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HIDDEN VEHICULAR ANTENNA TECHNIQUES

FINAL REPORT

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TABLE OF CONTENTS

LIST OF ILLUSTRATIONS

List of Illustrations, continued

I

INTRODUCTION

The objective of this study has been to investigate hidden vehicular antenna techniques. The antenna is to operate in both the AM and FM frequency bands. It was desired for the hidden antenna to have electrical performance equivalent to a standard 22 inch fender-mounted whip antenna in the above frequency bands. During the contract period an investigation was made into the feasibility of using various parts of the car (doors, windows, trunk lid, etc) as the antenna. In addition to the above, measurements have been made on the existing (Thunderbird) windshield antennas. As a consequence of these measurements, consideration was given to optimizing the signal-to-noise ratio of the windshield antenna. This report will also discuss considerations that must be given to hidden antenna design concepts. We will also describe several hidden antenna types that have been considered and discuss their limitations.

II

MEASUREMENT TECHNIQUES

The function of an antenna is to efficiently transform RF energy from free space to a transmission line. For AM and FM receiver applications, one is primarily concerned with how much of the energy-density, in the surrounding medium of the antenna, can be delivered to the receiver terminals. Evaluation of the FM antenna performance is accomplished by comparing the performance of the hidden antenna with a resonant monopole. The free space wavelength for the FM band varies from 109 inches to 133 inches so that a resonant monopole with a well-known impedance and an absolute gain can be utilized for a convenient standard.

Antenna performance for the FM band is then expressed simply as the ratio of the power received by the antenna under test to the power received by a resonant antenna. The ratio of these two received powers is referred to as the gain of the test antenna with respect to the reference antenna. Taking ten times the log to the base 10 of the ratio gives the antenna gain in dB, i.e.

$$G = P_t / P_r \quad , \tag{2.1}$$

where

$G_{dB} = 10 \log P_t/P_r$,

P_t = power in test antenna ,

P_r = power in reference antenna.

The reference antenna can be, (1) an isotropic source (a fictional antenna that receives equal amounts of power from all directions, i.e. a point source), or (2) a dipole. Since the isotropic source is only a mathematical model, the physical standard utilized for the gain measurements in the 100 MHz region is the dipole, although the gain of the test antenna may be recorded with respect to the fictional isotropic antenna. For a lossless half wave dipole the gain over an isotropic source is 2.15dB.

To make antenna gain measurements by the comparison method, the antenna pattern of the test antenna is obtained with an antenna pattern recorder. The half wave dipole replaces the test antenna and without changing either the transmitted power or the antenna recorder setting, a pattern is obtained for the dipole. If the patterns are recorded in dB, the gain in dB is simply the difference between the peak power of the two antenna patterns. This type of gain measurement, called the substitution method, takes into account all losses including any mismatch loss between the antenna and the receiver impedance. In the 100 MHz region, the receiver impedance is typically 50 Ω which is very close to the medium of the impedance for a resonant dipole (72 + j0), and a resonant monopole (36 + j0).

FM antenna patterns of the antenna on the vehicle were obtained by rotating the entire car on a modified vehicle display rotator. For monopole antennas such as those found on automobiles, the ground plane (car shape) may have considerable effect on the FM band antenna pattern and performance. Therefore, it is desirable for the FM performance of the antennas to be checked on the automobile on which they are to be mounted. In the AM band the dimensions of the car are small compared to the wavelength so that the car has little effect on the AM antenna pattern (Williams 1950). Signals for these antenna measurements were provided by the Radiation Laboratory's log periodic transmitting antenna on a 56 foot vertical tower. The transmitting antenna is variable in height permitting proper illumination of the antenna under test.

Unfortunately for the AM band with a wavelength of approximately 1000 feet, no convenient standard exists. A two foot whip is approximately 0.002 λ long so that the magnitude of the signal received by the antenna is quite reduced from a resonant element. Obviously for the AM frequency band a different method of performance evaluation is required. One could compare the power received by an antenna under test to that received by a whip of a specified length. This approach creates problems since the reactance of the antenna is very large compared with the resistance, and the receiver needs to conjugate match for maximum power transfer between the antenna and receiver. Conjugate matching is not possible in the AM frequency band as shown by the formulas for antenna impedance. The following formulas for resistance and reactance agree very well with theory (King, 1956).

$$\left.\begin{aligned} R_o &= 18.3\,\beta_o^2 h^2 (1 + 0.086\,\beta_o^2 h^2) \\ X_o &= -\frac{396.0}{\beta_o h} (1 - 0.383\,\beta_o^2 h^2) \end{aligned}\right\} \beta_o h \leq 1;\ \Omega = 10 \qquad (2.2)$$

$$\beta_o = \frac{2\pi}{\lambda} \quad .$$

The Ω term accounts for the height-thickness ratio

$$\Omega = 2 \ln (2h/a) \ . \qquad (2.3)$$

This gives $\Omega = 10.5$ for a two foot whip with a diameter of 1/4 inch which is a good approximation for the fender-mounted whip antenna. The above formulas give an impedance $z = 1.5 \times 10^{-3} - j\,1.6 \times 10^{+4}$ ohms for a monopole at 1 MHz. We see the real part of the resistance is minute so that even low loss reactive elements have a resistive component large in comparison. Since the antenna base, coaxial feed line and connector to the radio all present a shunt capacitance, this capacitance combines to significantly change the impedance that is presented to the receiver terminals (King, 1956). Both the resistance and reactance terms are greatly affected by a small shunt capacitance as is shown by the formulas below.

$$R_s = \frac{R_o}{1-2\omega C_s X_o + \omega^2 C_s^2 (R_o^2 + X_o^2)} ,$$

$$X_s = \frac{X_o - \omega C_s (R_o^2 + X_o^2)}{1-2\omega C_s X_o + \omega^2 C_s^2 (R_o^2 + X_o^2)} . \tag{2.4}$$

A shunt capacitance of 100 pico farads reduces the reactance of a two foot monopole at 1 MHz from $-1.6 \times 10^4\ \Omega$ to $-1.5 \times 10^3\ \Omega$. The figure of 100 pico farads is not unreasonable since RF cable such as RG-62 has 13 pico farads per foot and the average cable run would be 4 to 6 feet of coaxial cable giving approximately 50 to 80 pico farads of capacitance without taking into account the connector and antenna base capacitance. The receiver must conjugate match the modified impedance presented to its terminals, i.e. $Z_s = 0.7 \times 10^{-4} - j\,1.5 \times 10^3$. Obviously, any low loss reactive device cannot match this even lower resistance term. If one uses an inductor with a reactance of 1.5 K Ω at 1 MHz and a Q of 300 the resistance would be 5 Ω, which is over 10,000 times the radiation resistance at the receiver terminals. Signal strength is lost in several ways; some of the signal is shunted to ground by the capacitance, the receiver cannot conjugate match with low loss reactive elements, and the antenna is not terminated in the conjugate antenna impedance. In general the distribution of current in receiving antennas depends greatly on the load and this effect is more pronounced in antennas not near the self resonant length (King, 1956). It would be desirable to eliminate those losses due to the cable and connector shunt capacitances. Since the shunt capacitances can significantly change both impedance and signal level, one must be careful in minimizing cable length and account for variations in shunt capacitance when comparing different antennas under test.

Comparison of antenna performance by using a standard such as a whip has another problem in the "conjugate match" in the receiver. Conjugate matching for a fender-mounted whip is achieved by adjusting the trim capacitor which is in shunt with the inductor on the automotive receiver. Antenna types considered during the test program gave orders of magnitude differences in AM band impedance levels. The function of a trim capacitor, a small variable capacitor, is to account for minor impedance level variations seen at the receiver due to the variations in cable and antenna capacitances that will occur in mass production and for the inductor variation in the radios. Comparing antennas with different impedances requires considerable adjusting of the receiver input impedance by adding and removing capacitors to obtain this "conjugate match" for maximum power transfer with each antenna. Changing the receiver input impedance may also change the loss terms of the passive network so that a true picture of the antenna performance is not obtained. Also, the automotive receiver has an impedance transformation due to the turns ratio on the tuning coil. Different impedance levels therefore do not transform to the optimum impedance for the RF stage of the receiver. To test the antennas a transformer should be made for each impedance level of each type of antenna and the loss of each transformer should be checked. Design of such a matching circuit should be considered as part of the antenna design when working with electrically small antennas. The use of a standard antenna as a comparison reference is time consuming and does not appear to be a desirable means of testing. One needs a better means of evaluating AM band antennas and one which is independent of the receiver. As shown above small changes in the cable capacitance or loss terms in the receiver can be the cause for significantly different measured antenna efficiencies.

Measuring open circuit signal-to-noise ratio and comparing this to that obtained for a whip antenna is not a satisfactory means of evaluation. Open circuit voltages were measured on a series of different windshield antenna configurations, and were found to be similar although antenna performance in terms of station count was quite different. When comparing similar antenna configurations, such as the windshield antenna, a station count is a reasonable means of antenna evaluation if one recognizes the restrictions discussed in section IV. Impedance of the various windshield antennas does not vary to the extent that the same trim capacitor cannot be used. Figures 2-1 and 2-2 show the AM band impedance variation for a top loaded windshield configuration 1 inch and 2 inches down from the top of the windshield with a length of top loading section as a parameter. Table II-1 gives a sample of open circuit voltages for a top loaded monopole 2 inches from the top of the windshield. The 1/2 L windshield gave a station count of 26, while 1/3L gave a station count of 20 compared to 41 stations for a 22 inch whip. It is important to note that the open circuit voltage of an untuned antenna is not a sensitive indicator of its performance when used with a receiver.

When the added complexity of an active antenna is considered the magnitude of the signal delivered to the receiver terminals is not adequate to evaluate antenna performance. If the amplifier does nothing but amplify the signal and the noise at the antenna terminals this will only increase the observed signal without increasing the listenability of a weak station.

The problem of dealing with an electrically short antenna and developing meaningful figure-of-merit has led some to talk of the antenna noise temperature. This is especially true when one is dealing with an integrated antenna on which there is gain in the antenna itself. The noise temperature problem applied to integrated antenna techniques to the AM band for vehicular antennas is discussed in section V. It appears there is no easy measurement of AM band

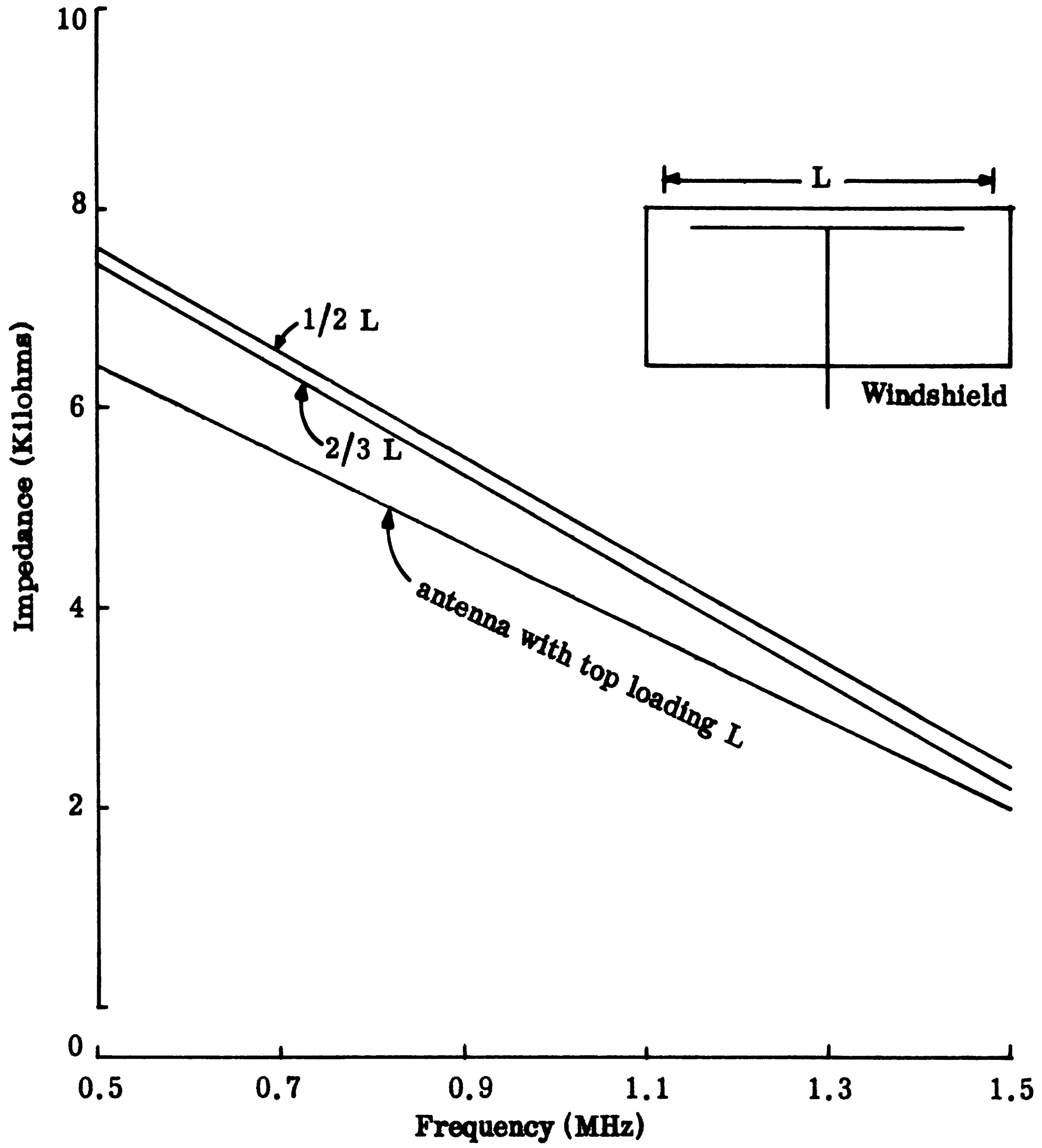


FIG. 2-1: **Impedance for Windshield With Top Wire 2 Inches From Top of Windshield. (Experimental)**

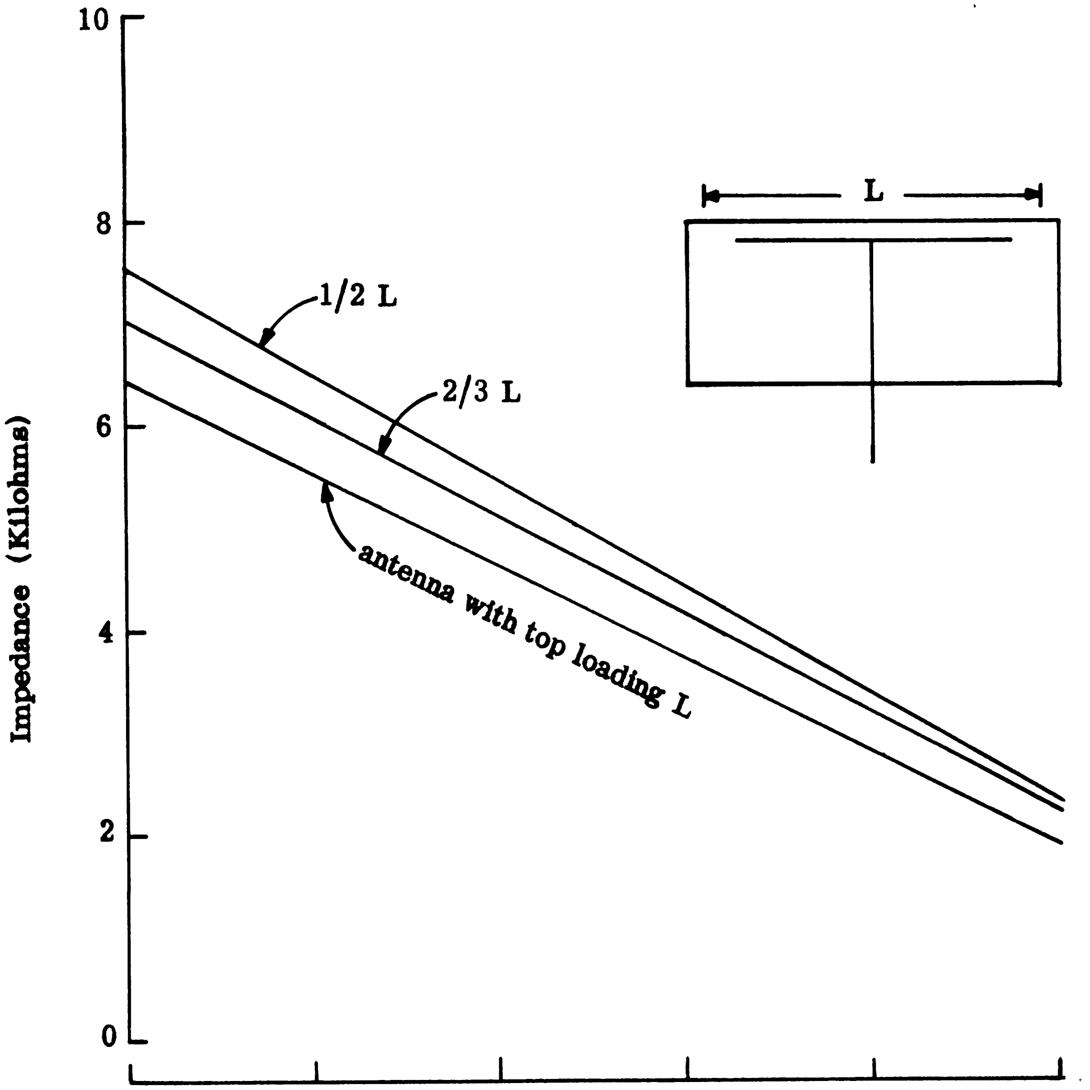


FIG. 2-2: Impedance for Windshield With Top Wire 1 Inch Down from Top of Windshield. (Experimental)

TABLE II-1

Station		Windshield Top 1/2 L	Windshield Top 1/3 L	Whip 22"
WJR	(760 Kc)	0.0060 V	0.0060 V	0.0060 V
WCAR	(1150 Kc)	0.0016 V	0.0017 V	0.0029 V
WSDS	(1480 Kc)	0.0016 V	0.0016 V	0.0115 V

linear antennas that would give a meaningful figure-of-merit that would realistically indicate a level of performance when combined with a properly designed receiver.

