VHF/UHF Filters and Multicouplers
To my mentor Pitt Will Arnold, the inventor and tireless promoter of the “dynamic empiricism”.
VHF/UHF Filters and Multicouplers

Applications of Air Resonators

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“Vacuum is a physical medium capable of transporting electromagnetic actions. It is characterized by three physical constants:

1. Permittivity of the vacuum, \( \varepsilon_0 = 8,854,187 \times 10^{-12} \text{ F.m}^{-1} \) (Faraday per meter),

2. Permeability of the vacuum, \( \mu_0 = 1,256,637 \times 10^{-6} \text{ H.m}^{-1} \) (Henry per meter),

3. Speed of light in the vacuum, \( c_0 = 299,792,458 \text{ m.s}^{-1} \).

Here we have, displayed once again in clear and precise terms, the extraordinary ambiguity that contemporary physics maintains with the meaning of the word “vacuum”. The above definition can be read in the volume dedicated to physics, *l'Encyclopédie scientifique de l’univers*, edited at Gauthier-Villars in 1981 by the *Bureau des longitudes*. We cannot doubt either the seriousness or the competence of the editor or the commissioner: Gauthier-Villars is to science what *le Goncourt* is to literature. Meanwhile the *Bureau des longitudes*, created in 1795 during the Revolution and which still today remains in the shadow of Henri Poincaré, has remained an institution that is as atemporal and symbolic as the *Ecole des mines*. But how is it possible to conceive that an individual of ordinary intelligence, only a little interested in physics, would not be puzzled by such a formulation? Vacuum, in the current understanding of the term, is the absence of everything, the nothingness, nothing, unless we specify it: vacuum of air, vacuum of matter, etc. To attribute an arbitrary property to something whose existence as a physical object we deny seems therefore not only surprising, but also illogical and ultimately shocking. However, the previous definition is of such precision that doubt is not even allowed: vacuum is not void, it is a medium.

It is not a question, in a work primarily intended for engineers and that is first conceived within the scope of practical application, of reviving some intellectual or
philosophical quarrel from the past and thus mixing the problems. But it is also not forbidden, when we find ourselves before a definition of this nature, to pose questions, if only out of mere curiosity. Like this one, for example: if the vacuum is a real medium, as stated very clearly after all — and we are insistent on emphasizing this — by the cited encyclopedia, should we continue to call it the vacuum, or should we give it another name, so that we no longer continue to confuse it with the true void, the absolute void, that in which nothing exists?

The notion of ether has accompanied physicists from Descartes up to the beginning of the 20th Century, when special relativity struck the final blow. Its abandonment has made of electromagnetic waves the only vibrations that propagate without support. This singularity imposed as a simplifying hypothesis by Einstein’s theory but also admitted without too much concern by all theoretical physicists, at the time represented by Poincaré and Duhem amongst others, is one of the curiosities of modern science. Moreover, the contemporary scientific community, educators included, does not seem to be disturbed by this: physics, like other human activities, coexists with its own contradictions.

Among all the names of reference, the most important in the domain of electromagnetism is undoubtedly that of James-Clerk Maxwell. The equations that carry his name, which no-one today has the slightest reason to question, govern invisible phenomena so well that they are, for example, at the basis of the most used and recent systems of simulations. But one must remember that Maxwell, whose idea of dummy current is generally regarded as the stroke of genius that permitted the real taking-off of the theory of electromagnetism, only introduced and developed this fundamental idea in a very particular context: its conception and reasoning constantly lean on the hypothesis, often suggested in his main work but never explicitly mentioned, that electromagnetic phenomena were in reality the visible manifestation of mechanical events in the conventional sense of the term, intervening in a fluid just as real as it is inaccessible to our senses.

Meanwhile, we have been able to note that the works of Maxwell, with the exception of the small group of fundamental equations that link the electric and magnetic fields and their corresponding inductions, are almost ignored, at least in their detail, by all those who use them either directly or implicitly, be they engineers, technicians or students, or even specialized lecturers. It is not very difficult today though to find a reprint of the *Treatise on Electricity and Magnetism* for an affordable sum, but it is still necessary to devote time and interest to the subject. Its reading, for those who are passionate about physics, is exciting. It provides access to the sources of the author’s inspiration, to the foundations of his intuition and allows us to better follow his reasoning. An entirely Cartesian prudence led him to give to his main work the mathematical presentation that we know and which it was recommended he respect, in order not to clash with the
scientific establishment. But some justifying allusions, here and there, leave no doubt to those who can read in-between the lines: ether is its major inspiring idea, the permanent backdrop and necessary support.

Along these lines we can also mention that there exists, in addition to this fundamental work, *An Elementary Treatise* by the same author, edited almost at the same time (1884), but even less well-known than his greater treatise. It contains a preface by Professor Garnett, from the University of Nottingham, where the Maxwellian conception of ether is explained in detail. Though no-one is obliged to share these views entirely, they shed light on the fundamental role that the hypothesis of the existence of such a medium played in the great physicist’s approach. This book is unfortunately very difficult to find in the French language, and it would maybe be a good thing if an inspired book shop studied the profitability of its reprinting, for the sake of the interested reader.

Maxwell had a boundless admiration for Faraday, whom he considered to be a master to his reasoning. The concept of lines of force, intimately related to that of ether, was transmitted in a natural way from one to the other, all the more easily given that both scholars, being contemporaries, also held each other in mutual esteem and maintained correspondence. This concept, reused a century later, has allowed us to tackle and solve with striking simplicity a significant number of practical problems, such as the exact determination of the electric length of quarter-wave resonators, or the development of new procedures such as multiple loops or secondary loops in combline filters. We used for that a mechanical model of these filters in which we gave to the so-called “vacuum” insulator a physical reality, by considering it as a fluid of an extremely large volumic mass and of a very small, albeit non-zero, compressibility.

In spite of everything, whatever we might think of a question constantly quashed over the centuries, it is forbidden today to speak of ether: it is a problem officially settled once and for all, a thing of the past, an obsolete and outdated notion of space that can only be of occasional interest to physics historians or some ill-inspired student of philosophy. Misfortune befalls those who dare pronounce its cursed name. But this does not prevent the “relativists”, quite troubled nowadays by this desperately void space about which they can do nothing, from hypocritically repopulating it with bizarre objects, in a way that only mathematical physicists know how to do, and to restructure it by recreating what was destroyed and giving back to it the beginning of a second life by means of a completely abstract, but consistent modeling.

There are two types of physics: theoretical physics and rational physics. Theoretical physics is the official one, the mathematical physics, that which is taught. It works through a systematic mathematical modeling of the phenomenon we
propose to study, and its best definition was given by its practitioner and most intransigent promoter, Pierre Duhem, in Physical Theory: “a physical theory is not an explanation; it is a system of mathematical propositions, deducted from a small number of principles, whose aim is to represent as simply, completely, and as exactly as possible a whole group of experimental laws”. Rational physics, meanwhile, is based on reasoning, correlations and the systematic search for causalities. It is not taught, it is an “unofficial” physics that is exerted in a natural and instinctive way, outside academia, but in industry, when the question is to understand a phenomenon whose modeling does not succeed in penetrating its mystery. It explains, or at least aims to explain, by extensive use of analogies. Depending on necessity, the engineer has the right, in the name of efficiency, to make a choice, which moreover can be a harmonious mixture of both.

It is perhaps in the definition of a wave that the difference between the two types of physics is made most clear and is best demonstrated. A theoretical physicist would say: “a wave is a physical quantity (scalar or vectorial), which depends on spatial and temporal coordinates and which is a solution to a partial differential equation called wave equation”. A rational physicist would say: “a wave is a deformation that propagates in a medium”. This definition evidently implies that it is not possible to have a wave without a medium of propagation. It is, however, a problem that will not be likely to trouble anyone but those theoreticians who have not renounced the philosophy of their science. The others, meanwhile, will evade this question, no matter how important it is, by means of formulas of which we will be able to appreciate the extreme caution (or the extraordinary hypocrisy), such as this one, found in a second-year electromagnetism exercise book: “a material medium is necessary to transmit mechanical waves (for example, sound cannot propagate in vacuum). On the other hand, two physical systems can interact at a distance, in another way than an elastic process between particles. Example: the interaction between radiation and matter by means of electromagnetic waves that can propagate in vacuum”. This is how we get out of an embarrassing position.

It seemed of interest to us, along these lines, to slightly come off the beaten track of purely mathematical expositions, and to offer the reader additional possibilities to reflect on and alternative points of view, other images, which they will be able to use as they wish: either engaging with them or leaving them aside. In the next pages it should therefore not come as a surprise if from time to time we come across the rational physics to which we are so attached. The work is carefully presented, from this point of view, in such a way that no criticisms of the theoretical physics will be found, not even veiled, since it has demonstrated its effectiveness and (one must recognize this) has lead us to wonderful discoveries, providing engineers with what they demand as a priority: the mathematical formulas that allow them to efficiently exercise their art. But we have been able to notice, after 40 years spent in the industry, that the mathematical presentation of electromagnetism is far from
satisfying everyone, and that something else other than formulas is necessary for the understanding of the working of, for example, a resonant cavity or a band-pass filter: a good analogy is sometimes better than a page of calculations. This is something the French Minister Lang would probably not have contradicted when he spoke of his desire to fight against the “dictatorship of mathematics”. Rational physics will thus be present, but discreetly. It uses similarities with abundance and exists in a complementary manner alongside theoretical physics. It is, in our view, not only the physics of engineers, but also the physics of true physicists, of those who are not satisfied with models and the rigor of numbers but who want to know the reason why and how things work; they cannot be satisfied by Einstein’s universe. Nature, to quote Pascal, abhors a void. It would be good to return someday to the fundamental questions. We are convinced that, revived as it should be and suitably revisited, ether, like the phoenix, will be reborn from its ashes. It has a bright future ahead of it, as does physics.
We can find on the roofs of certain office buildings veritable forests of antennas, sometimes so close to each other that we could ask ourselves how they can fulfill their role properly. We should not forget, in effect, that the radiation pattern of an antenna is plotted on specialized sites, with sufficient ground extension and generally outside cities, where the nearby conductive objects that are disruptive to propagation are carefully avoided. This means that the operating conditions, which are always different from the measuring ones, mean an antenna almost never radiates in the expected way. Hence the necessity, in order to try to remedy this proliferation at a time when public space has become a gigantic telephone cabin, of possessing weapons that allow us to fight against this phenomenon related to the development of modern telecommunication. One of these weapons is multicoupling.

Multicoupling is the group of techniques that allow us to connect several transmitters or several receivers, or often both, to the same aerial. It is thus an activity that concerns anything that puts a radio link in operation, in the most generic sense of the term: broadcast, radiotelephone, frequency modulation, DAB, military radiocommunications, satellite links, etc. It is present in network infrastructure and is consequently a technology and an activity reserved for professionals and completely ignored by the general public. On the other hand, all radio engineers, as well as all technicians in the field, today need to have knowledge of at least the basics.

Although probably born in the military domain like other branches of the high-frequencies technique, this particular application has seen its major development in the civil domain: television, initially, and then frequency modulation, but most of all, as of 1985, the mobile telephone. Behind a somewhat obscure name is thus hidden an activity that, although unknown, is today indispensable in the HF domain. Let us take as an example a TV station, which is usually presented under the appearance of a tower or a mast positioned at height. This emitter has to radiate over
a theoretically circular zone, the radius of which covers dozens or even hundreds of kilometers, with a certain number of frequencies corresponding to a certain number of channels. The powers involved are enormous, of the order of several dozen kW, and the transmission is omnidirectional, which demands an aerial of large dimensions composed of several basic antennas regularly distributed all around the radiating source. Due to its size and cost, this aerial is thus unique, but it must transmit all the transmitted frequencies. It is consequently indispensable that there exists, inside the building, a device that guides towards the exterior radiating system the signals from the group of transmitters in such a way that none of them is perturbed by the others. This device is called a transmitter multicoupler.

From a theoretical point of view, the concept of combiners, which is another name given to this same object, is a direct application from third-year university courses on microwaves. It makes use of the theory of lines, laws of propagation, resonance, resonator coupling, microstrip lines, and, generally speaking, all those basic concepts of the domain that until recently were still called microwave frequencies. From a technological point of view, it makes use of associated components such as resonant cavities, directional couplers, coaxial cables, matched loads, low-noise amplifiers, Wilkinson splitters, etc. For a time, radio was not yet part of the commercial target. Things have changed significantly since then.

