Prof. Dr.- Ing. H. Lindenmeier

9 Active receiving antennas

9.1 Introduction

An antenna is that part of a receiving system which converts the space wave into a guided wave with a well-defined wave-
resistor "$_L$ transformed. An antenna can therefore be regarded as a quadripole which, in the case of the transistorized or active antenna, has an "integra-
In contrast to a passive antenna with a downstream antenna amplifier, there is no interface with the impedance of the characteristic impedance of an ordinary line in an active antenna. However, the formation of an optimal unit of passive antenna parts and amplifier elements, avoiding an interface with restrictive impedance requirements, makes it possible to create broadband high-sensitivity receiving systems with small antennas [9.9, 9.10].

With the installation of active components, the antenna receives an internal gain and provides a contribution to the electronic noise of the receiving system [9.2, 9.3] and thus cannot be unconstrained linear [9.9, 9.10] and reciprocal as is the case with passive antennas.

Thus, the active receiving antenna has a much greater influence on the system parameters of a receiving system than the passive antenna. For the last 10 years, a larger number of different short, active E mp-
antennas in operation [9.4, 9.5, 9.9, 9.10, 9.13, 9.15 - 9.18]. In the meantime, these have passed their acid test.

In the following, the theoretical basics for the development of this kind of antennas are shown. More details about realized designs of active antennas can be found in the given literature. An overview is given in [9.13].

9.2 The disturbed receiving system

The decisive parameter for the performance of a communications channel is the c anal capacitance C_C, which is a function of *PEPE,* where *P is* the power of the wanted signal and P_U is the power of any unwanted interference in that channel at the output of the receiving system. For a channel bandwide B_C gives siCh the channel capacity to Shannon:

$$C_C / \frac{\text{bit}}{\text{s}} < \frac{B_C}{\text{Hz}} \; 3.32 \lg (1 + PEPE]. \qquad (9.1)$$

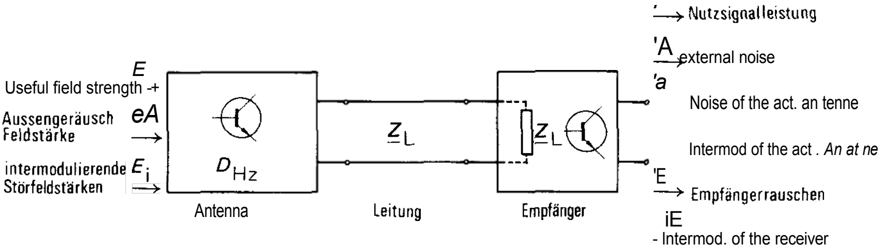


Figure 9.1: Receiving system with interference due to noise and nonlinear effects

In any receiving system, as shown in Figure 9.1, interference occurs that does not have the character of "white noise" but similarly reduces the capacity of the message channel.

Figure 9.1 lists the disturbances in question. Besides the received external noise power P_A, the el "ectronic noise of the active antenna P_a and the receiver noise P_E, the distortions *P,*
and P_E included, which are due to nonlinear effects in the active antenna and are caused in the receiver. Taking into account all disturbances receives one for the signal-to-noise ratio:

$$\frac{P}{P_U} \quad \frac{'é}{1 + \frac{a}{A} \left(t + \frac{E}{a} + \frac{}{a} + \frac{iE}{a} \right)} \qquad (9.2)$$

$$E^2_{\#2u} = \frac{E^2 / e_A}{1 + \frac{e^2}{eA}\left(1 + \frac{e_E}{e_a^2} + \frac{6_i^2}{e_a^2} + \frac{E_E^2}{e_a^2}\right)} \quad (9.3)$$

In equation (9.3), the useful signal is expressed by the signal field strength *E*. All interferences are described by equivalent fictitious field strengths of a fictitious plane wave incident from the direction of maximum reception at the receiving location. Interference of a statistical character is described by the root mean square value of its field strength values, while an intermodulation interference has the character of an unwanted carrier and is therefore described by the rms value of the field strength. This°. fictitious interfering field strength *E,* wi ll be generated by the interfering carriers as a result of non-linear processes [9.9]. From equations (9.2) and (9.3), indications can be obtained with respect to optimization of the channel capacity can be achieved.

9.3 Small antennas without directivity

9.3.1 Noise matching for active antennas

Assuming that an active antenna has a sufficiently high electrical gain, the noise contribution of the receiver becomes negligible. The signal-to-noise ratio *!SINj* is in this case:

$$SVN - \frac{E^2 / eA^2}{1 + e_a^2 /\!/'7A} \quad \frac{6^2 \; Ae}{k \, Tg_{BC}} \quad \frac{1}{1 + 7s} \quad (9.4)$$

In (9.4) $_{MA}$ and $_{AN}$ describe $_{the}$ noise temperatures of the external noise and the electronic noise of the active antenna, which are assigned to the radiation resistance of the passive part of the antenna. Zt120 x U is the field impedance of the free space, $_{Ae}$f the effective area of the passive part of the antenna and *k* the Boltzmann constant.