For the above reasons the whip antenna and the hidden versions of the antenna for the AM band could not effectively be measured on an antenna range as were the FM antennas for vehicular applications. One could transmit signals from a vertical wire of approximately 50 feet and turn the car on the turn-table, but the field distribution at the car would not be known as the car would not be in the far field and the resulting pattern would not be the true antenna pattern.

Another problem is that most of the hidden antennas investigated did not give signal response to commercial stations equal to the whip so the signals were too weak to be measured accurately. Initially we attempted to measure the antenna performance with an Empire Receiver model NF-205 employing the T-A/NF-205 tuning unit which covered the AM band. The model NF-205 receiver requires an AC power source for operation. For the purpose of these tests the NF-205 was placed in the car and a long power cable extended from the car to an AC outlet. This resulted in an inducted RF signal on the AC power line much greater than that induced on the hidden antennas under test. This is understandable since the line cords were in the neighborhood of 300 - 400 feet long which is close to the resonant length in the AM band. It is evident that for most antennas in the AM band region the measurement techniques used cannot be the automated measurement of an antenna pattern unless battery operated equipment is utilized. If a receiver is used, the receiver for such a measurement should have a high input impedance to power match the antenna, but after reviewing the above considerations it becomes evident that for these longer wavelengths, in the AM band, the antenna pattern does not present enough information. Due to the electrical size of the car, the antenna patterns would be nearly omni-directional. Therefore, additional information will be required from the measurement of the signal-to-noise ratio and possibly the noise temperatures of the system. At present it would appear that no reliable easily measured figure-of-merit exists for electrically short antennas.

III

DESIGN CRITERIA

The hidden antenna should perform as well on the AM band as a 22 inch whip and on the FM band as well as a $\lambda/4$ monopole which would approximate a 30 inch whip. Two approaches were taken in an effort to hide the antenna:

1) use of an existing part of the car for the antenna, and
2) hide a conventional antenna in the car.

Many problems are associated with using the first approach, i.e. utilizing an existing part of the car. The problems associated with this are created by periodic model changes. This would necessitate checking the antenna for each major model change to see if the antenna performance had been adversely affected. However, as a result of this investigation it was found that almost any part of the car will pick up a small amount of RF energy and could be considered as a form of antenna. Body contour shapes may greatly affect the FM performance of the antenna when one is utilizing an integral part of the car. Another problem that exists when using the integral part of the car is the large amount of noise generated by the electrical circuits and this noise is more noticeable if the antenna is an integral part of the car body than if it is on the exterior. The car body acts as a form of shield and reduces the amount of automotive noise that is received by an antenna mounted on the car exterior. Another type of noise affecting integral antennas is intermittent contact noise resulting from poor RF bonding of the various portions of the car.

Conventional automobile antennas are fairly well understood and it appears that it may be more desirable to incorporate them into the automobile structure than to attempt to design a new antenna. The windshield antenna is a type of conventional antenna being incorporated into an existing portion of the car.

IV

ANTENNA TYPES CONSIDERED

The antenna types considered were:

interior antennas,
trim antennas,
structural antennas, and
hidden conventional antennas.

Several types of car interior antennas have been investigated. However, it is to be noted that AM antennas inside of the car will not give comparable performance to an exterior antenna due to the RF shielding. Disregarding the aesthetic qualities of an automobile the car's interior forms roughly a rectangular box or cavity. This rectangular cavity is electrically small compared to the wavelength of the signal in the AM band and therefore does not permit RF energy to propagate. The windows are also electrically small so that the car's body forms a leaky Faraday shield. Because of these facts the signal strength inside the car body is significantly reduced compared to the external signal strength. To substantiate this statement the following observations have been made. If a whip antenna is slowly drawn into the car's interior through an open door or window, the signal strength observed on a micro voltmeter in the AGC circuit (of an automobile receiver) will decrease as the antenna passes through a vertical plane defined by the window or door opening. A second observation that may be made to demonstrate the reduced signal strength available inside a car is to slowly move a pocket transistor radio into the car's interior. From these experiments we see that neither a whip or ferrite loopstick antenna will give satisfactory performance inside the car. At the Willow Run experimental facilities only a few radio stations (approximately 12) are powerful enough to be received in the car's interior. If one's driving was limited to the urban area near the transmitting antennas (of lower power stations) the field strength would be of sufficient strength for interior antennas to perform satisfactory for local stations. However, the interior antenna would be unacceptable for more remote AM stations.

For the FM band the car's interior again forms a rectangular cavity. However, at the FM band the cavity of a full size car is close to resonance (116-120 MHz). The windows of the car form nearly resonant slots which efficiently couple energy from outside the car to the inside, and cause the interior to behave like a lossy resonant cavity (i.e. a low Q circuit). We have observed that resonant wires (approximately $\lambda/4$ long) attached to practically any part of the car, in the car's interior, will function satisfactorily as an FM antenna in the metropolitan area. Antennas of this type could be embedded in the crash pad on the dash, in the foam padding along the A pillar, or it could be attached to the steering wheel column. These antennas do not perform as well as the conventional whip antenna and the range (radio station to receiver) of such antennas is limited to 20 - 25 miles. Performance of the $\lambda/4$ wires is reduced by close proximity to metal which is a physical constraint inside the car.

Consideration has also been given to the use of the trim on the car's exterior. As the electric field approaches the car body boundary conditions require that the field approach zero at the car body. Therefore, the energy available for antenna excitation should be greatly reduced. Several configurations of chrome stripping have been tried for use as antennas. The top chrome strip on the Lincoln Continental Mark III front fender gave a station count of 20 - 22 stations which exceeded the expected signal level. Performance of such an antenna could possibly be improved by spending time on the proper impedance matching to the receiver. It is not known whether the chrome strip has the potential performance of a 22 inch whip.

Performance of the strip does improve as the chrome strip to fender spacing is increased. If the height of the chrome strip is increased to several inches above the ground plane (fender) the configuration approaches the over-the-car roof antenna popular during the late thirties. The writer does not know the performance level of that type of antenna, but it is believed to be below that of the whip. Only a limited time was spent on the chrome strip since the problems associated with such an antenna were felt to outweigh the desirable characteristics. It is believed that snow, ice and possibly even rain would considerably affect the antenna performance because of the close proximity to the ground plane (fender). If such an antenna were to be considered a thorough environmental testing would be required. Chrome stripping on a car normally collects road grime and salt. Such a buildup may possibly be reduced by using plastic chrome mounts that hold the chrome strip above the metal surface much like Chrysler Corporation's mounting of several years ago.

A trim antenna could take the form of the rear view mirror. Such an antenna is presently being marketed in Germany. A small piece of metal about the size of the rear view mirror which is being sold as an integrated antenna was investigated as a part of this study and was observed to pick up approximately 20 stations in the Willow Run area while a normal 22 inch whip will detect approximately 45-50 stations depending upon weather conditions. There are a large number of high powered stations in the Detroit Metropolitan area so that AM band performance is difficult to rate by station count. If there is not a sufficient gradient in field strength between stations in the station count, one can loose considerable performance before noticing a significant decrease in the station count. However, this procedure is still used for lack of a better means of evaluation.

Another form of hidden antennas employ structural members of the car, such as the trunk lid, car door or an isolated fender. Earlier research by Ford Motor Company indicated the trunk lid gave a larger effective area for AM than a 33 inch whip. However, it was rejected because of the reduced performance as a result of salt, snow, rain and slush forming around the seam of the trunk lid and car body. The above elements effectively shorted the antenna (trunk lid) to the main body thus making the antenna inoperable. During this investigation passenger doors were considered as a possible antenna. The passenger door of the Continental Mark III was isolated from the ground in an attempt to utilize it as an antenna. Its area is large and is similar to the trunk. We felt that it would not suffer from the undesirable affects of slush, snow and ice forming along the vertical seam between the car door and body. Figure 4-1 gives a comparison of the impedance of the trunk lid, and isolated passenger door of a Mark III for the AM frequency band. Impedance of both the trunk lid and the passenger door depend upon "antenna" body spacing and attachment location of feed cable as shown by the second curve for the trunk lid. The trunk lid and open passenger door displaced similar impedances. Closing the passenger door changed the impedance slightly with the resulting impedance being quite similar to that for the lower impedance trunk lid plot. Signal strength from the closed car door is reduced from that of the trunk although signal strength from the open car door was comparable. Both the door and trunk lid display a lower AM band impedance and a higher FM band impedance than would be found for the whip antenna. The impedances of both trunk lid and closed door would vary from 600 - 900 Ω at 0.5 MHz depending on the body spacing and feed position, but at the upper frequency limit of the AM band the impedance of both approached 270 Ω. The phase angle of the impedance across the AM band is approximately 90^o

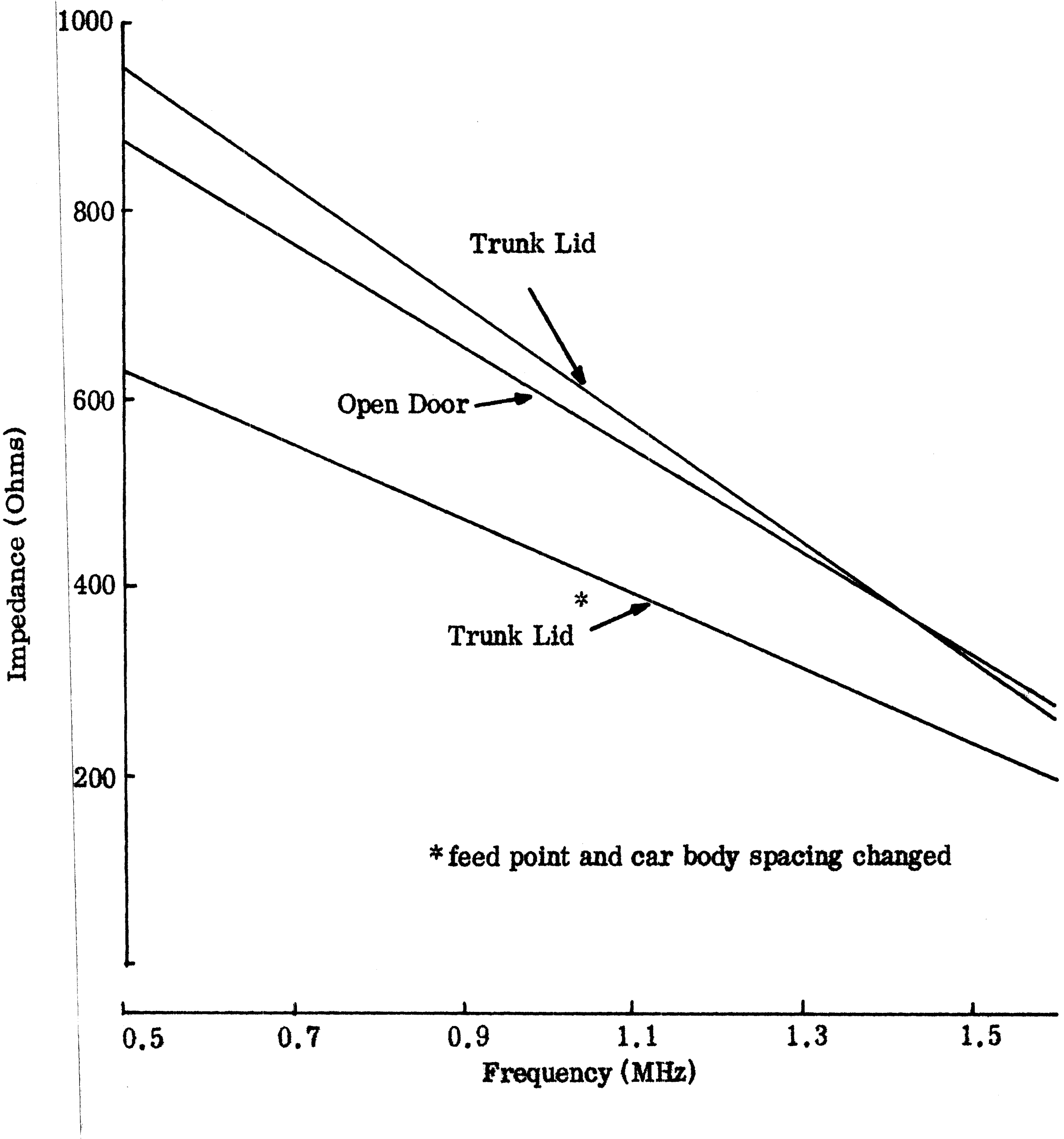


FIG. 4-1: Impedance of Isolated Door and Trunk on Continental Mark III.

capacitive. At Willow Run a station count for a signal level above 5 microvolts in the AGC circuit yields approximately 50 stations for the whip as compared to approximately 22 for the closed passenger door.

At the FM band both the trunk lid and passenger door have inductive impedances varying from approximately 300 Ω at 88 MHz to 100 Ω at 108 MHz. The 40 Ω FM receiver (used for these tests) utilizes a 35 pico farad capacitor to ground to give a 40 Ω input impedance which is capacitive. This capacitive impedance reacts with the inductive antenna impedance to cause a hole in the tuning band. FM performance of both the trunk lid and the isolated passenger door are below that of the $\lambda/4$ wavelength whip antenna. Impedance variation with spacing between the body, trunk, passenger door and signal level variations with lead attachment make it difficult to assign a level of FM performance for both the trunk lid and passenger door.

Utilizing the trunk lid as an antenna would present additional problems because of the year to year design changes associated with trunk lid configurations. As the trunk lid design changes so would the antenna performance, therefore, requiring re-checking the antenna. Similarly, isolating or using a part of a fender or any part of a car that could undergo considerable changes every two or three years would present serious problems to radio engineers in the design and evaluation of antenna performance.

The passenger door was considered as one of the more desirable integral antenna types because it is a component that does not change greatly in size or shape with the model changes. However, we do not feel that either the trunk or car door will make desirable antennas due to the effect on antenna performance by nearby objects. If one places their arm on or near the isolated car door the signal level is noticeably affected. The American car owner generally fills the trunk with luggage when going on a vacation and this would also change the antenna's characteristics probably detrimentally when the owner is most interested in the car radio's performance. Generally the car radio is in steady use and the owner can notice the station "pulling power" of the radio antenna combination as he travels large distances on todays turnpikes.

When one considers hiding a conventional antenna on a car they are severely limited by the car contour and the fact that almost all available space is being utilized at the present time. Much of what has been discussed above has been directed toward the upper portion of the car. It is known that in years past auto makers installed antennas under the car (wires under the running boards). Available information indicates that these were not satisfactory. However, we have investigated the effects of placing an antenna underneath the car. The test antenna was basically a metallic disk capacitively coupled to the metallic frame of the car.

When the 22 inch diameter disk with a 1 inch spacer was mounted on the roof of the Lincoln Continental Mark III 54 AM stations were above 5 microvolts on the meter in the receiver AGC circuit. Increasing the spacer to 1-1/2 inches the disk received 60 stations and a 2 inch spacer received 62 stations. That day the whip mounted on the front fender received 52 stations. Mounted under the floor behind the right front wheel the 22 inch plate with a 1 inch spacer received 49 stations, which is comparable to the whip. A 15 inch diameter disk roof mounted received 49 stations for a 1 inch disk to ground plane spacing and 44 stations from a 1/2 inch spacing. From this data we can conclude, 1) a larger diameter disk will increase performance, and 2) for a specified diameter increasing the ground plane to disk spacing will increase performance level. For a hidden antenna it is desirable to reduce the overall height but Figure 4-2 shows the impedance is dropping along with signal level as the disk to ground spacing is reduced.

