative phase of this wave, $-\theta_{i,P}$, is just right for retransmission toward the west terminal near the frequency $\omega_{P(\mathrm{W})}$, using the same antenna element as for receiving. By mixing this steering signal with the modulation $\varphi_{(\mathrm{E\text{-}W})}(t)$ derived in a similar fashion from the right-half of the repeater, together with a microwave local oscillator LO_5, the retransmission function is complete.

The optional interconnection among amplifying elements, shown by dashed lines on Fig. 7, provides for the situation encountered with low-altitude satellites serving widely separated earth stations. In this case, the body of the satellite "shadows" some of the elements, and the part of the satellite surface seen by both earth stations is a small part of the total. Since it requires a signal from each earth terminal to generate the retransmitted signal, without interconnection only a small number of elements are effective. With interconnection, all satellite elements visible from the transmitting earth station will receive the modulated signal; the contributions from the various elements will then be added in phase at IF and the sum impressed on the outgoing carrier. All satellite elements visible from a receiving earth station will then emit information-bearing waves which will add in-phase in the direction of this earth station receiver. Also, it should be noted that since not all branches receive the same signal levels, some weighting[10] of voltage levels must be accomplished before combining the outputs or there will be a loss in signal-to-noise ratio. This may be accomplished by operation of mixers in a strictly square law region or by auxiliary means beyond the scope of this paper.

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* The magnitude of this residual is also affected by the size of the satellite.

The satellite repeater can have an arbitrarily large antenna array gain for transmitting or receiving or both, regardless of satellite altitude or orientation and independent of earth terminal separation. This is not true for any other scheme of which the authors are aware.

There is another factor which should be mentioned at this point. It is important that the pilot frequency be filtered from the surrounding noise or modulation and desirable that it be enhanced to a level well above that of the modulated signal before the second mixer shown in Fig. 6. This is to keep the pilot from bringing noise into the final modulation band and to assure that the desired products predominate over higher-order products.

V. ARRAY SCALING AND FREQUENCY MULTIPLICATION

In the foregoing scheme, it is phase rather than time delay that is compensated by the frequency conversion operation. If the application permits the receive and transmit frequencies to be nearly the same, as was assumed in discussing Fig. 2, a single array of antennas can be used for both transmitting and receiving. However, if a rather large change in frequency is required between repeater input and output, as is frequently the case in both terrestrial and space systems, additional phase compensation must be provided. One possibility is to use a different array for transmitting than for receiving, the arrays to be similar but scaled in proportion to the wavelength. Alternatively, one may compensate by using a step of frequency (and phase) multiplication, as we shall see. Let us consider the four waves associated with reception of information from one distant terminal and retransmitted to the other. The total phase shift ψ_i of a signal wave in passing from a reference wave front (Fig. 8a), through the ith branch of the circuit and to a reference plane perpendicular to the path to the other distant terminal is:

$$\psi_i = \underset{\text{(receiving)}}{2\pi\left[\frac{S_i}{\lambda_R} - \frac{S_i}{\lambda_{P_1}}\right]} + \underset{\text{(transmitting)}}{2\pi\left[-\frac{d_i}{\lambda_{P_2}} + \frac{d_i}{\lambda_T}\right]} \tag{9}$$

where

$2\pi S_i/\lambda_R$ = phase shift of the received modulated signal caused by delay between a reference plane wave front and the ith receiving element,

$2\pi S_i/\lambda_{P_1}$ = phase shift of the pilot associated with the above signal and path,