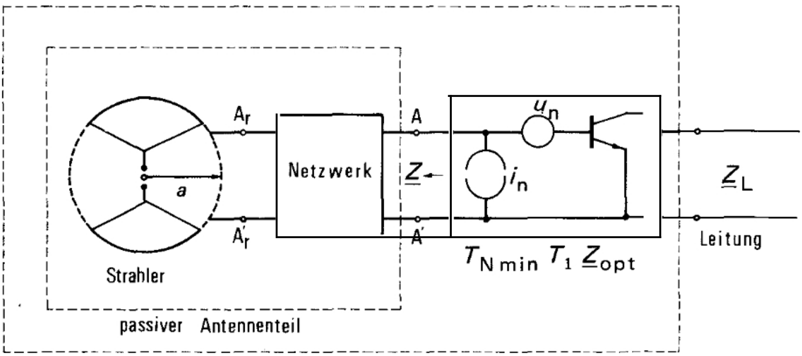
Obviously, even in the theoretical case of an ideal receiving system with $_{Pa}$--PE = *P,* -- PiE 0 the maxi mum ratio *PHP*]*cannot be* greater than *I E*² / *Ț I* $_{Ae}$ I($_{fi0}$ *kBCI* become. The latter is exclusively a func tion of the Ratio of the useful field strength *E* to the equivalent interference field strengtheA , which exists due to the external noise at the location of reception, because for short antenna is constant. For a real antenna, a maximum signal-to-noise ratio is obtained if the impedance ?' of the passive part of the antenna in

Figure 9.2 corresponds to the noise matching impedance ZOpp of the active circuit [9.4]. The signal-to-noise ratio is maximum when the passive part of the antenna and its passive feed are lossless and when, according to Gl. (9.4) $_{AN}$ = $_{AN\,iPi}$
is selected. Locations of constant R ausch temperature $_{AN}$ (Z') lie in the impedance-level on circles. This also applies to the system noise temperature
7 = $_{MA}$ + $_{AN}$ and according to Gl. (9.4) also for *SIN* (?'). If we denote by y the efficiency of the passive part of the antenna with respect to the terminals A - A'
in Fig. 9.2 and taking into account Gl. (9.5), an equation in which the normalized I mpedance deviation from the noise matching impedance ' $_{pt}$ is represented
\ *R - - r e a l* part *o f* 2"), we obtain relation (9.6) for the corresponding system-temperature [9.8]:

' lü) l6-6$_{O}$ t l^{2} /(/t /?$_{yt}$), (9.5)

$$\frac{T_{s\eta}}{T_A + T_{N\,min}} - 1 + \frac{1}{\eta}\;\frac{AN\,min}{T_A}\left[i1 - '' \quad \frac{1 + 7o\,ANm\,i\,n}{1 + ANm\,in \quad A} + \frac{T_1/T_{N\,min}}{1 + T_{N\,min}/T_A} \cdot \chi(\underline{Z}) \right] \quad (9.6)$$

7_0 : Ambient temperature.



active reception s antenna

Figure 9.2: Active receiving antenna

9.3.2 Signal to noise ratio bandwidth and power bandwidth

Since the effective area of an electrically short, ideal, lossless radiator is independent of the antenna dimensions, the achievable efficiency and the required bandwidth of the real, electrically small antenna are the decisive parameters. and the required bandwidth are the decisive parameters. The known bandwidth limitation b_o' (2 x a/?.$_0$) of a radiator was given by Chu, where a represents the largest dimension of the radiator in Fig. 9.2. Even under the assumption of an arbitrarily complex antenna str ucture, the limit for the achievable relative bandwidth 2 z /?, /I n 2 ermi ttelt, within which *PEPE* $_a$ m> 1/2 is.

In a receiving system, however, the signal-to-noise ratio is the only decisive variable instead of the power. Consequently, the bandwidth of the signal-to-noise ratio is decisive here and not the power bandwidth.

9.3.2 I Passive antenna part without losses

Figure 9.3 shows the complex impedance plane with the impedance z ?' of the passive part of the antenna.