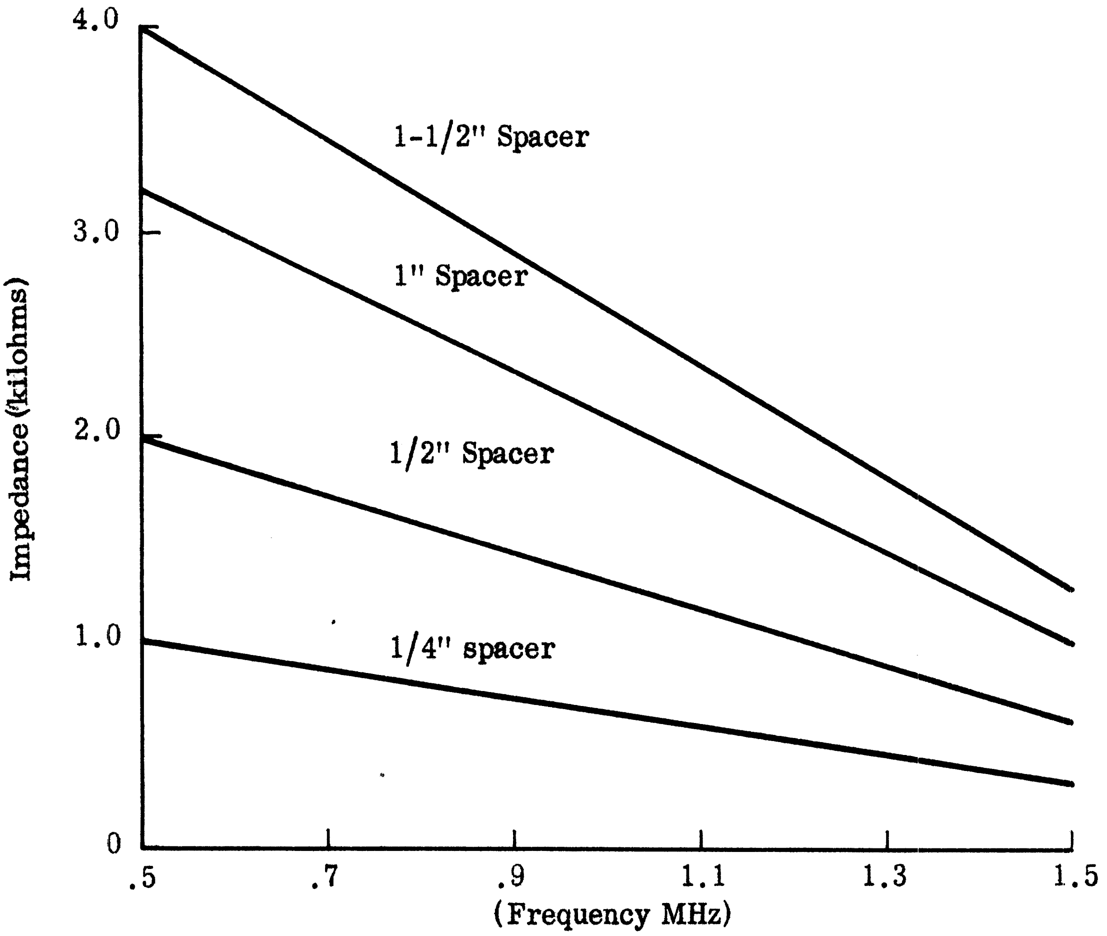


FIG. 4-2: Impedance of a 22" Disk as a Function of Ground Plane Spacing.

Radio waves in the AM band penetrate the earth for several meters so that signal level under the car is not significantly different from that at fender level. It is believed that properly protected with a waterproof cover the AM reception of the disk antenna would be relatively unaffected by weather and road grime buildup. However, steel reinforcing in concrete highways or wet pavements may reduce signal strength for under-car antennas one-third or more, (Williams, 1950). Perhaps the greatest problem for antennas under the car is the higher noise level. This noise stems from ignition noise and partly from the discharge of static fields set up by action of the rotating wheel as the rubbing action on the road charges the tires. This charge leaks onto portions of the wheel which are insulated from the main body by oil films. Noise fields are generated when such capacitors are discharged by the potential being great enough to puncture the oil films or by sparking across to the brake drums. The under-car antenna is especially prone to the intermittent contact noise caused by poor bonding of metal portions of the car (Williams, 1950). In our testing the plate under the car the "chopper" noise of the gas gauge in the dash was noticeably evident in the received signal although this was not a problem for the roof mounted disk.

Currently the Mercury Cyclone and the Continental Mark III have sufficient space for mounting a 22 inch diameter disk antenna underneath the passenger side of the vehicle behind the front wheel. This proved to be a good location for the AM band frequencies as the antenna gave a station count of approximately 49 stations and was comparable to that of the 22 inch whip in overall performance for the AM band.

The plate above the ground plane antenna is not a desirable configuration for the FM band. Reduced FM performance was due to two factors; 1) the antenna itself was not particularly well suited for the FM frequency band, and

(2) as one lowers an antenna towards the ground (at the FM frequency band) the signal strength decreases. As was noted for AM reception the performance increases with disk diameter but in this case improvement with ground plane spacing is not as important a parameter. This trend is supported by the impedance which ranges from 7 - 9 Ω at 88 MHz to 16 - 22 Ω at 108 MHz. This impedance mismatch cannot explain the reduced FM band gain which for a 22 inch diameter disk was 15 to 20 dB below a whip antenna when mounted on a ground plane simulating a car roof. Another factor in reduced under-car performance of the disk is the sharp gradient of FM signal level near the ground. There is a signal difference of approximately 10 - 15dB between the signal levels of an antenna mounted on the fender and one near or slightly above ground level. This reduced signal level near the ground indicates that it is undesirable to mount any FM antenna underneath the car. The performance of the 22 inch disk for FM varied considerably depending on whether it was on the roof or under the car. FM antennas under the car will be more severely affected by ice, snow and slush forming on the antenna in the driving conditions encountered in the northern states.

The disk antenna does have desirable features for a hidden antenna if the injection molded trunk lid becomes financially feasible. A different configuration of such an antenna, an annular slot, in which the disk is flush with the surrounding ground plane and is mounted in a shallow cylindrical cavity would appear to improve the FM performance (Jasik, 1961). For the reported work the cavity depth (4.5") is too great to be located in the trunk lid, but this depth should depend upon disk diameter. Loaded foam dielectrics with high dielectric constants could further decrease the apparent cavity depth. Other work (Rhodes, 1949) has indicated the slot depth could be reduced to 1 - 1-1/2 inch without dielectric loading. The annular slot should work well in the AM band region. The cavity backed antenna has the advantage of being insensitive to trunk luggage or other type of trunk loading. However, this antenna or any type utilizing a horizontal surface has the disadvantage of ice and snow accumulation on the surface. Such accumulation would not "short" the antenna to the ground plane as in the case of the isolated trunk lid but may reduce signal level in the antenna vicinity. A solid state amplifier matching network could be used to solve the problem of reduced performance from the increased capacity of the longer cable run to the trunk.

The windshield antenna is difficult to evaluate in the AM band because of the high field gradient that exists across its interface. It is known that the production windshield antenna does not have an effective length equal to a whip antenna. This was thought to be due to a reduced induced voltage caused by a polarization mismatch, i.e. $\vec{E} \cdot \vec{h}_{eff}$. Subsequent tests have shown that the fields at a level near the fender height are not truly vertically polarized but that reception of a 22 inch whip differs little when mounted horizontally normal to a fender surface or mounted vertically on the fender. It is possible to talk of a pure electromagnetic wave polarization only in free space conditions. Energy propagation for local stations in the AM band is composed of two main types; 1) the direct wave, and 2) the surface wave. Surface topography affects both modes of propagation. For the direct wave trees, telephone poles and buildings re-radiate the energy incident upon them. Since the surface wave re-radiates the energy along the contours of the earth, such contour irregularities as large buildings and hills radiate a wave with a horizontal component. As a result the electromagnetic field has an unpredictable almost random polarization which may vary with topography. In the Willow Run area the fields are sufficiently randomly polarized that the antenna polarization is not important.

Because of the large buildings in a city, it would appear that the random polarization will occur there. Unanswered is the question of polarization in the level rural areas of Illinois, Iowa, Kansas, etc. The polarization of AM energy is an important aspect of the future of windshield antennas in terms of their ability to receive AM stations as the windshields of newer models become increasingly horizontal.

If the received wave is truly vertically polarized top loading will have a limit in increasing the effective height. A short monopole has a triangular current distribution with a resulting effective height one-half the actual length. Top loading increases the antenna capacitance and with increasing length of top section changes the current distribution from triangular toward uniform distribution. Therefore, a vertical monopole in a purely vertically polarized field could at most double the effective height with top loading.

A more important problem is that the voltage induced by a true vertically polarized field corresponds to $E \cdot h_{eff}$. As the windshield contour is increasingly streamlined the horizontal projection increases and the voltage induced from a vertically polarized wave is reduced, thus reducing the power received by the windshield antenna.

Placing a "top loaded" antenna in a randomly polarized field, which is believed to be a more accurate representation of the AM field, creates a much more difficult antenna to analyze especially when it is placed in a windshield with metallic boundaries and a steep field gradient across the windshield area. Performance due to the "top loading" of the antenna in a randomly polarized field is not limited to a doubling of the effective height as the horizontal portion of the antenna is no longer a non-radiating element. Increasing the length of the horizontal portion will be effective in increasing the effective height. Theoretically close looping of the wire to get the added length will not give the desired results in terms of added signal strength for the number of linear feet of wire added. Consideration of the transmitting mode of operation will give insight into the reason for this. In the top section the currents in these parallel wires of the closely spaced loops will be approximately of the same magnitude and opposite in direction with adjacent parallel wires and this will appear like a two-wire transmission line. Any radiation occurring results from; 1) the reduced currents from one line to another due to increased length of a distributed resistance of the wire (which is considerable for the thin wire or silk screening process utilized in windshield antennas), 2) the taper of current that is imposed by the boundary condition of zero current at the tip, and 3) the small phase variation (AM band) due to distance along the wire. As the length of the top loading increases the importance of the second effect will be reduced. Very tight looping of the wire may have the affect of adding a solid conductor the width of the total horizontal portion. This can be checked by placing aluminum tape on a separate windshield to represent the horizontal portion of the antenna and connecting a thin vertical wire being careful to obtain continuity between tape and wire at the junction. One would expect the wire loop to be more effective but the above test might show the diminishing returns from working to get increased length of the wire in the tinted windshield gradient band by compressing the looping configuration.

V

NOISE

Antenna engineers have been intrigued for many years by the theoretical possibility of an electrically small antenna being nearly as efficient as a resonant half wave dipole. We can see how this theoretical efficiency comes about by looking at the maximum effective aperture. First consider the maximum amount of power that can be delivered into a load resistance at an antenna terminal. For this case, consider a lossless transmission line between an antenna, which can be represented by a voltage generator and an impedance, and the load which can be a complex impedance. This circuit representation is shown schematically in Fig. 5-1. For such a circuit the current as shown in Eq. (5.1) is simply expressed as the voltage divided by the sum of the two impedances, the antenna, and the load impedance.

$$I = \frac{V}{Z_t + Z_a} \qquad \begin{array}{l} Z_t = R_t + j\,X_t \\ \\ Z_a = R_a + j\,X_a \end{array} \tag{5.1}$$

Equation (5.2) shows the general form of the equation for the current when we substitute in the values for the two complex impedances and include a loss term in the antenna resistance.

$$I = \frac{V}{\sqrt{(R_r + R_\ell + R_t)^2 + (X_a + X_t)^2}} \tag{5.2}$$

$R_a = R_r + R_\ell$

R_r = radiation resistance

R_ℓ = antenna loss resistance

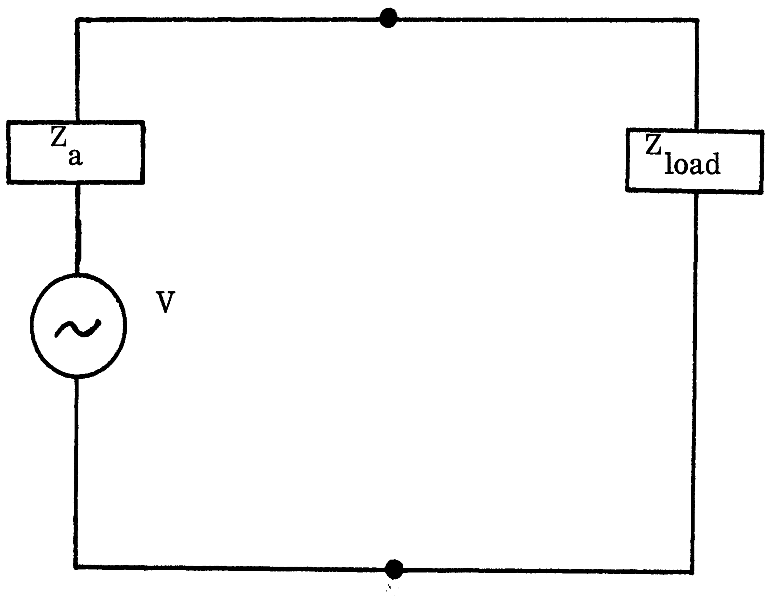


FIG. 5-1: Circuit Representation of Antenna with Load.

A general expression for the power is given in Eq. (5.3).

$$W = I^2 R_t = \frac{V^2 R_t}{(R_r + R_\ell + R_t)^2 + (X_a + X_t)^2} \quad . \tag{5.3}$$

The effective aperture of an antenna is defined as the ratio of power in the terminating impedance to the power density of the incident wave, and this is shown in Eq. (5.4).

$$\text{effective aperture} = \frac{W}{P} = A_e \, , \quad \begin{array}{l} W = \text{power in terminating impedance} \\ P = \text{power density in incident wave.} \end{array} \tag{5.4}$$

For such a definition the effective aperture has a dimension of area if W is in watts and P is in watts per square meter, the resulting effective aperture will have the dimension of square meters. The maximum effective aperture occurs when maximum power is delivered to the terminated impedance. The power to the termination can be maximized by having the impedance of the termination be the complex conjugate of the antenna impedance. This causes the second term in the denominator of equation (5.3), the quantity $(X_a + X_t)^2$, to go to zero. If we neglect the loss term or loss resistance (R_ℓ) in the antenna, the maximum power delivered to the antenna is simply given by Eq. (5.5).

$$W_{max} = \frac{V^2 R_t}{(R_r + R_t)^2} = \frac{V^2 R_t}{(2R_t)^2} = \frac{V^2}{4R_r} \qquad R_t = R_r \ \text{for maximum power transfer to terminal impedance} \tag{5.5}$$

The maximum effective aperture is defined as the value obtained for the maximum power delivered to the load, i.e. with an expression using W_{max} obtained in Eq. (5.5).

This leads to the expression for the maximum effective aperture as shown in Eq. (5.6).

$$A_{e\,max} = \frac{W_{max}}{P} = \frac{V^2}{4P\,R_r} \quad . \tag{5.6}$$

The voltage induced in an antenna is equal to the dot product of the E-field and the effective height of the antenna. Current distribution on a properly terminated electrically short dipole is approximately linear, i.e. 0 at the tips of the antenna with the current increasing in a linear fashion to a maximum current at the center of the dipole. When the current distribution is linear the effective height is one-half the physical length (Schelkunoff and Freize, 1952). The power density of the incident wave at the dipole is given by the square of the E-field divided by the intrinsic impedance. For a short dipole with a linear current distribution the input resistance is given by Eq. (5.7), (Schelkunoff and Freize, 1952).

$$R_r = 80\,\pi^2 \left(\frac{\ell}{\lambda}\right)^2 \quad . \tag{5.7}$$

Although the equation for the radiation resistance given by Schelkunoff and Freize is not as complex as that given by King (Eq. (2.2), which includes the expression for the thickness ratio, the two equations give similar values of radiation resistance for thin antennas. As was shown earlier the value for the radiation resistance was 1.5×10^{-3} Ω for a two foot monopole at 1 MHz. This simpler equation for radiation resistance gives the same value when one takes into account that the radiation resistance of a monopole is one half that of a dipole.

Substituting the values for the voltage, the power, and the radiation resistance into equation (5.6) we are left with a simple expression for the maximum effective aperture which is shown in Eq. (5.8).

$$A_{e\,max} = \frac{E^2 \left(\frac{\ell}{2}\right)^2 (Z)}{4\,E^2\, 80\pi^2 \left(\frac{\ell}{\lambda}\right)^2} = \frac{\frac{1}{4} \times 120\,\pi}{320\,\pi^2 \left(\frac{1}{\lambda}\right)^2} = \frac{3\,\lambda^2}{8\,\pi} = 0.119\,\lambda^2 \quad . \tag{5.8}$$

This maximum effective aperture of $0.119\,\lambda^2$ is very close to the theoretical maximum effective aperture which is obtained for a resonant half wavelength dipole $0.13\,\lambda^2$ (Krause, 1950). The difference between these two maximum effective areas is negligible and if one is able to match the short antenna properly at 1 MHz the theoretical effective area of slightly over 1000 square meters would be possible. The more general equation for the effective aperture given in Eq. (5.9) indicates the reason why this maximum effective aperture cannot be obtained.

$$A_e = \frac{V^2 R_t}{P\left[(R_r + R_\ell + R_t)^2 + (X_a + X_t)^2\right]} \tag{5.9}$$

Loss in the coil needed to "conjugate match" the antenna causes the terminating resistance to be over 1000 times the radiation resistance. Also the ohmic loss resistance of the antenna is not negligible for small wire. For example, two feet of No. 38 copper wire has a loss resistance almost 1000 times greater than the radiation resistance. Shunt capacitances of (1) the antenna base mounting, (2) the coaxial cable between the antenna and the receiver, and (3) the shunt capacitance of the connector to the radio prevents the antenna from being conjugate matched since the transformed antenna reactance at the receiver terminal is approximately one-tenth of the antenna capacitance. For these reasons we see by comparing Eqs. (5.8) and (5.9) that the actual effective area is greatly reduced from the theoretically possible area shown in Eq. (5.8).

