AFCRL 63-13

THE COAXIAL DIPOLE
PARAMETRIC AMPLIFIER ANTENNA

Albert D. [illegible]

Contract AF19(628)-307
SCIENTIFIC REPORT NO. 1

DECEMBER 1, 1962

ANTENNA SYSTEMS LABORATORY
UNIVERSITY OF NEW HAMPSHIRE
DEPARTMENT OF ELECTRICAL ENGINEERING
DURHAM, NEW HAMPSHIRE

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Antenna Systems Laboratory, Dept. of
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THE COAXIAL DIPOLE PARAMETRIC AMPLIFIER ANTENNA, by A.D. Frost, 1 December 1962. Scientific Report No. 1 36 pp incl. illus.
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A study has been made of the theoretical and applied problems associated with the design of a parametric amplifier antenna in coaxial dipole form. An analysis is made of the coaxial interior which serves to provide the resonances needed for parametric amplification, and an expression is derived for the complex impedance seen by the varactor diode in terms of the inner coaxial geometry, the relative electrical length of the output loop the receiver input resistance and an adjustable terminating reactance. This latter element permits the tuning of the system for a desired pair of signal and idler frequencies. A completed tuning and bias system for a quarter wave dipole at 108 Mc. is shown

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1. Antennas
2. Amplifiers

I. Frost, A.D.

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The University of New Hampshire
Durham
New Hampshire
Attn: A.D. Frost

Microwave Associates, Inc.
South Avenue
Burlington, Mass.

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ELECTRONICS RESEARCH DIRECTORATE
AIR FORCE CAMBRIDGE RESEARCH LABORATORIES
OFFICE OF AEROSPACE RESEARCH
UNITED STATES AIR FORCE
BEDFORD, MASSACHUSETTS

by

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ABSTRACT

A study has been made of the theoretical and applied problems associated with the design of a parametric amplifier antenna in coaxial dipole form. An analysis is made of the coaxial interior, which serves to provide the resonances needed for parametric amplification, and an expression is derived for the complex impedance seen by the varactor diode in terms of the inner co-axial geometry, the relative electrical length of the output loop, the receiver input resistance and an adjustable terminating reactance. This latter element permits the tuning of the system for a desired pair of signal and idler frequencies.
A completed tuning and bias system for a quarter-wave dipole at 108 Mc. is shown.

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List of Illustrations

Introduction

The invention and initial development of the parametric amplifier antenna was carried out with the support of the Air Force Cambridge Research Laboratories, Electronics Research Directorate, as a portion of the work performed by members of the staff of the Department of Electrical Engineering, University of New Hampshire under contract AF19(607)-3892. Initial phases of this work were presented in Scientific Report #2 and #3 and in the Final Report prepared under this contract as well as in a number of other publications[1]. From a rudimentary beginning the amplifier antenna has progressed through numerous steps of evolution both in physical design as well as in the development of technique. The circuit design problems have been those common to most parametric amplifiers; the need to embed a varactor element into a mutually compatible configuration of circuit elements which provide independent resonant loops for the signal frequency and the idler which are common to the diode, while at the same time allowing efficient access to the diode for the pump signal. To this we have added, in the case of the "parant" the need to integrate these designs within the confines of a free half-wave dipole or a grounded quarter-wave dipole.

The present contract for which this is the first Scientific Report provides direct support for the further development of this novel form of low-noise amplifier. In line with the basic concern of the previous contract in the reception and analysis of satellite signals , initial parametric amplifier antenna work was carried out at 54 Mc. using a grounded quarter-wave dipole and at 108 Mc. with a half-wave dipole. In either case the signal frequency was such as to require partial lumped and partial distributed tuning networks while the idler frequency circuits were dominated by geometrical and distributed parameter considerations. Circuit configurations used for these early designs are shown in the references. As was noted at that time excessive cross-coupling between signal and idler tuning adjustments was noted in addition to numerous circuit complications arising from the need to introduce the pumping signal without serious impairment to the tuning and selectivity of the other circuits. It was not possible to resolve these problems into a satisfactory design procedure beyond an outline of the general objectives of circuit performance since the combinations of lumped and distributed circuit elements , assembled in close proximity , could not be scaled to other frequency combinations. Technical problems associated with the introduction of the pump signal were largely eliminated by the addition of an additional small loop at the ground plane end of the dipole as illustrated in the Final Report.

With inception on 1 February 1962 of the present contract parametric amplifier work has progressed in the following phases.

I. Development of a theoretical analysis of the coaxial dipole ($\frac{1}{4}$ and $\frac{1}{2}$ wave) form of amplifier-antenna including

(a) linear,distributed circuit analysis of the coaxial interior region of the antenna including the effects of tuning and the output coupling mechanism and leading to a general statement regarding the multi-resonant properties of the region and their control.

(b) improvement and simplification of tuning procedures with particular emphasis on methods useful over a wide frequency range with only linear modifications.

(c) approximate calculations on the current distribution on the exterior of the antenna cylinder and the proper length for optimum signal capture and transfer to the interior active amplifier.

(d) consideration of the coupling present at the interface between the exterior antenna and the interior amplifier and its relation to optimum varactor operating impedance.

II. Extension of the integrated amplifier - antenna concept to higher frequencies through the use of ground plane slots combined with resonant cavities. This work to begin with an exploration of possible slot shapes, location of output coupling loops or probes, biasing , use of parasitic sub-cavities for idler storage.

THE COAXIAL AMPLIFIER DIPOLE

Exterior Antenna Surface

During the preliminary phases of the work on parametric amplifier antennas the electrical length of the external antenna structure was selected on the basis of conventional antenna practice to be slightly less than $\lambda/2$ for the "half-wave" dipole and less than $\lambda/4$ for the grounded quarter wave structure. An empirical relationship commonly quoted in this connection is the expression for the length

$$L = 0.48 \frac{L/D}{L/D + 1} \lambda$$

D = dipole diameter

This expression assumes that the antenna is resitively terminated in a matched load at the center point. We have explored the current distribution along a cylinder of finite diameter for the case in which there is no central load when placed in plane wave field polarized along the axis of the antenna. Using the sucessive integral approximations of King and Harrison[2] a digital computer program (Fortran) was written for use on the IBM 1620 for the calculation of axial current distribution as a function of the relative total length of the cylinder. This treatment of course assumed a simple capacitative loading at the cylinder ends. Charge concentration at the ends of the antenna can be calculated from derivative of axial current with respect to axial position evaluated at the end points. To simplify the search routine the intended antenna diameter(1.5") and the operating frequencies of 54Mc. and 108Mc. were directly introduced into a specific size term during the computation. This resulted in a best length designation of 0.238 λ for the quarter-wave antenna and 0. 468 λ for the half wave antenna. where the latter unit , since it uses the same diameter cylinder is relatively more stubby.