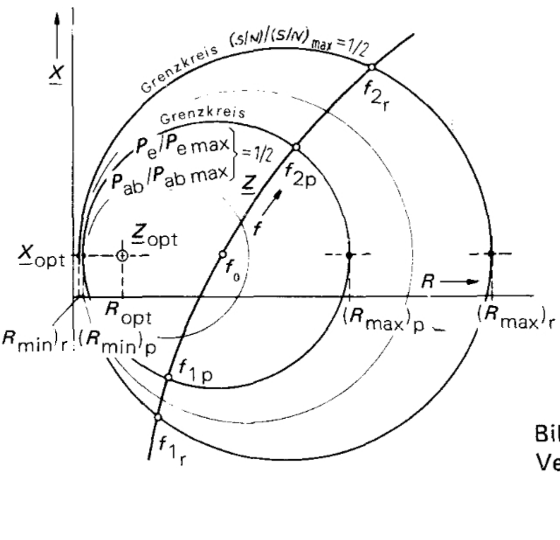


Bild 9.3:
Verlauf der Antennenimpedanz

A maximum of the received power $_{PC}$ $_a$ or a maximum of the out-power $_{Paba\ is}$ obtained for2 ' 2 p . Here 2 $_{pt}$ is the k onjugated complex value to the load impedance or the generator impedance. With Z (Z') from Gl. (9.5) results for the normalized power:

' ee ma x ' 'a b "a b m a x " $\dfrac{1}{1 + \chi(\underline{Z})/\vartheta_p}$, (9.7)

where t'p/ ' is 4.

If the passive antenna section in Fig. 9.2 is connected to the input impedance of a transistor, the noise characteristics of which are determined by its minimum noise temperature AN , which can be achieved during noise matching, and by its If a characteristic temperature *T,* can bedescribed, an active receiving antenna is obtained.

With the help of Gl. (9.4), the normalized signal-to-noise ratiocan be given in a form similar to Gl. (9.7) similar form:

$$\frac{S/N}{(S/N)_{max}} = \frac{1}{1 + \chi(\underline{Z})\vartheta_n} ,$$

wobein ' (1+ MA / AN m i n!'t T1 AN m in 'S!

The power bandwidth or the bandwidth of the signal-to-noise ratio are determined by the frequency range for which Z' in Fig. 9.3 is defined within the boundary circle for *P/P* a, = 1/2 and (*SIN* V(*SIN* \ a = 1/2. The normalized diameter of the boundary circle ö is obtained by Gl. (9.9) .

'n, p $\dfrac{max \; - R_{min}}{R} = \delta_{n,p} \cdot \sqrt{1 + 4/\delta_{n,p}}$; $\delta_p = 4 \cdot \sqrt{2}$. (9.9)

Since TA/TN i n increases rapidly with decreasing frequency, ögi changes within the order of 30 at 100 MHz to about 10*6 at 10 k Hz, andis much larger than dy = 4 .Therefore, the bandwidth of the signal-to-noise ratio is much larger for an electrically short radiator of a given height than the power bandwidth. On the other hand, this means that the dimensions of a receiving antenna can be selected several orders of magnitude smaller than the corresponding transmitting antenna.

With a short antenna, in the noise matching at center frequency \ *R* -- / ppt) is achieved by means of a single-circuit resonant circuit, the associated signal -noise bandwidth is obtained according to Gl. (9.10).

$$b_n/b_{ro} = \sqrt{2\,(\sqrt{(1 + (\delta_n/2)^2} - 1)} \qquad (9.10)$$

This results in the ratio ön//?,O i n for large values of d. However, a maximum of the signal-to-noise ratio bandwidth can be obtained with the following resonance impedances *Rp* and USA, if a circuit with parallel resonance character or series resonance character is assumed.

("P o' , opt) b m ax ' 1/(so ' opt b m ax' 1 + (/2) .

(9.11)

In this case, the relative signal-to-noise bandwidth is given by

(bn'm ax lbro n 2

(9.12)

Assuming an arbitrarily complicated network with an infinitely large number of lossless blind elements, Gl. (9.13) can be derived:

$$\frac{b_{n\infty}}{b_{ro}} \leqslant \frac{\pi}{\ln \frac{ö / 2}{/ 2 \quad + 1 - 1}} \tag{9.13}$$

For large values of $_{dn,}$ this gives /?$_{F\backslash}$//7,$_{O}$ " .v d $_{Pl/2}$. Thus, the relative bandwidth will never be greater than -multiple the maxi mum signal-to-noise ratio.
bandwidth with a single resonant circuit.

9.3.2.2 Verfuste within the passive part of the an.tenne

In practice, the passive part of an electrically short antenna can never be considered lossless. With Q as the quality of the matching circuit, one obtains as efficiency:

$$q = \frac{b_{ro} \cdot Q}{1 + b_0} \leqslant \frac{(2\pi a/\lambda_0)^3 \cdot Q}{01 + (2 \times a/2\,)\,0} \tag{9.14}$$

Antenna losses, which are included in the efficiency, reduce the size of the limiting circuit, but also increase the impedance bandwidth of the passive part of the antenna. If ü Piy is calculated from Gl. (9.15):

$$ny \quad \frac{T_A \cdot}{T_1} \quad [1 + \frac{2\, T_{N\,min} + T_0}{T_A} - \frac{T_{N\,min} + T_0}{\eta \cdot T_A}] \tag{9.15}$$

i n Gl. (9.9) is inserted, the normalized diameter dig of the I mpedance 2y of a lossy antenna in the impedance plane is obtained. To
Determination of the associated signal-to-noise bandwidth *big* must be replaced by *b,* y and $_{ön}$ by ö ny. Normally, the characteristic temperature *Tg* is in the range 0.25 < *T,* / $_{AN\ in}$ < 0.5. Therefore, for low
frequencies *Tfifi-/ T* << 1. Since, for feasible antennas, the quality Q in the range