The first French mobile radiotelephony system worthy of this name was called Radiocom 2000 and it saw the light of day in 1985, in terms of its deployment and infrastructure – that is, everything that the user does not see. The goal to be achieved was that of covering the whole country with omnidirectional cells, laid out in such a way as to ensure total national coverage. A little more than 600 base and relay stations were thus installed. Each station routinely comprised 8 to 24 analog radio channels, each equipped with a 50 W transmitter, the group of emitters being connected to a unique antenna by way of coupling racks. There was even an attempt to couple up to 64 emission bays, but the overvoltages generated lead to incompatible filter dimensions. The receivers were likewise connected to a single antenna. With the conversations being held in duplex, there were the same number of receivers as there were transmitters, each transmitting frequency being separated from the associated receiving frequency by a constant quantity called “duplex separation”, in accordance with its function. The guiding principle of the network was to ensure maximum coverage for each cell in such a way that a moving device (at the time these were essentially vehicles) could keep its line for long enough to guarantee the continuity of a conversation of standard duration, knowing there would be a disruption when the mobile reached the limit of the cell and that it would be necessary to restart the call on a different frequency upon entering the neighboring cell. One of the conditions associated with these conception principles was that radio stations should be located at high points, owned at the same time by TDF (French television channel) and the water distribution companies, who thus
were able to start getting involved in the business of radiotelephony. To return to the subject of multicoupling, each station was first set with two omnidirectional antennas: one for transmitting and the other for receiving. But the aforementioned constraints, which were rapidly added to by the appearance of Nokia’s rival network system, meant that there was a rapid switch to the single antenna, and thus to the production of power duplexer which thus became part of the standard range.

Today, the mobile telephone has become portable and no longer resembles its original ancestor: in place of a device weighing several kilos, which a token handle allowed it to hypocritically be called “portable”, we now have a handset small enough to fit in our pocket. We have increased in frequency, analog has given way to digital, antennas are sector coverage instead of omnidirectional, we change from one cell to another without noticing, we can call and be called from anywhere... At the base stations however, multicoupling, which remains indispensable, has not undergone any major changes. Everything is smaller, since powers are weaker and the channels of a higher frequency, and everything is cheaper, but we continue to use the same principles and the same devices: coaxial cavities and hybrid couplers for transmitter couplers, and filters, splitters and low-noise amplifiers for receiver couplers.

Inside the south pillar of the Eiffel Tower, inaccessible to the public and today enclosed by an enormous bunker that Louis de Broglie himself, prestigious tenant of the place in wartime, would not recognize, we can discover, should we be the happy bearer of the right to enter, an impressive display of the main broadcast techniques. There we have an extraordinary condensed summary of radio in its broad sense: TV transmitters, free-radio emitters and radio-telephone stations, controls, labs, etc. Each family possesses its own coupling system. The powers per channel are organized according to their use, from 50W to 50 kW, and the corresponding cavities have a volume that goes from about 5 dm³ for radio-telephony up to about 0.25 m³ for broadcasting. Low-loss coaxial feeders transport the signals three floors higher, to the top of the Tower, where relatively discrete antennas do not give the general public reason to suspect what exists in the underground. In fact, whenever we spot an antenna on the roof of a technical building we can be almost certain that below there exists a coupling device managing a sometimes unrealistic number of transmitters and receivers.

Nowadays, the systematic usage of digital has allowed us to perform all of the filtering of mobiles by means of direct signal engineering. Concretely, this means that there are no longer duplexer in a mobile, which along with progress made in battery technology, has made it possible to give them the dimensions by which we know them today. The multicoupling techniques are no longer applied generally or in radiotelephony other than in fixed installations.
After this glance at the main domains of application for multicoupling, it remains for us to mention, in an attempt to be a little more complete (despite being certain of not being exhaustive), the case of mobile devices of large dimensions, such as warships or trains, in which we also find multicouplers, or the locations where radiocommunications are indispensable, such as airports, surveillance sites, satellites, etc. We thus hope to have a sufficient panorama of the activity in order to correctly establish its framework that is in fact very extensive.

In this guise, a remark should be made regarding the simultaneous evolution of surveying methods. In the high-frequency domain there exists a measuring device that has long become indispensable to laboratory personnel, which is to HF specialists what scales are to chemists: the network analyzer. In comparison with the referential components of a “calibration kit”, this appliance directly provides the value of the S parameters of a measuring device, that is, its identity card, as it were. It is the performances of the analyzer that in fact leads to the values demanded by the specifications of the products which we use in multicoupling: “we can only specify that which is measurable”. An immediate consequence of this inevitable rule is that the specifications harden as and to the extent that the accuracy of the measuring equipments increases: in 1985, passive-filter manufacturers were required to produce adaptations of 15 dB, whereas today we have moved to 23, then 26, or even 30 dB, essentially because we can now measure them, more than for any other real necessity. And yet, accurately measuring a 26 dB return loss not only requires an analyzer of the latest generation, but a set of procedures, precautions and tips that unfortunately few people, even amongst professionals, know and systematically apply. This is the reason why an entire chapter will be dedicated to this problem that often poisons relations between the ordering party and manufacturers, due to controversy related to disagreeing measurements. Along these lines, we have discovered that there exists, in the large family of the radio and high-frequencies, a new illness that we could call “dB mania”. In the hushed and protective environment of major operating companies, in the comfortable offices where engineers work who are responsible for writing the specifications of the systems about which we are talking, strange things happen. Young or not so young, a certain number of them, and they are hardly to blame, have no idea, through lack of practice, what one modification or another of a technical specification, which they consider as minor, can represent for the manufacturer. Certain people handle decibels without any qualms, with the vigor of a lumberjack and the serenity of a Buddhist, and without the least consideration for the producer; and since it is often the case that the relationship between the ordering party and the contractors is of the “master and slave” type, it is always rather difficult to make the former understand where the limit is between the easy, and therefore cheap, and the difficult or the impossible. What exactly is meant by a 160 dB TX-RX isolation? Does it lack 10 dB to equilibrate the link balance? Not a problem, we change the specification: that will prevent suppliers from resting on their laurels, assuming they have laurels!
A little before 1900, there was a delusion of grandeur in France. On the occasion of the *Exposition Universelle* in Paris, we decided to build an astronomical telescope that was supposed to allow visitors to see the Moon as if they were a meter away from it. In fact Deloncle, the deputy of the Basses-Alpes, protractor of the project with the National Assembly, had only evoked a magnification of 5 or 10,000, figures put forward by the experts, but it was too late and the journalists had already gone about their work of amplification. We thus constructed an enormous piece of equipment, at great cost, with a steel tube 60 m long and 2 m in diameter, a Foucault siderostat of 70 tons whose mirror alone weighed 4 tons, the whole being sheltered in a building equipped with a passageway where the public was to discover the unknown details of our satellite projected onto a 50 m² screen. In fact, visitors contented themselves with a pale image of about a meter and a half in diameter, which was already miraculous. The whole was dismantled, recast, and today nothing remains but two silverless mirrors, which lie at the Paris Observatory. Forgive us the anecdote and let us come to what is of interest: approximating the Moon to the equivalent of a meter represents a linear magnification of 172 dB, or a quadratic magnification of 86 dB. What is the relationship with decibels? Today we can find on the market, for less than €1000, UHF duplexers of a few dm³ whose attenuation from one band to another is above 90 dB with a duplex separation equal to the passing band. In the case of our duplexer, this means that only a thousand millionth of the emission signal arrives at the receiving channel: it is like, in terms of decibels and to come back to the preceding comparison, seeing the Moon from 40 cm away! Of course the comparison is subject to discussion; but we are not creating a thesis, it is simply intended, by speaking of the Moon, to have us return to Earth so to speak, trying to provide a clearer image of what a power ration in dB represents.

Using the decibel flattens the real scalar values. We can moreover say that it was defined, to a great extent, for this reason. But its generalized use ends up dangerously distancing us from the notion of the orders of magnitude involved, and if we are not attentive enough, in us losing contact with the reality of things to the point of not exactly knowing what the numbers we write truly represent. And this is precisely what happens in the territory about which we are talking. It is harmful to everyone for it is a source of incomprehension and pointless discussions. We would be very satisfied if this book could contribute to encouraging engineers and technicians to take up the good habit (but do they not already have it?) of recycling themselves from time to time, and to never forget, when exercising their profession, to ask themselves about the consequences involved in changing, in a specification, the value of one quantity or another evaluated in dB. We feel supported in this sincere wish by all filter and HF equipment manufacturers.
Chapter 1

Reminders and General Points

1.1. Lines

We use the term “line” to refer to all systems of two conductors capable of propagating a high-frequency signal. There are two types of line theorized: the bifilar line, formed of two parallel conductors, and the coaxial line. The microstrip line is connected to one or the other depending on the problem to be solved. The waveguide, which is a high-pass device (whereas lines are low-pass), can be considered as the transition between the line and the free propagation.

1.1.1. Bifilar lines

![Figure 1.1.](image-url)

Figure 1.1.
We use these lines little in the industry because they radiate a lot. Their practical application is limited to the fabrication of FM antennas, for example, where this characteristic becomes an advantage. From a theoretical point of view, on the other hand, it is the model we use in lessons to teach the Telegraphists’ equation and the rest of the theory. The line is the typical example of a circuit with distributed constants. Each element of length $dx$ is electrically characterized by four basic quantities, a serial resistance $dR = R \cdot dx$, a parallel resistance $d\text{Rp} = \text{Rp} \cdot dx$, a serial inductance $dL = L \cdot dx$ and a parallel capacitance $dC = C \cdot dx$. Since it is a low-pass structure, a HF line possesses a high cut-off frequency.

Telegraphists’ equation: if we do not take into consideration the ohmic losses represented by $Rs$ and $Rp$, we have the D’Alembert equation:

$$\frac{\partial^2 V}{\partial x^2} = LC \cdot \frac{\partial^2 V}{\partial t^2}$$  \hspace{1cm} [1.1]

with

$$c = \frac{1}{\sqrt{LC}}$$  \hspace{1cm} [1.2]

and

$$Z_0 = \frac{L}{\sqrt{LC}}$$  \hspace{1cm} [1.3]

1.1.2. Coaxial lines

Unlike those mentioned above, these are a device of universal usage in HF in the form of coaxial cables. It is a line that radiates very little, where the external conductor plays the role of an insulator, consisting of either a single or a double
braid, or of a cylindrical, continuous (semi-rigid) layer. The industry produces a large number of adapted connectors that enable the creation of all the cable types that we need. The formulas that one must know are given below.

**Characteristic impedance:**
\[
Z_0 = \frac{1}{\sqrt{\frac{2\pi}{a} \log \frac{b}{a}}} \quad \text{with } \mu = \mu_0 \text{ and } \varepsilon = \varepsilon_0 \varepsilon_r \quad [1.4]
\]

**Linear capacitance:**
\[
C = \frac{2\pi \varepsilon}{\log \frac{b}{a}} \quad [1.5]
\]

**Linear inductance:**
\[
L = \frac{\mu}{2\pi} \log \frac{b}{a} \quad [1.6]
\]

**Serial linear resistance:**
\[
R = \frac{R_s}{2\pi} \left( \frac{1}{a} + \frac{1}{b} \right) \quad [1.7]
\]

In this expression \(R_s\) designates the surface resistance, \(R_s = \frac{1}{\sigma\delta}\), where \(\sigma\) is the conductivity of the metal that constitutes the two conductors of the coaxial, and \(\delta\) the skin depth: \(\delta = \sqrt{\frac{2}{\omega \mu \sigma}}\). Let us recall that the skin depth is the depth of the conductor where the field is reduced to \(\frac{1}{e}\), that is, to about 37\% of its surface value \((e=2.71828)\). This means, for example, that a depth of 3\(\delta\) allows 95\% of the field to pass through, or indeed that a metallic layer of 5\(\delta\) is needed to allow 99\% of the power to pass.

**Parallel linear resistance:**
\[
R_p = \frac{\log \frac{b}{a}}{2\pi \sigma_d} \quad [1.8]
\]

In this last expression \(\sigma_d\) represents the dielectric conductivity, which is supposed imperfect. It is expressed, as \(\sigma\), in ohms/m.
1.2. Adaptation and stationary waves

![Diagram of VHF/UHF Filters and Multicouplers](image)

We say that there is a perfect adaptation when the transmitted power is maximum, which means that there are no reflected waves. In the general case, when the line length is large compared to the wavelength, and for a characteristic load impedance $Z_t$, there is the creation of a reflected wave at this load that combines with the incident wave to produce a stationary regime. We then note that the measured value for the voltage (or the current) varies sinusoidally when we move along the line, with a succession of maxima and minima. The distance separating two successive extremes is $\lambda/2$.

If we call the two extreme values of the measured voltage $V_{\text{max}}$ and $V_{\text{min}}$, we have:

$$S = \frac{V_{\text{max}}}{V_{\text{min}}} = \text{standing wave ratio (SWR)} \geq 1$$  \[1.9\]

The expression VSWR (voltage standing wave ratio), is also used as an equivalent name for this quantity.