The Output Coupling Region

The long rectangular coupling loop used inside the coaxial region of the antenna produces a significant effect on the impedance and resonant behavior of this region as a whole. In order to determine the available spectrum of resonances as well as the coupling efficiency a study must be made of the effect of this region and how its physical and electrical parameters can be included into the design of the amplifier-antenna system.

The appearance of this coupling or interaction region is shown in Fig. 1 and Fig. 2 . Neglecting for the present discontinuity effects this region can be defined by reference to Fig. 2 . The larger diameter center conductor is coaxial with the axis of the antenna cylinder while the output loop is composed of the off-center conductor of lesser diameter. The right hand boundary of the coupling region represents the lower or "ground" end of the quarterwave dipole. At this plane (b-b') we have assumed that the center conductor (transmission line "A") is terminated , with reference to the inner surface of the antenna cylinder, in a reactance X ; while the coupling loop line (transmission line "B") is terminated at this same plane in a resistance R . The left boundary (plane a-a') of this region where there is an abrupt transition to the simple single coaxial region contains a radial short circuit for line "B" .

A discussion of the effect and implications of coupled distributed transmission lines was presented by Fuchs[3] as related to two wire open lines and later extended by Karakash and Mode[4] to symmetrical closed coaxial coupling. In this report it has been necessary to extend the latter treatment to include the case of unequal characteristic impedances in the individual lines. The characteristic impedances involved have been designated Z_A and Z_B and the continuous

Fig. 1

Cut-away view of base of amplifier-antenna showing center conductor and output coupling loop. It is this lower bi-coaxial region that is termed in this report as the interaction or coupling section.

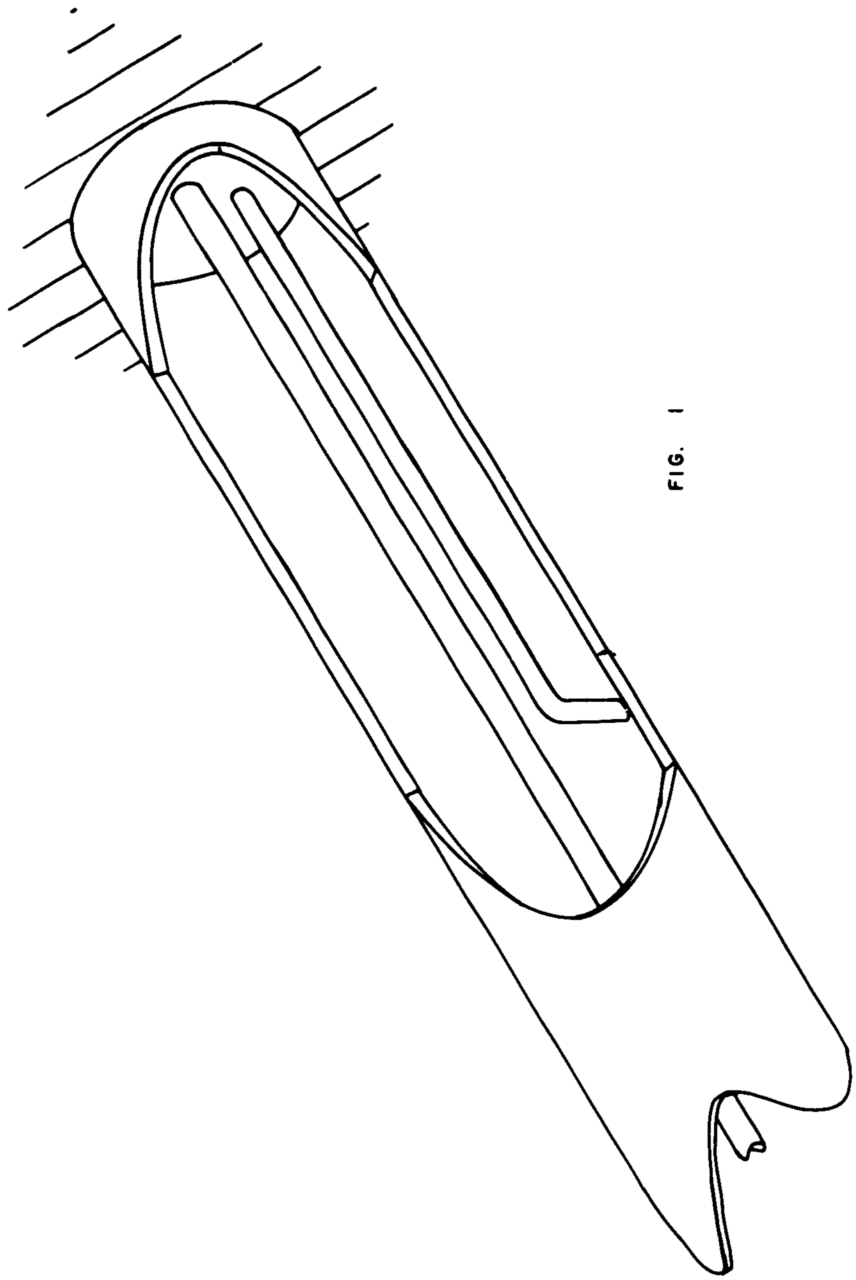


FIG. 1

Fig. 2

Cross-section of the coupling region indicating the termination of each line at the b-b' plane in impedances X and R respectively. These are two-terminal driving point impedances with respect to the inner surface of the coaxial region. The discontinuities introduced by the vertical end portion of the loop and the connectors used at b-b' have not been included in the preliminary calculations of system resonance.