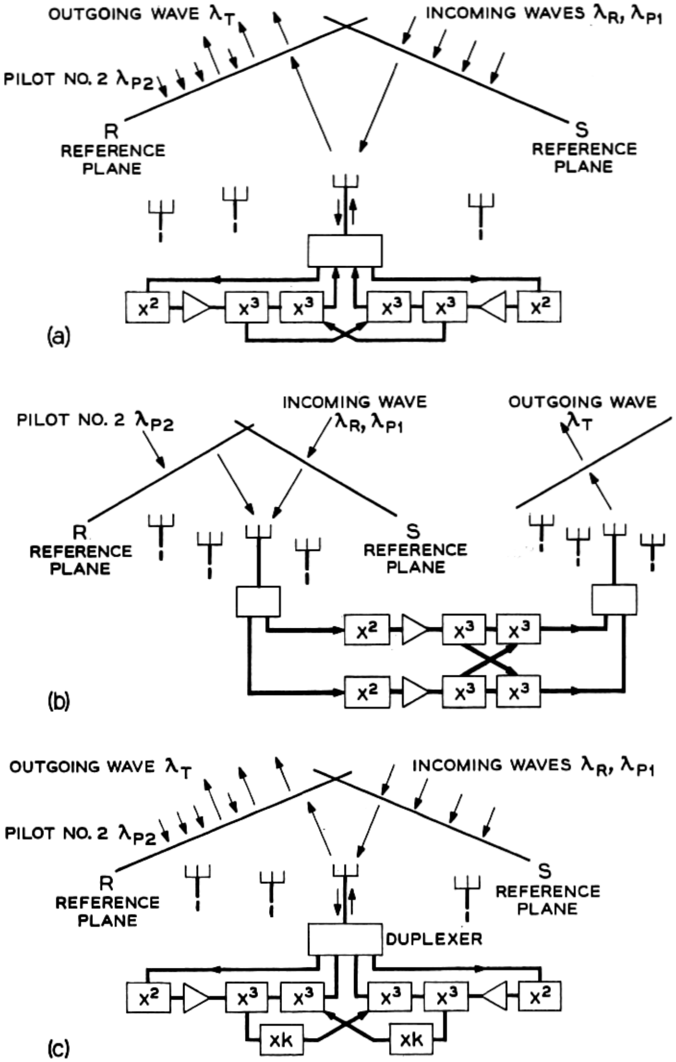

Fig. 8 — Methods of compensating for frequency change in the repeater: (a) reference diagram, (b) array scaling, (c) frequency (and phase) multiplication.

$2\pi d_i/\lambda_{P_2}$ = phase shift of the received pilot signal from the outgoing path, caused by delay between a reference plane wave front of this wave and the ith antenna element, and

$2\pi d_i/\lambda_T$ = phase shift of the retransmitted signal between the associated reference plane and element.

There are a number of possible ways to make $\psi_i = 0$. Let us suppose that all of the waves received are near the same frequency; then

$$\lambda_R \approx \lambda_{P_1} \approx \lambda_{P_2}$$

and the first term in (9) is very small. The transmitter in many applications will be considerably removed from the receiving band, in which case $\lambda_{P_2} \not\approx \lambda_T$. To make $\psi_i \approx 0$, we may scale the transmitting array as in Fig. 8(b), in proportion to the wavelength. Then, letting the primes indicate the scaled dimensions.

$$\frac{d_i'}{\lambda_T} = \frac{d_i}{\lambda_{P_2}}$$

and

$$\psi_i \approx 0.$$

It may not be convenient to scale the array. As an alternative, we can operate on the pilot signal before mixing. Suppose that after the second step of frequency conversion we pass the pilot signal through a frequency multiplier, as shown in Fig. 8(c). This multiplies the pilot frequency term, including phase, by a factor k. The signal out of the multiplier is of the form [from (8)]

$$\exp j[k(\omega_{LO_3} + \omega_{LO_1} - \quad)t - \omega_{P(\mathrm{W})}\, k\theta_{i,P} + 2\pi nk]. \tag{10}$$

Now, adding the phase contributions through the repeater we get

$$\psi_i - 2\pi nk = \underset{\text{(receiving)}}{2\pi \left[\frac{S_i}{\lambda_R} - \frac{S_i}{\lambda_{P_i}}\right]} + \underset{\text{(transmitting)}}{2\pi \left[-k\,\frac{d_i}{\lambda_{P_2}} + \frac{d_1}{\lambda_T}\right]} \tag{11}$$

and k can be chosen to make $\psi_i = 0$. In general, however, k will not be an integral and there is a phase ambiguity. A possible way of removing this ambiguity is to lightly couple the frequency multipliers in adjacent channels, so that they prefer to be nearly in phase, and limit the array design so that adjacent elements are not more than $\lambda/2$ apart in the direction of transmission* at both the transmitting and receiving frequencies.

Of course, combinations of array scaling, phase shift multiplication and a judicious choice of pilot, transmitting, mixing and receiving frequencies will be important and interrelated parts of the design of a practical system.

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* Elements need not be physically less than $\lambda/2$ apart, but the distance along the direction of propagation should not differ by more than $\lambda/2$ when the line of sight is within the beamwidth of the element.