It is the value of $Z_t$ in relation to $Z_0$ that will determine the proportion of reflected wave, hence the following definitions:

$$z_t = \frac{Z_t}{Z_0}$$  \[1.10\]

$$\Gamma = \frac{z - 1}{z + 1} = \frac{S - 1}{S + 1}$$  \[1.11\]
This relationship is valid anywhere on the line. At the ends, \( z = z_t \) and \( \Gamma = \Gamma_t \). If the impedance at the end of the line is a pure resistance, which is generally the case in practical applications, it becomes:

\[
\gamma = \left| \frac{1-r}{1+r} \right|, \text{ with } \gamma = |\Gamma| \leq 1 \]  
\[[1.12]\]

In this expression, \( r \) is the reduced value \( R/R_0 \).

Standing wave ratio:

The definition was given above, but it is of interest to link it to the reflection coefficient, which is a more easily accessible value:

\[
S = \frac{1+\gamma}{1-\gamma} = \text{VSWR} \]  
\[[1.13]\]
Likewise in power: \[ S_p = \frac{1 + \gamma^2}{1 - \gamma^2} = \text{PSWR} \] \[ \text{[1.14]} \]

1.3. Smith chart

Derived from the system of Cartesian coordinates by a conformal transformation, that is, that conserves the angles of intersection, it is by far the most used chart in HF because it very much takes into account the periodicity of the electromagnetic phenomena along a line. One lap of the chart corresponds to a distance of \( \lambda/2 \) on the line. The perpendicular lines \( R = \text{Const} \) and \( X = \text{Const} \) in Cartesian coordinates are transformed into families of orthogonal circles.

Figure 1.4 recalls the appearance and essential characteristics of the chart, for information purposes. In fact we will use it hereafter in a very simplified form, represented below (Figure 1.5), but that will suffice to resolve the problems related to the length of lines, during harness calculations, for example.

The Smith chart will give us a graphical representation of the reflected power and its phase, but only that, and when one must characterize a quadrupole completely, it will be necessary to associate an amplitude response curve to it. The two diagrams are given by the network analyzer. The reflection coefficient is represented by the vector \( \Gamma \), which rotates clockwise towards the generator and in the trigonometric direction towards the load. The circle annotated \( \rho = \sqrt{2}/2 \) corresponds to a reflected power equal to half the incident power, in the same way that in transmission we define the 3 dB passing band that corresponds to it.

1.4. Power in a line

When the internal impedance of the generator is equal to that of the line, we say that it is adapted, based on what was defined previously. The same definition applies for the load. In almost every application dealt with in this work, we will consider it to be as such.

In addition, the impedance in question will be supposed to be a pure resistance: \( Z_0 = R_0 = 50 \) or \( 75 \, \Omega \), according to the adopted standard. In the adaptation conditions, the generator can deliver a maximum power \( P_0 = \frac{\gamma^2}{2Z_0} \) to the load. If the load is not adapted, it is characterized by a reflection coefficient of module \( \gamma \).
We then define the following values, $V$ being the voltage of the generator in vacuum, and the line being assumed to be without losses:

- **Incident power** \( P_i = \frac{V^2}{2Z_0} \) or \( 10 \log P_i \) in dB \[1.15\]

- **Transmitted power** \( P_t = \frac{V^2}{2Z_0} (1 - \gamma^2) \) or \( 10 \log P_t \) in dB \[1.16\]

- **Reflected power** \( P_r = \frac{V^2}{2Z_0} \gamma^2 \) or \( 10 \log P_r \) in dB \[1.17\]

When we speak of transmitted power, it is implied that it transmitted by the line and to the load. If this load is a quadrupole, one must not confuse it with the power transmitted by this quadrupole, which generally dissipates a certain power \( P_d \) and thus transmits a total power \( P = P_i - P_r - P_d \). We may point out that this relation, not surprisingly, is simply a particular formulation of the principle of energy conservation.

### 1.5. Line sections

We always assume, in this section, that the lines are without losses. The module of the reflection coefficient is constant along the whole line, and the phase varies linearly with the abscissa. In the chart, this translates into the fact that the vector \( \overrightarrow{\Gamma} \), which represents the load impedance, has a constant length, and so its end describes a circle centered in \( O \) when we traverse the line. The origin of the phases itself corresponds to a specific place in the line that we arbitrarily choose according to the problem to be solved, and that could be, in particular, the plane of connection to the load, that is, \( \overrightarrow{\Gamma} \) on the chart (this plane, or this place, must be geometrically defined with the greatest precision). If we define the distance that separates a point in the line from the load as \( l \), \( Z_x \) the impedance returned to this point by the load and \( Z_t \) the impedance of the load, we have:

\[
Z_x = \frac{Z_t + j \beta Z_0}{1 + j \frac{Z_0}{\beta} \frac{l}{c}} \frac{l}{l} \text{, where } \beta = \frac{\omega}{c} = \frac{2\pi}{\lambda} \]

\[1.18\]
A line section is then an impedance transformer. If we set to $l$ the two noteworthy values $\frac{\lambda}{4}$ and $\frac{\lambda}{2}$, $Z_x$ takes the values $\frac{1}{Z_r}$ and $Z_r$ respectively.

The quarter-wave is an impedance reverser, whereas the half-wave is transparent, and does not change the impedance. Particularly regarding the quarter-wave, in practice we will abundantly use its property of changing a short-circuit into an open circuit, and vice-versa. We will verify, on this occasion, the great utility of the Smith chart to graphically solve this type of problem: to have the transform of an open circuit by a quarter-wave, for example, we start from point A (Figure 1.5, $Z_\infty$), and we describe a half great-circle ($\rho = 1$) passing through D (towards the generator), arriving at C where $Z = 0$. 

![Figure 1.5.](image-url)
1.6. Lines with losses

The coaxial cables used in practice are not perfect. This is the reason why, in addition to their characteristic impedance, the constructor also specifies their linear loss. This is a factor that can become very important when it is a case, for example, of transporting strong powers towards an aerial located at great height, and thus far from the source. Despite that, in the majority of projects, we prefer to start by considering that they have no losses, which simplifies the models considerably, provided that corrections are applied afterwards.

In the case of a line with losses, the reflection coefficient is no longer a constant module, which translates on the chart as the fact that the end of the vector no longer describes a circle but a logarithmic spiral that winds towards the centre as one goes along it and as we distance ourselves from the load, starting from a certain value determined by its impedance. Thus, there exists the implicit possibility of adapting a wrong load by a loss-cable that is sufficiently long. We use this property to protect an amplifier in a set-up, or to manufacture loads of linear power, that is, with a low level of intermodulation.
Chapter 2

Measurements in HF

2.1. Material

No HF laboratory can exist without a network analyzer. All technicians have their own; it is the characteristic and personal device of the profession and it is indispensable, though it cannot do everything itself because it works at low level. Once a prototype is developed, at some point its power behavior must be studied to verify that it satisfies the specifications. This is the role of the power bench, which itself is also an essential element of the laboratory. Given its cost, its usage restrictions and working rate, the power bench forms a part of the laboratory’s collective equipment, and constitutes its most significant investment. When possible, it is located inside a Faraday cage, in order to shield the equipment to be tested from the external radiation of a place usually extraordinarily polluted by radiocommunication emitters in rampant proliferation.

On the other hand, we are often led to possess several benches in order to be able to treat multiple cases simultaneously at different frequency bands.

2.2. The power bench

2.2.1. The measurements to carry out

The measurements performed at the bench may be separated into two main categories: on the one hand, those whose goal is to verify the correct sizing of the prototype and thus its capacity to work for a long time at maximum power, and on the other, those which make it possible to ensure that it will not pollute the hertzian space beyond the limits authorized by the regulatory bodies, the CCIR for example.
The following is a standard list:
– insertion loss,
– VSWR in,
– VSWR out,
– thermal frequency drift,
– third-order intermodulations,
– higher-order intermodulations,
– harmonics,
– attenuation (or protection) outside the usage band,
– noise factor.

Performing each of these measurements requires a given number of devices, some of which, such as the power amplifiers, are of relatively narrow bandwidth for technological reasons, which requires us to have a sizeable range should we want to treat all the VHF s or the UHFs, for example: an HF laboratory is expensive.

The indispensable ones are the following:
– signal generator (2 per bench),
– power amplifiers (typically 50 to 100 W) (2 per bench),
– dual-channel wattmeter (Power Meter),
– spectrum analyzer,
– noise meter.

To this electronic measuring equipment we must add some accessories that, despite being passive, are no less necessary than those just mentioned:
– band-pass filters (anti-harmonics),
– low-pass filters (anti-harmonics),
– circulators (valve, amplifier protection),
– power chargers,
– power attenuators,
– connectors (cables, tees, shorts, etc.).
2.2.2. Bench configuration

There is no real standard configuration capable of taking all the measurements. Figure 2.1 shows a fairly complete conventional layout, except that we have not represented the two-way coupling device before the DUT. It can be used for several measurements involving two frequencies, such as, for example, that of intermodulations. The latter form part of spurious lines, which are undesirable, but inevitable. The designer of the device under test (DUT), can only be limited – and this is already very difficult at times – to maintaining the level of these spurious lines below the specified limits. But it is essential at the outset to repair those which are due to the set-up itself, in order to exclude them from the balance-sheet when interpreting the final result. It is for this reason that we find elements in this device that are not theoretically part of the basic diagram, – which could be simpler in functional terms – but which are practically indispensible: these are the circulators, mounted as insulators, and the filters. The role of these passive accessories is usually multiple.

The band-pass filter, set according to the frequency of the generator, reduces the noise around the carrier and thus increases the dynamic of the measurements. Furthermore, it suppresses the even-numbered harmonics. The filter band-pass suppresses all the harmonics, and in particular, therefore, the odd harmonics that pass through the band-pass.

The two circulators absorb in their loads everything that is reflected by the filters and the DUT. For this reason, in the jargon we apply the term “dust-bin” to them. In addition, they protect the power amplifier against incorrect maneuvers, which happens frequently when, to give an example often experienced in the day-to-day, the operator forgets to reconnect a cable downstream after a change in configuration.
For this reason the loads of 1-b and of 2-b must be dimensioned so as to be able to dissipate the maximum power from the amplifier.

The directional couplers C1, C2 and C3 enable the different measuring devices to be connected to the relevant spots. C1 and C2 are dedicated to the incident power per line, and they can be bidirectional, as C3 if we also want to evaluate the reflected power (measurement of the DUT entry return loss at nominal power). The coupling is of the order of 20 dB and the directivity of 30 dB.

All these elements must be widely dimensioned, so as not to create spurious lines themselves: the levels at which the “spurious” measurements are made in radio-telephony are such, nowadays, that no device can be assumed to be linear anymore. Duplexers in particular are, in this regard, the object of attentive supervision. The band-pass filters are generally constituted by single or double coaxial cavities. We have not represented them in the common part that goes from the DUT to the load, so as not to overload the diagram, but in practice we are often obliged to use them. In fact, the practical rule is to start by the simplest possible set-up, and then to complete it according to the problems that come up. It is worth saying, in fact, that the more components there are, the more possibilities there are for the bench to create its own harmful elements due to the large number of connections alone. Figure 2.2 shows one of the most simplest apparatuses, which may be used for a single measurement of power insertion loss, in which we have replaced the load with an attenuator.

2.3. Measurements on the network analyzer

2.3.1. Measurement principles

There are two major families of network analyzers: the vectorial and the scalar. The latter give values on transmission and on reflection without evaluating the phase: no Smith chart, these are economic models that simply enable us to trace the amplitude/frequency response curves. The vectorials are the only ones capable of completely characterizing a quadrupole and of providing readings of quantities
dependent on the phase, such as the group propagation time, for example. Both types are summarized today.

In a vectorial analyzer, three directive couplers placed in the measurement circuit, inside the device, will separately $P_i$, $P_r$ and $P_t$, with phase information which will allow the software program to deduce $\Gamma$, the insertion losses and the return loss (return loss in if we are connected to the entry, return loss out if we are connect to the exit). The device is represented in Figure 2.3.

The analyzer is pre-programmed regarding the losses and the return loss, but it is absolutely necessary to calibrate it before every measurement using the indispensible complement that is the calibration kit. This is composed of four elements, in its minimum version:

- a calibration load,
- a parallel short-circuit (short),
- an open circuit (open),
- a serial short-circuit (through).

The calibration is thus performed in four steps, in any order. The analyzer is equipped with two flexible cables, in general in RG 214 with connectors N, and it is this that we are going to suppose from now on, except when stated otherwise; and likewise, we will suppose that $Z_0 = 50 \, \Omega$. The procedure is that known as “one-port calibration”, concerning only one of the DUT accesses, that which we will refer to as the “entry”. In order to measure the exit parameters, it will be necessary to swap the cables so that the exit becomes the entry.

![Diagram of data analysis](image-url)
We will refer to as the entry cable that which goes from the analyzer (RF access) and injects the signal into the DUT, and exit cable that which returns to it. Once the calibration sequence is started, the analyzer displays the three configurations to be reported. It is necessary to successively connect the short-circuit ($Z = 0$), the open circuit ($Z \rightarrow \infty$) and the calibration load ($Z = Z_0$) to the end of the entry cable, recording the data taken each time, and then end the process by joining both cables with the shortest possible short, to set zero transmission. We can then connect the DUT and start the measurements.