Some of the physical effects that reduce effective apertures can be overcome by placing an active device in the antenna. Using the active device to transform the large antenna impedance to a lower value reduces the problem of shunt capacitances in the antenna base, coaxial cable between the antenna and receiver, and the cable connector. Notice that the theoretical maximum effective aperture in Eq. (5.8) is independent of antenna height and it is theoretically possible to make a very short antenna with the same gain as a longer antenna and thus reduce also the line loss term in Eq. (5.9).

In the early 1950's the radio astronomers began moving the first amplifier of the receiver to the base or to the output of the antenna to eliminate the problems with line loss. This brought an improvement by amplifying the signal before it was attenuated through the line thus reducing the overall system noise. This is still not the desired solution for the electrically short antenna since the base capacitance of the antenna would still be present in the new configuration although the problem of the shunt capacitance of the cable is reduced since the transistor can be used as an impedance transformer. In the early 1960's an engineer at Wright-Patterson AFB proposed putting the transistor inside the monopole itself, i.e. fabricating the antenna with the transistor as an integral part of the antenna, not adding it to the antenna base. Shortly after this funding was obtained for Meinke in Germany to work on the problem of reducing the losses in short antennas and impedance matching networks by utilizing integrated transistor antenna techniques. The following section is a summary of the Meinke work with active antennas. Material contained in the rest of section V represents information contained in three reports by Meinke on his work from 1966 to the fall of 1969. The purpose of this section is to form an introduction into integrated antenna techniques. These reports have generated much controversy among antenna engineers with some disputing the theory and data presented. Regardless of whether one argues for or against the concept the technique works and will have an increasing influence on electrically small antennas in the near future.

In working with this concept of active antennas Meinke found an additional bonus, not only did one overcome the problem of base capacitance of the monopole and ohmic losses in the antenna, but as one increases the distance above the ground plane for the installation of the transistor, the radiation resistance also increases as shown in Fig. 5-2. Although this data is for a spiral wound monopole, theory and data by Meinke, King (1956), and Schulkunoff (1952) indicated that this type of increase in radiation resistance is also found for a linear monopole. Maximum radiation resistance for such antennas is found when the slot for the transistor placement is at the top of the antenna rather than base as had been earlier tried by those who had merely moved the first amplifier of the receiver to the base of the antenna. Placing the transistor in the antenna creates a whole new set of problems on measurement of antenna performance. A new type of figure-of-merit had to take into account the amplification of the device so that one would not be measuring just an increase in signal level without an increase in signal noise ratio. This led to the study of the noise temperature for the antenna and for the system.

The terminology and concept of noise temperature for an active antenna along with a receiver may seem confusing when the concept is first encountered but the following discussion concerning the measurement of noise temperature of both the antenna, receiver and active element should help clear the issue. There are several ways that the antenna noise measurements can be obtained to get an overall system measurement as we can see from Fig. 5-3. We can break the network at terminals 2 and 2' and use noise measurements to measure the noise of the receiver with the impedance presented to it by the antenna matching network. The other source of noise is the antenna combined with the matching network which has an internal gain of its own. Note that the gain of the matching network can include a loss term if the matching network is passive. The second measurements corresponding to the B part of the figure could be made at terminals 1 and 1'. In this case the one would obtain a

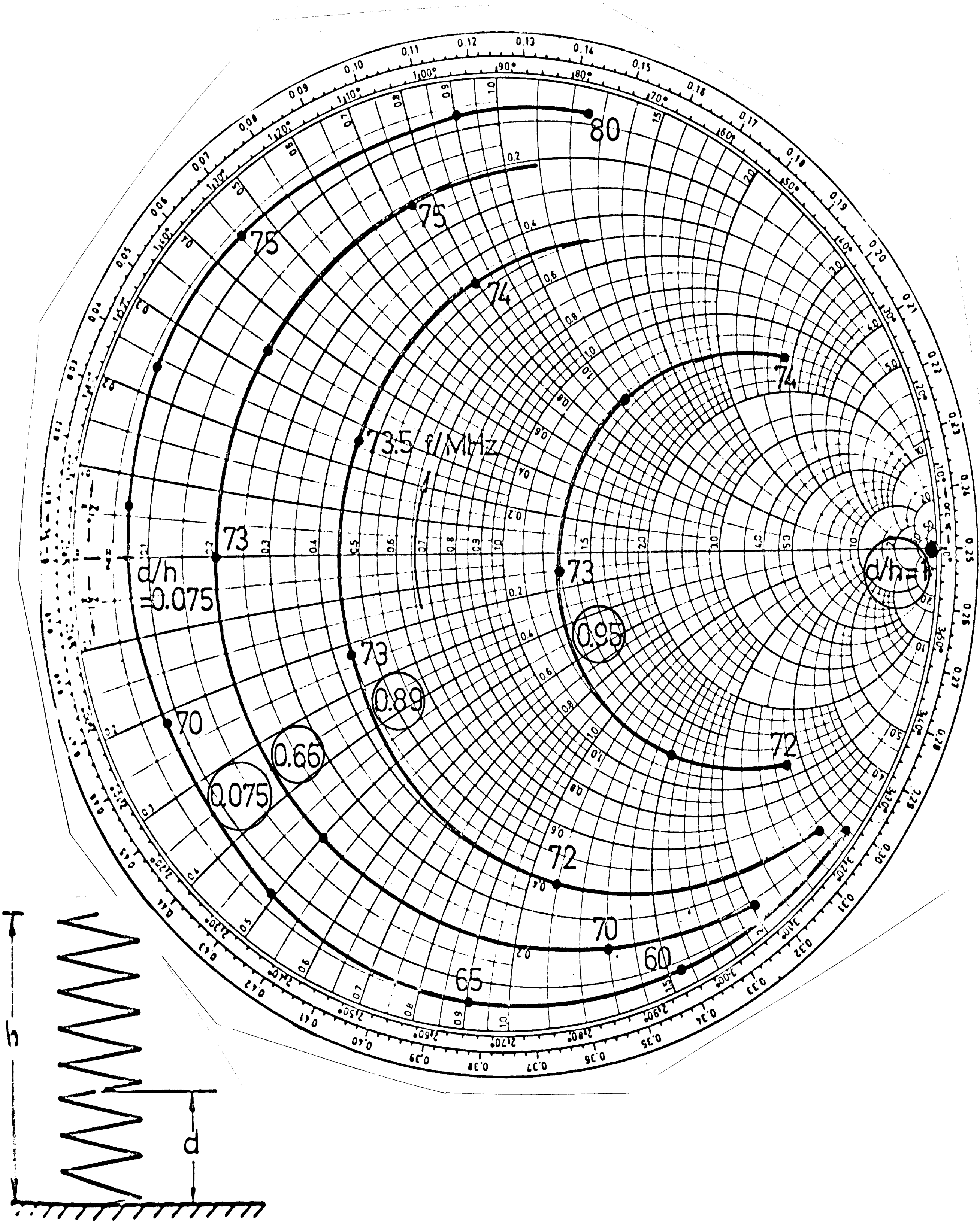


FIG. 5-2: Impedance of Spiral Wound Monopole Showing Effect of Moving Feed Point Toward Tip, F = 60 - 80 MHz, (after Meinke).

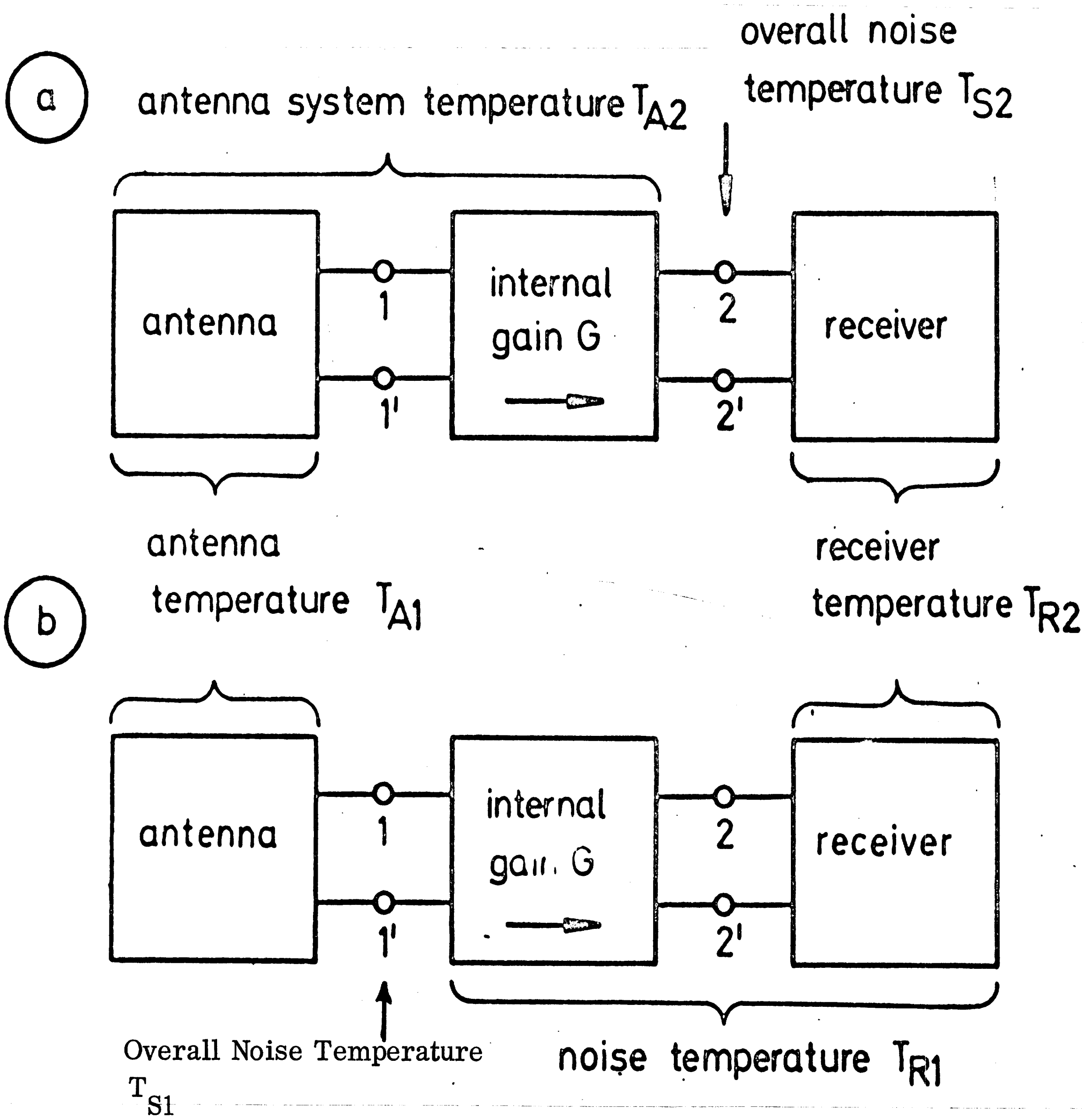


FIG. 5-3: Break Points in Antenna-Receiver System for Measuring Noise, (after Meinke).

noise temperature T_{R1} which includes the noise temperature of the matching network and the receiver and also the noise temperature of the passive antenna by itself (T_{A1}).

First let us consider the case of measuring the noise temperature at terminals 2 and 2'. Figure 5-4 shows how the overall noise temperature T_{S2} is measured at terminals 2 and 2'. The antenna system of Fig. 5-4 is replaced by a calibrated noise generator which offers an adjustable noise temperature T_N with an impedance of 50 Ω. A low loss network transforms this impedance into the complex impedance Z_i which the antenna system offers between terminals 2 and 2' of Fig. 5-4, Part A. As a result the antenna impedance is substituted by the transformed noise generator with an equal impedance Z_i. The low loss transforming network does not influence the calibrated noise temperature of the generator. The receiver output is terminated by a square law indicator Q. By varying the noise temperature of the calibrated noise source a linear relationship is obtained for the variable noise temperature and the square law output of the receiver as shown in Fig. 5-5. Noise temperature T_N of the adjustable noise source cannot practically be lower than the room temperature T_0 which is approximately 290^o Kelvin. However, the noise temperature of the receiver can be found by extending the curve by the dotted line which intersects the vertical axis at a value $Q = Q_2$ and the horizontal axis at a point $-T_{R2}$. Q_2 is the noise output of the receiver at a noise temperature of $T_N = 0$, therefore, it is a measure of intrinsic noise of the receiver. T_{R2} is the equivalent noise temperature of the receiver at terminals 2 and 2' if the receiver is fed by a source with a source impedance Z_i which is the antenna impedance transformed by the matching network N. T_{R2} is the magnitude of the noise temperature that we would have to translate the vertical axis to the left to obtain 0 noise temperature so it becomes clear that T_{R2} is the intrinsic noise temperature of the receiver at noise temperature $T_N = 0$. If the noise

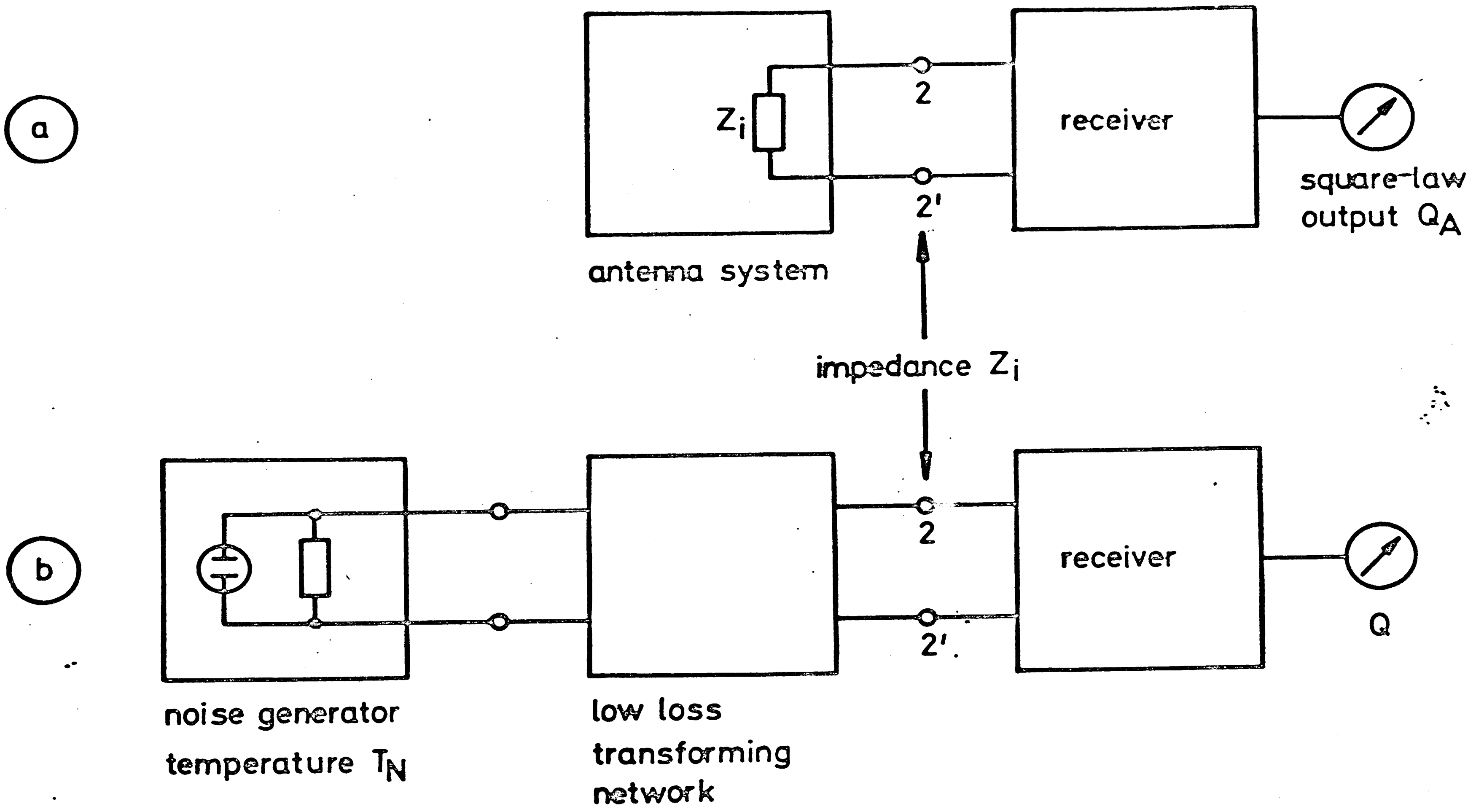


FIG. 5-4: Method of Noise Measurement T_{S2} at Terminals 2 - 2' (after Meinke).