VI. ARRAY GAIN

Up to this point no limits have been placed on the form of the antenna array. The elements do not have to be arranged with any particular form or symmetry. However, the presence of the satellite itself, in the case of satellite repeaters, and mutual coupling between array elements and bandwidth considerations in either space or terrestrial systems, provide some limits to the form of the array. Since the phase between elements varies, strong coupling between elements would have serious consequences in impedance mismatch. However, element directivity and separation can be used to reduce coupling and also to reduce the shadowing of elements one by another. If array elements are mounted on a conducting surface, an element gain of two is inherent in that the element can only radiate into a hemisphere. A gain of three to five is more practical, can be obtained from small elements, and results in relatively small coupling between elements.

In the case of a satellite without orientation control, elements must point in all directions; but an element having a gain (g) can illuminate only $1/g$th of the total solid angle. Thus, if N elements are distributed more or less uniformly over the surface of a spherical satellite, only $(1/g)N$ will contribute to the received (or transmitted) signal. Also, of the total power radiated, only a fraction $(1/g)$ is delivered to the array elements forming the beam. The remainder is not utilized. Net effective array gain for transmission is the product of element gain and the number of elements effective and the fraction of the total power which is useful, i.e.,

$$G = g\ (N/g)\ (1/g) = (N/g). \tag{12}$$

Thus the net array gain for transmission is equal to the number of elements in the array divided by the element gain. Evidently one should use little element gain on nonoriented satellite repeaters. In the case of terrestrial repeaters and oriented satellites:

$$G = Ng \tag{13}$$

and element gain is limited by more customary factors. In other circumstances which we will not elaborate, $G = N$.

One would like to get equivalent performance in all directions from an unstabilized satellite. This can be accomplished with the present scheme by covering the outside of a sphere or polyhedron with small radiators. It is best that the radiators be close together to reduce side lobes and possible interference.

If each element is assigned one square wavelength, the satellite diameter must be

$$D \geqq \lambda\sqrt{N/\pi}. \tag{14}$$

Alternatively, the elements can be grouped in a ring on a great circle around the satellite, each element having a fan beam with a maximum in a radial direction and radiating with a gain of $(1/g)$ in a direction 90° from the plane of the array. All of the elements of such an array would contribute in the polar direction with an array gain of (N/g). In the equatorial plane, even though only a fraction of the elements contribute, the gain is also (N/g) by the argument used in deriving (14). In intermediate directions, the gain depends upon the detailed characteristics of the elements, but it should be possible to keep it very near (N/g) in all directions.

One should not confuse the round-trip performance of the array with the radiation pattern obtained with a fixed excitation. The former can be truly isotropic but the latter cannot be, and may be a multilobe affair. If the elements are in a ring, as described above, the re-radiation will in general have two large lobes, one above and one below the plane of the array; and if the elements are widely spaced, some minor lobes may be as large as the major. This is of little consequence to the transmission performance of a satellite system, however, because the phasing of the elements automatically assures that a maximum is always directed toward the appropriate earth terminal. The shape of the pattern depends drastically upon the distribution of elements, but to a first order the strength of the major lobe does not. In any case, spacing between array elements and the element gain should be chosen to minimize side lobes in order to lessen the likelihood of interference.

VII. NUMBER OF ARRAY ELEMENTS

How many array elements is it practical to consider for a repeater of the type proposed herein? As has been shown, the directivity gain for a nonoriented repeater for reception and transmission is equal to the number of elements used divided by the element gain, thus the effective radiated power (ERP) or the power which an isotropic source would have to radiate to produce the same received signal is

$$ERP = (N/g)P_R \tag{15}$$

where N = number of elements in array,
g = gain of individual elements,

P_R = total power radiated = NP_2, and
P_2 = power radiated per element.

The power P_2 which must be radiated by each element to provide a given ERP is, from (15)

$$P_2 = P_R/N = ERP(g/N^2). \tag{16}$$

Thus there appears to be an advantage in using a large number of elements. However, there will be a component of the dc input power used for local oscillators, low-level amplifiers, etc., which is directly proportional to the number of elements and is nearly independent of the RF power output per element. Hopefully, this can be made small through development of suitable solid-state devices. For the minimum dc power consumption consistent with a given repeater performance, there is evidently an optimum number of array elements. If the low-level mixing and amplifying operations can be accomplished with a power consumption p_1 watts per element, and if a high-frequency output power P_2 watts per element can be obtained with a power amplifier efficiency of η, then the dc input power required to radiate a beam having a stated ERP is