It is necessary to always keep in mind that the analyzer, of which we speak as if talking about a person since it possesses intelligence, considers the “standards” that we have just presented as perfect. Even if there is little chance that the short-circuits or the open circuit are faulty, this is not the case with the load, which is more vulnerable. Now, whatever its value, its return loss trace will indicate $-\infty$, once the calibration sequence is considered. It is therefore down to the operator to verify it often, on a daily basis, and a good precaution is to have a second one for comparison.

2.3.2. Measurement errors

Errors due to imperfections in the analyzer itself are corrected as much as possible by calibration, whereas the thermal drifts are nowadays negligible given the quality of the synthesizers. We will refer to the construction notes to obtain the values of the residual errors.

![Figure 2.4.](image)

The aim of this paragraph is to better define the uncertainties that the elements of the external circuit can bring due to their imperfections. It concerns, for example, the quantification of the effects of a measuring cable, or of an imperfect load, on the measured value of a return loss, in accordance with the specification to be complied with. With regard to insertion losses, there is no problem of reproducibility owing to the fact that the calibration consists of registering as a reference a loss of 0 dB, materialized by a short, the impedance of which can truly be regarded as null: the uncertainty will be measured in hundredths of dB. For
Measurements in HF

return loss measurements, however, the situation is not the same. We have progressively observed a hardening of the specifications, with divergence between results from different benches for the same product being more and more significant. When at some point there is an input control put in place in the serial production process of a product, possibly by an automatic bench, the consequences of diverging results between a client and his supplier can be very serious in industrial terms, and generate damaging situations. There are, however, very simple solutions to this problem. Let us try, to begin with, to analyze it more precisely from a practical point of view.

In the simplest case, the measurement is done according to the following schematic layout (Figure 2.4).

If we suppose that the DUT and the measuring cable have negligible losses, and if we refer to the reflection coefficient of the DUT as \( \Gamma_1 \), that of the cable as \( \Gamma_2 \) and that of the load as \( \Gamma_3 \), the analyzer sees a total reflection coefficient \( \Gamma \) such that:

\[
\Gamma = \Gamma_1 + \Gamma_2 + \Gamma_3
\]

Which we can represent vectorially in the following way (Figure 2.5): the analyzer is supposed to measure \( \Gamma_1 \), but finds itself in the presence of a reflected signal that is the sum of three signals whose phases are arbitrary. The modulus of the apparent reflection coefficient is thus comprised between a maximum value, which is necessarily \( \gamma_1 + \gamma_2 + \gamma_3 \), and a minimum value which is the smallest algebraic combination of the three.

We can already establish that this scenario, which appeared simple, is no longer so, and that in order to quantify it, it will be necessary to simplify the situation even more. We are thus going to suppose, to start with, that the load is perfect, that is, \( \gamma_3=0 \), which will leave us the matching of the cable as the only free parameter. The module of \( \Gamma \) is now comprised between \( \gamma_1 + \gamma_2 \) and \( \gamma_1 - \gamma_2 \). Let us therefore calculate the uncertainty on the measurement of the DUT return loss:

a) return loss DUT 20 dB, return loss cable 35 dB.

We have \( \gamma_1 = 0.1 \) and \( \gamma_2 = 0.0178 \). The return loss measured will then be between:

\[
20 \log (0.1 + 0.0178) = 18.58 \text{ dB}
\]

and \( 20 \log (0.1 - 0.0178) = 21.70 \text{ dB} \)
b) return loss DUT 26 dB, return loss cable 35 dB as previously.

We have $\gamma_1 = 0.05$ and $\gamma_2 = 0.0178$. The return loss measured will be comprised this time between:

$$20 \log (0.05 + 0.0178) = 23.38 \text{ dB}$$

and

$$20 \log (0.05 - 0.0178) = 29.84 \text{ dB}$$

Figure 2.5

In this way we can perform a large number of calculations of this type by making the number of parameters and their values vary each time, but we can already guess from the above example that the uncertainty associated with the measurement bench is almost always far greater than that of the analyzer. Now a realistic example: 26 dB is the matching commonly specified for a band-pass filter in radiocommunications, and 35 dB is the matching of a 1 m long RG214 cable, equipped with two N-type male connectors, an object that we should be able to find in abundance in any HF laboratory.

In the basic calculations above, we neglected $\gamma_3$ for simplicity’s sake. In fact, the terminal load of Figure 2.4 is usually that of the analyzer, and it is thus not unrealistic to make this supposition. But to complete the study, we can also envisage the case where the load is external, which leads us to recall some of the properties of the calibration load.

In a calibration kit N, the matching of the reference 50 $\Omega$ is either of 49 dB, or of 52 dB according to the sex of the connector (Hewlett-Packard equipment). Let us say it is of the order of 50 dB, which corresponds to a reflection coefficient of...
0.00316 that we will round to 0.003. In example b) above, the limits of the return loss would become:

\[ 20 \log (0.05 + 0.0178 + 0.003) = 23.34 \, \text{dB} \]
\[ \text{and} \ 20 \log (0.05 - 0.0178 - 0.003) = 29.92 \, \text{dB} \]

The difference is not only measurable, but significant. If we now place the calibration load directly in the DUT exit, the uncertainty on the return loss, assumed to be 26 dB, spreads between the two limits:

\[ 20 \log (0.05 + 0.003) = 25.51 \, \text{dB} \]
\[ \text{and} \ 20 \log (0.05 - 0.003) = 26.56 \, \text{dB} \]

This is an interval of results that is perfectly tolerable in this case and that gives us a first solution for performing a correct measurement: charging the DUT with a calibration load. In practice, this method is used very little, because it has two major disadvantages. Firstly, a calibration load costs a lot, and in general we prefer to dedicate it to its essential role of reference to spare its connector, which is its weak point, by not using it unless it is indispensable.

In addition, this is a process that impedes the simultaneous measurement of the insertion loss, obliging us in this case to perform another sequence of measurements, which most of the time is incompatible with the imperatives of production associated with tight schedules. Despite these disadvantages, we must keep this in mind because it is a simple manipulation that allows us to alleviate doubt in the case of contrasting measurements.

It is thus necessary to find another procedure, more industrially convenient, to help us avoid the ambiguity we have tried to highlight: this will be the measurement attenuator.

### 2.3.3. The measurement attenuator

In the preceding section we dealt with questions of reflection by means of the reflection coefficients and their vectorial representation. There is another way of doing this, which consists of considering the reflected powers. If the diagram in Figure 2.5 seems too abstract, we can replace it by another that is practically equivalent, but closer to reality (Figure 2.6).
The load we call “total” comprises all that is found after the DUT: cables, couplers, attenuators, etc. It returns a reflected power $P_{r2}$ to the analyzer, which re-crosses the DUT, which is still assumed to be without losses. $P_{r1}$ is the intrinsic reflected power of DUT. The analyzer then an apparent returned power $P_{r1} + P_{r2}$ that it attributes to the DUT as being uniquely due to $P_{r1}$. Under the assumption that the $\gamma$ are sufficiently small ($\leq 0.1$), there results a relative error of $100 \frac{P_{r2}}{P_{r1} + P_{r2}} \%$ to this measurement. From this perspective, given that we are now dealing with scalar rather than vectorial quantities, the information on the phase disappears, so that if we perform the measurement with the aid of a thermal device such as a bolometer, the DUT-reflected power will always seem larger than the true value. It will then lead us to the least favorable value of the return loss, corresponding to the case where the reflection coefficients all add up. If we come back to example a) of section 2.3.2, we can now write:

\[ P_{r1} = P_i - 20 \text{ dB or } 0.01 P_i \]
\[ P_{r2} = P_i - 35 \text{ dB or } 0.00316 P_i \]
\[ \text{giving } : P_{r1} + P_{r2} = 0.01316 P_i \text{ or } P_i - 18.8 \text{ dB} \]

We find, save calculation approximations, the value previously determined, which constitutes the upper limit, the least favorable, of the measured return loss. Should we want to be more rigorous, we will take into consideration the fact that the return loss matching is not perfect, and that what we obtain is not $P_i$, but $0.99 P_i$, and we then recover the $18.58$ dB of the first calculation. This way of looking at the problem of course prevents us from accessing the low value, due to the principle adopted, but it is faster and, above all, more physical and leads directly to the essential.

In all that has been said, we have supposed the losses of the various quadrupoles found in the circuit external to the analyzer to be null. What happens if this is not the case? Let us take the example of an attenuator (Figure 2.7).
It is a purely resistive, passive quadrupole, intended to reduce the incident power $P_i$ by dissipating part of it in the form of heat. It is characterized by its attenuation $A$, given in dB, and its maximum power, which for now does not concern us. If the attenuator is not loaded or if it is terminated by a short-circuit, there is total reflection of the incident power. The incident signal thus crosses it twice, once in the direct direction where it undergoes an attenuation of $A$ dB, and a second time in the reflected direction, where it undergoes a new attenuation of $A$ dB. The return power $P_r$, due to the return loss in is thus increased in this hypothesis of total reflection by a quantity $P_i - 2A$. We usually refer to as “protection” this property of the attenuator which guarantees that the return loss in will not be less than $2A$ dB no matter what its load, disregarding, to begin with, its own entry matching, which will have to be excellent in order not to destroy this protection. We can readily see the interest of this property in problems concerning the accuracy of the measurements that we have highlighted previously: the attenuator is capable of masking a mal- or imperfectly-matched load, and of ensuring that it will no longer perturb the return loss in of the DUT. We are thus in the process of defining a new product in the panoply of measurement devices: the attenuator. We now need to know what attenuation value to attribute it. If it is too small, the protection will be too weak. If it is too large, the dynamic range of the analyzer risks reducing to the point of not seeing, forgetting, the traces off-band. It is also desirable to choose a value that leads to standard components. A well-tried solution, although we have the right to use another, consists of choosing an attenuation of 12.5 dB, performed according to the diagram in Figure 2.8.

This attenuation value preserves a good dynamic range in loss measurements, given that the network analyzers, even if scalar, have minimum values of at least 110 dB. The installation is performed on a small printed circuit placed in a connection board and covered by a deformable internal case, which fine tunes the capacitors. The components are of the “chip” type in order to minimize the inductive components. We easily manage to obtain a return loss in of 40 dB up to 3 GHz.
Usage: the measurement attenuator is permanently connected to the return cable of the analyzer (Figure 2.9). It is taken into consideration in the calibration sequence of the insertion losses (through), that is to say, it will afterwards be transparent in this measurement. We simply see the background noise increase by 12.5 dB at the bottom of the screen, leaving a dynamic range of about 100 dB for the analyzer user, largely sufficient for most ordinary plottings of amplitude response. In return, we are freed of the influence of the return cable in the return loss in reading of the DUT, even if the cable is manufactured with a coaxial cable of medium matching such as the RG 223 U, but which possesses versatility and flexibility qualities that make it extremely convenient for repetitive measurements. On the other hand, it is evident that this set-up prevents the calibration mode called “2 port calibration”: it is necessary to choose between rigor and rapidity, although the proposed method constitutes an interesting compromise.

The measurement attenuator allows those who use it to have coincidental plots in the measurements where the return loss specifications do not exceed 26 dB. If we want to fix stricter values, we will progressively encounter the problems exposed before, and it is important to bear in mind that the more we approach the value of the
calibration load, that is around 50 dB, the more the measurement loses its meaning: to ask for 30 dB of return loss is an imprudent demand! It is thus strongly advised for writers of specifications to clearly define their exact needs, to only request what is indispensible, and to do so, not to hesitate in quantifying the risks by means of a small calculation.

**Example**

Let us take the DUT with an assumed return loss of 30 dB, that is, \( P_r = 10^{-3} P_i \) and the measurement attenuator of a return loss in of 40 dB, that is, \( P_{ra} = 10^{-4} P_i \). It is connected to a cable of return loss 35 dB, whose return power is reduced by \( 2\Delta = 25 \) dB at the DUT exit; we can overlook it. What remains is that the apparent reflected power at the DUT entry is \( 0.0011 P_i \) instead of \( 0.0010 \). The corresponding return loss is 29.58 dB. We could be satisfied by this precision, yet we are not able to be, because the 40 dB of matching of the attenuator themselves result from an erroneous measurement, since made relative to a standard that is only 10 dB better. If the attenuator is only matched to 35 dB, we go down immediately to 28.86 dB regarding the corrected return loss of the DUT. We can thus see the practical limits of the return loss measurement appearing with clarity and in a justifiable manner, translated by the following conclusion: in fact, all measurements are false, but the engineer’s solution lies in being able to determine by how much.
Chapter 3

Resonant Cavities

3.1. Resonance

![Circuit Diagram](image)

Figure 3.1.

