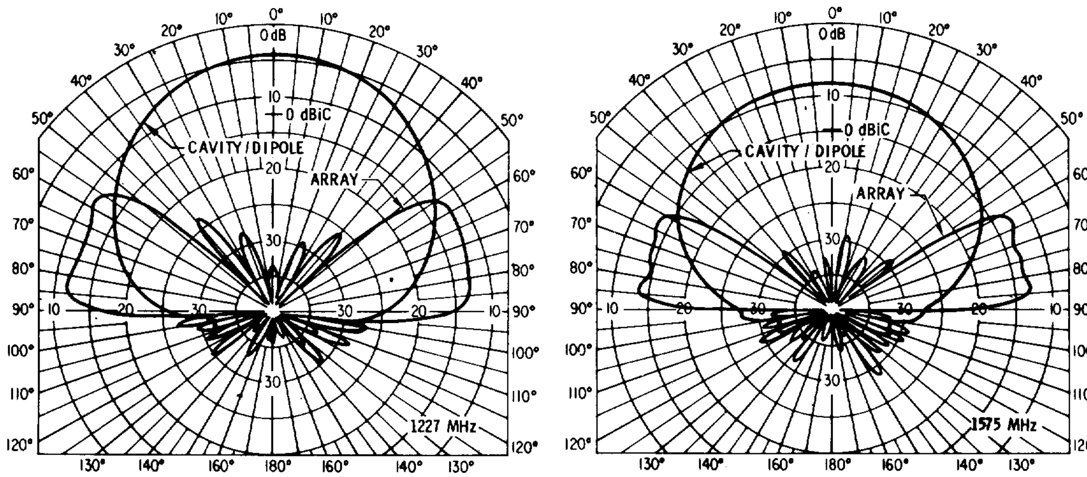

Fig. 6. Measured gain patterns at 1227 and 1575 MHz.

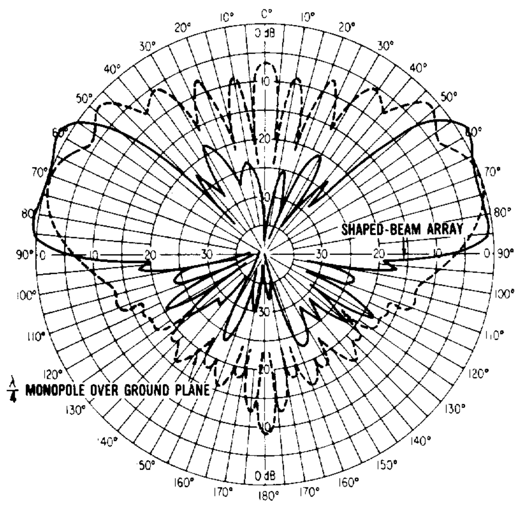

Fig. 7. Comparison of monopole and phased array measured patterns.

angle coverage, cavity-backed turnstile antenna. Although the design of this antenna is straightforward [1], it should be pointed out that a 12-in diameter plate, placed at approximately 0.5 in below cavity aperture, was used to optimize the patterns at both 1227 and 1575 MHz (see Fig. 1).

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Active Antenna Performance Limitation

ROBERT A. SAINATI, MEMBER, IEEE

Abstract—**Many amplifier designs have been proposed for use with electrically small antennas in receiving applications. A study of pertinent noise figure and available power gain equations shows that performance is strongly influenced by antenna resistance. Low values of antenna resistance limit noise performance, and a study of antenna Q shows that only a limited increase in resistance can occur.**

I. INTRODUCTION

Over the years there has been a continued interest in obtaining wide-band performance from electrically small antennas. Recent efforts have been involved with receiving systems and have entailed designing preamplifiers that couple the antenna to a 50 Ω input impedance receiver. This communication discusses a factor which can limit receiving system noise performance when electrically small antennas are used.

RECENT DESIGNS

A survey of recent literature [1]–[3] indicates that present design efforts involve the use of a field-effect transistor (FET) amplifier in conjunction with a monopole antenna, typically over the frequency range 10 KHz to 30 MHz. Monopole antennas are selected for their high efficiency when compared to loop antennas, their unbalanced characteristic, and their desirable mechanical properties. An analysis of common amplifier configurations [4] shows that with a highly capacitive source impedance, a common source FET amplifier provides the best noise figure and available power gain.

In one design [1], extensive use of feedback is employed to provide required performance. Another approach [2] uses several stages of FET amplifiers in cascade, while a third concept [3] employs a commercially available 50 Ω bipolar transistor amplifier as a second stage. It is interesting to note that all the approaches offer essentially the same sensitivity (noise figure). This suggests that something other than the circuit design is limiting performance.

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The author is with the Defense Systems Division, Honeywell, Bloomington, MN 55420.