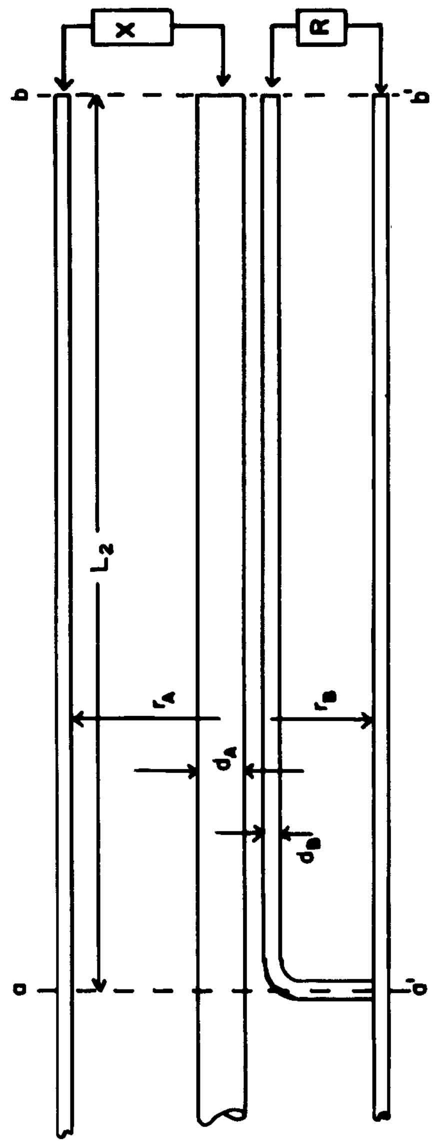


FIG. 2

distributed mutual inductive and capacitative coupling between them as Z_m .

For the steady state case with an assumed sinusoidal excitation of constant frequency we can write a general form equation for the voltage distribution in each line as,

$$V_A = C_A e^{j\theta} + D_A e^{-j\theta} \tag{1}$$

$$V_B = C_B e^{j\theta} + D_B e^{-j\theta} \tag{2}$$

$$I_A = \frac{Z_B C_A - Z_M C_B}{\Delta} e^{j\theta} - \frac{Z_B D_A - Z_M D_B}{\Delta} e^{-j\theta} \tag{3}$$

$$I_B = \frac{Z_A C_B - Z_M C_A}{\Delta} e^{j\theta} - \frac{Z_A D_B - Z_M D_A}{\Delta} e^{-j\theta} \tag{4}$$

where $\theta = \gamma X \qquad \Delta = Z_A Z_B - Z_M^2$ (5)

γ , complex propagation constant

X , distance along line from plane b - b'

Applying the boundary conditions stated above

$\theta = 0$ at plane b - b'

$$V_A = C_A + D_A \tag{6}$$

$$I_A = \left[Z_A(C_A - D_A) + Z_M(D_B - C_B)\right]/\Delta \tag{7}$$

$$V_B = C_B + D_B \tag{8}$$

$$I_B = \left[Z_A(C_B - D_B) + Z_M(D_A - C_A)\right]/\Delta \tag{9}$$

At plane b - b' the terminating impedances X and R serve to define the ratio of voltage and current in the two lines.

$$V_A/I_A = X \tag{10}$$

$$V_B/I_B = R \tag{11}$$

At plane a - a' $\theta = \phi$

$$V_A = C_A e^{j\phi} + D_A e^{-j\phi} = V_s \tag{12}$$

$$I_A = \frac{Z_B C_A - Z_M C_B}{\Delta} e^{j\phi} - \frac{Z_B D_A - Z_M D_B}{\Delta} e^{-j\phi} = I_s \tag{13}$$

$$V_B = C_B e^{j\phi} + D_B e^{-j\phi} \tag{14}$$

By the substitution of equations (6) and (7) into (10) and of (8) and (9) into (11) we obtain two equations involving the four line amplitude parameters C_a, C_b, D_a and D_b, the terminating impedances and the length of the interaction region. Together with (12) and (14) we have then four equations as shown.

$$\begin{bmatrix} 0 \\ V_s \\ 0 \\ 0 \end{bmatrix} = \begin{bmatrix} 0 & 0 & e^{j\phi} & e^{-j\phi} \\ e^{j\phi} & e^{-j\phi} & 0 & 0 \\ (XZ_B - \Delta) & -(XZ_B + \Delta) & -XZ_M & XZ_M \\ -RZ_M & +RZ_M & (RZ_A - \Delta) & -(RZ_A + \Delta) \end{bmatrix} \times \begin{bmatrix} C_A \\ D_A \\ C_B \\ D_B \end{bmatrix}$$

The resulting solutions for the line constants will include the arbitrary magnitude V_s .

$$Z_{A-A'} = \frac{V_A}{I_A} = \frac{\Delta \left[C_A e^{j\phi} + D_A e^{-j\phi}\right]}{(Z_B C_A - Z_M C_B)e^{j\phi} - (Z_B D_A - Z_M D_B)e^{j\phi}} \tag{15}$$

$$= \frac{G}{K} = \frac{g + jg'}{k + jk'} \tag{16}$$

Where $Z_{a-a'}$ as defined above is the impedance seen by line "A", and

$$g = \Delta\left[2\Delta^2 \cos(2\phi) - 2\Delta X' Z_B \sin(2\phi) - 2\Delta^2\right] \tag{17}$$

$$g' = \Delta\left[2(X'RZ_A Z_B - X'RZ_M^2)\cos(2\phi) + 2\Delta RZ_A \sin(2\phi) + 2X'RZ_A Z_B - 2X'RZ_M^2\right] \tag{18}$$

$$k = Z_B\left[2\Delta RZ_A \cos(2\phi) + 2(X'RZ_M - X'RZ_A Z_B)\sin(2\phi) + 2\Delta RZ_A\right] - 4Z_M^2 \Delta R \tag{19}$$

$$k' = Z_B\left[2\Delta X' Z_B \cos(2\phi) + 2\Delta^2 \sin(2\phi) - 2\Delta X' Z_B\right] \tag{20}$$

are the real and imaginary components of the numerator and denominator as indicated in equation (16). In order to collect the vector terms we have substituted

$$X = jX'$$

with X' the positive or negative magnitude of the reactance terminating line "A" at plane (b - b').