The phenomenon of resonance is one of the most important and one of the most general that exist in physics, and we find it in all its branches: we can think, or rather hope, that one day a resonant model of the atom will replace the planetary models in force today, thus dissipating the fabricated mystery of quantization; but it is taking time to arrive. Resonance takes place when a device under periodic oscillation presents a geometry or an arrangement such that this signal, by means of a reflection or a particular configuration, becomes in phase with itself and its amplitude increases to a maximum. This maximum is limited by a damping factor, which is exerted by friction in mechanics and by resistance in electromagnetism. The resonance can be both useful or a hindrance. In general, we can say that in mechanics we try to avoid it, even if there are examples of the opposite, whereas in physical electronics we make abundant use of it.
It is accepted in physics that all resonant systems can be modeled by an inductance and capacitance coupled together – be it in series or in parallel – and a resistance that represents the damping.

A resonant circuit is characterized by its quality coefficient in vacuum $Q_0$ and its resonant frequency $F_0$:

$$Q_a = \frac{L\omega}{R} \quad [3.1]$$

$$Q_b = RC\omega_0 \quad [3.2]$$

$$F_0 = \frac{1}{2\pi\sqrt{LC}} \quad [3.3]$$

Any closed volume is susceptible to resonate, either acoustically or electrically, or both simultaneously. From a practical point of view, that is, industrially, it is evident that we will choose forms that are easy to produce, which usually lead to the use of prismatic elements. The air resonator most frequently used is the coaxial cavity, whereas in the ceramics domain we will essentially find squared or, more rarely, circular section bars, as well as discs.

From an analytical point of view, there is a fundamental difference between the functioning of the oscillating circuit by reel and capacitor, called a resonator of localized constants, and that of a cavity or more generally of a system of distributed constants. In the first case, we are in a similar resonating system to that of a mechanical mass-spring system, where the volume occupied by the elements is of no importance. In the second case it is a matter of producing, in a space where there is propagation of a guided wave, reflection conditions such that the excitation will be found in phase with itself, which means that in this case it is, on the contrary, the dimensions that are of primary importance.

In equation [3.3], we see that the resonance of an LC circuit is a function of two quantities independent of the frequency. It follows that this type of resonator only works at a single frequency: if we want to make a periodic squared signal or a signal of any other shape perfectly sinusoidal, it is enough to make it cross a self/capacitor circuit. The harmonics are systematically eliminated from the scene. In a resonant system of distributed constants, on the contrary, there is a periodicity of the phenomenon as a function of the wavelength and of the frequency, which means that a filter studied for a given frequency will be equally transparent to all harmonics for which the phase condition is verified. A quarter-wave resonator, for example, allows all the odd-numbered harmonics to pass and cannot replace a low-pass.
In Figure 3.2, which we can imagine to be either a filled or an empty bar where either a sound or an electromagnetic signal can propagate, we see that there are three mutually perpendicular directions that are propitious to multiple reflections. It is enough to choose one of them, in general the longest one, and to adjust the corresponding dimension to a convenient multiple of $\lambda/4$ in order to have resonance.

![Figure 3.2.](image)

In fact, even if the possible combinations are infinite, the choice will be much more limited for three major reasons. The first is that amongst all possible resonance there is one which is more naturally favored than the others, leading to a better yield, that is to say, to the greatest overvoltage: it is the fundamental mode, that which corresponds to the lowest resonant frequency and thus to the shortest path in the longest dimension. It is this one, in principle, that we will choose. Afterwards, the imperatives of rationality and profitability to which every industrial project is subject, will eliminate a great number of theoretical solutions that would lead to unsalable, or even unrealizable products. It is enough, moreover, to consult the catalogs of manufacturers of couplers and filters to establish that there is a strange similarity between them. Finally, if we can a priori envisage half-wave resonators as much as quarter-wave ones, temperature stability considerations, which will be detailed later, mean that the former are almost never used in practice.

The choice between resonators of localized constants and resonators of distributed constants depends firstly on the frequency used. Even though the latter are always better than the former for equal values of $L$, $C$ and $R$, the difference appears progressively as we increase in frequency. Above all, below a certain usage frequency limit, the dimensions of basic elements such as cavities become incompatible with certain practical applications, such as for example the size of the ensembles where they have to be placed, or that of the places where we must install them. Low VHF, between 30 and 100 MHz, roughly represents the limit where we are still allowed to hesitate. This does not prevent the existence, below this practical limit and for very long-distance radiocommunications, of line and cavity devices of colossal dimensions, but their usage is extremely restricted. As regards UHF, air
coaxial resonators constitute the technological basis for filtering and emission multicoupling.

### 3.2. Coaxial cavities

When a very selective filtering is needed, we must choose a resonator that has the largest possible quality coefficient. It is quartz that then comes to mind, with values of $Q_0$ that can be calculated in the hundreds of thousands. But for certain applications, they have a major disadvantage: their insertion loss, associated with their small dimensions, is too high, which implies that they cannot support a strong power. We will use them mostly in receivers or oscillators. Ceramics come after, with excellent quality coefficients and high relative permittivities, which lead to reduced dimensions. They support a certain amount of power, however their price and limited options of available shapes avoid them being used universally. In addition, and we would say above all, their adjustment is delicate and cannot be performed by the user unless they are a specialist themselves.

![Figure 3.3.](image)

Air cavities offer a compromise solution as they are a more voluminous product, but relatively cheap, of flexible design, presenting the possibility of being adapted to numerous variants, with an $Q_0$ of interest (15,000 to 30,000 in UHF, for a volume of about 15 dm³), and especially, easily adjustable by means of simple mechanical arrangements. In this category, the coaxial cavity is by far the most used, particularly the quarter-wave model, which is considered standard.

The coaxial cavities propagate the TEM (Transverse Electric Magnetic) mode, which is the natural mode of plane waves under free propagation: the electric and magnetic fields, perpendicular to each other, lie on a plane itself perpendicular to the propagation direction, that is to say, to the axis of the cavity. In the half-wave cavity there is reflection under a short-circuit, without any change of sign and thus without phase displacement, whereas in the quarter-wave cavity there is reflection under an
open circuit, with a change in sign and thus a phase change of $\pi$, equivalent to a step change of $\lambda/2$. In both cases we verify that the total return path is equal to $\lambda$, the reflected wave being in phase with the incident wave.

Although there is the possibility of a theoretical choice between a half-wave coaxial resonator and a quarter-wave one (Figure 3.3), we almost never use the former, despite its simplicity of construction, because it is not adjustable over a large frequency range on the one hand, and because it is not easily compensated with respect to variations in ambient and internal temperature on the other hand. The quarter-wave cavity has no such disadvantages; we can, on the contrary, add to it simple mechanical devices that allow the length of the central conductor to be varied, and thus the frequency, to a sizeable proportion, rendering it practical to use and stable in temperature.

3.3. Quarter-wave cavities

3.3.1. Excitation

![Figure 3.4.](image)

We now understand, from all that has been said, that a cavity will constitute a basic element to bring a certain selectivity to a circuit by obtaining a pass-band function. What remains is to insert it in the circuit, for which it is necessary to equip it with an entry/exit connecting system and an excitation system that will allow the penetration of the signal into the interior of the cavity and its exit. It will be necessary, in particular, to simultaneously pass from a low impedance $Z_0$ to a high internal impedance, a condition of a high $Q_0$. We generally represent this double interface by two identical impedance transformers that complete the representation of Figure 3.1, which gives the electric scheme of Figure 3.4. It is a question now of knowing what the two transformers represent, knowing that the cavity, as a piece of laboratory equipment, will be equipped with two sockets on which two cables will be connected, one coming from a source of impedance $Z_0$ and the other going towards a load circuit of impedance $Z_0$. 
There are two procedures or, if one prefers, two devices: probes and loops. Probes are rectilinear threads that extend the core of the coaxial to the interior of the cavity, in the zone said to be the electric field, or high impedance, or voltage zone, level with the end of the piston. We use the term “piston” to name the central conductor of the cavity when it is equipped with a system for adjusting its length, which is the general case. The probes are radially oriented, according to an electric field line, and adjust themselves by deformation via a hole provided in the lining. The loops, on the other hand, are placed in the head of the cavity, that is, on the fixation of the piston, in the zone said to be a magnetic field, or low impedance, or current zone. They adjust themselves by deformation like probes, or more commonly by rotation around their own axis with a positioning reference. They work by capture of the magnetic field, whose lines, in the Faraday sense of the term, are concentric and cross the loop with a greater or lower density according to their orientation (Figure 3.5).

![Figure 3.5.](image)

Figure 3.5 shows a diagram of the two systems. The probes to the left and the loops to the right are represented in the position where the capture of the electric field and that of the magnetic field, respectively, are maximum with respect to the configuration drawn: the probes are directed towards the axis of the cavity and are collinear to an electric field line, whereas the loops are oriented on the sketch in such a way that the captured field is maximum. In fact, by examining in detail Figure 3.5, we see that one can further increase the captured field by hoisting the probes to the level of the upper end of the piston or by approaching the loops to the axis of the cavity, where the magnetic field is most intense. We have likewise
understood the bases of the tuning of the excitation or, if one prefers this term, of the external coupling of the cavity: from the optimal dispositions represented in Figure 3.5, we can reduce the coupling by deforming the probes in whichever direction, what will reduce their effective length. Regarding the loops, we can choose, in order to make their surface vary, to make them rotate around themselves, or to deform them. In the latter case, it will be necessary to pay particular attention to the procedure: in fact, if for example we have decided to increase the coupling by approaching a loop to the central conductor by deformation, we will not be able to do so in general, except by modifying its geometry. Now, depending on the chosen deformation point, this second effect can either add itself to or remove itself from the former. We can see very clearly through this simple example what will be the helping hand in this work of specialists, particularly as they always perform this task blindly, as the tuning holes are the only continuity solutions in the envelope of the cavity.

To finish, let us note that it is not forbidden to use both types of coupling simultaneously. We do not do this ordinarily, but it is not impossible. What is essential is to adjust $T_1$ and $T_2$ in the same way, so that both accesses are symmetric and present an impedance that is as close as possible to $Z_0$ at the resonant frequency. This will depend on the choice we will have made between the two interdependent parameters which will comprise the adjusted cavity: passing band and insertion loss.

3.3.2. Possible shapes

The theory of coaxial cavities is traditionally based on the cylindrical model. This is the school model, that allows us to directly apply the results already obtained on the properties of the coaxial line, and which we find in all university courses. In fact, all right prismatic models are suitable for industrial production: it is simpler, for example, to make a squared section cavity than one with a circular section, and it will lead to a better filling factor if, for example, we must accommodate a certain quantity in a 19” standard bay, which is very often the case. We can also think of rectangular or trapezoidal sections or even stranger configurations suggested by the possibility of converting previously existing containers for a completely new task. We have already seen, in our profession, couplers manufactured in beer casks, barrels, ventilation conduits and other objects with no a priori relation to high frequencies, but that once suitably reinstalled and internally silver-plated, have obtained results by all means comparable to those of conventional products, with a cost that is more attractive in large serial productions.
The few schematic sections displayed above illustrate the diversity of forms that we can give to a cavity, as well as the practical limit of these possibilities, which is highlighted in the model furthest to the right, expressed in the form of the following empirical rule: it is pointless, from an electrical point of view, to extend a dimension beyond a value which is of the order of 1.5 times the distance between the central conductor and the closest surface. This does not necessarily lead to the suppression of the redundant part; simply useless.

3.3.3. Lines of force

The concept of lines of force was invented by Faraday, who named them lines of electric induction. We can also call them Faraday lines or field lines. It was Maxwell who introduced the idea of force into this representation of the field. A line of force starts perpendicularly from a positively loaded conductor, to end perpendicular to a second conductor, negatively loaded. It corresponds to the shortest electric path to go from the departure point to the arrival point; that is, that of the minimum work performed by an elementary positive load following one of these lines and whose displacement is supposed infinitely slow. The lines of force enable us, in a way, to “draw” the static electric field, tangent to every point of the line; its intensity is represented by the density of lines at a given point in space, that is, the quantity of lines that perpendicularly cross a unitary surface element. Surfaces perpendicular to all the lines are called equipotentials, and they also represent the geometric place of the magnetic field lines when we go from a static to an alternating regime.

It is the idea expressed by Faraday that the voltage exists not only at the surface of the conductor but all the way along the lines of force, which led Maxwell “to conceive the electric action as a phenomenon of deformation of a medium” (Electricity Treatise, Volume 1, section 48). The rejection of the notion of a medium by special relativity disfavors this idea, but the representation of the field by lines is so practical that it still has its supporters, which allows us to use it to illustrate, according to a certain conception of the propagation space, what happens in a cavity.
Figure 3.7 represents the two possible structures of the central conductor: open-end or closed-end. The first, simpler to produce, is found mainly on reception filters, the second mainly in cavities and power filters, where it is necessary to avoid the weak curvature rays that increase the field intensity. Drawing these lines makes it the existence evident of the field beyond the end of the central conductor, which leads to a consequence being stated that can seem insignificant at first sight, but that will in fact prove very important in the practical calculations of filters: the length of the central element is always less than $\lambda/4$.

