Reduction of Mutual Coupling Between Active Monopoles: Application to Superdirective Receiving Arrays

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Abstract—Theoretical results and measurements of mutual coupling between active monopoles show that the incorporation of a field-effect transistor (FET) with a shunt input inductance leads to large reduction of interactions at reception. Each active monopole then works as an isolated element, and feed requirements of active array become easier to obtain. Realization of a superdirective array with an appreciable bandwidth confirms the main possibilities of active antennas.

I. INTRODUCTION

IN A PREVIOUS paper [1] a definition of mutual coupling between antennas was given; it was proved theoretically that an optimal load exists for a null coupling and that the knowledge of constant coupling circles allows us to choose proper loads for low coupling. Then an experimental method [2] was developed and tested for measurements of coupling between two monopoles (using passive loads).

Here, constant coupling circles of two monopoles enable us to choose a field-effect transistor (FET) (with a shunt input inductance) which results in a large reduction of interaction. Experimental results are in good agreement with theoretical predictions. Owing to these suppressed interactions, design and realization of superdirective arrays become easier, even though passive arrays may remain unusable because of the extreme sensitivity to small changes in frequency.

An experimental active array which permits superdirective performance with an appreciable bandwidth has already been described for transmission [4]. For reception, bipolar transistors do not offer any coupling improvement [1], whereas field-effect transistors provide good reduction; so using Schelkunoff's theorem [3] a four active element end-fire array has been investigated for reception.

II. MUTUAL COUPLING OF TWO ACTIVE MONOPOLES

Coupling is given by the following formula [1]

$$C_r = C_0 \left| \frac{1 - \dfrac{\Gamma_T}{\Gamma_0}}{1 - \dfrac{\Gamma_T}{\Gamma_M}} \right|,$$

where C_0, Γ_0, Γ_M are parameters that depend on the physical dimensions of monopoles, frequency, and a normalization impedance (50Ω generally).

Γ_T is the complex reflection coefficient of load. When one uses a transistor, $\Gamma_T \approx S_{T_{11}}$ (input reflection coefficient) for reception and $\Gamma_T \approx S_{T_{22}}$ (output reflection coefficient) for transmission. In Fig. 1, different circles for two monopoles (—height = 7.5 cm and radius = 0.15 cm, spacing = 3.75 cm) at four frequencies are plotted. It appears that Γ_0 always remains inductive between 0.8 GHz and 1.1 GHz, when $S_{T_{11}}$ of FET (Plessey type GAT1 in common source) at 0.9. GHz exhibits a capacitive reactance for any bias (V_{DS} = 5 V and I_D = 1, 2, 5, 8, 12.5 mA). To get low coupling Γ_T must be close to Γ_0, for instance, inside a −10-dB circle. So one can put a shunt inductance L_c, its value being chosen in such a way that $S_{T_{11}}{}'$, (the new input reflection coefficient) moves toward Γ_0.

The experimental $S_{T_{11}}{}'$ parameter measurement device is given in Fig. 2(a); the shunt inductance L_c is a short conductor wire ended with a 1000-pF capacitor. $S_{T_{11}}{}'$ has been measured for four biases between 0.8 GHz and 1.1 GHz (Fig. 2(b)). The knowledge of constant coupling circles and $S_{T_{11}}{}'$ allows us to draw coupling (C_r) in term of frequency.

In Fig. 3(a), it appears that C_r reaches a very small value (−30 dB) for each drain current I_D between 0.9 and 0.95 GHz and keeps practical values between 0.85 and 1 GHz (C_r lower than −15 dB).

To verify experimentally the theoretical predictions, two printed monopoles were developed; they behave identically to the cylindrical monopoles and facilitate contact with the feeding networks. The photographs of Fig. 4 show the antennas above the ground plane (Fig. 4(a)) and the electronic circuit below (Fig. 4(b)). Each transistor is biased with two high-impedance quarter-wave lines, which do not disturb the 50-Ω transmission line. At the input of the two FET's, one can recognize the shunt inductance. Measurements of coupling have been performed using the experimental method previously developed and tested [2]; owing to the guide, which is visible on the right of Fig. 4(b), it was easy to remove antenna 2 to find a good null of the exterior interfering field, and then to put it back for measurement of coupling.