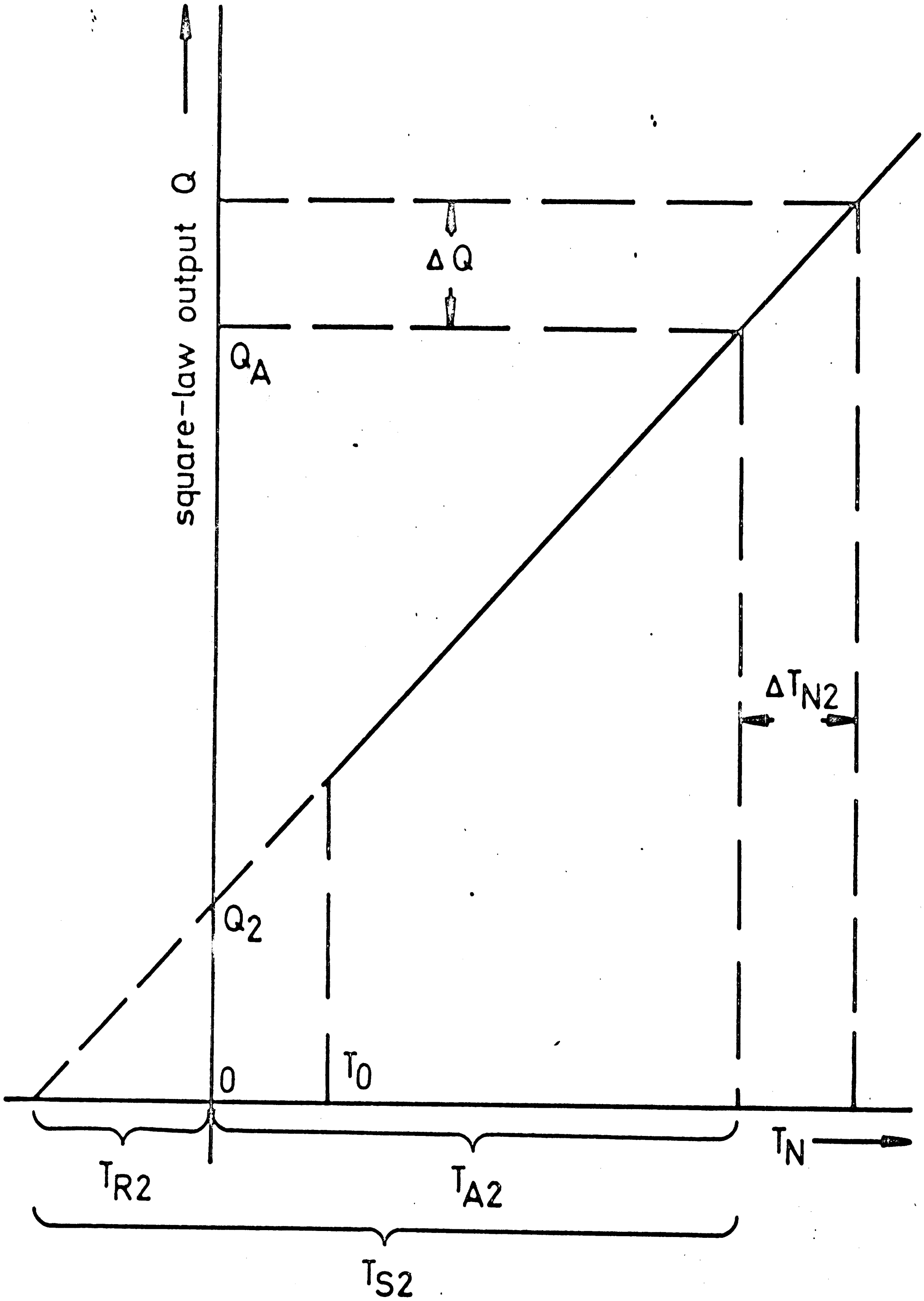


FIG. 5-5: Square Law Curve for Variable Noise Temperature at Terminals 2-2', (after Meinke).

generator is replaced by the real antenna system at terminals 2 and 2' the square law receiver output will be Q as shown in Fig. 5-5. Using the calibration curve obtained by the variable noise source (Fig. 5-5) Q_A corresponds to a noise temperature T_{A2} of the antenna system at terminals 2 and 2'. The overall noise temperature of the receiving system, i.e. the antenna system plus receiver, is given by Eq. (5.10).

$$T_{S2} = T_{A2} + T_{R2} \tag{5.10}$$

Figure 5-6 shows how the overall noise temperature T_{S1} is measured at terminals 1 and 1'. The antenna is now replaced with a noise generator capable of varying the noise temperature T_N. Again the source impedance of the noise generator is transformed by a low loss network but this time it is transformed into the complex impedance Z_A that the antenna offers between terminals on 1 and 1' and Fig. 5-6. Notice that in the previous case when we entered at terminals 2 and 2' the low loss network transformed the noise generator impedance to the impedance that was presented to the receiver at the output of the matching network, i.e. the complex impedance which the antenna system offers, not the antenna alone. Again the receiver output is terminated by the square law indicator as before. Figure 5-7 is the measured relationship between noise temperature T_N and the noise generator in the square law output Q of the receiver. Extending the linear curve by a dotted line below the room temperature this line intersects the vertical axis at Q_1 and the horizontal axis at $-T_{R1}$. Q_1 and T_{R1} describes the noise of the receiver plus the noise of the antenna matching network N. That is, T_{R1} is the equivalent noise temperature of network N plus the receiver at terminals 1 and 1'. If the network is fed by a source with an impedance Z_a, the antenna impedance, at terminals 1 and 1' replacing the noise generator by the

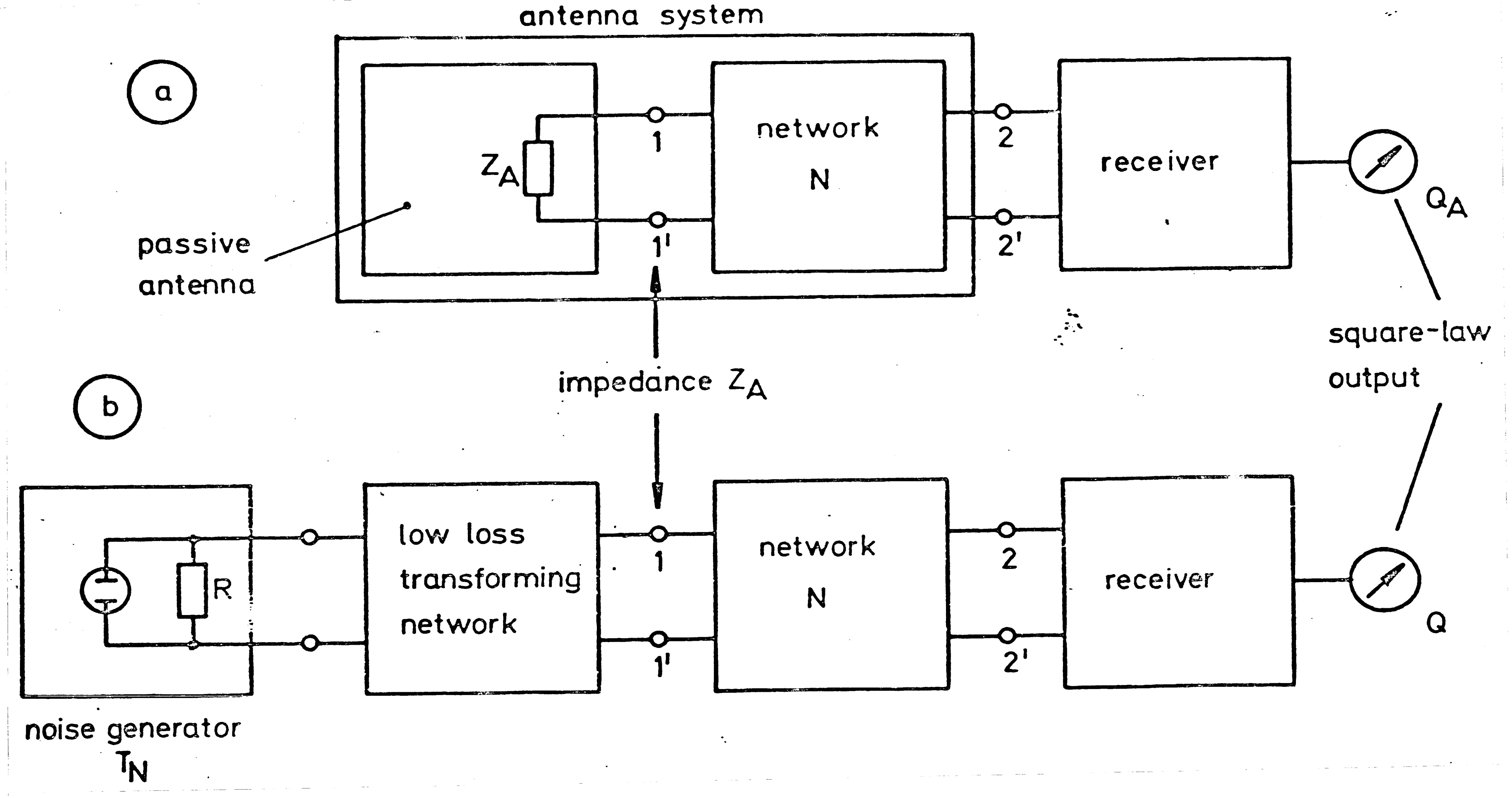


FIG. 5-6: Measurement of Noise Temperature T_{S1} at Terminals 1-1'. (after Meinke).

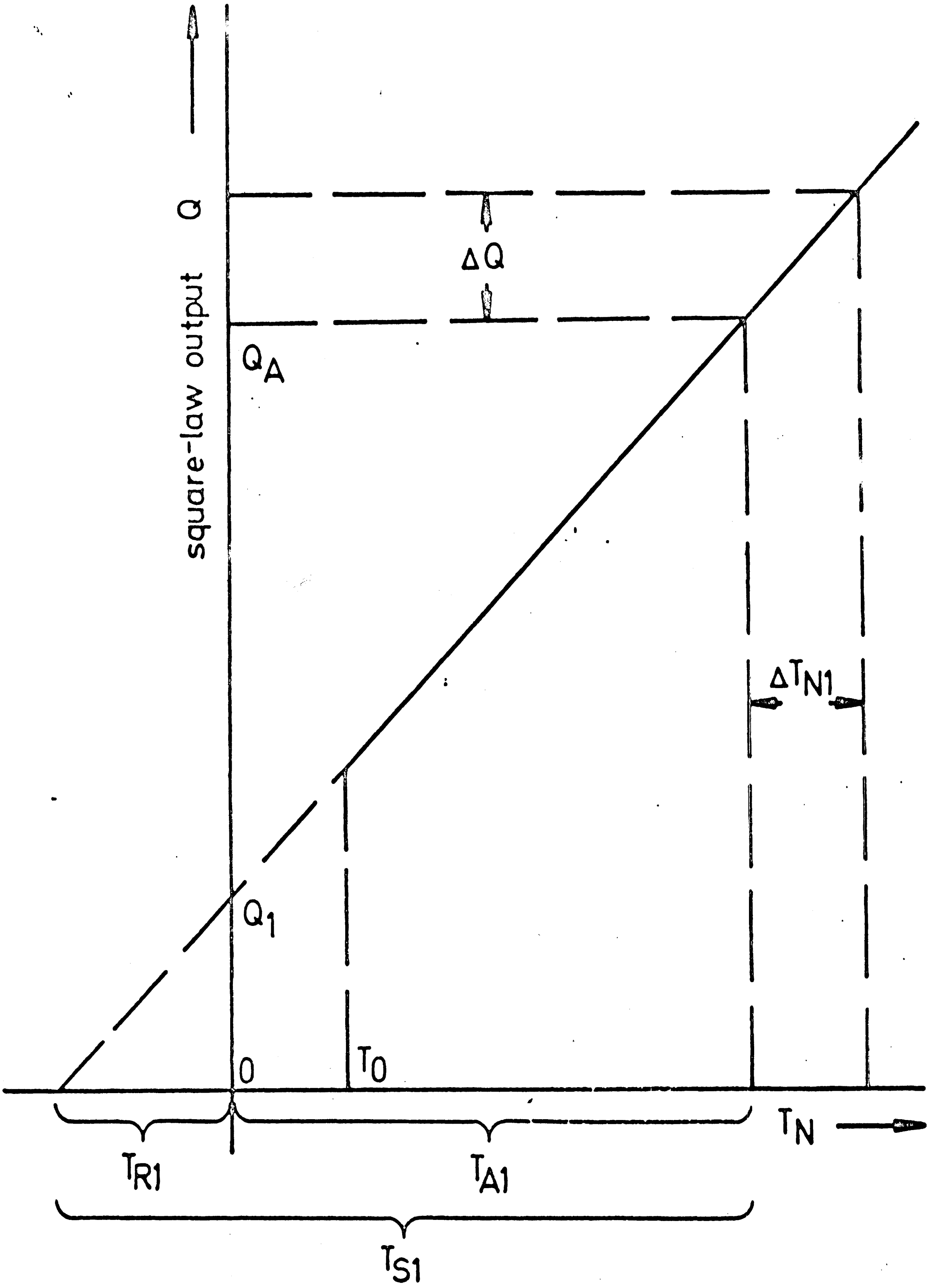


FIG. 5-7: Square Law Curve for Variable Noise Temperature at Terminals 1 - 1', (after Meinke).

real antenna, the square law receiver output will have a measured value Q_a. This measured level of square law response with the antenna replacing the noise generator corresponds to a noise temperature T_{A1} which is the noise temperature of the antenna at terminals 1 and 1'. The antenna noise temperature is due to cosmic noise, atmospheric noise, and man-made noise. Overall noise temperature of the receiving system, antenna, the matching network and receiver is given by Eq. (5.11).

$$T_{S1} = T_{A1} + T_{R1} \tag{5.11}$$

We can now express the antenna system response by replacing the noise of the antenna with a noise source of temperature T_{S1} and source impedance Z_A at terminals 1 and 1' as in Fig. 5-8b or by a noise source temperature T_{S2} with an impedance equal to the transformed antenna impedance at the terminals 2 and 2'.

By comparing Eqs. (5.10) and (5.11) it is possible to determine the receiver noise alone and the antenna noise alone and by manipulating these equations it is possible to determine the noise of the matching network N. There are three important sources of the noise, (1) the antenna noise T_{A1}, (2) the noise of the matching network N, and (3) the noise from the receiver T_{R2}. From noise theory of cascaded amplifiers we realize that if the matching network contains an amplifier, the noise figure of the receiver is reduced by the gain of the matching network. Therefore two important noise temperatures remain, i.e. the noise of the antenna and the noise of the active matching section which can be called the transistor noise.