![closed tube](closed-tube.png) ![open tube](open-tube.png)

**Figure 3.7.**

### 3.3.4. Equivalence criterion

Whilst having the choice of the section’s form gives a great degree of freedom, it is still useful to know what the aims of the project are before making this choice, as well as to always be capable of characterizing a cavity which is not built according to the school model. How can we adapt the methods of calculating impedance or the quality coefficient to these unforeseen and possibly complicated models?

In fact, we adapt nothing: we will simply define an equivalent cylindrical cavity, about which we already know everything. The only problem is to know what equivalence criterion to adopt. This criterion is almost evident if we remind ourselves of the meaning of the quality coefficient, defined in section 3.1 by means of the model of localized constants of the resonant cavity, and try to give it a more
physical significance. In fact, equation [3.5] can be written, multiplying nominator and denominator by \( i^2 \), as: \( Q_0 = \frac{L\omega_0 i^2}{R, i^2} \), which we can also write:

\[
Q_0 = \omega_0 \frac{Li^2}{Ri^2} \tag{3.4}
\]

We recognize in the quantity in the numerator, the energy contained in the R, L, C circuit, which alternates, in the course of a pulse, from the inductive and kinetic form \( W_1 = \frac{1}{2} Li^2 \) to the capacitive and potential form \( W_2 = \frac{1}{2} \frac{Q^2}{C} \). The total energy of the circuit is \( W = W_1 + W_2 = 2W_1 = Li^2 \). As for the denominator, it expresses the resistive losses under the form of a power. By noting that expression [3.4] can also be written:

\[
Q_0 = 2\pi \frac{Li^2}{Ri^2 T_0} \tag{3.5}
\]

and considering things on a more physical level, we arrive at another definition of the quality coefficient, which will apply to all cases, whether they be localized or distributed constants:

\[
Q_0 = \omega_0 \frac{\text{stored energy}}{\text{dissipated energy}} = 2\pi \frac{\text{stored energy}}{\text{dissipated energy per period}} \tag{3.6}
\]

This other formulation of the quality coefficient will give us the key to solving the problem of the equivalent cavity, that is, one possessing the same \( Q_0 \). Once the frequency is given, the height of the cavity is also determined because it always represents a quarter-wave, and the dissipated power is the same because the central conductor, which is the main ground of resistive dissipation, is also the same.

Two cavities will thus be considered equivalent if their stored energy is the same and, as it is not possible for this energy to be located anywhere other than in the dielectric (whose volume is given), as a result the surface of the right-hand section must be the same: the cylindrical cavity equivalent to that of a cavity of any shape is that which has the same section.
In fact, this is not entirely the case. The quarter-wave cylindrical cavity, taken as a reference, is the only one presenting a homogenous radial distribution of the electric field, and thus of the magnetic field, represented by concentric circles. For all the others it will be necessary to take into consideration the specific features of their spatial distribution, and the Faraday lines will once again be a practical way of materializing the electromagnetic field. Let us take as an example the squared-section cavity, to which we will attribute, to begin with, a side equal to the diameter, $a$, of the circular section (Figure 3.8). We see that the lines of force of the magnetic field, if circular in the vicinity of the central conductor, progressively take a squared form as they approach the external envelope, which is entirely natural given they should be perpendicular to the electric field lines, which in turn are perpendicular to the walls. We also note that the field is null in the corners, which indicates that there is space that is electrically unused, meaning that the effective surface of the squared section will be less than its real surface, $a^2$. Therefore, we generally apply a correction coefficient of 0.9 to it. In that respect, when using a cavity with squared section in place of a cavity of circular cavity with the same dimensions, $Q_0$ will be increased by a factor $0.9 \times \frac{4a^2}{\pi a^2} = 1.146$ or 14.6% at best, which is not negligible if we consider that a stacking of an equal number of cavities of one type or the other occupies the same volume. The side of the squared section equivalent to a circular section of diameter $a$ is thus $1/\sqrt{1.146} a = 0.934 a$. Regarding other shapes, the equivalent circular cavity will be determined largely by intuition and experimentation, using the small number of rules suggested above, but nevertheless with a well-defined path to follow, even if largely empirical: we have to draw the circular section of the same surface (by weighing, calculation, programming, etc.).
and apply a correction coefficient that in principle is part of the know-how of the manufacturer, and so can vary from one shape to another, but is usually between 0.8 and 0.9.

### 3.3.5. Electric length

We mentioned in section 3.3.3 the fact that in a quarter-wave coaxial cavity, the length of the central conductor is always less than $\lambda/4$, although the mathematical conditions of resonance demand that the forward path of the wave be rigorously equal to this length. The fact that we can draw Faraday lines, even though it does not prove anything, nevertheless helps highlight the fact that the field also exists in the section comprised between the top of the central conductor and the back of the cavity, which we usually express by defining a “back capacity”. Given that, we can see that the length of the central conductor will depend on the geometry of the cavity, or in other words, of its proportions, with the exact place of the phase inversion always being located at $\lambda/4$ from the base. This knowledge is extremely important because not only does this datum determine the resonant frequency, but also be almost indispensible for calculating the coupling coefficient when we are dealing with coupled resonators, as it depends on it directly (see Chapter 6).

![Figure 3.9](image)

Figure 3.9 indicates a simple geometrical construction which consists of tracing, from the end of the central tube, the straight segment AB, making an angle $\alpha$ of 27° with the horizontal line, a value which is not noteworthy in itself except for the fact...
that its tangent is equal to 1/2. The point B is then such that \( \frac{d}{c} = 0.5 \) and gives the proportionality of the path \( \lambda/4 \) in relation to the end of the central tube. This empirical construction is not based on any mathematical demonstration but, applied to cavities of the most varied forms – from the widest to the narrowest – it has proved to be of remarkable exactness. The laws of physics are rarely complicated, and the fact that there exists a method as simple and precise as this, and widely tested, to determine the electric length of the central conductor, that is, given in degrees of phase, suggests the possibility that it could be mathematically justified with the aid of an adequate modeling.

\( \lambda/4 \) corresponds to 90°. Let us call the maximum electric length (we can choose a smaller value) of the piston \( \theta \). We have: \( \theta = 90 \frac{h}{\lambda/4} \)

But \( \lambda/4 = h+d = h+c \tan \alpha \) with \( c = b-a \), thus:

\[
\theta \degree = 90 \frac{h}{h+c} \frac{b-a}{2} \tag{3.7}
\]

We observe that there is an unknown segment on the drawing: the distance \( H-h \), between the top of the tube and the back of the cavity. This distance is never less than 2a; once this condition is fulfilled we have noted that its value has a negligible influence on the final result.
Figure 3.10 shows the internal proportions of the two extreme cases of geometry between which any cavity will be situated. As indicated, the electric length is between 72°, for the widest cavity, and 89° for the most narrow: the difference between the value determined with equation [3.7] and the measurements is less than 1°. The length of the tube can be chosen arbitrarily and in practice go down to $\lambda/8$, or even $\lambda/10$, which we can consider to be an extreme limit. Below this value, there is the risk of undesirable superior modes appearing, which have the effect of deteriorating the yield and stability of the cavity.

Another remark: the formula is valid both in the case of an open or a closed tube, although Figure 3.7 shows different field distributions for both cases. This experiment-based finding, added to what was said before, leads us to conclude that the stored energy is contained – if we suppose its density to be uniform – in a prismatic volume which we will call effective volume, materialized by the side segment $\lambda/4$ of Figure 3.9, which is always larger than that which contains the central conductor and which does not depend on anything other than the diameter of the latter.

3.3.6. Conditions for optimal $Q_0$

The ratio $b/a$ between the internal and external radius of a coaxial cavity, considered as part of a particular line, is present in the expression of its linear constants. We suspect, as a consequence, that it will also affect the value of the quality coefficient. If $Q_0$ is a function of $a$ and of $b$, it is thus natural to look for values of these two parameters for which there exists a maximum of $Q_0$, which was supposed in the preceding sections.
Let us briefly remind ourselves of the characteristics of the coaxial line:

\[ L = \frac{\mu}{2\pi} \log \frac{b}{a}, \quad C = \frac{2\pi\varepsilon}{\log \frac{b}{a}}, \quad R = \frac{R_0}{2\pi} \left( \frac{1}{a} + \frac{1}{b} \right), \quad G = \frac{2\pi\sigma}{\log \frac{b}{a}} \]

\( G \) represents the losses in the dielectric. In the case of air, which is an excellent insulator, we will neglect them: \( G = 0 \).

\[ Q = \frac{L\omega}{R} \] can be written:

\[ Q = k \cdot \frac{\log \frac{b}{a}}{1} + \frac{1}{b} = kb \cdot \frac{\log \frac{b}{a}}{1} + \frac{1}{b} \]

If we fix \( b \) and if we put \( \frac{b}{a} = x \), looking for the maximum of \( Q \) (or of \( Q_0 \), which is of course the same) is reduced to the problem of studying the function:

\[ y = \frac{\log x}{1 + \frac{1}{x}} \]

Its derivative is

\[ y' = \frac{x}{\left( 1 + \frac{1}{x} \right)^2} \]

which is zero for

\[ \log x = 1 + \frac{1}{x} \]

We can verify by the usual methods that this is a maximum, and the solution of the equation is \( x = 3.6 \). This demonstrates the condition that was introduced in the preceding section. It is interesting now to know the shape of the curve of \( Q \), by giving some values to \( x \).

<table>
<thead>
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<th>x</th>
<th>2</th>
<th>3</th>
<th>4</th>
<th>5</th>
<th>6</th>
<th>7</th>
<th>8</th>
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<tbody>
<tr>
<td>y</td>
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<td>0.277</td>
<td>0.268</td>
<td>0.256</td>
<td>0.243</td>
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<tr>
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<td>0.989</td>
<td>0.996</td>
<td>0.964</td>
<td>0.921</td>
<td>0.874</td>
<td>0.831</td>
</tr>
</tbody>
</table>

Table 3.1.

We establish that the curve is very “smooth”, that is, with small variations around the maximum. In particular, we see that between the values 3 and 5 of \( x \), there is almost no variation of \( Q \), which means that from a practical perspective (and only from this point of view) we could choose a central tube of relatively low diameter – in any case, lower than that indicated by the rule \( b/a = 3.6 \) that all industrials scrupulously respect. But experience shows that when the insertion losses are judged too extensive, on a filter or cavity prototype, it is enough to increase the diameter of the central conductor to gain the few tenths of dB that we might be lacking for a given selectivity.
This brings us to cast some doubts on the foundations of the method detailed above, which we have nevertheless explored because it is commonly used.

3.3.7. Refutation of the method

The word “refutation” is maybe excessive, to the extent that it is not a question here of opposing, in an absolute and definitive way, the conventional model, but rather of suggesting that there are other paths or other ways of reasoning that produce alternative approaches, which, using another logic, leads to other results – sometimes different and complementary. It is common in electricity courses to see comparisons, analogies, or attempts at identification between certain electrical phenomena and mechanical equivalents that, viewed with the necessary detachment, present striking resemblances, which a number of physicists have highlighted and used, J.C. Maxwell in particular. Examples are the parallel between self-induction and the conventional concept of mass, or that between a capacitor and a spring (which, as we know, explains the electricity origin of the word “voltage”). It is thus not forbidden to borrow these views and to make reference to such examples to suggest that the vacuum, home of the stored energy, could itself have a property analogous to that of a rotating or vibrating mass that stores mechanical energy. If we adopt this hypothesis, it is necessary that the quantity of stored energy must be proportional to the dielectric volume.

Let us return to the configuration proposed previously: b/a = 3.6. Let us now apply an increment of 20% to the diameter of the central tube: a becomes 1.2a, which implies that the linear resistance of the central tube becomes $R_s/1.2 \times 2\pi a$. The linear resistance of the whole line is thus $\frac{R_s}{2\pi} \left( \frac{1}{1.2a} + \frac{1}{b} \right)$ instead of...
\[ \frac{R_{m}}{2\pi} \left( \frac{1}{a} + \frac{1}{b} \right) \]. We suppose for now that the central conductor and the external conductor are of the same nature, otherwise we would have to put \( R \) in the form

\[ \frac{R_{m}}{2\pi a} + \frac{R_{m}}{2\pi b} \]. If we refer to the general definition of the quality coefficient given by expression [3.6], the increase in the central diameter translates into a decrease of the resistive losses, and thus a proportional increase of \( Q \) by the ratio:

\[ \left( \frac{1}{a} + \frac{1}{3.6a} \right) / \left( \frac{1}{1.2a} + \frac{1}{3.6a} \right) \], giving \( \frac{4.6}{3.6} / \frac{4.8}{4.32} = 1.15 \). Similarly, the dielectric section, which was \( \pi (b^2 - a^2) = 11.96 \pi a^2 \), becomes \( \pi (b^2 - 1.44 a^2) = 11.52 \pi a^2 \), the ratio between both being of 0.96.