Experimental results (Fig. 3(b)) confirm theoretical values of Fig. 3(a) and prove that a very low mutual coupling can be obtained for a good choice of active device relative to the positions of mutual coupling circles.

III. SUPERDIRECTIVE ARRAY AT RECEPTION

The mathematical theory of linear arrays developed by S. A. Schelkunoff [3] states that an array may be made more

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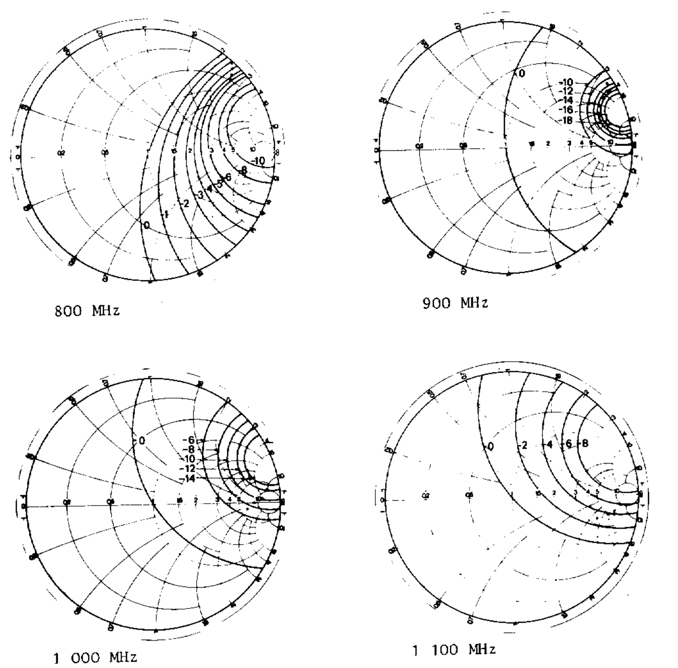
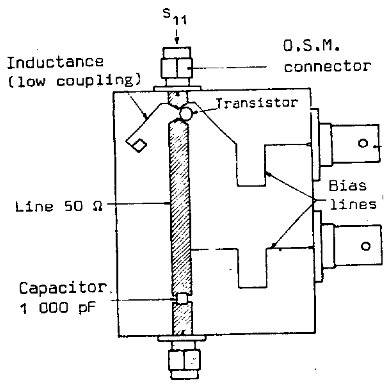
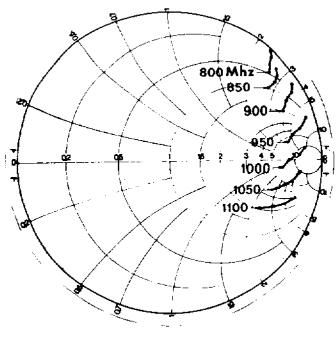


Fig. 1. Mutual coupling circles of two monopoles. Height = 7.5 cm, radius = 0.15 cm, spacing = 3.75 cm. Four frequencies: 0.8 GHz: $C_0 = -11.4$ dB, 0.9 GHz: $C_0 = -8.4$ dB, 1 GHz: $C_0 = -8.2$ dB, 1.1 GHz: $C_0 = -8.4$ dB. C_0 = value of mutual coupling for a load Z_0 = 50 Ω.