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ANALYSIS

A review of the pertinent noise performance equations suggests a possible limiting factor. The noise figure of an antenna-preamplifier combination is

$$f = f_a - 1 + \frac{f_p}{\eta} + \frac{f_r - 1}{\eta g_p} \tag{1}$$

where

- f_a "antenna" noise figure
- f_p preamplifier noise figure
- η antenna efficiency
- f_r receiver noise figure
- g_p preamplifier available power gain.

f_a represents the effects of external noise (atmospheric, man made, or galactic) received by the antenna. For systems operating at frequencies below about 30 MHz, good noise performance requires an externally noise limited system, i.e.,

$$\frac{f_p}{\eta} + \frac{f_r - 1}{\eta g_p} \ll f_a - 1. \tag{2}$$

Because of the large variability of noise levels, some care must be used in choosing a value for f_a [5]. The use of median values can imply that a system satisfying (2) will not be externally noise limited for significant periods of time.

The noise performance of the preamplifier can be found using the chain parameter equivalent circuit and its associated noise sources [6], [7]. The impedance Z_a of an electrically small monopole is $Z_a = R_a - j(1/\omega C_a)$, but $1/\omega C_a \gg R_a$ so that the noise figure and available power gain become

$$f_p \cong 1 + \frac{\langle e_n^2\rangle + (1/\omega C_a)^2 \langle i_n^2\rangle + 2\,\mathrm{Re}\,\{\langle e_n i_n^*\rangle(-j/\omega C_a)\}}{4kTBR_a} \tag{3}$$

$$g_p \cong \frac{R_a}{\mathrm{Re}\,(AB^*) + (1/\omega C_a)^2\,\mathrm{Re}\,(CD^*) + \mathrm{Re}\,\{(AD^* + BC^*)(j/\omega C_a)\}} \tag{4}$$

where $\langle e_n\rangle$ and $\langle i_n\rangle$ are equivalent noise sources, A, B, C, and D are chain parameters, and k, T, B have their usual meanings. Since $1/\omega C_a \gg 1$, some of the difficulty in obtaining good noise performance involves trying to reduce the contribution of terms containing the antenna reactance. Note also the strong dependence of f_p and g_p on antenna resistance.

When the FET noise sources are evaluated [3], [4], (3) becomes

$$f_p = 1 + \frac{\alpha_d}{g_m R_a}\left\{1 + \frac{G_L}{g_m \alpha_d} + \left(\frac{C_{IN}}{C_a}\right)^2 \cdot \left\{1 + \frac{G_L}{g_m \alpha_d} + \frac{\alpha_i}{\alpha_d} - \frac{2\alpha_c}{\alpha_d}\right\} + 2\frac{C_{IN}}{C_a}\left\{1 - \frac{\alpha_c}{\alpha_d} + \frac{G_L}{g_m \alpha_d}\right\} + \left(\frac{1}{\omega C_a}\right)^2 \left\{\frac{G_B g_m}{\alpha_d} + \frac{q I_g g_m}{2kT\alpha_d}\right\}\right\} \tag{5}$$

where C_{IN} is the FET input capacitance, g_m is the transconductance, G_B is the unbypassed gate bias conductance, G_L is the effective load conductance, and α_i, α_c, α_d are constants.

With the exception of the last two terms in (5), the frequency variation of $\langle i_n^2\rangle$ and $\langle e_n i_n^*\rangle$ cancels that of the antenna reactance. The last two terms in (5) are noise contributions from the gate bias resistors and shot noise caused by gate leakage current. The latter is negligible for metal-oxide-semiconductor field-effect transistors (MOSFET). In any case, the last two terms typically only affect the noise figure at frequencies below 100 KHz [3].

In a similar manner, common source available power gain for frequencies below about 100 MHz is [4]

$$g_p = \frac{g_m R_a}{\frac{g_0 + G_L}{g_m} + \frac{C_{IN}}{C_a}\left\{C_{gd} + \frac{(g_0 + G_L)}{g_m} C_{IN} + \frac{G_B}{g_m}(C_{gd} + C_d)\right\} + \frac{2\,C_{IN}}{C_a}\frac{g_0 + G_L}{g_m} + \frac{C_{gd}}{C_a}} \tag{6}$$

where g_0 and C_d are the drain conductance and capacitance and C_{gd} is the gate to drain capacitance. Again the frequency dependence of the antenna reactance is cancelled by the FET. It should be mentioned that for usual parameters, the available power gain is typically less than one; therefore, the noise performance of the following stage must be carefully considered.

When (5) and (6) are evaluated, good design practice usually results in extremely small values for G_B. The FET parameters are such that g_0 is negligible, and all the capacitances, C_{IN}, C_{gd}, C_d, and C_a are between $2pf$ to $10pf$. In wide-band designs, G_L is of the order of g_m. Also [3], $\alpha_i/\alpha_d \cong \alpha_c/\alpha_d \cong 0.18$. There are, therefore, no overwhelmingly dominant terms in (5) and (6) (except at very low frequencies). The performance is set by the denominator of (5) and the numerator of (6). Typically g_m ranges from five to 20 mS and if $g_m R_a \gg 1$ (for low noise figure and high gain), then R_a must be around $1K\ \Omega$ or more. This agrees with usual optimum source resistance values for FET amplifiers. For electrically small antennas, R_a will be less than 10 Ω, and even this value is only approached as the antenna length becomes about an eighth of a wavelength. From the above it is apparent that one major limiting factor in overall noise performance is the low antenna resistance. This statement is true for amplifier configurations which can deal with the antenna reactance, i.e., common source FET, and from (3) and (4), it also holds for less optimum amplifiers.