The preceeding example considered the noise that would be found in the system for a given input noise i.e. the antenna impedance was not considered a variable and the overall system noise was found for a given antenna impedance. Consider what would happen if the antenna impedance were a variable and one were trying to minimize

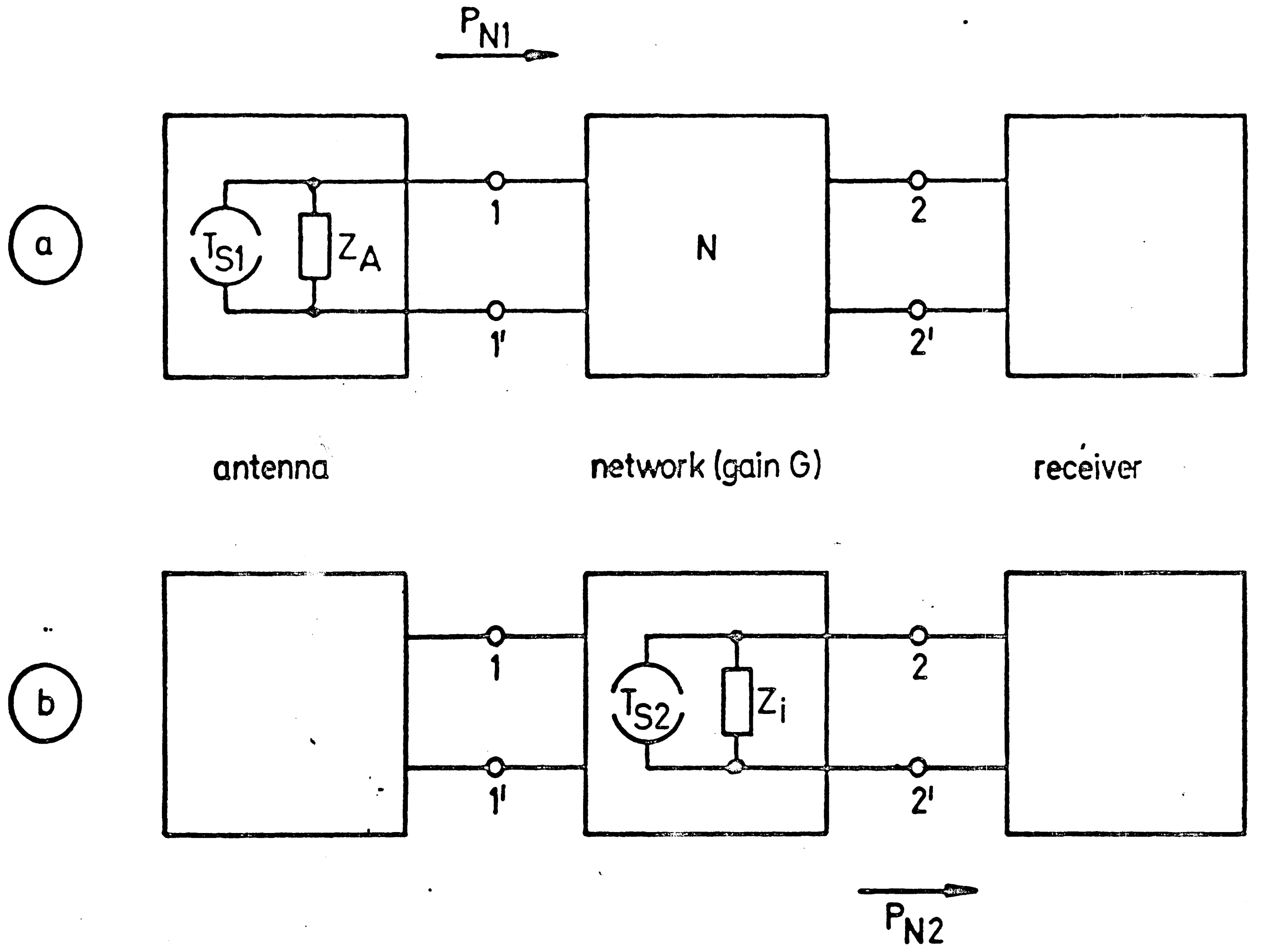


FIG. 5-8: Antenna Replacement by Noise Generators of Proper Noise Temperature and Impedance Will Give Equivalent Noise Performance, (after Meinke).

the noise from the amplifier section, the transistor noise. The noise measurements on the transistor would be made similar to that in the previous section, except now the transforming network between the noise generator and the transistor is varied to present different impedance levels to the transistor and the noise level of the transistor is measured for each of these impedance levels as is shown in Fig. 5-9. The relationship between the transistor noise temperature and the input impedance can now be displayed on the Smith chart as in Fig. 5-10 for a transistor at 1 MHz. In this chart there is a point on the Smith chart where a certain impedance that will give a minimum noise temperature with that transistor. In this case the transistor has a minimum noise temperature for an input impedance that is real with a magnitude of approximately 235 Ω. The circles on the Smith chart represent impedances that will give a constant noise temperature. That is, any impedance inside the circle will give an equivalent noise temperature with the transistor that is less than the noise temperature represented by the circle boundary. By plotting such a Smith chart we are now talking in terms of impedance noise mis-match rather than impedance power mis-match, and the most important consideration is proper noise match rather than power impedance match now. Figure 5-11 is a similar plot for a transistor operating at 100 MHz.

The antenna noise temperature is due to the cosmic noise, atmospheric noise, and man-made noise and is greatest near the lower frequencies decreasing in the 50 MHz and above region. At 50 MHz and above regions can be considered as the cross-over point between the importance of antenna noise and transistor noise. Figure 5-12 is a plot of the antenna noise temperature and transistor noise temperature as a function of frequency. From this plot it would appear that the cross-over point from where the antenna noise and the transistor noise become about equal is considerably above 100 MHz. One must remember this is a plot of transistor noise

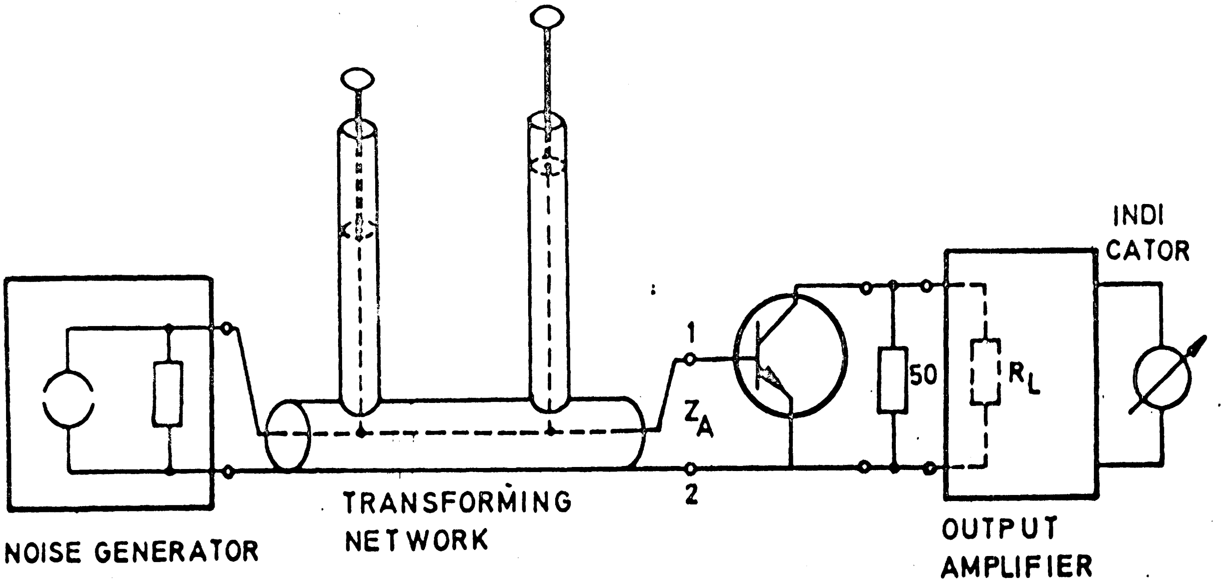
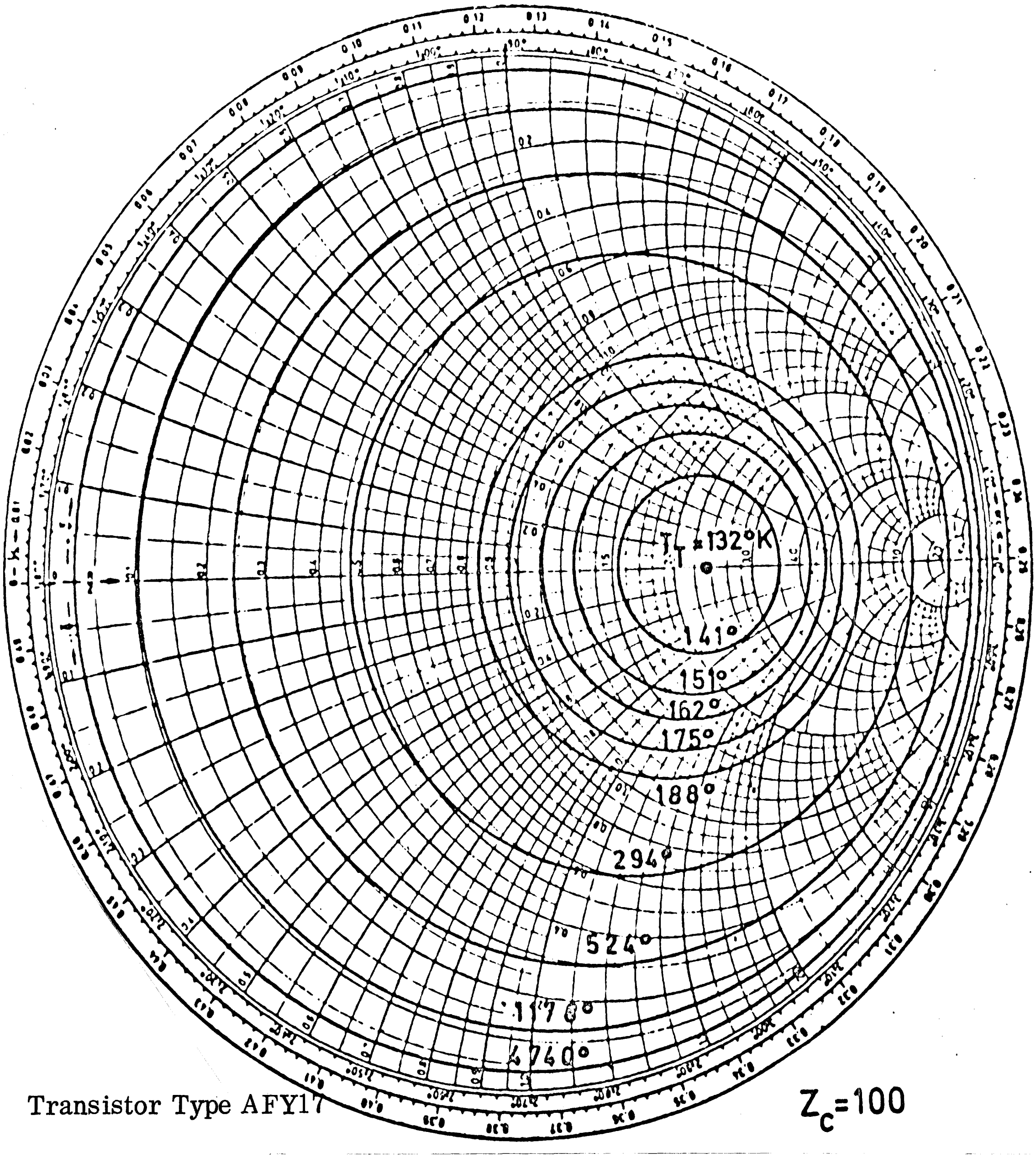
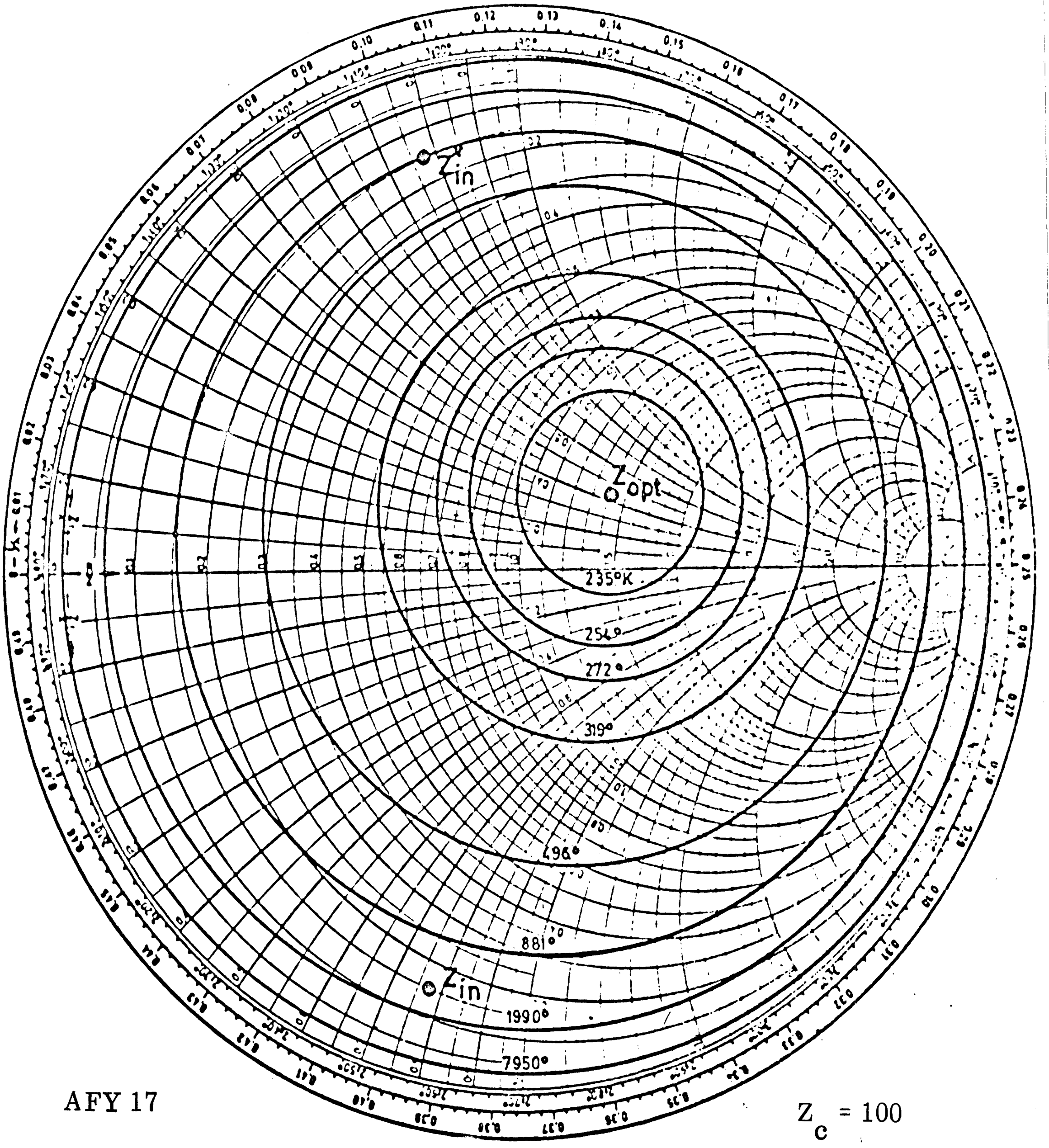


FIG. 5-9: Measurement of Transistor Noise Temperature as a Function of Input Impedance, (after Meinke).



Frequency = 1 MHz

FIG. 5-10: Noise Temperature of a Transistor as a Function of Input Impedance, F = 1 MHz, (after Meinke).



Frequency = 100 MHz

FIG. 5-11: Noise Temperature of a Transistor as a Function of Input Impedance, F = 100 MHz, (after Meinke).