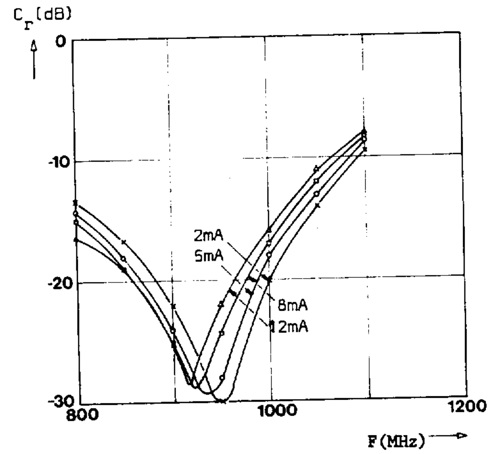


(a)

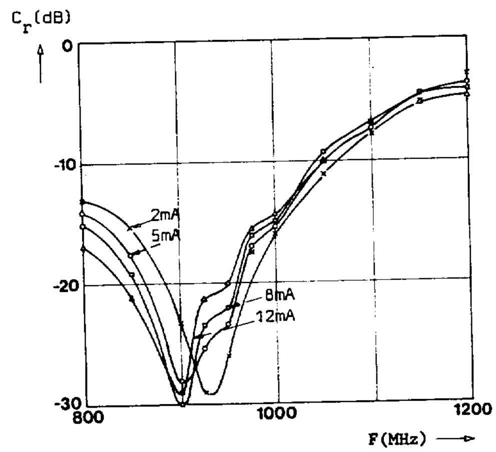


(b)

Fig. 2. (a) S_{T11} measurement fixture. (b) Measured values of S_{T11}' of FET GAT1 with a shunt inductance for different bias: V_{DS} = 5V I_D = 2, 5, 8, 12.5 mA.

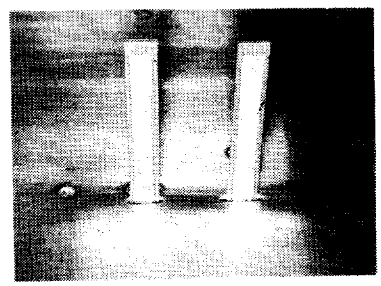


(a)

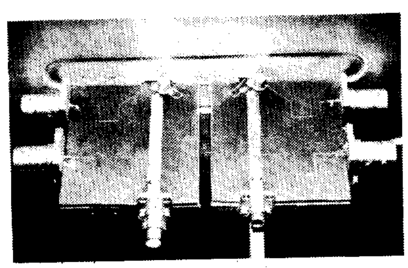


(b)

Fig. 3. (a) Theoretical coupling of two active monopoles versus frequencies for different bias (height = 7.5 cm; radius = 0.15 cm; spacing = 3.75 cm). (b) Measured values of coupling between two active monopoles versus frequencies (h = 7.5 cm, a = 0.15 cm, d = 3.75 cm).



(a)



(b)

Fig. 4. (a) Active printed monopole above the reflector plane. (b) Electronic circuit of the active monopole (under the reflector plane).

directive if its total length is kept constant, but the number of elements increases. This improvement of directivity can be secured only for a suitable feeding. However, design and realization of a superdirective array with passive antennas are quite difficult because of low or negative resistance due to large coupling. Integration of suitable active devices (such as FET's) with antennas permits a reduction of coupling. Then each active monopole works as an isolated element, the properties of which can be represented by the following formula [5]

$$I_i = Y_i V_i + I_{0_i}, \qquad \text{with } I_{0_i} = T_i \frac{E_i}{\beta},$$

where I_i and V_i are output current and voltage of active antenna and Y_i and T_i are coefficient depending on geometrical dimensions and frequency (expressions of $[Y]$ and $[T]$ are reduced to diagonal matrices because the off-diagonal elements become negligible in comparison with diagonal terms. E_i is the incident field on monopole i. β is the free space propagation constant.

Let us suppose now that these n equispaced sources (spacing $= d$) are joined by suitable lines of characteristics admittance $Y_{c_i}(Y_{c_i} = Y_i$ for Y_i real) and deliver power in the load Z_c.