Low antenna resistance is a limiting factor for two reasons. First, it is doubtful that the FET (or any other active device) parameters could ever be improved enough to significantly overcome the problem. The FET capacitances would have to be reduced to values much less than a picofarad. Also the transconductance would have to become much greater than a Siemen. Second, and most important, significantly increasing antenna resistance is difficult. It is well known that the antenna has a minimum Q which is directly related to its electrical size [8]. Standard expressions for (electrically small) antenna resistance and capacitance show that the monopole Q is typically 10 to 20 times higher than its minimum value. This implies that some increase in R_a is possible.

The results in [8] show that minimum Q occurs with a uniform current distribution. This increases antenna resistance by a factor of four. Only when

$$\frac{f_p}{\eta} + \frac{f_r - 1}{\eta g_p} \cong f_a - 1 \quad (7)$$

does the increase offer any performance improvement. Over a large portion of the frequency range, (7) does not hold, and since the antenna resistance is so small, a four-fold increase produces no significant change in performance. Antenna resistance thus limits system performance.

CONCLUSION

The continued interest in obtaining wide-band performance from electrically small antennas has produced a number of preamplifier design approaches, all offering about the same performance. An investigation of the noise figure and available power gain for typical amplifier configurations shows a strong dependence on antenna resistance. Much larger values than normally occur with small antennas are needed, but analysis of antenna Q suggests that only a fourfold increase is possible. Over much of the frequency range, the resistance is so low that the increase offers no significant performance improvement. The low antenna resistance thus becomes a performance limiting factor.

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Comments on "Polarization Dependence in Electromagnetic Inverse Problems"

G. C. MCCORMICK, FELLOW, IEEE, AND A. HENDRY, MEMBER, IEEE

In the above paper[1] the concept of "optimal polarizations" in scattering theory is introduced. Two of the optimal polarizations (CO-POL) are those which, through the appropriate unitary transformation, separately zero the diagonal elements of the related scattering matrix; the other two (X-POL) are those which diagonalize the matrix. There is some reference in this and a subsequent paper [1] to meteorological applications. A more recent presentation [2] deals specifically with radar meteorology.

We are frankly puzzled by some of the statements in these papers and by claims with respect to optimal polarizations that we believe to be excessive. The importance attached to the theory of optimal polarizations may be judged from the following quotations.

"... recently considerable efforts have been expanded on using polarization in radar meteorology, though it must be stated here that decisive improvements can be made by fully exploiting the optimal polarization concepts introduced in Section III."[1]

"... optimal polarization properties of a scatterer in isolation and/or distribution are of paramount importance for the complete description of radar targets ..." [2].

On the subject of measuring the elements of the scattering matrix there is the warning that "... it is absolutely necessary to measure the relative phase between the co-polarized components in addition to the relative phase between the co/cross-polarized components ..." [2]. We made no such measurements in our determinations of the scattering matrices of spheroids and finite cylinders [3], [4] and we believe that they are not needed for reciprocal bodies.

A detailed analysis of optimal polarizations in terms of our formulation of polarization theory [5] will be the subject of a separate paper. We shall limit ourselves here to a few general comments. In order to obtain all of the backscatter properties of an assembly of scatterers it is necessary to make a complete measurement of the polarization and amplitude characteristics of the backscattered radiation. This can, for example, be done by means of Stoke's parameters or the coherency matrix. Such measurements require two separate sequential orthogonal transmissions, the polarizations of which determine the base vectors for the measurement. Part of the information contained in this system of measurements can also be obtained from a knowledge of two optimal polarizations, say one CO-POL and one X-POL. Optimal polarizations do not in themselves constitute a complete backscatter description but they can be made complete by a measurement of the cross section and the degree of polarization. Alternatively, the measurement can be completed by obtaining the power in the CO-POL channel and in its orthogonal.

The latter alternative suggests the possibility of a new complete measurement system in which optimal polarizations and the related channel levels are obtained experimentally. However any polarization scanning technique suffers in comparison with a polarization switch from the disadvantage that the data rate is determined by the stochastic nature of the target which may well have undergone macroscopic changes before the scanning cycle is half complete.

There is then the question of whether optimal polarizations provide a superior representation revealing insights not otherwise obtainable. This may be true for some scattering systems, for example, for a target body in the presence of clutter. We see no reason for expecting any such enhancement for meteorological targets per se. With regard to our own observations optimal polarizations could easily be calculated from existing data, but we see no advantage in doing so.

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G. C. McCormick is at R.R. 2, Granville Ferry, NS, Canada B0S 1K0.
A. Hendry is with the Division of Electrical Engineering, National Research Council of Canada, Ottawa, ON, Canada K1A 0R8.

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