50 < 0 < 500, g1 is 1 in the case of low antenna height. If we assume that y $_{Tfi}$ / Tg1, the equations for *R,* ö and the signal-to-noise bandwidth can be simplified as shown in the table in section 9.5.
is given. It is noteworthy that under the assumptions made above, the maximum achievable signal-to-noise bandwidth is not reduced by antenna losses if the antenna impedance is selected as indicated in the table. In case of transmission, however, a poor efficiency results in a strong reduction of the power bandwidth. If one tolerates a reduction of the radiated power by 3 dB compared to the achievable transmitter power, one obtains the following for the normalized circuit aver- age
knife:

py 2 4 1) 1 . (9.16)

If Gl. (9.16) instead of ö $_{F\backslash}$ in Gl. (9.13), we obtain the maximum achievable power bandwidth *biß /b y* for an arbitrarily complicated matching circuit.

9.3.3 Required minimum height of narrowband antennas

In practice, the required signal-to-noise bandwidth, the achievable
Q and the outside noise temperature $_{MA}$ are known at the receiving location.
Thus, according to Gin. (9.14), (9.15), (9.9) and (9.10) the minimum required
The minimum height hf n of a rod radiator was investigated in [9.12] for different external noise temperatures. The minimum height *hf* $_{n}$ of a rod radiator was investigated in [9.12] for different external noise temperatures.

Figure 9.4 shows that the required height of a receiving antenna is not exceeded for any frequency exceeds the value of 80 cm, if mean values of $_{MA}$ are taken as a basis.

At low frequencies, their height can be much lower than that which would be required for transmitting antennas. A capacitance C connected in parallel to the imaginary junctions *A* and *A,* (Fig. 9.2) of a capacitive radiator *C*, must be considered as part of the antenna. The effective bandwidth is reduced by the parallel connection to

b ro' $^{/?}$ro 1 + *C/C* \ (9.17)

is reduced. This effect reduces both the efficiency and the achievable bandwidth of the signal-to-noise ratio. From this point of view, inductive tuning of a capacitive antenna is preferable to capacitive tuning. This applies accordingly to inductive antennas, which should be capacitively tuned.

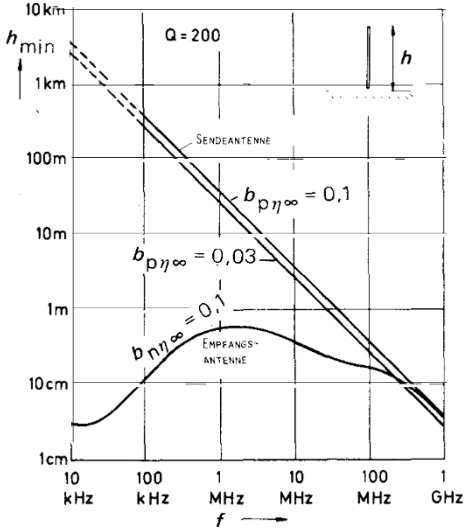


Figure 9.4:
Required minimum height of narrowband antennas

A similar effect of bandwidth reduction occurs when the antenna is loaded by a feed line (see Fig. 9.5). The influence of the line increases with increasing *l/h.* To compensate for this effect, an increase in antenna height would be necessary, as can be seen from the right scale of the diagram in Fig. 9.5. A minimum of antenna height therefore only results if the feed line to the antenna base is avoided.
will.

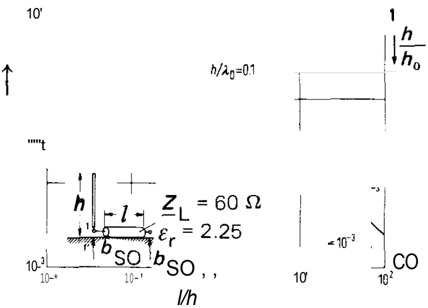


Figure 9.5:
Reduction of the radiator bandwidth, caused by the
antenna feed line

9.3.4 Antennas with bandpass characteristics [9.4 - 9.6, 9.14, 9.15]

With a suitable design of the matching circuit, the course of the antenna impedance Z' in Fig. 9.2 can be shaped in such a way that it forms a loop uFn2 it of the transistor in a desired frequency band; instead of such an antenna with a two-circuit filter, a multiple resonance antenna can also be used.
whose impedance curve contains several loops around Zapt. The gain in signal-to-noise bandwidth that can be achieved by an arbitrarily complicated struc- ture lies between the values from Gl. (9.12) and Gl. (9.13), taking into account § from Gl. (9.14) and ó" from Gln. (9.15) and (9.9). This principle of noise matching with bandpass characteristics is mainly applied at frequencies above 50 MH z where, due to the low values of A, the permissible impedance deviation from ZOpt k leiner. For a given frequency band to be covered, a fixed ratio of A /7N ifj of the active element and a certain quality of the matching elements, the minimum required antenna height can be determined. The optimum quality of the matching circuit and thus also the minimum possible antenna dimensions are obtained if the matching circuit is integrated with the antenna in such a way that spatially large and thus lossless dummy elements can be used.