Parametric Amplifier

To maximize the flow of current in the varactor at the signal and idler frequencies it is necessary that the varactor element see a conjugate reactance match at these frequencies, at the o-o' terminal plane , as shown in Fig.3 . The impedance $Z_{o-o'}$ is a function of $Z_{a-a'}$ as seen through the length L_1 of line "A" which extends to the left of the a-a' boundary of the interaction region discussed in the previous section.

We will assume that the length of the interior region , the sum of L_1 and L_2 , is equal to external cylinder length L. It is mechanically possible to put plane b-b' at any point within the volume ,thereby reducing the interior length or conversely one can introduce dielectric materials which would increase its effective electrical length. In the case of the grounded quarter wave dipole it would also be possible to increase the coaxial region length by extending the antenna cylinder below the ground plane. A considerable time was spent in the analysis of such techniques with the conclusion that all suffered from a common difficulty in accomplishing the adjustments required to allow for variations in varactors. The technique for the control and adjustment of the resonance spectrum of the coaxial cavity through the selection of a proper form for reactance X as presented in this and following sections has proved to be simple,reliable and subject to close control through direct measurements prior to final tuning steps.

Examination of the equations developed in the previous section devoted to the coupling region confirm the expectations based on magnetic field distribution considerations that at those frequencies at which the output loop length is a half wavelength (or multiples thereof) the impedance $Z_{a-a'}$ will be a pure reactance equal to X and there will be no output at R .

Fig. 3

Lateral cross-section of quarter-wave parametric amplifier antenna dipole. The pumping signal input loop has been omitted for clarity. It is about 20% the length of the output loop and loacted diametrically opposite at the base of the antenna.

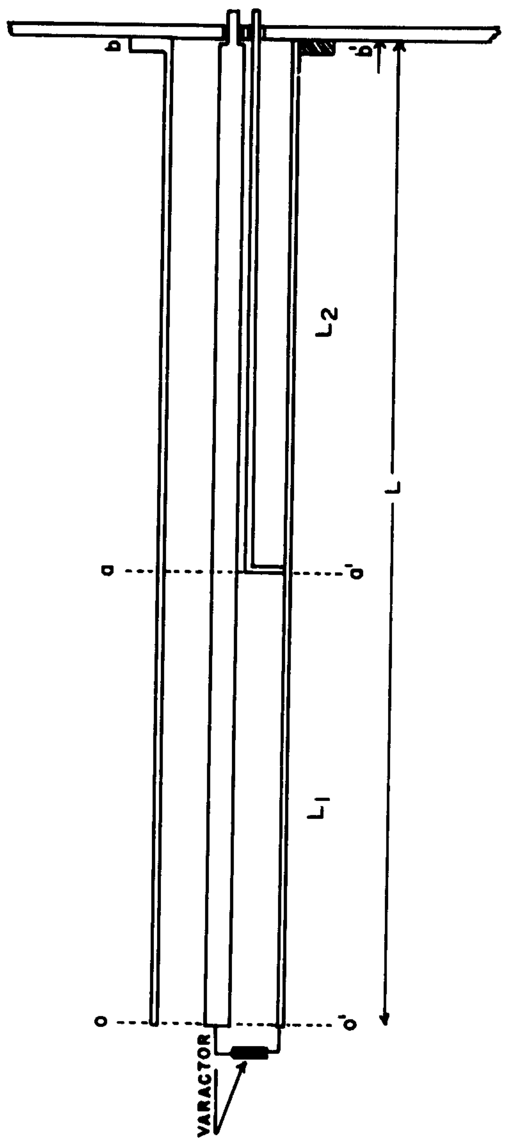


FIG. 3

System Design

In a number of cases in the following section specific design parameters have arisen as a consequence of design choices regarding component dimensions and the selection of operation at 54 Mc. and 108 Mc. At this point in our work these values are not to be considered optimum selections or recommendations. We will endeavor to indicate these parameters as they arise in the discussion and how values appropriate to other choices could be obtained.

As noted previously the relative antenna dipole diameter at the operating frequency establishes the optimum length L. For the quarter wave dipole this was 0.238λ . The size of the output loop and in particular its length L_2 is determined by (a) the need to provide adequate output coupling of the amplified signal and (b) the need to minimize coupling to the idler resonance mode. If we choose an idler frequency (f_i) at five times the signal frequency (f_s) then,

$$L \equiv .238\lambda_s \equiv 1.190\lambda_i$$

$$L_1 = .138\lambda_s = .690\lambda_i$$

$$L_2 = .100\lambda_s = .500\lambda_i$$

We now have the information necessary to proceed with the computation of $Z_{a-a'}$ using the equations on page 12. Since the value of $Z_{o-o'}$ depends directly on $Z_{a-a'}$ as seen through the length of line "A" L_1 we can include this further transformation directly in our calculations.

If we designate

$$\phi = \phi(s) \quad \text{electrical length of output loop at signal frequency } f_s$$

$$= \frac{2\pi L_2}{\lambda_s}$$

and

$$\theta = \theta(s) \quad \text{electrical length of upper portion of inner coaxial region at } f_s$$

$$= \frac{2\pi L_1}{\lambda_s}$$

Using the values suggested on page 16

$$\phi(s) = 0.2\pi$$

$$\theta(s) = 0.276\,\pi$$

$$Z_{o-o'}(s) = \frac{Z_{A-A'}(s) + jZ_A^* \tan\theta(s)}{Z_A^* + jZ_{A-A'}(s)\tan\theta(s)}\,(Z_A^*) = R_o(s) + jX_o(s)$$

while at the idler frequency f_i

$$\phi(i) = \pi$$

$$\theta(i) = 1.38\pi \quad \text{for a similar calculation}$$

of $Z_{o-o'}(i)$, using in both cases an assumed resistive load R of 50 ohms. For a range of values of X' from 2000 to -2000 ohms we can determine and plot $X_o(s)$ and $X_o(i)$ as a function of X'. With this information we can then proceed to determine the required values for X' at the signal and idler frequencies. Note that Z_A^*, the impedance of the upper coaxial line is not equal to Z_A.