On one side, we thus gain 15% on \( Q \); on the other we lose 4% by the decrease in the dielectric volume, seen as the reservoir of stored energy. This globally gives a theoretical improvement of 11%, while the official method predicts a small loss, if we refer to the graph of Figure 3.12. This different result must thus sound suspect, but in its favor it has a decisive argument: it corresponds to reality. In fact, we have never seen, in a cavity or a combline filter that the insertion losses increase (for a given selectivity) when the diameter of the tubes increases from the value \( b/a = 3.6 \). This leads us to contest the position of the maximum of \( Q_0 \) for this particular value of \( b/a \).

We should remember that we have implicitly used the hypothesis that the quantity of stored energy, and consequently the quality coefficient, are proportional to the volume of the dielectric. This implies that we replace the concept of stored energy (a term consecrated by habit) by that of “storable” energy, more unusual but that nevertheless attributes to the dielectric, that is, the vacuum, special properties that are usually justified using linear constants of the circuit equivalent to the resonator. Put another way, we now state the hypothesis that it is the space itself that is intrinsically capable of storing energy, with a density that we will suppose uniform. In addition, the construction of Figure 3.9, whose legitimacy is given by experience, clearly shows that calculating linear constants based on the measured length of the central conductor, instead of considering its effective electric length, leads to a wrong value, unless we interpolate the boundary effect by another bias.

If, for the sake of simplicity, we temporarily neglect this boundary effect by considering it solely as a corrective factor, the volume of the air dielectric is \( V = h \pi (b^2 - a^2) \), \( h \) being the height of the central tube. The expression of resistive losses being the same, the search for the maximum of \( Q_0 \) comes back
to studying the function \( y = \frac{1-x}{1+x} \), the derivative of which is

\[
y' = \frac{2x^2(1+x)-(1-x^3)}{(1+x)^2} = \frac{2x^3-x^3}{x^3(1+x)^2},
\]

which is zero for \( x = 2 \) in the interval of interest (said here to be \( x = 1 \) to \( x = 8 \)). All that is left now is to draw the curve representing the function \( \frac{Q}{Q_m} \) (Figure 3.13), and to compare it to the previous one. We remark a similarity that suggests to us that both relate to the same phenomenon, but we can see that the variations are larger and that there is a displacement of the maximum from 3.6 to 2.

### Table 3.2.

<table>
<thead>
<tr>
<th>x</th>
<th></th>
<th></th>
<th></th>
<th></th>
<th></th>
<th></th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>1</td>
<td>2</td>
<td>3</td>
<td>4</td>
<td>5</td>
<td>6</td>
<td>7</td>
</tr>
<tr>
<td>y</td>
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<td>0.250</td>
<td>0.222</td>
<td>0.188</td>
<td>0.160</td>
<td>0.139</td>
<td>0.122</td>
</tr>
<tr>
<td>Q/Qm</td>
<td>0</td>
<td>1</td>
<td>0.888</td>
<td>0.752</td>
<td>0.640</td>
<td>0.556</td>
<td>0.488</td>
</tr>
</tbody>
</table>

Conclusion: only experiments can decide between two ways to treat a problem that lead to two different results. Nevertheless, things are not so simple. Curiously, there is in fact no database that allows us to make this comparison. It would seem that generations of disciplined engineers have never cast doubt on a mathematical formula demonstrated from logical hypothesis. Moreover, there are material considerations which have meant that the need for such experimental tests, despite being important, is never imposed. When we talk about cavities, we are talking about high selectivity and often strong powers that lead to objects of large dimensions. While the external envelope is usually made to measure, the central tube is, on the other hand, chosen almost always from tube-factory standards, where we find raw material that is properly manufactured, and with attractive prices for small diameters. We are, however, rapidly limited by the availability and cost for the most significant diameters, so that we have the tendency to choose the smallest possible which does not take us too far from the norms. From this point of view, the first method of calculation is good enough. On the other hand, the space one must reserve for the installation of the loops, as well as the constraints of expansion associated with the peak voltage to be supported, disallow the small ratios \( b/a \): we rarely go down to 3, and never to 2. To conclude, the frequent difficulties in production, especially concerning surface treatments, obscure small variations in the calculations. In summary, all these effects do not support experimentation, and there is ultimately a shortage in numerical results that prevents us from deciding between the two theories, at least with regard to coaxial cavities.
For the case of reception filters, on the contrary, we have access to a great number of models, of measurements and figures that enable us to tend towards the second theory. Dishal, the developer of a calculation method for interdigital filters which we will discuss later, indicates that the ratio between the width of a filter and the diameter of the resonators can vary without any inconvenience from 5 to 2. It does not say that a ratio of 2 leads to a better filter than a ratio of 4, but experimental results nevertheless allow us to do so. It remains to be known, of course, what the correspondence is between the ratio $b/a$ of an isolated cylindrical cavity, for which there is no ambiguity in the definition, and the ratio $b/a$ of a cavity equivalent to one of the resonators of a multipole filter. This we cannot determine with exactitude.

What should we do then, now that the doubt has been introduced, to avoid upsetting the habits and destabilizing conceivers? The simplest approach is to continue to take, as the basis of construction of a coaxial cavity, a ratio $b/a$ of about 3.5 and, if the $Q_0$ is insufficient, increase the internal diameter to verify whether or not there is any improvement. In the event that there is, it is advisable to try to find a new definition for $Q_0$.

3.3.8. Calculation of $Q_0$

When calculating the field in the interior of a cavity, the equations of propagation and boundary conditions allow us to calculate the quality coefficient in the vacuum, as a function of the dimensional parameters and of the frequency, starting from a general definition of $Q_0$ as given previously. We evaluate both the stored energy and the power dissipated as a function of the quantities defined in
Chapter 1, in order to use them in the expression \( Q_0 = \omega_0 \frac{W}{P} \). To begin with, it is habitual to define the two modes, even and odd, corresponding, respectively, to a reflection with or without a change in sign. The even mode will thus concern cavities where the electric length is of the form \((2k+1) \frac{\lambda}{4}\), whereas the odd mode will correspond to \(2k \frac{\lambda}{4}\). For a signal of the form \( V = V_1 e^{i(\omega t - \beta z)} \) where \( V_1 = A \log \frac{b}{a} \), the expression of the field inside the cavity is written:

\[
E_r = \frac{A}{r} e^{i(\omega t - \beta z)} \quad \text{and} \quad H_\theta = \frac{A}{\eta r} e^{i(\omega t - \beta z)}
\]

After making all the calculations we find, for any given mode:

\[
W = 2\pi A^2 \varepsilon_0 \varepsilon_r \frac{\lambda}{4} \log \left( \frac{b}{a} \right)
\]

\[
P = 2\pi R_0 \frac{A^2}{\eta^2} \left( \frac{\lambda}{4a} + \frac{\lambda}{4b} + 2 \log \frac{b}{a} \right), \quad \text{with} \quad \eta = \sqrt{\frac{\mu_0 \mu_r}{\varepsilon_0 \varepsilon_r}},
\]

which leads to the general expression:

\[
Q_0 = \frac{\mu_r}{\varepsilon_r} \frac{4\pi R_0}{F b \mu_0 \mu_r} \left( \frac{\lambda}{b} + \frac{\lambda}{a} + \frac{8}{N} \log \frac{b}{a} \right)
\]  \[3.8\]

where \( N \) is the number of \( \lambda/4 \) and \( R_0 = \frac{1}{2\pi} \log \frac{b}{a} = 60 \log \left( \frac{b}{a} \right) \). This formula is applicable to all types of coaxial cavities.

It is simplified in the case of copper air cavities, where \( \mu_r = 1 \) and \( \varepsilon_r = 1 \), so that we can ultimately write:

\[
Q_0 = 9.0791 \cdot 10^4 \cdot 0.85 \cdot \frac{\log \frac{D}{d}}{B.10^3 + \frac{8 \log \frac{D}{d}}{N} + \frac{1}{\sqrt{F}}} \]
\]  \[3.9\]
with:

– D, external diameter = 2b in mm;
– d, internal diameter = 2a in mm;
– B = 600/F  F in MHz.

For a silver cavity we multiply the result by 1.0445.

This formula, which is extremely practical, is fast to program on a pocket calculator. The coefficient 0.85 that appears is an empirical correction which allows us to adjust the result to that which we find from the measurement of Q. We can verify the programming with the following example:

\[
D = 200 \text{ mm} \quad d = 50 \text{ mm} \quad F = 400 \text{ MHz}
\]

We must obtain \(Q_0 = 11009\) for copper, and 11499 for silver.

We can see on the theoretical formula that the frequency figures in the denominator, under a fairly complex form that does not make the variation of \(Q_0\) as a function of \(F\) obvious. In fact, the measurements show that \(Q_0\) increases when \(F\) increases, and that it is appreciably proportional to \(\sqrt[3]{F}\).
4.1. Standard structures

4.1.1. Band-pass

There are three main types of coaxial cavities: band-passes, rejects and pass-rejects. The specific features and the functions will be detailed later on; the basic architecture is the same for all three, and it is that of the band-pass cavity that is schematized in Figure 4.1.

![Figure 4.1.](image)

On the cavity body the head and the back are screwed or extruded. The electric continuity between the body and the back is less of a concern than that of the head,
which must be perfect, with the body and the tube at the same time, and this is the reason why extrusion is always preferred to screwing. The mobile piston is the frequency regulating element; it is made accessible from the exterior by means of an invar rod which is blocked by a lock-nut or a more sophisticated device according to necessity (demultiplication, fine tuning, remotely-operated cavities, etc.). The usage of the invar is indispensable to assure frequency, but as a general rule it is not sufficient and must be associated with a mechanic compensation system making use of the intervention of several metals. It is obligatory for the head, the tube and the piston to be silver-plated so as to minimize the surface resistance. For the body and the back this is optional, and in fact it depends on the metal used and the intended performance levels.

4.1.2. Reject

![Diagram of Reject](image)

The reject cavity is a band-pass in which we have integrated a small section of line 50 Ω placed not far from the central connector and thus coupled to it by proximity, which is equivalent to saying that we have created a capacity between them. Figure 4.2 shows one of the possible layouts, and we can also place the line over the envelope, parallel to the central conductor. On the figure we have schematized a small return of the line, which can consist of a simple strap, of which
the distance to the piston can be adjusted by deformation. Another solution, which can be complementary to the first, is to equip the central tube with an offset piece that we make more or less distant to the line by rotation, etc. These devices transform the resonator into a perturber of the line, which will create a rejection whose response is the dual of the band-pass function: the rate of the transmission curve of one becomes the return loss curve of the other, and vice versa (Figure 4.3).

![Figure 4.3.](image1)

The operating principle is simple: the capacitive coupling connects by a by-pass to the line, at the coupling point A, a quadrupole (the cavity) that has an impedance that is almost null, or at least very low if the cavity is of good quality, and that is in addition a pure resistance $r$ to the resonance (Figure 4.4). The capacitor $C$ exerts the coupling by proximity. If we neglect the resistance $r$, which represents the intrinsic loss of the cavity, the signal sees a drift where it undergoes the following phase

![Figure 4.4.](image2)
rotations: $\pi/2$ when crossing $C$, a reflection in a short-circuit and thus without a change in sign, then again $\pi/2$ while re-crossing $C$, giving a total of $\pi$. In other words, there is phase opposition between the signal that returns in $A$ and the incident signal, and thus a weakening that depends on the strength of the coupling, represented by the value of $C$. In practice, with a strong coupling, it is difficult to surpass 30 dB because beyond that, even if the cavity is sufficiently voluminous to be able to attain a greater attenuation, the rejection band becomes too narrow and its stability decreases. If we really need a greater attenuation, we can place additional cavities in series, coupled by quarter-wave cables, displacing them in frequency if necessary to obtain the desired compromise between rejection and band. We thus make a rejecting filter.

4.1.3. Pass-reject

![Figure 4.5.]