In the output plane

$$\sum_{i=0}^{n-1} I_i' = V \sum_{i=0}^{n-1} Y_i + \sum_{i=0}^{n-1} I_{0_i} e^{j\phi_i},$$

where ϕ_i is the phase increment of line i, and for match conditions ($Y_c = 1/Z_c = \Sigma_{i=0}^{n-1} Y_i$) the received power is

$$P = \frac{1}{8} \left| \sum_{i=0}^{n-1} I_{0_i} e^{j\phi_i} \right|^2.$$

If the direction of incident waves makes an angle θ with the line of sources, P may be expressed as follows

$$P = \frac{1}{8} \left| \sum_{i=0}^{n-1} A_i e^{j\alpha_i} e^{j(i\beta d \cos\theta - \alpha)} \right|,$$

where $A_i = |I_{0_i}|$, u_i is the phase of I_{0_i}, $\alpha_i - i\alpha = u_i + \phi_i$, α is a progressive phase delay, and α_i is the phase deviation from the above progressive phase delay. So the active linear array obeys [3, theorem 1]; it can be represented by a complex polynomial

$$P = \frac{1}{8} \left| \sum_{i=0}^{n-1} a_i z^i \right|, \qquad \text{where } z = e^{j(\beta d \cos\theta - \alpha)},\ a_i = A_i e^{j\alpha_i}.$$

It is then possible to choose the $(n-1)$ roots equispaced between 0 and $-2\,\beta d$ to obtain a superdirective end-fire array with a major lobe narrower than the classical end-fire array (with$(n-1)$ equispaced roots between 0 and 2π.)

Theoretical and Experimental Results

Since the passive monopoles studies for coupling were resonant at 915 MHz, the frequency band has been centered on this value. The array consists of four identical elements with spacings equal to 3.75 cm. In Table I, two sets of complex coefficients of end-fire arrays ($\alpha = \beta d$) are given. Realization of superdirective feeding appears to be quite difficult for reception with passive elements because of the large variations of amplitude and phase from one element to one another. On the other hand, bias possibilities of transistors enable us to fit good amplitudes; phases are adjusted with different length of line. An example of active array is presented on Fig. 5. Antenna and circuitry have been made using the usual printed-circuit method; the removal of the copper ground plane (for antenna only) does not cause a significant change. A shunt inductance L_c (for a low coupling) associated with an FET, follows each monopole. To obtain a quasi-unilateral active device, a reactive cell composed of an inductance in series with a capacitor is put between the drain and gate; it is then easy to match the transistor output to 100 Ω with a simple network (series inductance L_s plus a shunt inductance L_p); each transistor is properly biased vias a λ/4 short-circuited line of high impedance and a 100-Ω transmission line of correct length give good phase distribution. A 50-Ω output is obtained using a quarter-wave transformer. Calculations have been performed with FET Plessey GAT 1 for the values defined in Table II.

Reductions of coupling effects appear clearly from a comparison of $[Y_A]$ and $[T_A]$ of the passive array and the $[Y_T]$ and $[T_T]$ of the active one in Table III. Thus each active

TABLE I
EXAMPLE OF COEFFICIENT OF COMPLEX POLYNOMIAL RELATING TO SUPERDIRECTIVE OR CLASSICAL END-FIRE ARRAY

Element	Superdirective end-fire array A_i	Superdirective end-fire array α_i	Classical end-fire array A_i	Classical end-fire array α_i
0	1	15.3	1	0
1	2.775	-109.8	1	0
2	2.775	125.1	1	0
3	1	0	1	0

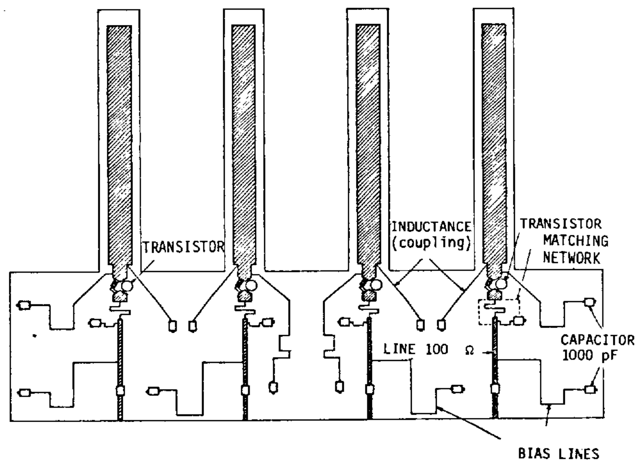


Fig. 5. Structure of a four active monopoles array (aerials and electronic circuit).