(a) Select a trial value of varactor capacitance C_v
Under actual operating conditions the effective capacitance for tuning considerations will be a mean value of diode capacitance <u>as pumped</u>. This will differ from the DC static value for a specified bias point.

(b) calculate varactor reactance at f_s, $X_v(s)$; and at f_i, $X_v(i)$.

(c) determine that value of $X'(s)$ so that

$$X_{o-o}'(s) = -X_v(s)$$

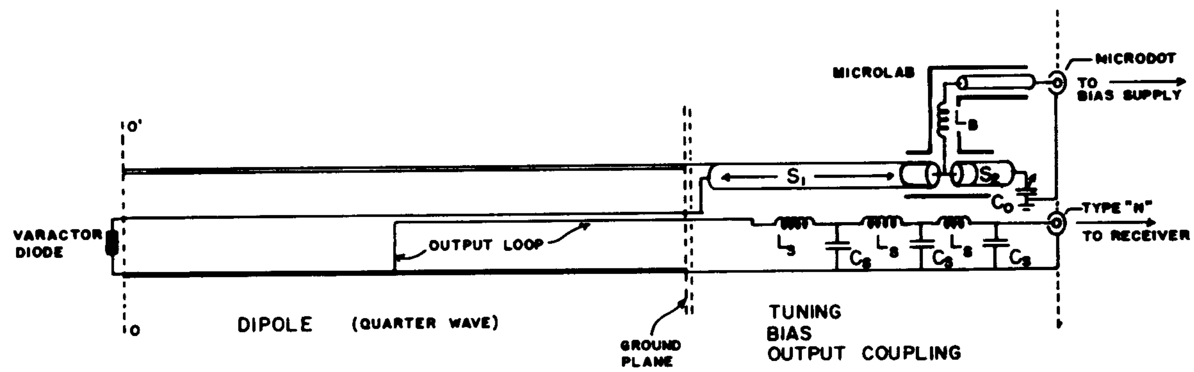
(d) determine that value of $X'(i)$ so that

$$X_{o-o}'(i) = -X_v(i)$$

It now remains to determine a circuit configuration which, when connected to the "X" terminals as shown in Fig. 2 will exhibit these impedances $X'(s)$ and $X'(i)$. At the same time we must be able to provide the desired DC bias to the varactor. These objectives have been accomplished in the manner shown schematically in Fig. 4. A commercial DC bias isolator "T" (Microlab HW-02N) has been combined with a length s_1 of coaxial line and a variable capacitor C_o. By the appropriate selection of the total <u>effective</u> line length between the ground plane end of the coupling region and the lumped capacitor it is possible to meet the impedance conditions noted above. With a suitable correction for the RF fitting (modified BNC) at the dipole base it is possible to verify these impedance conditions using a bridge and slotted line prior to final tuning. Final adjustment is achieved using a signal source set f_s and a pumping source set to $f_i + f_s = f_p$. Starting with a low level of pump power we mutually adjust C_o and the DC bias for maximum gain, increase and repeat.

Fig. 4

Schematic presentation of the combination of bicoaxial resonant region within the antenna dipole and the combination lumped and distributed element region required for tuning and bias connection.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DIPOLE (QUARTER WAVE)
GROUND PLANE
TUNING
BIAS
OUTPUT COUPLING

FIG. 4

Experimental Results

In accordance with the design considerations outlined in the preceding sections the antenna pictured in Figs. 5 to 8 was constructed. The interior coaxial region of the dipole was in the usual form as shown in Fig. 3 with a small additional coupling loop added for the pump signal input. This construction is shown in detail in Fig. 5 , 6 and 7 in the Final Report of Sept. 30 1961 for AF19(604)-3892. The signal/idler tuning arrangement diagramed in Fig.4 of this report, was fabricated in a cast metal base box enclosure as pictured in Fig. 5. In the model pictured here a three section low-pass filter was included in the signal output circuit. With its critical frequency set well above the signal frequency but below the idler it served to further prevent loss of idler or pump frequency energy into the receiver. The filter is located inside of a partition on the left of the interior of the box. Fig.6 shows the tuning elements concerned with the determination of the parameter X',as seen by the coaxial region through a modified BNC connector, prior to final installation within the base box. The system is composed of the combined effect of the monitor "T" (Microlab HW-02N) , two right angle adapters(UG-306/U),BNC cable connector plug (UG-88/U), a length of RG 58/U, a second UG-88/U, a right angle connector (UG-535/U) and a terminating variable air capacitor(MACP-50). The units in this case were selected for convenient assembly and were suited to use in the frequency range of 108 Mc. (signal frequency) and 492 Mc(idler frequency). At higher frequencies the use of connectors would be less satisfactory and this whole system could resolve into a very simple strip-line structure , etched to approximate size and tuned by dielectric or conducting posts.

In the upper right of the inclosure can be seen a type N bulkhead

Fig. 5

Enclosure at base of antenna showning signal/idler tuning line and capacitor, output loop filter and pump signal input. Standard BNC connectors are used to connect the special tuning section to the center conductor so that a preliminary measurement of this impedance (X) can be made before final assembly.