9.3.5 Extremely broadband active receiving antennas [9.7, 9.9 - 9.11] It

is impossible to realize an electrically short antenna in such a way that its impedance Z approaches the noise matching impedance Zgpt of an amplifier in a wider frequency range. Even if an arbitrarily complicated matching network is used, the bandwidth of the signal-to-noise ratio remains limited to the value according to Gl. (9.13). However, the considerations in section9.3.1 show that this complicated dimensioning is not necessary.

It is only required that ż' within a certain frequency band in the, interior of the boundary circuit is *(S//N)/(S//V)* ã = 1/2 at the respective working fre- quence. The fact that the boundary circuits at low frequencies are very I suggests that the active element can be connected directly to a rod radiator without transforming its impedance by a complicated antenna structure or any network. In the case of the capacitive antenna (see 8ild 9.6), the optimum signal r oad bandwidth is achievable by selecting I a suitable active element with the appropriate operating point, so that the noise characteristics satisfy the I mpedance condition for an antenna of minimum size(A in Gł {9.8)). This I dee is supported by the fact that according to the last row of the middle column of the table in section 9.5, the required antenna impedance is *Rd* ' UOP " A/2 *T,* at low frequencies is much larger than *R* p *.*

Figure 9.6 shows the equivalent circuit of a short active rod antenna with antenna capacitance C'_A and directly connected FET amplifier. C represents the unavoidable input capacitance of the amplifier, which should be as small as possible.
The length of the bar required should be proportional to the factor (1 + Ca/CA). Based on this principle, broadband antennas have been developed which cover the frequency range from 10 kHz to 100 MHz. The equiva- lent noise field strength, the cause of which is the noise voltage $_{oa}$ of the amplifier

is obtained from $\sqrt{e_a^2} = *\sqrt{\;}^2$ *a (1 " C /Q A!' eff Whereby

heff is the effective height of the passive antenna part.

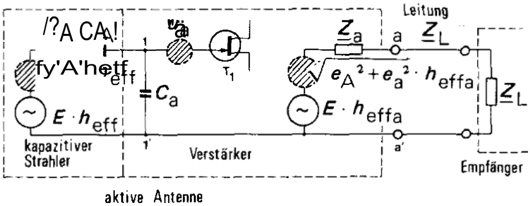


Figure 9.6: Equivalent circuit diagram for a short, active rod antenna

9.3.5 I Minimum size of the antenna

If Ua2 is expressed by an equivalent noise resistance /?e, then for hfine active rod antenna of height *h:*

$$\frac{h_{min}}{\lambda_0} \approx \frac{h_i}{\lambda_0} \cdot \left[1 + \frac{1}{2} \cdot \left(\sqrt{1 + \frac{4\,C_a}{c\,h'_i}} - 1\right)\right] \tag{9.18}$$

Where c = CA/li is the capacitance of the antenna per unit length and *h,* denotes the height of the antenna that would be required for Ca = 0.

$$\frac{h_i}{\lambda_0} = \sqrt{\frac{T_0}{T_A}} \cdot \sqrt{\frac{Re}{Z_0}} \cdot \sqrt{\frac{3}{\pi}} \;;\quad Z_0 = 120\pi\,\Omega\,. \tag{9.19}$$

This dependency of *hb* on Tg/7A is fundamentally different from the dependency of the minimum required height of a short passive rod antenna.
ne, the base of which is connected directly to the inner conductor of a coaxial line. and which is a receiver with noise figure *F* Here *hf* jn changes with $\sqrt[4]{T_0/T_A}$:

feeds.

$$\frac{h_{min}}{\lambda_0} = 0{,}43 \cdot \sqrt[4]{T_{MA} \frac{(F_r - 1)}{(Z'L/50\ \Omega) \cdot (c/10\ pF/m)^2}} \qquad (9.20)$$

Here, $2'_L$ denotes the world resistance of the line and the input resistance of the receiver.
A comparison of the solid curves in Fig. 9.7 shows that the minimum required antenna height of the active antenna is much lower than that of a passive antenna. Although the active antenna is broadband, the required height does not differ significantly from the curve in Fig. 9.4. The curve shown there applies to narrowband receiving antennas tuned to the receive frequency.