TABLE II
THEORETICAL DATAS USED FOR COMPUTING SUPERDIRECTIVE END-FIRE ARRAY PROPERTIES

Element	*Bias* (V_{DS}=5V) I_D (mA)	L_{e}(nH)	L_{s}(nH)	L_{p}(nH)	*Length of line* (cm)
0	1.1	15	30	9	0
1	12.5	13	29	22	16.18
2	12.5	13	29	22	2.04
3	1.1	15	30	9	20.14

TABLE III
MATRIX $[Y_A]$ AND $[T_A]$ OF PASSIVE ARRAY AND MATRIX $[Y_T]$ AND $[T_T]$ OF ACTIVE ARRAY (WITH AN OUTPUT MATCHING) AT 0.915 GHZ

ELEMENT	*PASSIVE ARRAY*				*ACTIVE ARRAY*			
ij	$\|Y_A\|$	PHASE (Y_A)	$\|T_A\|$	PHASE (T_A)	$\|Y_T\|$	PHASE (Y_T)	$\|T_T\|$	PHASE (T_T)
11	137.3039	36.16	.1313E 00	35.09	10.2884	-2.26	.1425E-01	131.58
12	113.0505	-141.91	.1099E 00	-142.16	.0927	102.93	.9366E-03	-84.06
13	103.1434	-168.88	.1006E 00	-168.98	.0666	72.79	.7727E-03	-110.60
14	121.9821	36.76	.1183E 00	36.99	.0003	49.49	.3751E-03	-148.11
21	113.1296	-141.95	.1094E 00	-141.88	.0016	99.45	.1501E-02	-109.11
22	121.2691	37.56	.1159E 00	35.99	10.0890	.18	.3889E-01	115.37
23	89.0219	-17.90	.8676E-01	-17.69	.2510	86.74	.2532E-02	-100.22
24	103.2318	-168.92	.1001E 00	-168.71	.0012	69.31	.1340E-02	-131.80
31	103.2318	-168.92	.1001E 00	-168.71	.0012	69.31	.1340E-02	-131.80
32	89.0219	-17.90	.8676E-01	-17.69	.2510	86.74	.2532E-02	-100.22
33	121.2691	37.56	.1159E 00	35.99	10.0890	.18	.3889E-01	115.37
34	113.1296	-141.95	.1094E 00	-141.88	.0016	99.45	.1501E-02	-109.11
41	121.9821	36.76	.1183E 00	36.99	.0003	49.49	.3751E-03	-148.11
42	103.1434	-168.88	.1006E 00	-168.98	.0666	72.79	.7727E-03	-110.60
43	113.0505	-141.91	.1099E 00	-142.16	.0927	102.93	.9366E-03	-84.06
44	137.3039	36.16	.1313E 00	35.09	10.2884	-2.26	.1425E-01	131.58

monopole behaves as an isolated element. Computed patterns of the active end-fire array are plotted in Fig. 6 for same frequencies; superdirectivity is well-demonstrated if one compares the classical end-fire and the superdirective pattern. Moreover the superdirective performance can be maintained over an appreciable frequency band. Measurements confirm the theoretical prediction (Fig. 7); in fact the experimental 3-dB beamwidth and frequency bandwidth are a little larger than calculated values because of differences between experimental and theoretical feedings (amplitude and phase). However, the array remains very directive (around 80° beamwidth at the 3-dB points) for such a small length structure (the overall length equals 0.34 λ at 0.915 GHz).