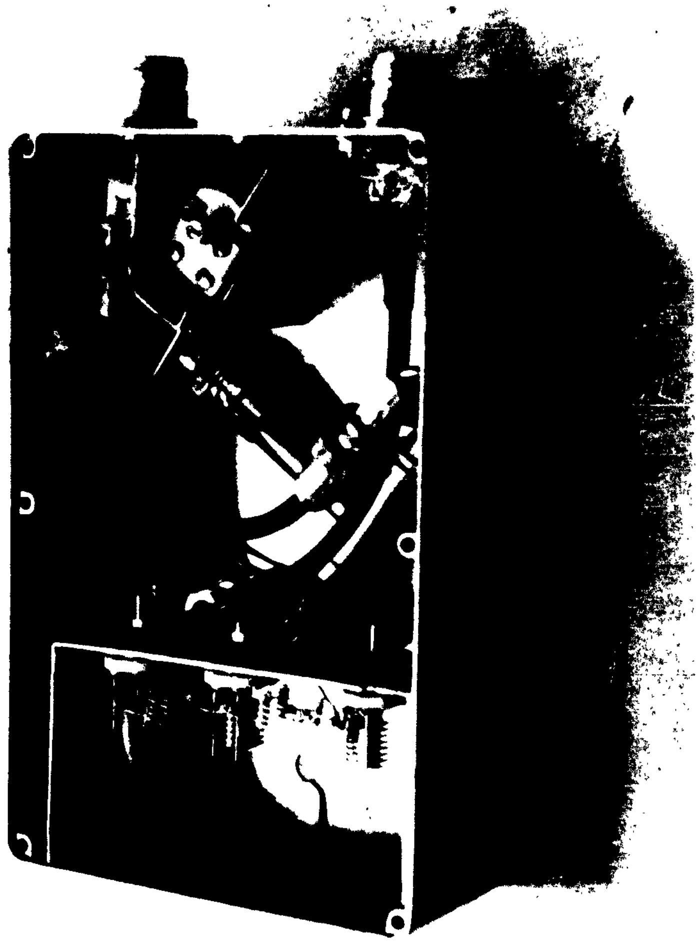
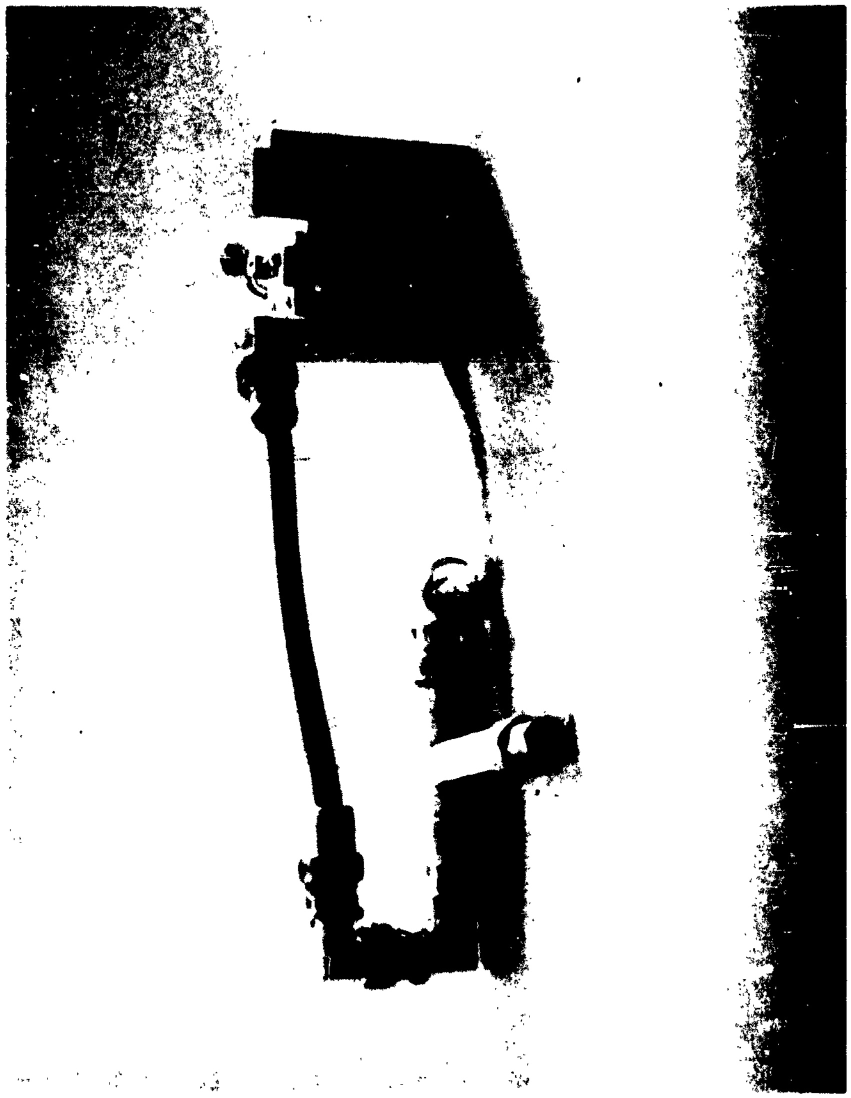


Fig. 6

Coaxial tuning section and bias input "T" prior to assembly in tuning box inclosure. The air capacitor is located on the right.



receptacle(UG-680/U) used for the amplified RF output. At the lower end of the same side can be seen the Microdot receptacle(31-01) used for the DC bias and a second UG-680/U for the pump signal input.

A top view of the lower portion of a quarter wave dipole is shown in Fig. 7 . In operation the hexagonal head machine bolts which can be seen on the flange base are used to support the dipole which is attached to the ground plane sheet with the tuning enclosure below the surface as shown in the sketch at the right. Preliminary adjustment of the tuning section, involving the length of the cable section and a tentative setting of the capacitor, was accomplished using a Hewlett-Packard VHF bridge. As a consequence satisfactory operation was obtained by control of bias and pump level alone. Operating gains of 8 to 14 db were measured at 108 Mc. These results were not necessarily optimum in terms of efficient pumping or noise figure. Our capability for measurements in these two areas is being improved and we expect to include these important parameters in future design reports.