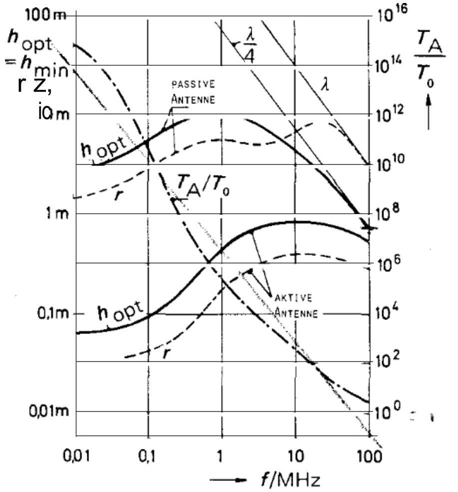


Bild 9.7:
Optimale Antennenhöhe einer aktiven und einer passiven Breitbandantenne

9.3.5.2 Optimum positioning of the antenna interferer [9.7, 9.11, 9. J 9] .
inside a rod antenna

It is known that for a rod antenna with a given total height *h,* the effective height h_{ff} does not assume the maximum possible value if the feed scfilitz is located at the base of the rod. At this point, however, the maximum antenna capacitance C_A is measured. Looking at the source voltage in connection with the voltage division between the passive antenna part and the amplifier input in Fig. 9.6, it becomes clear that there is an opti male height /7MOp
for optimal signal-to-noise bandwidth at the active antenna output.
Approximately applies to RMop .

$$\frac{h_{M\,opt}}{h_t} = \left(1 + \frac{C_a}{C' \cdot h_t}\right) - \sqrt{\left(1 + \frac{C_a}{C' \cdot h_t}\right)^2 - 1} \qquad (9.21)$$

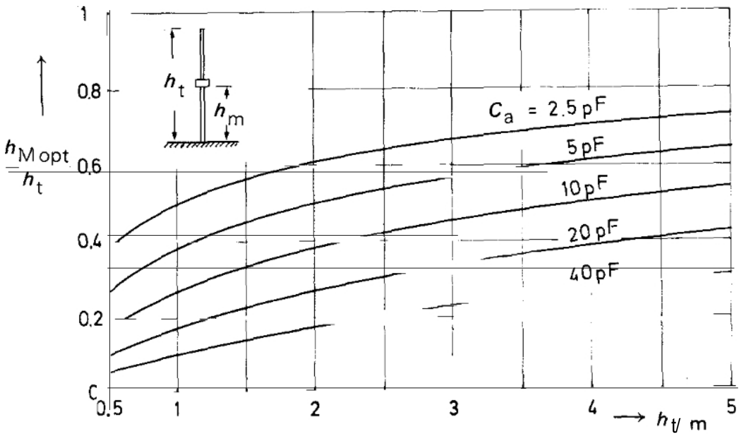


Bild 9.8: Optimale Höhe $h_{M\,opt}$ für die Anbringung des Antennenverstärkers

9.3.6 Nonlinear effects in active antenna elements [9.9 - 9.11, 9.13].

Non-linear effects can be avoided either by selectivity or by a high negative feedback factor in the amplifier. The negative feedback must be designed in such a way that the sensitivity of the active antenna is not degraded. In the immediate vicinity of a transmitting station, electric field strengths of up to 100 V/m can occur. To ensure undisturbed operation in such cases, the active antenna must be shielded.
can be tuned. In [9.21J the dimensioning guidelines of a hbchselective active antenna are shown. If such an antenna is tuned by premagnetization of ferrite elements, the linearity of the antenna is limited by diP linearity of these ferrite elements. In general, however, an active antenna should be broadband. In this case, a broadband negative feedback is required in the amplifier.

9.3.6 I Sensitivity of peg-coupled amplifiers

In the circuit in Fig. 9.9, the negative feedback resistor /?$_ä$ + Z'$_L$ contributes to the noise of the transistor T,' whose forward slope is *gg.* The equiva- lent noise field strength $_{ea\ of}$ the active antenna is:

$$\sqrt{\frac{ea}{\beta c}} \quad \frac{1}{bright} \cdot (1 + C_a/C_A) \cdot \sqrt{4\,k\,T_0/g_m} \cdot \sqrt{1 + \frac{G,}{\beta_2 \cdot \beta_{34}}}, \qquad (9.22)$$

where the last factor describes the influence of the ge9c coupling resistance on the noise. With increasing negative coupling factor $_{Gr}$' L/\ \ / $L/\_2$ the ratio of the voltages L/t_1 between the terminals 1 - 1 ' and the control voltage 2 at the transistor *Tg* increases. The influence of / is, however, reduced by a factor of +2 " h34, whereand Jhg represent the current gain factors of transistors 7s and 73 , *T4* , respectively.