$$P_{\text{dc}} = p_1N + (P_2/\eta)N = p_1N + (g/N)ERP/\eta \tag{17}$$

where the symbols are as defined above. This has a minimum when

$$\frac{\partial}{\partial N}[p_1N + (g/N)ERP/\eta] = 0 \tag{18}$$

$$p_1 - (g/N^2)ERP/\eta = 0 \tag{18a}$$

or

$$N = \sqrt{(g/p_1)ERP/\eta}. \tag{19}$$

We note that from (18a) and (16)

$$p_1 = g/N^2(ERP/\eta) = P_2/\eta \tag{20}$$

which says that the minimum dc power will be required when the number of elements is chosen to make the power supplied to the output amplifiers equal to that consumed by the low-level devices.

If, in the case of a nonoriented satellite repeater, it is possible to miniaturize the circuitry enough so that the power supply is the principal source of weight, then this is a real optimum. Otherwise, it represents a sort of design objective, and minimizing satellite weight and complexity will require a smaller number of elements. In any case, it is clear that the practicality of the scheme depends heavily upon the degree of miniaturization and power efficiency which can be achieved with solid-state devices and circuits.

VIII. EXPERIMENTAL VERIFICATION

The basic principle upon which the self-steering array depends is the coherent in-phase addition of randomly phased inputs. The principle has been demonstrated by the simple laboratory experiment shown in Fig. 9. The outputs of two oscillators at 6200 and 6034 mc, which may be thought of as representing the received modulation and pilot signals, are combined and fed together to two receiving circuits, and an adjustable phase shifter or line stretcher is inserted in one branch. Each receiving circuit separates the incoming frequencies, heterodynes the 6200 mc with a 6274-mc local oscillator which is common to the two receiving circuits, to produce a 74-mc intermediate frequency, which in turn is amplified and recombined with the 6034-mc signal in a second mixer to produce a 6108-mc output. Now, it is asserted that the phase of the

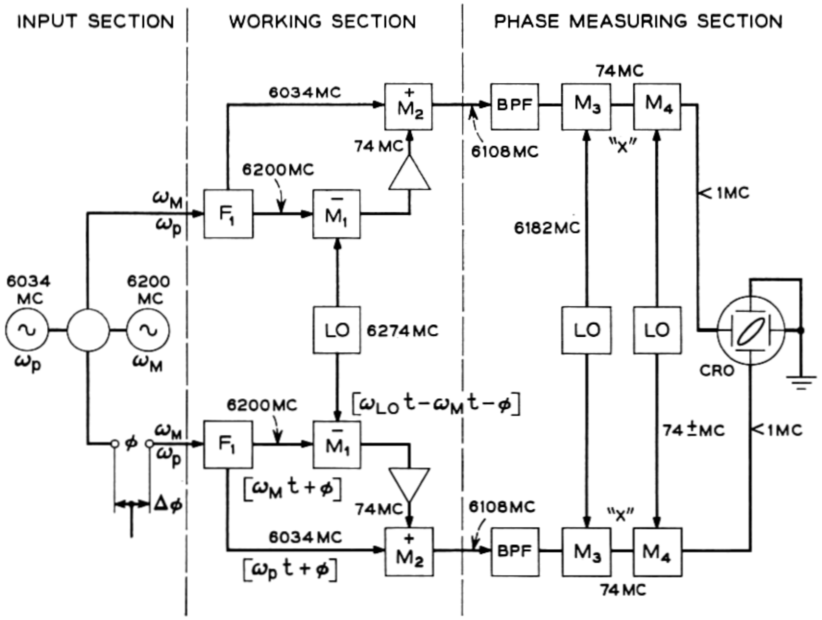

Fig. 9 — Experimental arrangement for testing the principle of phase compensation.

6108-mc wave should be to first order independent of the phase of the combined signals at the input to the branching filter. To check this, the signals from the two branches are compared in the circuit shown to the right of Fig. 9. If the idea is sound, the phase of the output should change very little with movement of the piston in the input section.

The degree of phase change correction or cancellation was observed visually by noting the stability of a Lissajous figure formed by the output sine waves as received over the two branch paths of the test circuit, and was measured using a phase meter. Some change in the phase between the two output branch signals was to be expected because of the difference in frequency between the pilot and the signal. Fig. 10 shows the measured and calculated phase change in the output produced by large changes of phase in the lower branch of the circuit. It will be noticed that the resultant measured output phase change, $\Delta\theta$, is not exactly a linear function of changes in input phase. These variations from linearity were shown to be caused by mismatches in impedance and leakage between various parts of the circuit and were partially corrected with isolators. Calculation of the ratio of wavelengths for WR159 rectangular waveguide for the two frequencies 6200 and 6034 mc gives a value of 1.04 or a difference of 15° for each wavelength, which is in very good agreement with the average measured value. The difference frequency was later reduced to 40 mc, and the output phase variation was reduced proportionally.