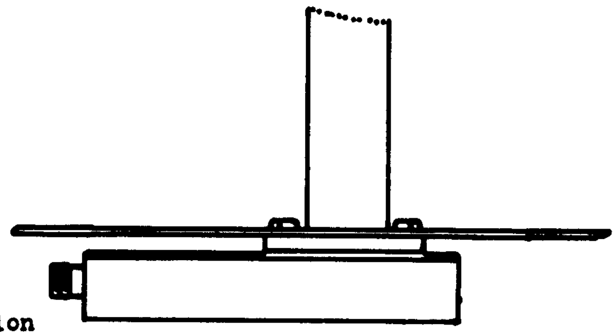
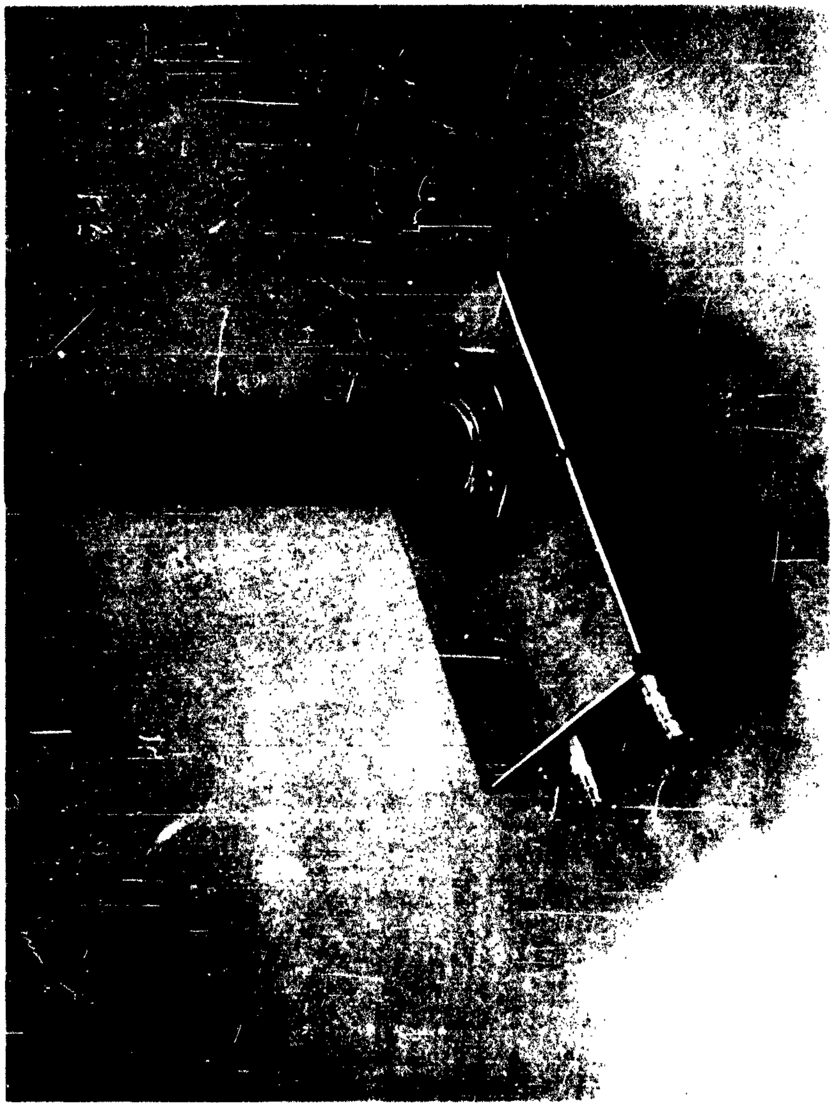


Fig 7

Fig. 7

View of the base of the assembled dipole and tuning unit.



Calculations

The calculations required for the determination of the X' termination of the interaction region to provide resonant reactance conditions at the upper of the dipole have been indicated as equations (16) to(20) on page 12. These expressions involve simple,direct evaluation but require a large number of computations for each finally determined coordinate point. They have accordingly been programmed for machine computation in the form of a table of $Z_{o-o'}$ values as a function of X'. This table or series of tables correspond to a decision regarding the relative length of the entire coaxial region and of the coupling region, combined with information on the three geometrical parameters representing the characteristic impedance of the symmetrical coaxial structure consisting of the center conductor and the inner surface of the antenna cylinder, Z_A; an asymmetric systems comprising the output loop and the inner cylinder surface, Z_B ; and the mutual impedance between them, Z_M .

Through a conformal transformation it is possible to determine an equivalent two-wire-over-ground-plane configuration for a balanced two wire coaxial line such as RG 22/U. From the former it is then possible to compute the self and mutual impedances . We have sought to develop a modification of this transformation which would transform the structure at hand to one subject to direct calculation. Pending the outcome of this study for an explicit expression for the needed parameters preliminary computations were made using estimated values based on single wire systems and measurements of mutual capacity. While indicating in a general way the validity of the approach, they are too approximate for the needs of the low loss and hence high Q systems in the present designs. In view of this limitation we have supplemented our information on an actual antenna model by direct measurement of the existing pattern of resonant frequecies for various known terminations of the "A" and "B" lines.

A recent report by Matsumoto[5] has presented two new avenues for further progress in this important area. These are (a) his suggestion of a transformation which presents an asymmetric bi-coaxial structure, in which neither conductor is on the central axis, equivalent to the cross-section geometry which characterizes the antenna amplifier, that is one central and one off center conductor, and (b) the presentation of a multiple image solution to the self and mutual impedance matrix. It is anticipated that both of these contributions will be significant. The latter,in allowing approximate solutions for structures of specific design and the former in furthering the search for explicit forms needed for optimization.

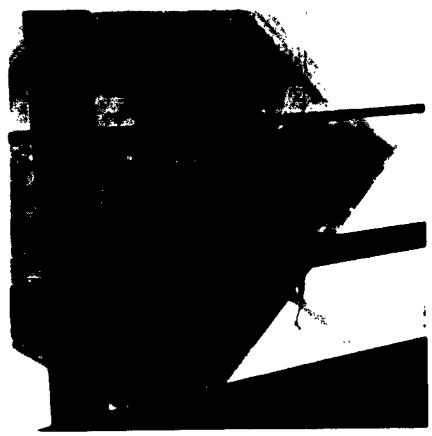
THE SLOT AMPLIFIER ANTENNA

The upper frequency limit for the practical application of the coaxial dipole form of "parant" has not been explored in detail. It would appear,however,that somewhat above 350 to 400 Mc. the necessary physical size of a "thin" dipole would seriously complicate the mechanical fabrication of the inner region and in particular the output loop. In view of recently reported copper-clad materials with reduced loss at high frequencies it is possible that some form of printed or etched configuration could extend this limit somewhat higher. We have according considered other forms of antenna and in particular the ground plane slot backed by a resonant cavity. This work is in a preliminary stage and has involved the evaluation of input, output coupling methods,diode location and the use of parasitic cavities or resonator arrays for idler storage. Slots used have been of the typeshown in Fig. 9(a) built for experiments at 365 Mc. Measurements have been made using the outdoor ground plane screen shown in Fig. 9(b) This screen has been erected for this purpose on the rear wall of the Engineering building at the University. Convenient access to the rear of the slot and the cavity walls is possible through a laboratory window as can be seen in Fig 9(c); which leads to a special research area devoted to this work.