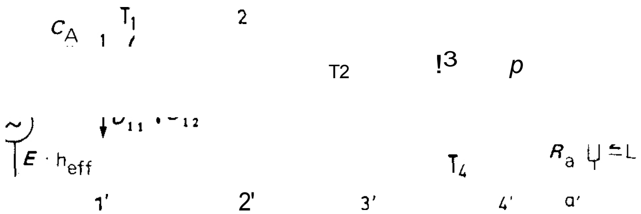


Figure 9.9: Equivalent circuit diagram of a highly linear antenna amplifier with negative feedback

In the following, the small-signal linear ity characteristics of the active antenna are investigated. The intermodulation distances ay_3 i n dB are considered with respect to 2nd and 3rd order effects caused by the two intermodu- lating field strengths - $_1$ = $_{6f2}$' *E;* in Fig. 9.10. There are no theoretical limit of the available $_{gr}$ and the achievable a_2 , this is completely determined by the circuit and the available semiconductors. Another very important parameter of an active antenna is its dynamic range, which indicates the distance between the noise level on the one hand and the maximum tolerable intermodulating level on the other. The maximum tolerable intermodulating field strength is the value $E,_{t2}$ = E/ $_1$ ' *E q,* whose 2nd order intermodulation product is equal to the noise level per t of the active antenna. A similar value in terms of intermodulation products 3rd order on the frequencies 2 f, * respectively f, + 2 is $E,_{t3} = \sqrt{Ep_1 \text{ " } E_{fZ}}$ bZW. Ej_{t3} -- $\sqrt{E' 1- Eff}$. Thus the dynamics of the nth Order def ined as follows:

$$d = 20 \cdot \log \frac{E_{itn}}{\sqrt{\overline{e_a^2}}} \qquad (9.23)$$

The dynamics are entered in Figure 9.10.

In many practical cases, the receiving system is more vulnerable to large-signal nonlinearity, such as broadband cross-modulation due to a nearby transmitting antenna. In this case, the tolerable rms value $E_{,C}$
of an unwanted amplitude-modulated signal is considered, which as a result of cross modulation causes a 20 dB reduction in the modulation depth on the useful carrier.

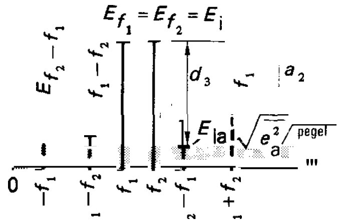


Figure 9.10: Intermodulation products due to unwanted signals

9.3.7 Optimal height of a broadband antenna

For a given antenna amplifier, the minimum required height $hf;_n$ according to section 9.3.3 is identical to the optimum antenna height h p , if nonlinear effects are taken into account. This can be calculated with the help of Figure 9.11, where the power contributions $P,$ $_{PA}$ and $_{Pa}$ of GI. (9.2) are plotted for a given field strength situation versus the height h of an active rod antenna. In addition, the intermodulation perturbations $P_{;1}$, $_{Pi}2$ and
Pi_3 for different values of the intermodulating signals $E;_1 > Ed_2 > E_{i\,3}$
are shown. The dash-dotted curves each describe the sum of the
The distance from the useful signal level P represents the current signal-to-noise ratio that is set for the corresponding value £;. This signal-to-noise ratio is represented by the curves in the lower part of the image.
h iii describes the point for $_{Pa}$ --PA. If the antenna height is chosen to be greater than h ¡pl, P/PjJ can be greatly reduced by nonlinear effects. Thus, $hf;_{fi}$ represents the height for which maximum PHP
is reached at maximum values of the intermodulating field strength Ed, i.e.:
$hf; \;\; = h\,y$.

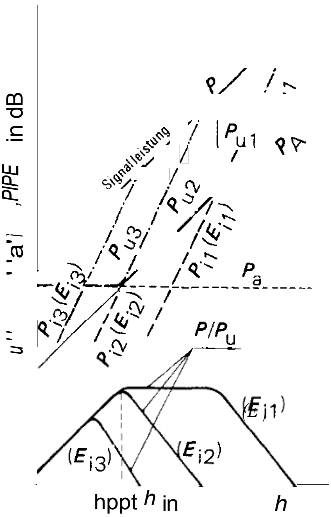


Figure 9.11:
Signal to interference ratio as a function of log Antenna height

If the antenna is mounted on a slim mast as shown in Fig. 9.12, the equivalent noise field strength and the maximum tolerable intermodulating field strength are reduced by a factor of *b* ",'n dB due to the height effect. This means that the sensitivity of the receiving system is increased by this factor, but the linearity of the antenna is reduced by the same factor. According to GI. (9.23), the dynamic range of the antenna is influenced by this factor.