Most of the tests described were made under a condition of large signal-to-noise ratio and low gain. The phase correction was found to be

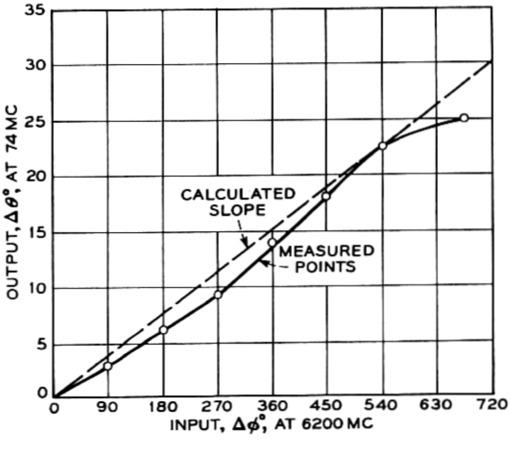

Fig. 10 — Comparison of calculated and measured phase compensation.

very stable with change in time, level and frequency. In order to determine how the scheme would operate for low values of signal-to-noise and with over-all gain typical of an actual repeater, attenuation was added in the lower branch of the circuit to reduce the level of ω_M. This loss was then compensated by adding about 100 db of IF amplification to bring the level back to its former value. When the signal-to-noise ratio was measured at 2 db, the pattern on the scope was much more ragged due to the large random noise present, but it was still stable and indicated the desired cancellation of phase.

IX. CONCLUSIONS

An antenna beam-steering scheme using a pilot tone and phase inversion makes possible large antenna gain even for nonoriented satellite repeaters or movable terrestrial repeaters. A large number of low-power, elemental repeater amplifiers with inputs and outputs connected to like elements in similar arrays, or diplexed onto common elements in a single array, are used. The scheme is particularly suited for use with solid-state devices since the (low) power output of many units is effectively added in-phase. Reliability is provided by the many parallel paths through the repeater; failure of individual units will only slightly degrade performance. Although the radiation is not isotropic, the radiation or sensitivity toward distant terminals can be independent of array orientation, and thus the idea is well suited for use with unoriented satellites. A simple experiment has been performed to demonstrate the basic steering property of the phase inversion scheme.

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REFERENCES

1. Van Atta, L. C., Electromagnetic Reflector, U. S. Patent No. 2,908,002, October 6, 1959.
2. Sharp, E. D., and Diab, M. A., Van Atta Reflector Array, I.R.E. Trans. on Antennas and Propagation, **AP-8,** July, 1960, pp. 436–438.
3. Hansen, R. C., Communication Satellites Using Arrays, Proc. I.R.E., **49,** June, 1961, pp. 1066–1074.
4. Rutz, E. M., and Kramer, E., Modulated Array for Space Communications, NEREM Record, **4,** November, 1962, pp. 16–17.
5. Davies, D. E. N., Some Properties of Van Atta Arrays and the Use of Two-Way Amplification in the Delay Paths, Proc. IEE (London), **110,** March, 1963, pp. 507–512.

6. Miller, Barry, Self-Focusing Antenna Arrays Developed (News Item), Aviation Week and Space Technology, Aug. 21, 1961, pp. 54–59.
7. Adams, R. T., and Mindes, D. N., Evaluation of Intermediate-Frequency and Baseband Diversity Combining Receivers, I.R.E. Trans. on Communication Systems, **6,** June, 1958, pp. 8–13.
8. Bello, P., and Nelin, B., Predetection Diversity Combining with Selectivity Fading Channels I.R.E. Trans. on Communication Systems, **9,** March, 1962, pp. 32–42.
9. Phase inversion by this means was known and used by Friis in the 1930's; see Friis, H. T., and Feldman, C. B., A Multiple Unit Steerable Antenna for Short-Wave Reception, B.S.T.J., **16,** July, 1937, pp. 337–419, particularly Fig. 28 and accompanying legend.
10. Kahn, L. R., Ratio Squarer (Letter) Proc. I.R.E., **42,** November, 1954, p. 1704.