Fig. 9(a) Test slot prior to mounting .
Constructed of sheet copper the
slot aperture is / at 365 Mc.

Fig 9(b) Twelve by twelve foot vertical reference
ground plane shown as erected on the
rear wall of Kingsbury Hall. U.N.H.
A test slot is in position near the
center of the plane.

Fig. 9(c) Interior view of ground plane shown in
Fig. 9(b) above showing ready access to
the rear terminals of the slot structure.



CONCLUSIONS

A consideration of the elements involved in the effective design of a parametric amplifier antenna in coaxial dipole form has lead to an analysis of the impedance character of the interior region of the antenna. In the dipole as presently used this region is characterized by a simple single center conductor coaxial line section in the outer or end region and an asymmetric dual coaxial section in the center or lower portion. This interaction or coupling region involves mutually coupled, distributed parameter lines and is in effect a three-port. The varactor diode views this region as a driving point impedance seen through the upper port and rotated through the electrical length of the upper line section. In order that the varactor , as pumped , be tuned to the signal and idler frequencies a significant measure of control must be had over the magnitude and sign of this impedance. Assuming that the output loop is most conveniently terminated in a simple resistance equal to that of flexible transmission cable, control of the reactance-frequency pattern at the remaining port will serve to provide the necessary tuning adjustment and incidently provide access for DC bias. Realization of this tuning reactance in the form of a length of coaxial line terminated by a lumped capacitor has made a significant improvement in the practical aspects of tuning.

Further work is now in progress to improve our evaluation of the self and mutual impedances involved in the coupling section. While conventional antenna theory has been used as a preliminary guide in the selection of an antenna length a consideration must be made of the perturbation in this value due to the active impedance of the amplifier and varactor.

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Project Staff

During the period of this report 1 February 1962 to 31 October 1962 the following persons were employed on a part time basis by the project and participated in various phases of the work described in this report.

Albert D. Frost	Project Director, Professor of Electrical Engineering.
William Borkowski	Research Assistant, Programming and machine computation
Ray Harrington	Research Assistant, Electronic Technician
Ernest Nichols	Research Assistant, Measurements and Computation.

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2430 E Street, N.W.
Washington 25, D.C.

ASTIA (TIPAA)
Arlington Hall Station
Arlington 12, Virginia

Scientific and Technical Information Facility
Attn: NASA Representative (S-AK/DL)
P.O. Box 5700
Bethesda, Maryland

Director
Langley Research Center
National Aeronautics and Space Administration
Langley Field, Virginia

Chief, Bureau of Naval Weapons
Department of the Navy
Washington 25, D.C.
Attn: DLI-31

Director (Code 2027)
U.S. Naval Research Laboratory
Washington 25, D.C.

Director, USAF Project RAND
The Rand Corporation
1700 Main Street
Santa Monica, California
THRU: AF Liaison Office

AFCRL, OAR (CRXRA - Stop 39)
L.G. Hanscom Field
Bedford, Mass.

Technical Information Office
European Office, Aerospace Research
Shell Building, 47 Cantersteen
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U.S. Army Aviation Human Research Unit
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P.O. Box 428, Fort Rucker, Alabama
Attn: Maj. Arne H. Eliasson

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Bedford, Mass.

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Research Laboratory
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Director, U.S. Army Ordnance
Ballistic Research Laboratories
Aberdeen Proving Ground, Maryland
Attn: Ballistic Meas. Laboratory

Pickard & Burns, Inc.
240 Highland Avenue
Needham 94, Mass.

Standard Oil Company
Chemical and Phys. Rsch. Div.
440 Warrensville Cntr. Road
Cleveland 28, Ohio

Philco Corporation
Weapons Dev. Lab.
3875 Fabian Way
Palo Alto, California

Sylvania Electric Products, Inc.
Electronic Systems Division
Missile System Laboratory
100 First Avenue, Waltham, Mass.

General Electric Company
General Engineering Laboratory
1 River Road, Schenectady, N.Y.

Commanding Officer and Director
U.S. Navy Electronics Lab. (Lib)
San Diego 52, California

California Institute of Technology
Jet Propulsion Laboratory
4800 Oak Grove Drive
Pasadena, California
Attn: Mr. I.E. Newlan

Raytheon Company
Box 555, Hartwell Road
Bedford, Mass.
Attn: I. Goldstein, Manager
Solid State Physics

Commander
Pacific Missile Range
Attn: J.N. Shellabarger
U.S. Naval Missile Center
Point Mugu, California

The University of New Hampshire
Durham
New Hampshire
Attn: A.D. Frost

Microwave Associates, Inc.
South Avenue
Burlington, Mass.

AD

University of New Hampshire
Antenna Systems Laboratory, Dept. of
Electrical Engineering, Durham N. H.

THE COAXIAL DIPOLE PARAMETRIC AMPLIFIER ANTENNA, by A.D. Frost, 1 December 1962.
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A study has been made of the theoretical and applied problems associated with the design of a parametric amplifier antenna in coaxial dipole form. An analysis is made of the coaxial interior which serves to provide the resonances needed for parametric amplification, and an expression is derived for the complex impedance seen by the varactor diode in terms of the inner coaxial geometry, the relative electrical length of the output loop the receiver input resistance and an adjustable terminating reactance. This latter element permits the tuning of the system for a desired pair of signal and idler frequencies. A completed tuning and bias system for a quarter wave dipole at 108 Mc. is shown

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