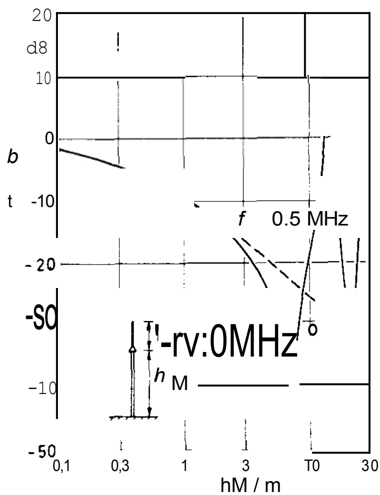


Figure 9.12:
Reduction of interference field strengths depending on the height of the mast

It goes without saying that the mast effect must be taken into account when selecting the optimum antenna height. Sol ange h_M
is short compared to the wavelength, the factor *b* is independent of the frequence. To compensate for the height effect, the antenna rod can be shortened accordingly. Too high and very slim masts should be avoided in practice because of their narrowband resonances [9.7].

9.3.8 Increased linearity due to selective negative feedback [9.13, 9.17].

The negative feedback achieved in the amplifier in Fig. 9.9 with the aid of R_a + 2 "$_L$ can also be achieved in the form of resonance negative feedback.
the. For this purpose, the input of the receiver in Fig. 9.13 can be used very advantageously. Instead of the impedance /?$_a$ + $_L$ in Fig. 9.9, a lead is connected which is connected at its end to the series resonance circuit at the receiver. The input of the active antenna is terminated. In contrast to a passive antenna,the output of the active antenna has a low impedance between terminals a and a'.

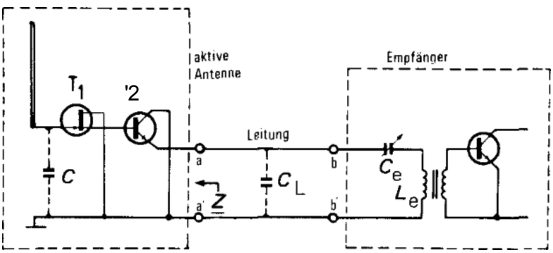


Figure 9.13:
Active antenna with broadband amplifier and selective feedback

Accordingly, the capacitance C_h of the line has no influence on the resonance frequency of the circuit as long as the line is short compared to the wavelength. The input impedance of the line at terminals a and a' is low impedance only in the vicinity of the resonant frequency of the resonant circuit and is high impedance at all other frequencies. Signals away from the resonant frequency should be kept away from the receiver. The selective negative feedback gives the amplifier extremely good cross modulation and intermodulation characteristics. This allows operation without nonlinear interference even in the immediate vicinity of powerful transmitters. If the circuit is suitably dimensioned, there is no loss of sensitivity and the considerations in section 9.3.5 apply unchanged. This active antenna solves the sensitivity problem in a broadband way in combination with all advantages regarding the linearity of a tuned input stage. It has
The particular advantage of this is that the tuning element is still housed in the receiver. This principle has so far been successfully applied in an active auto-

antenna. The input resonant circuit of the standard AM receiver part is used for selective negative feedback at the antenna output [9.13, 9.17 J.

In section 9.3.3 it was shown that capacitive tuning of a capacitive antenna worsens the sensitivity. For this reason, tuning of AM car radios must be done inductively, i.e. with technically inconvenient variometers. With the circuit in Fig. 9.13, the sensitivity and the selection problem are solved separately. Therefore, capacitive tuning can be introduced without any disadvantages [9.17].

9.4 Antenna groups [9.18 - 9.201

9.4.1 Reciprocal coupling [9.18, 9.19]

An important feature for the use of transistorized receive antennas in antenna arrays is the small mutual coupling between adjacent antenna elements. In symmetrical antennas, the coupling between neighboring antenna elements is limited to the coupling through the
backscattered field OR is limited. With monopoles above ground, a further coupling contribution is added because of the finite conductivity of the ground. In [9.18, 9.19], both coupling effects between active antennas and passive antennas, respectively, are compared for an Adcock DF antenna system.
The ratio of the required distances between two antenna elements for 10% back-radiated field strength for short active and short passive antennas is as follows:

$$\frac{r_{passive}}{r_{active}} \quad \frac{2}{3} \quad \sqrt[4]{\frac{T_A}{T_0}} \quad \sqrt[4]{\frac{F_r F_T\, 1}{L/50}} \quad \sqrt{\frac{Z_0}{e}} \quad \frac{1}{1} \quad \frac{1}{\sqrt{c/(10\ pF)}} \qquad (9.24)$$

with r from GI.
(9.18) .

Figure 9.7 plots the minimum required distance *r* for less than 10% back-radiated field intensity at the location of the neighboring antenna. The lower dashed curve applies to active antennas and the upper dashed curve to passive antennas.
rod monopoles if both have optimum height / $?_{pät}$. The results presented show the superiority of the active antenna in this respect.
The required distance between adjacent active antennas is all the more
k leaner the larger the boundary circle in the impedance plane, and is larger the larger the power mismatch achieved under these circumstances. Because of the small mutual coupling and the very broadband radiation patterns associated with the small optimum height, the