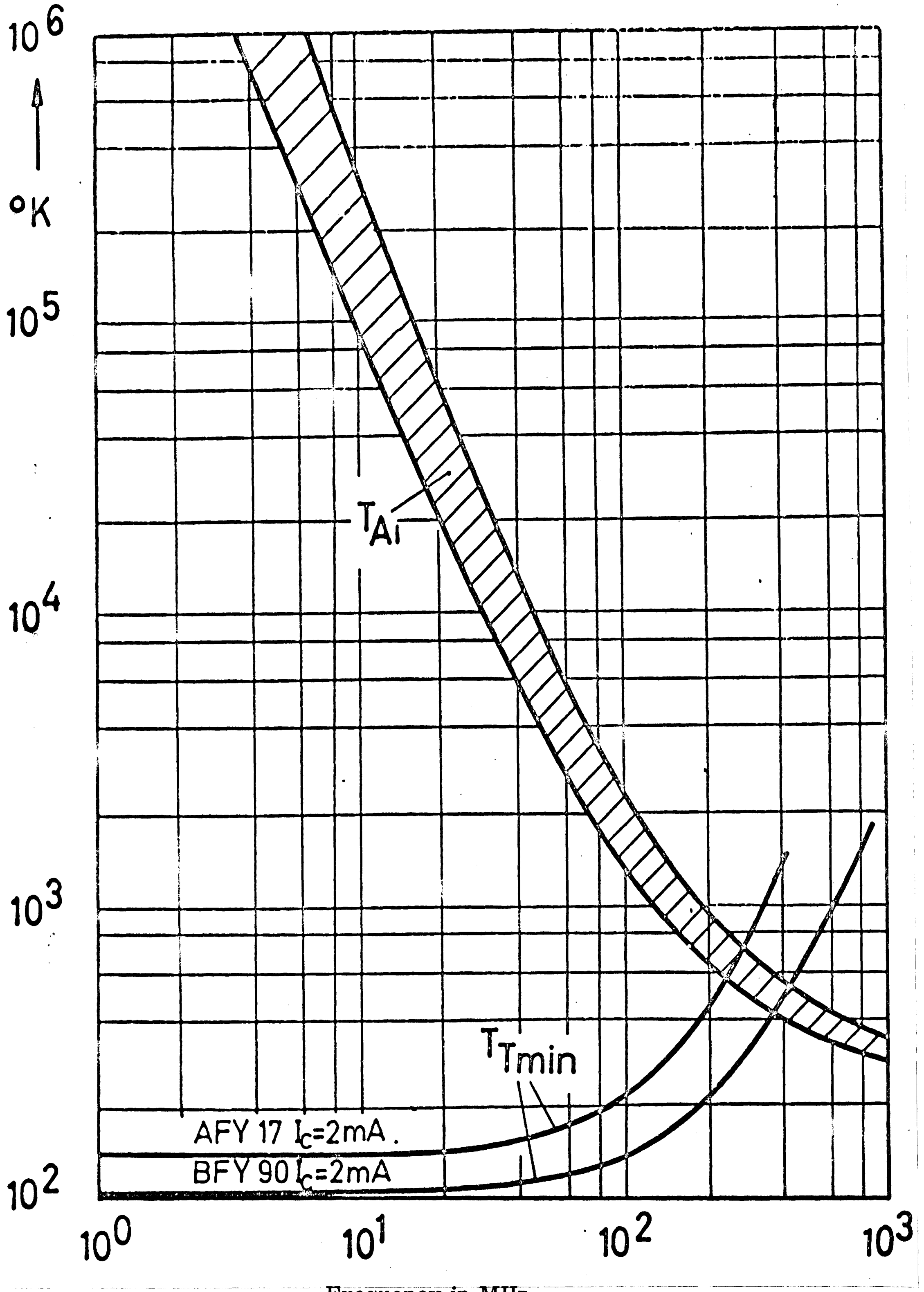
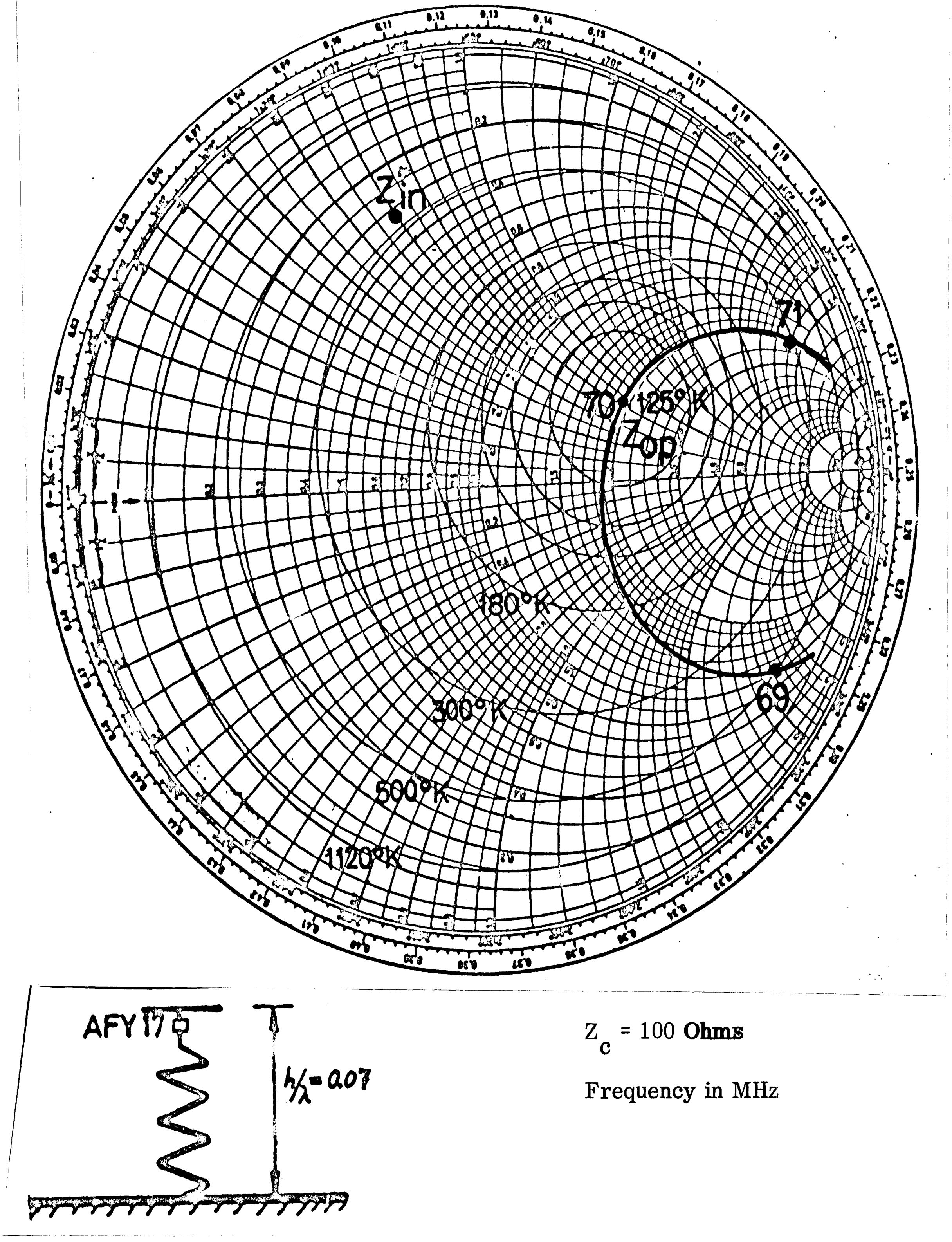


FIG. 5-12: Antenna and Transistor Noise as a Function of Frequency; note Values for Transistor Noise are Given for Minimum Obtainable Noise with Proper Input Impedance Match, (after Meinke).

minimum with a properly matched input impedance. If the proper match is not obtained the transistor noise becomes quite important in the 50 MHz region and therefore we could consider this as a design criteria between low and high noise antennas.

First, consider the problems of noise matching in the high frequency (above 50 MHz) region where the transistor noise is or can be large compared to the antenna noise. As the antenna impedance varies with frequency the noise temperature of the antenna with the transistor matching network will cross these circles of constant noise temperature. One now can talk of the overall permissible noise level as being the noise bandwidth and from the number of circles it crosses we can find the allowable noise temperature by plotting the impedance superimposed on the Smith chart for the antenna noise temperature of the transistor as is shown in Fig. 5-13. Rapid variation of antenna impedance with frequency is the primary limitation on the noise matching in the higher frequency band. Since the antenna temperature is low as soon as the transistor imperature increases from the resulting impedance mis-match, the overall system noise increases quite rapidly. In this discussion we see that we are not considering modeling or changing the transistor impedance to fit the antenna impedance, but instead the antenna impedance is designed to give minimum noise temperature of the transistor which is taken as a fixed quantity and not a variable. We should remember at this point that the signal to noise ratio can be expressed by noise temperatures so that by specifying the allowable noise temperature one is specifying the allowable signal to noise variation with frequency of the antenna and matching system.

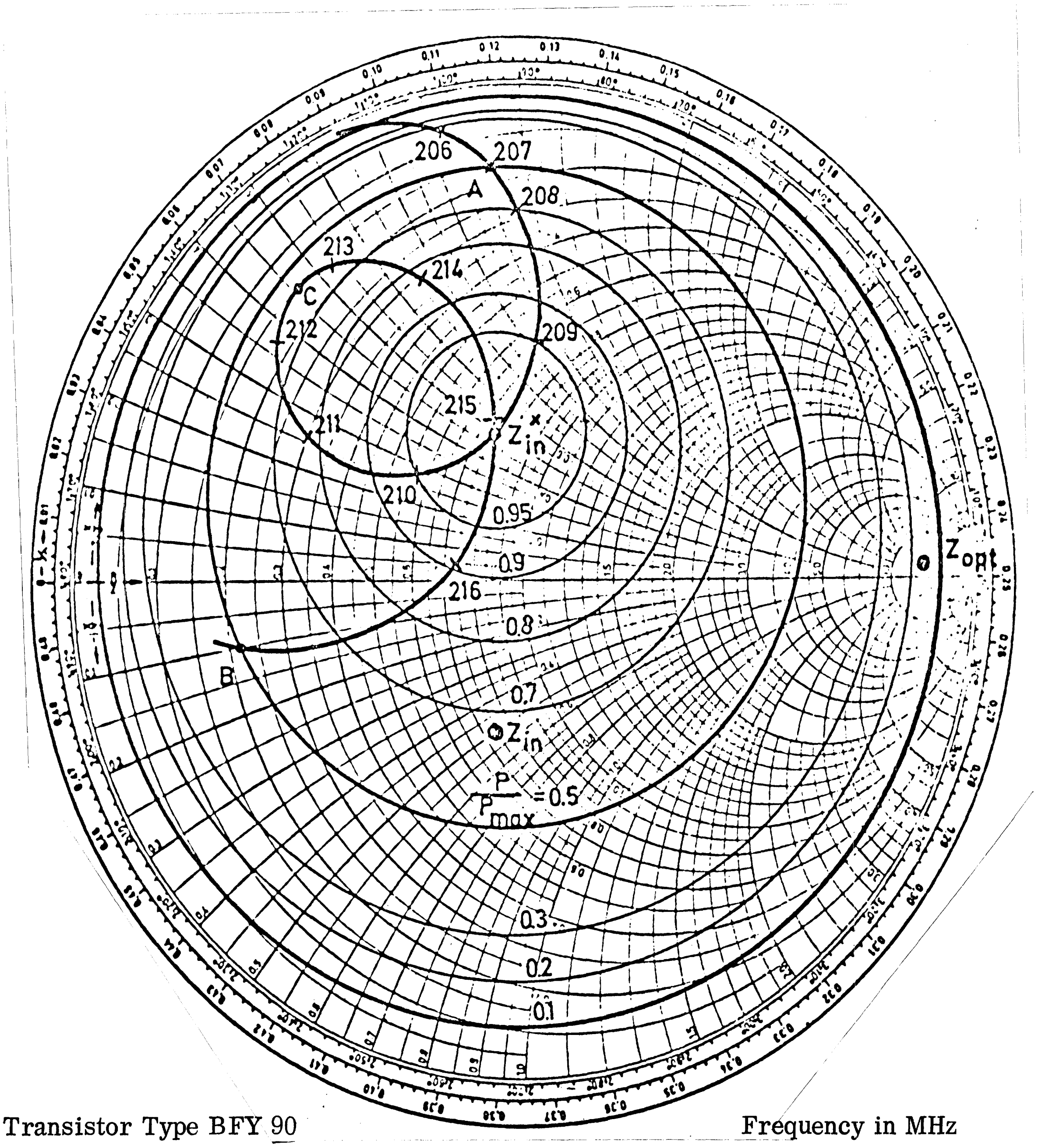
To get more bandwidth a "filter antenna" is used which generally consists of two or more coupled resonant circuits. The antenna impedance depending on frequency forms a loop in the Smith chart as shown in Fig. 5-14. If the antenna impedance can



Z_c = 100 **Ohms**

Frequency in MHz

FIG. 5-13: Impedance of Spiral Wound Monopole Superimposed on Noise Temperature as a Function of Input Impedance for the Transistor. This Defines Useful Bandwidth of Active System, (after Meinke).



Transistor Type BFY 90
I_{co} = 4mA
V_{co} = 10V Z_c = 10

Frequency in MHz

FIG. 5-14: Broadbanding Active Device by Designing an Antenna with a Loop in Impedance Near the Optimum Impedance for the Transistor, (after Meinke).

be made to loop close to the value of impedance which is optimum for minimum transistor noise the bandwidth is extended as the impedance now moves more slowly as a function of frequency through the optimum input impedance to the transistor. The impedance for the low noise operation of the transistor amplifier is then broadbanded much as a band pass filter is broadbanded by stagger tuning separate high Q circuits. The number of possible loops inside the circle of desired noise temperature increases with increasing complexity of the circuitry which forms the antenna impedance. It is for this reason that much of the circuitry involved in the Meinke rear view mirror antenna is for the high frequency FM band to get this looping for optimum signal to noise ratio to broadband the antenna's operation. This circuitry compared with the discrete capacitors and the printed circuit inductors on the right portion of the Meinke antenna is shown in Figs. 5-15 and 5-16. These two figures are of the amplifier section of the rear view mirror antenna and is being marketed in Germany for AM and FM radios.* This amplifier is part of the antenna and the rear view mirror body shell forms only a top loading capacitance to this monopole section which is part of the amplifier.

Now consider the frequency band around 1 MHz where the noise temperature of the antenna far exceeds the noise temperature of the transistor. For this case some people say that the transistor noise match is no longer important but we must remember that the input impedance of the average antenna presents a great mis-match between the optimum impedance for noise match of the transistor and the actual impedance of the antenna. In transistor noise theory the transistor noise temperature T_t is expressed by Eq. (5.12).

$$T_t = T_{t\,min} + T_1 \frac{(R_i - R_{opt})^2 + (X_i - X_{opt})^2}{R_i R_{opt}} \tag{5.12}$$

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*Marketed by the Fuba company in Germany from a design by Meinke.

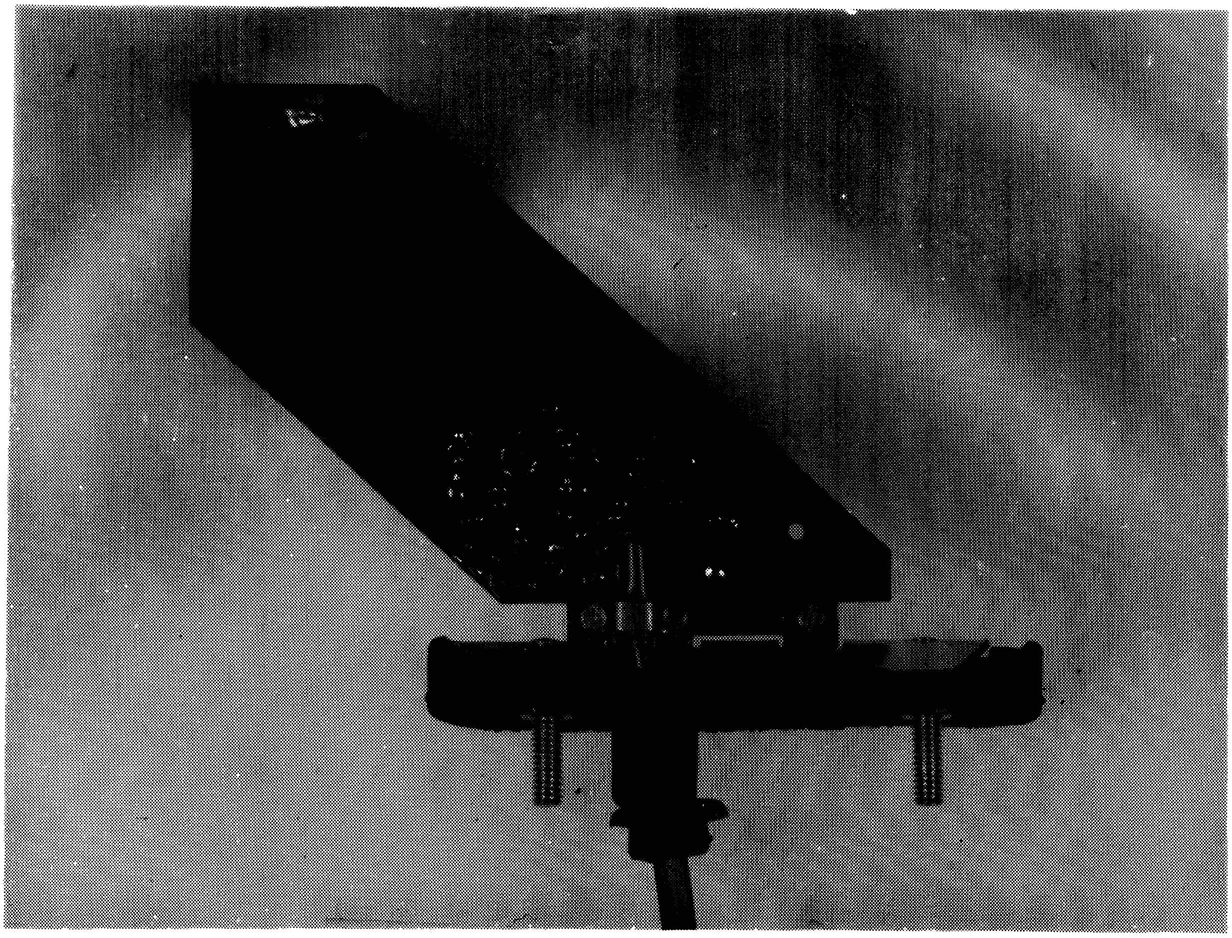

FIG. 5-15: Meinke Antenna - Right Side of Printed Circuitry.