IV. CONCLUSIONS

The performance of receiving monopoles loaded by transistors has been studied theoretically as well as experimentally. A

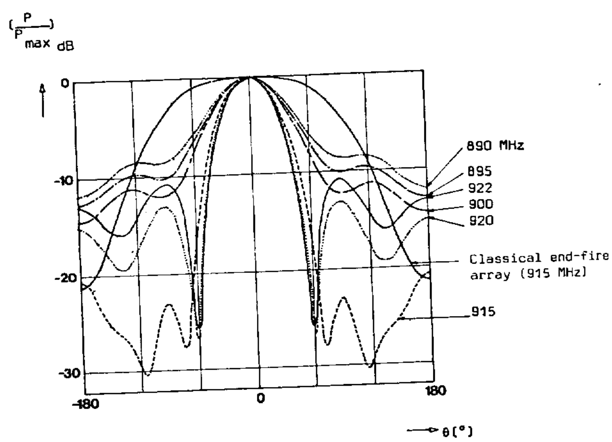


Fig. 6. Theoretical patterns of the end-fire superdirective array for different frequencies.

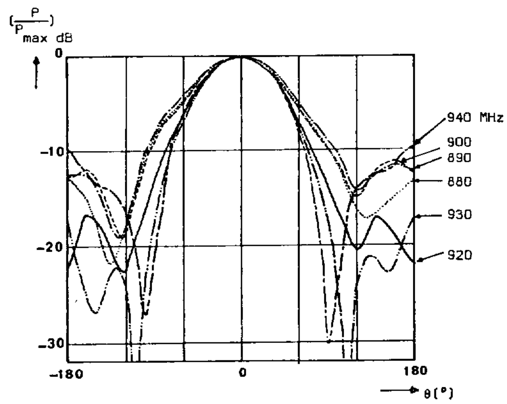


Fig. 7. Experimental patterns of the end-fire superdirective array for different frequencies.

knowledge of constant coupling circles enables us to choose appropriate microwave transistors for suppression of mutual effects. On reception, field-effect transistors exhibit interesting parameters; however, a shunt inductance is necessary at the transistor input to obtain the important improvement of coupling.

Since mutual coupling effects can be suppressed, design of superdirective array (using active antennas) becomes easier than in the passive case. The required current distribution is adjusted by a correct set of transistor bias and phasing lines. Experimental results confirm theoretical predictions for receiving end-fire active array.

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Mutual Coupling Coefficients in Collector Arrays on Circular Cylindrical Concave Surfaces

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***Abstract*—An analysis is presented of mutual coupling in collector arrays on a concave side of a large conducting circular cylindrical surface. The collector elements are uniformly spaced narrow axial slits which are parallel-plate-guide-fed in the TEM mode. Each element is equipped with a matching network appropriate to broadside scan in the corresponding planar slit array. The properties of coupling coefficients are studied numerically. It is shown that for spacing d less than $\lambda/2$, past a certain curvature dependent neighborhood of the excited element, the rate of decay of the E-plane coupling coefficients is considerably slower than in a planar array with the same elements and spacing. The slower decay of coupling coefficients should be considered in evaluation of the efficiency of collector arrays on concave surfaces. For $\lambda > d > \lambda/2$ the dependence of coupling coefficients on the distance exhibits considerable fluctuations which are attributed to grating-lobe formation. It is also shown that the collector element mismatch may be enhanced by the collector cavity resonances, so that the criteria for admissible voltage standing-wave ratio (VSWR) level of collector elements may be more stringent than in planar arrays.**

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I. INTRODUCTION

A WIDE-ANGLE electronically scannable feed-through lens of a dome antenna [1] type consists of a radiator and a collector array which are space-fed from a planar phased array. The radiator array is located on the convex exterior of the dome, while the collector elements are imbedded in a concave interior dome surface. A typical feed-through module consists of a radiator and a collector element connected back to back via a fixed phase shifter. To achieve an efficient power transfer from the collector array to the radiator array or vice versa, the collector and radiator elements are equipped with separate matching networks appropriate to their respective array environments and to the extreme scan or incidence angles. Matching of the radiator elements is sufficiently well understood; not so is the effect of mutual coupling on the design of the matching network and on the efficiency of collector arrays on concave surface. Experiments have shown [2], [3] that in such arrays the decay of mutual coupling is considerably slower than in planar arrays. For this reason the element match and its scan variation on a concave surface is expected to be affected by a generally larger environment than in a planar array with the same lattice. As a result, the minimum size of a "small" test array for measurement of coupling