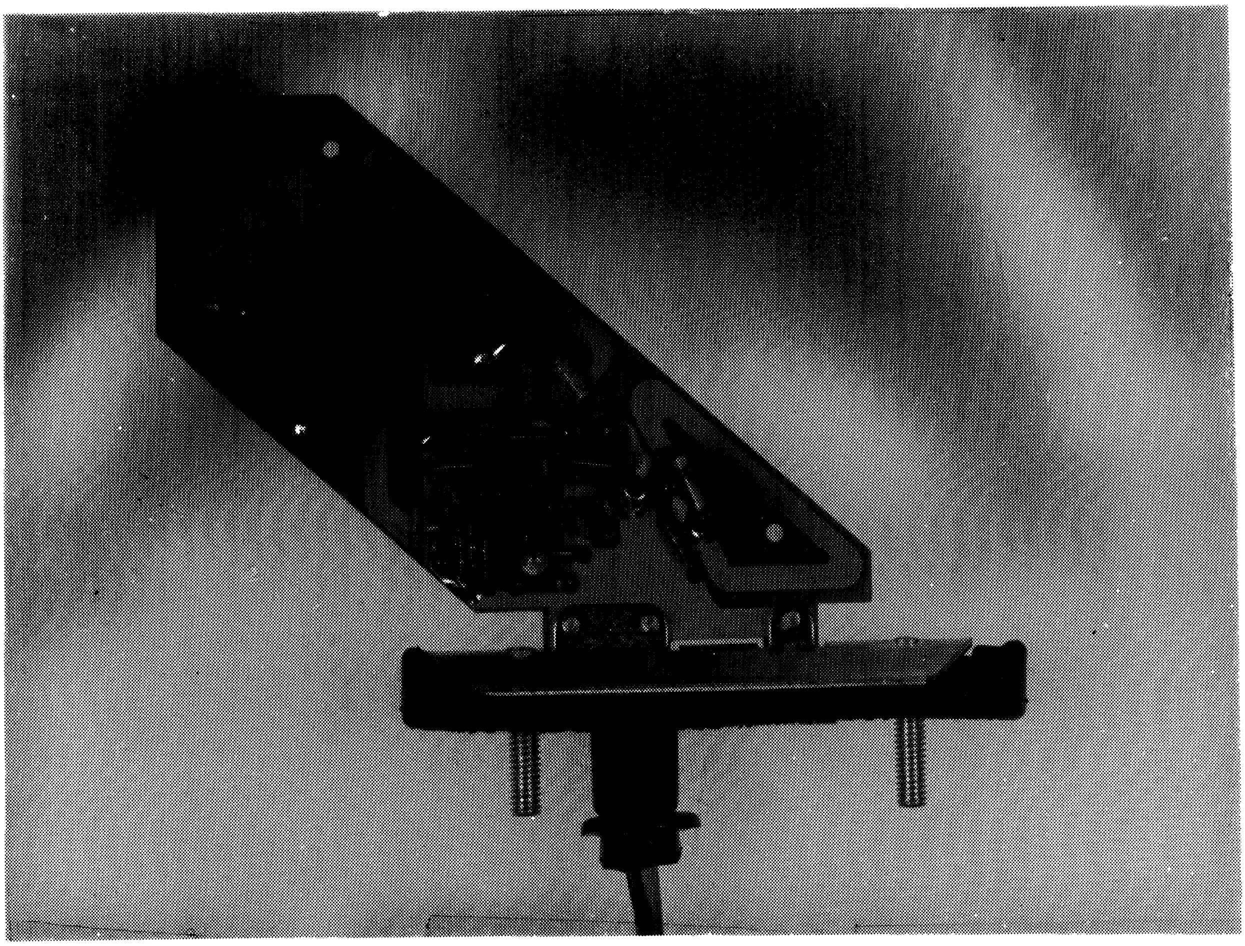

FIG. 5-16: Meinke Antenna - Left Side of Printed Circuitry.

The term T_1 in Eq. (5.12) is a characteristic noise temperature which describes the influence of noise mis-match on the noise temperature T_t. Examining this equation we see that the transistor noise temperature does increase as one places an impedance at the input terminal that is further and further removed from the optimum impedance. At the lower frequencies for short antennas the source impedance Z_i will give R_i and X_i terms which will be far from the optional impedance and the noise temperature will be far from the minimum. It is true low transistor noise will not be required at the low frequencies because of the external noise T_A is high, but at the lower frequencies the reactive component of the antenna impedance is very high while on the other hand X_{opt} for the transistor is minute at low frequencies. X_{opt} is nearly the conjugate impedance to the input capacitance of the transistor. At low frequencies for short antennas the transistor temperature T_t is far greater than T_{min} so we can neglect T_{min} in Eq. (51.2) and get a good approximation of transistor tmeperature by Eq. (5.13).

$$T_t = T_1 \left\{ \frac{R_{opt}}{R_i} + \frac{X_i^2}{R_i R_{opt}} \right\} \qquad (5.13)$$

Meinke reports that the half power bandwidth of an active antenna is of only minor importance but what is important is the noise bandwidth, i.e. the bandwidth of the optimum noise ratio. The theoretical limit of field strength sensitivity is given by the external noise and this noise field strength T_A. Meinke defines the noise bandwidth according to the optimum signal-to-noise ratio as in Eq. (5.14).

$$\frac{E_S}{E_A} = \geq \sqrt{2} \qquad (5.14)$$

and this criteria for low operation of the short antenna is the signal strength for a signal to noise ratio of 1 (E_S) divided by the noise field strength (E_A) must be greater than or equal to the square root of 2. This ratio of the E_S/E_A can be shown to be related to the ratio of the transistor temperature divided by the antenna temperature as shown in Eq. (5.15).

$$\frac{E_S}{E_A} = \sqrt{1 + \frac{T_T}{T_A}} \tag{5.15}$$

Equations (5.14) and (5.15) show the design criteria for the transistorized antenna AM band is related to the noise temperature and this noise temperature of the transistor is related to input impedance match as we can see from Eq. (5.13). Therefore, we see that the criteria for broadband low noise transistor antennas does include the transistor match. In other words there is a form of noise matching, although this is not as critical as it is for the higher frequencies where the noise temperature of the antenna is much lower. This high noise temperature of the antenna allows a larger mis-match between the antenna and the transistor which is essential to be able to operate over the complete AM band (a 3:1 bandwidth). Where the antenna temperature is lower in the higher frequencies broadbandwidth such as octave bandwidths do not appear to be possible at this time with the transistor antenna. The reason for this is the antenna impedance, even for the tuned antennas, varies too much over the frequency bandwidth to maintain the impedance needed in the temperature circle on the Smith chart representing the impedance essential for low noise. It is interesting to note that the antenna will be designed even in the AM band with a particular impedance for a particular transistor type even a particular transistor number. Therefore, this transistor amplifier cannot be used with another antenna and get optimum performance. The performance may be greatly reduced by placing another antenna on the transistor amplifier. Meinke goes on to show that there are three different types of transistors that can be considered in terms of their noise source equivalents. The noise source of a transistor can be represented as a voltage source and current source which are partially correlated. For high impedance transistors such as the field effect transistors the voltage source predominates and the current source may be neglected. Other types of transistors will have a current source that predominates and the voltage source may be neglected. There is also a third class of transistors in which the voltage source and the current source contribute almost equally and both of these must be considered in the design of the antenna. It is an interesting result that the field effect transistor in the transistorized antenna application gives nearly the same signal-to-noise ratio as the cheaper NPN transistors although the optional impedance is very different for the two types. This is the reason that the Meinke antenna for the rear view mirror utilizes a cheaper NPN transistor as shown in Fig. 5-17.

If the effective height of the antenna is made shorter the bandwidth generally becomes smaller. This can partially be overcome by adding artificial loss by using stripline techniques. Of course artificial losses give lower signal to noise ratios or speaking in terms of noise temperature the transistor noise temperature will become greater. Meinke found that certain amounts of artificial additional losses are necessary to optimize the integrated antenna i.e. to get a prescribed noise bandwidth with the selected low antenna height. It is obvious that very many parameters must be varied and measured to get the optimum antenna or one that will perform satisfactorily. Meinke makes the point that one who attempts to develop a transistorized antenna should not be disappointed if he is not successful within a few months; Meinke feels that the development of these antennas comes as much from experience as from theoretical knowledge.

Meinke found that by adding a feed-back to the amplifier the input impedance can be modified and this effect can be used to obtain resonance at low frequencies with

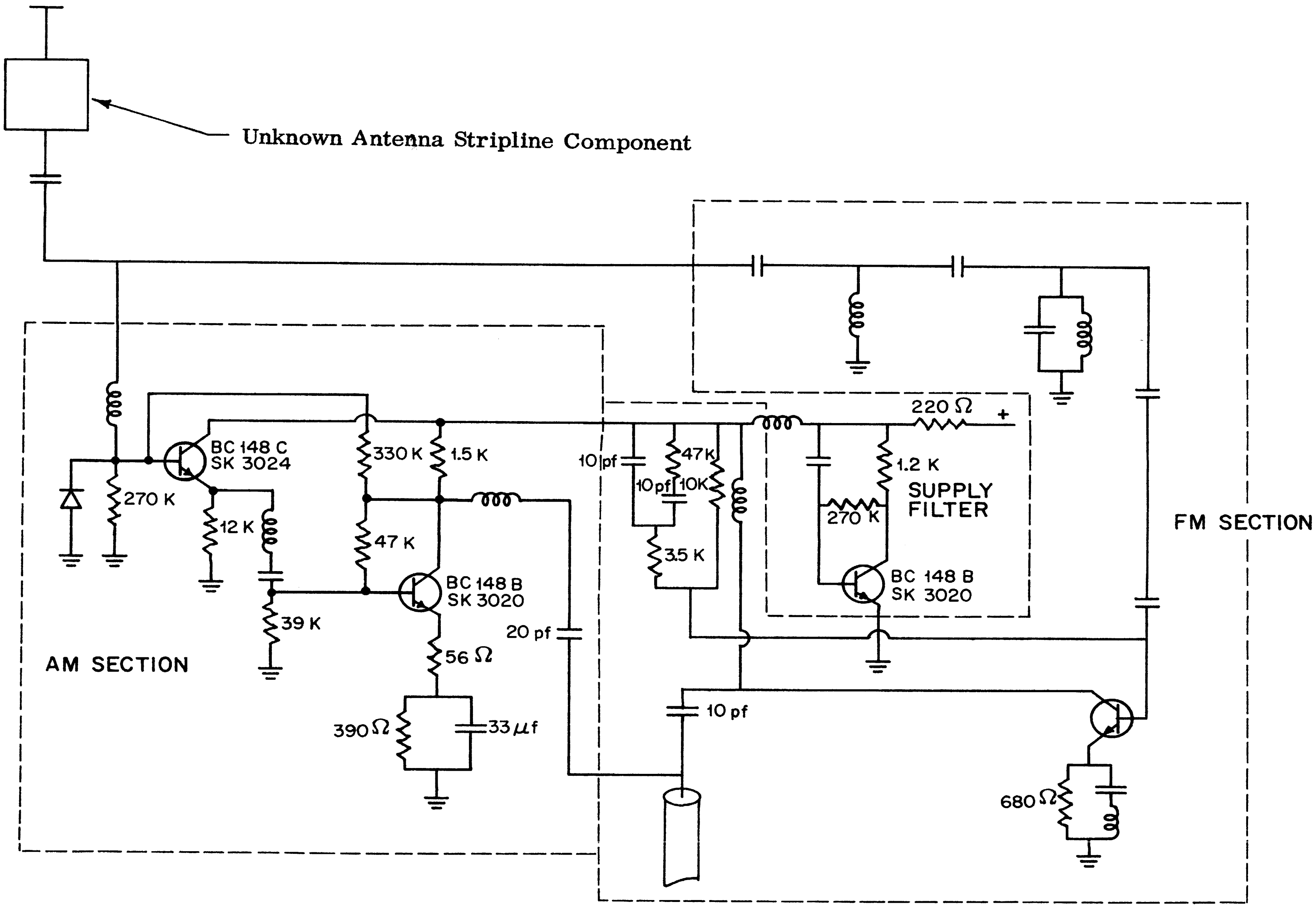

FIG. 5-17: Schematic Diagram of Meinke Rear View Mirror Hidden Antenna, (as deduced by writer).

a small inductance. It is desirable for matching the output impedances to add to another transistor so that the amplifier consists of two transistors. The noise behavior of these antennas is quite different from the noise behavior of a single antenna. The second transistor aids to the overall gain of the amplifier and effectively reduces the noise of the transistor amplifier since it can also be included in the feed-back. Exact details of the construction of such antennas is not given in the reports.

When we started on the building of a few simple transistor matching networks in January we did not have the Meinke reports to follow as a guide. Other reports we had read indicated that the noise match was not a serious problem. In fact, they stated one could ignore the noise match between the antenna and the amplifier and merely strive for a high input impedance on the amplifier by using a FET. However, the above discussion indicates this is not the case and that one still has to be quite sophisticated in design of an AM band transistorized antenna. This is not an impossible task but the design of such an antenna requires a systematic procedure where one first selects a transistor that would be desirable and then makes noise measurements on this transistor. The antenna impedance would then have to be designed to fall within the criteria for signal to noise bandwidth such that the antenna would be operating across the entire AM frequency band.

VI

FM POLARIZATION MEASUREMENTS

Polarization measurements summarized in Table VI-1 were made by the Radiation Laboratory at three different locations: 1) Willow Run area by Hangar 2, 2) the Irish Hills area near Wamplers Lake at the Hays State Park, and 3) at the Ford Motor Company proving grounds in Dearborn. The FM fields suffer from the same problems of reflection found in the discussion on the AM fields of polarization. However, the polarization is not felt to be quite as random at the FM fields as it is for the AM frequency bands. Tests especially at the Wamplers Lake area in the Hays State Park indicated that the polarization does change as a function of position and that one cannot make accurate and extremely exacting measurements on the ellipticity of a field because if one moves just a few feet these values will change. For this reason at the three different measurement sites, the major polarization is given and the notation of ellipticity will mean that the vertical and horizontal components will appear to be within 6dB of each other. These measurements were made at a test height of 48 inches which is felt to be a reasonable height to represent the car antenna. In the metropolitan area the signal is quite affected by reflections and it is not surprizing that a vertical component will be picked up for a horizontally polarized transmitted signal and at certain times and in certain places the vertical component can in fact be larger than the horizontal component radiated from the station. This is especially true in areas where there are large buildings to give quite a random polarization to the re-radiated field. It appears that in the metropolitan areas the higher power levels and the more random polarization would make the polarization of the receiving antenna less important. However, as one considers the more remote reception for the weaker signals such as in the Wampler's Lake area a trend does seem to appear in which the horizontal component is predominant even for the circularly polarized transmitted signals.

Polarization measurements at Willow Run may present some ambiguity due to reflections from Hangar 2. At first we did not realize the amount of depolarization that could occur from building reflections.

Table VI-1

Station	Frequency	Trans. Pol.	Willow Run	Dearborn Test Track	Irish Hills
TV station Jackson	88.1	H	~0°		0°
CKWW	88.7	C	90°	10-15° elliptical	
WDTR	90.9	H	0°		
WPHS	91.5	H		7°	0°
WUOM	91.7	H	16°	0°	
WCAR	92.3	V	0°	35°	
WJBK	93.1	C	90°	55° elliptical	
CKLW	93.9	C	90°	57° elliptical	0°
WHFI	94.7	C	90° ellip.	63° elliptical (slight)	
	95.1				0°
WLDM	95.5	H ✓	61°	45° elliptical	
	95.8				0°
WJR	96.3°	C	75° ellip.	75° elliptical	
WWJ	97.1	H	0°	5°	0°
WMZK	97.9	C	73°	50° elliptical (slight)	0° (elliptical)
	98.1		52°		45°
WBFG	98.7	C	90°	45° elliptical (slight)	0°
WADS	99.1		0°		0°
WABX	99.5	H		20°	0°
WKNR	100.3	H	90°	70°	60° elliptical
WXYZ	101.1	H	15°	22°	0°
WSTV (Toledo)	101.35		23°		0-28° elliptical
WDET	101.9	H	0°		50°
WBRB	102.7	C	0°	60° elliptical	
WOIB	103.2		0°		0°
WMUZ	103.5	H		14°	
Adrian	103.9				14-24° elliptical
	104.0		0°		
WOMC	104.3	C	48°	5° elliptical (slight)	0-10° elliptical
	104.8		165°		
WQRS	105.1	C	90°	45° elliptical (slight)	
WTLS (Ohio)	105.2		0°		26°
WCHD	105.9	C	0°	0°	
Jackson	106.0				0-34° elliptical
WWWW	106.7	C	90°	75°	90°
WPAG	107		44-136° ellip		37° elliptical
	107.2		55°		0°
WGPR	107.5	C		42° elliptical	

VII

CONCLUSIONS

From the studies so far the following conclusions can be made: 1) The reduced signal level inside the car body (passenger compartment) makes the location of an AM antenna in this area undesirable. This field gradient is sharp and would cause even dash-mounted antennas to operate below the level expected exterior to the car. 2) Location of antennas under the car is undesirable for the AM band antennas due to the increased field intensity of the noise fields under the car and this location is undesirable for the FM antenna due to the decreased field strength that exists underneath the car. 3) Utilizing parts of the existing car, such as an isolated trunk lid, car door or fender does not appear to be desirable as these will also suffer from an increased AM band noise due to the circulating currents and induced currents on the car due to transients from stop lights, chopped signals, and other problems of an RFI nature. Besides the problems of the RFI from the car itself these antennas will suffer from a sensitivity to surroundings such as people or luggage. 4) The best approach to the hidden antenna problem would appear to hide a conventional antenna or to blend this antenna in with the car's natural lines. This would take the form of a windshield antenna trim antenna, or an annular slot or flat plate in the trunk lid if the fiberglas trunk lids become available. 5) The electrically small rear view mirror antenna marketed by Fuba in Germany does give performance approaching that of a whip but is somewhat dependent upon atmospheric conditions and will not give the overall whip performance under all conditions. Performance is slightly degraded by wet weather causing corona on power lines and in high noise interference areas such as neon signs.

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