The pass-reject is the third variant of coaxial cavities. We can consider it a rejecting cavity where we have replaced the added line section by a series resonant circuit, the latter always being coupled by proximity. The addition of this simple self/capacity circuit enables us to create a series resonance close to the parallel anti-resonance and to obtain a response curve analogous to that of a quartz or of a piezoelectric ceramic, which leads us to offer a brief reminder of the properties of one or the other, both having the same electric scheme, equivalent to those of...
Piezoelectricity, discovered by the Curie brothers in the late 19th Century, manifests itself by the appearance of electric loads on a capacitor constituted by a quartz cut into fine strips perpendicular to the optical axis and submitted to pressure or exterior traction. It is a reversible phenomenon that transforms the object into a sound emitter when we apply an alternative tension to it. We can now represent it by the diagram of Figure 4.5: the capacity $C_b$ is the blocked or static capacity, $R$ represents the dielectric losses and the central branch regroups the motional values, which have no existence or significance except when the quartz vibrates, from where they derive their name. We thus see that there are two resonances, one in series leading to a minimum impedance, and the other parallel, transforming the quartz into a rejecting circuit. The quantities $r$, $L$ and $C_m$ represent mechanical entities: friction, mass and elasticity respectively. What is interesting and concerns us more particularly, is the impedance response curve (Figure 4.6). $C_b$ is very large compared to $C_m$, so that the parallel resonance frequency $F_p$, which makes the equivalent capacity intervene in $C_m$ in series with $C_b$, is very close to $F_s$ or higher. We are thus provided, by means of a physical object provided by nature and arranged a little by man, with a low-pass structure with high rejection, which is a second way to arrive at the pass-reject cavity and better understand its operating principle, but with one essential difference: in a quartz or a piezoelectric ceramic, the passing frequency is always lower than the rejected frequency. The pass-reject cavity is only of practical interest if we are able to choose and regulate at will the direction and the value of this gap. To do so, we use the principle set-up of Figure 4.7.
Between the two connectors where we would fix the 50 Ω line of a reject cavity, we now have a reel and a variable capacitor (trimmer), connected in series, to which we add a tuning element of the proximity coupling, represented in the diagram by a deformable strip which we bring closer to or distance from the central conductor, but that in practice can assume a large variety of forms. We have chosen to adjust the impedance of the circuit by a variable capacitor for practical reasons, bearing in mind that multi-cylindrical air trimmers exist on the market, which conventionally allow us to have a capacity whose value can be adjusted from 1 to 10 or 14 pF. But nothing theoretically opposes us varying the inductance. Some constructors even use the deformation of the reel connection strap to regulate the coupling. Others build variable capacitors to increase its tuning range.

In fact, the industrial production of this set-up is always a little complicated because it constitutes the weak point of the assembly, mechanically speaking. It is necessary in fact that the components be sufficiently sized so that they can be of a good quality, but not excessively so as not to excessively perturb the cavity. On the other hand, they are isolated from the body, and they do not have anything apart from the fixed connectors as an attaching point. In addition, they must be accessible from the exterior. Despite these minor difficulties, having long been mastered by specialists, the pass-reject cavity offers a unique filtering possibility that is very particular in making it possible to strongly reject a frequency near to that which must not be attenuated, something that a simple rejecting cavity cannot do, since its attenuation reduces symmetrically and progressively around the resonance value. Its principle will also allow the conception of duplexers, as we will see later.

![Figure 4.8](image-url)
Figure 4.8 shows the typical design of the response of the high-pass model, which is perfectly symmetrical to the low-pass. What now remains is to see how the LC circuit we have inserted in the main line really works. In fact, we could not obtain an abrupt slope between $F_1$ and $F_2$ and a very low loss in $F_2$ by simply adding the direct effects of a series resonant circuit – with very “soft” variations around the resonance value – to that of a rejection that is strong but with a very progressive recovery around its own frequency. As is the case for quartz, which contains in itself all the components of its equivalent electric scheme, the LC circuit that we add to the reject cavity will, while providing a phase rotation and a particular impedance, modify the intrinsic characteristics of the latter, which will in turn achieve its series resonance and its parallel resonance by using electric components of very good quality.

It is thus necessary to see the series LC circuit as an ensemble that we can render either capacitive or selfic. For that, we adjust the capacitor value from zero (or a nearby equivalent value) to a value allowing us to obtain an impedance double that of the self. In this way we will be able to choose the side of the rejection band in which the passing frequency will be, on one hand, and to regulate the gap between $F_1$ and $F_2$ on the other hand. The high-pass function will correspond to a capacitive coupling with the line, the low-pass function to an inductive coupling. The separate determination of $L$ and $C$ is theoretically complicated, and is usually made by successive trials from the previously defined relation:

$$\frac{1}{C\omega} = 2L\omega, \text{ that is, } L\omega^2 = 1/2.$$  

4.2. Materials

Although filters can be considered, by extension, as cavity assemblies, in this chapter we will only deal with connectorized individual models whose dimensions are measured at least in dm$^3$. To build the outer envelope, the technological choice is made with respect to two main directions: the thin metallic sheet and extrusion.

4.2.1. The thin metal sheet

According to the base cross-section we have chosen, circular or squared, the shaping tools are different. It is necessary to roll the metal sheet for the first, and to fold it for the second. We must then join, often with the help of “pop” rivets which we can leave in place or not after extrusion. We mainly use metal sheets made of copper or brass.
The latter is more easily extruded, if we can say that, given that the operation demands particular care, given the property of large metallic surfaces of cooling rapidly: it is necessary to preheat and use the extrusion torch. Subsequently it is necessary for the brass to be silvered, whereas a copper envelope left as it is will not alter the final performance much, given that it is the second conductor after silver, and with a conductivity value which is very close to that metal. Table 4.1 summarizes comparatively the metals or alloys most commonly used in HF.

<table>
<thead>
<tr>
<th>Material</th>
<th>Volumetric mass (T/m³)</th>
<th>Coefficient of linear dilation (ppm/°C)</th>
<th>Resistivity (Ω/m)×10⁸</th>
<th>Skindepth at 1GHz (microns)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Silver</td>
<td>10.5</td>
<td>18.8</td>
<td>1.58</td>
<td>2.00</td>
</tr>
<tr>
<td>Copper</td>
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<td>16.2</td>
<td>1.72</td>
<td>2.09</td>
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<td>Gold</td>
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<td>14.3</td>
<td>2.21</td>
<td>2.37</td>
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<tr>
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<td>23.8</td>
<td>2.72</td>
<td>2.62</td>
</tr>
<tr>
<td>Chrome</td>
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<td>8.5</td>
<td>2.8</td>
<td>2.66</td>
</tr>
<tr>
<td>Zinc</td>
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<td>26</td>
<td>5.95</td>
<td>3.88</td>
</tr>
<tr>
<td>Brass</td>
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<td>21</td>
<td>6.5</td>
<td>4.05</td>
</tr>
<tr>
<td>Iron</td>
<td>7.9</td>
<td>12.3</td>
<td>9.9</td>
<td>-----</td>
</tr>
</tbody>
</table>

Table 4.1.

In order to calculate the skin depth \( \delta \) at another frequency \( f \) expressed in GHz, it is enough to multiply the number of the last column by \( \frac{1}{\sqrt{f}} \). There is no value quoted for iron, and this is true for all ferromagnets, because the relative permeability, on which \( \delta \) depends, varies in proportions that are too large according to the mode of manufacturing or the kind of alloy. Moreover, iron and steels cannot be used as they are to constitute conducting surfaces, nor can brass, aluminum and zinc: they have to be coated with a thick-enough layer of silver or copper (between 3
Within the framework of a policy of drastic minimization of production costs, an attempt was made, in the United States around 1985, to manufacture large VHF and UHF cavities in copper-plated steel sheets. Although the electric performances obtained were acceptable, chemical reactions related to residues of the extrusion flow and the humidity of night condensation caused, in the course of prolonged storage, catastrophic degradations that made it necessary to abandon the process. Brass and aluminum, on the other hand, are currently used with a silver plating. But it is pure copper, without internal surface treatment, which is now the standard material for the envelopes of large cavities in radio-telephony. Even though we could \textit{a priori} fear progressive oxidation and an increase in surface resistance, we have observed upon opening copper cavities that had been enclosed for five years or more, that the surface state in the interior, contrary to the exterior, had suffered no alteration and were still shiny. They were made of pure copper, which is not the case for the standard plumbing pipes we use for the central conductor; it is obligatory for this to be silvered.

4.2.2. Extrusion

We refer essentially to aluminum extrusion, which has progressively replaced the extrusion of copper alloys, for of cost weight reasons, but also because the techniques for extruding aluminum have progressed a lot. We have specialists manufacture models that correspond to the profiles we have drawn and that will be delivered in the form of bars, which will be divided according to necessity. It is evident that we choose this way when we have the perspective of a large series, either because we have before us an important market, or because we have a common product in the catalog that is in great demand. There is nevertheless, in this situation, significant disadvantages linked to the fixing of the head: either we have chosen the screwing, and the standard mode of tightening will lead to intermodulations in power, or we weld it before silvering and this will require possession of an argon chamber whose dimensions will additionally set the limits of those of the cavities.

For commercial reasons, industrial styling no longer tolerates, in modern telecommunication equipment, the appearance of the coupling systems destroying the aesthetics of the rest of the equipment. That is, the computer-science system on one hand, and the purely radio part on the other, which comprises the synthesizers, power amplifiers and modems, these sub-systems needing to present themselves in the form of built-in assemblies in cabinets of the 19” standard. The same is also desired for the power filtering. In this type of cabinet we can accommodate, side-by-side and horizontally, two squared cavities with 223 mm sides. As it happens, this is more or less at the technological limit of patterns of a squared profile, with an
extrusion layer (that is, wall thickness) of 5 mm. We thus see the appearance of another limitation to the employment of extruded profiles for large cavities.

That said, we hardly use except for multiple or grouped small cavities which constitute the basic architecture of the reception filters and duplexers of small power. The flexibility of the fabrication of cavities in thin sheets gives the latter an almost definitive advantage, reinforced by the performance of automatic folding machines.

4.3. Assembly

Concerning the fixing of the head of the cavity to the body, it is important to bear in mind that the location concerned is the region where the magnetic field is most intense, and it is thus necessary to have a perfect link. In HF, every slit, every crack and every badly-performed extrusion is a source of parasite emission, of instability and of non-linearity, and thus of intermodulation. This is even more true for the mechanical link between the tube and the head.

Figure 4.9 shows the cross-section of a cavity head in thin sheet-metal, where we can see the three constitutive parts: the central tube, the plate and the envelope. The parts in contact at A and B are extruded to the wall, without any weld excess, in such a way that the surfaces are fully perpendicular and the angles are precisely right angles. Rounded off welds are not recommended. The plate was formed by pressing, in such way that there is an expulsion of material of at least 5 mm onto the central gap and onto the perimeter to assure a good support both on the tube and on the
envelope, and the weld only has a filling and metallic-continuity role. Next, the tube and the plate are extruded, and then silvering is performed. The HF contacts are then extruded to the end of the fixed tube. We do not weld or silver them along with the rest because electrolytic silver-plating on beryllium-bronze weakens it and makes it lose its elasticity. Besides their electric action, the contacts also assure the mechanical support and the concentricity of the mobile tube.

It is preferred to use the type of contact provided in the form of a strip represented in Figure 4.10, where each pattern comprises a finger which obliges the contact to be made at that precise place and impedes the twist when the piston is made to turn, during the frequency tuning. There exists a great variety of these contacts, and the sole issue we have is having to choose between them. It is also possible to do without the direct contact between the fixed and mobile tubes: it is first necessary that the latter does not prolong the former by more than about 1/5 of its length, and subsequently it is necessary to install guide pins made of Teflon (or a similar material) on the mobile part, to replace the mechanical action of alignment of the HF contacts. We thus obtain a totally smooth operation through the absence of metallic friction; we also save on a component that is delicate to install, but the quality coefficient, despite not being completely degraded, is nevertheless not as good as in the presence of a metallic contact.

Once the head of the cavity is completely assembled with the envelope, there only remains to attach the back. We can weld or not, and the choice to do so is totally free from the moment the electric contact is frank. The assembling by pop rivets (popping) is generally sufficient for a fixed cavity installed in a station, but if we choose to weld it, it is important not to forget to allow for a decompression hole so as to assure the continuity between the air in the interior and exterior. A cavity made of thin sheet, if completely airtight, would deform under the action of
atmospheric pressure variations and at best would have its resonance frequency modified. In fact, a cavity can travel by plane or be rapidly transported in altitude, subjected to large pressure differences in a very short time, and resist them. But even the slightest local variations in atmospheric pressure could result, in the event of total air-tightness, in an additional cause of degradation of a radio system by shifting of the resonance frequency.

There is another reason for the decompression hole, as important as the previous reason, which is linked to atmospheric humidity. HF specialists know it is practically impossible to prevent air humidity penetrating a volume sealed by screwing: whatever the quality of the screwing and the care and the attention given to the mechanical finishing, water vapor might take time, but it will eventually enter. Humidity has an influence on the value of $\varepsilon$ and on the refractive index, and thus on the velocity of propagation and, ultimately, on the resonance frequency. It is an additional complication, but for sure the toughest because we do not of know any simple method by which to avoid it. The nightmare for radio systems manufacturers is to have to install their equipment in a tropical zone at altitude, and not far from the sea, because it is there that we have the greatest and fastest hygrometric variations. In this case the decompression hole will maybe accelerate the penetration of humidity, but it will also allow it to leave the cavity just as rapidly and so minimize the period necessary for it to return to the initial conditions. The theoretical solutions to the humidity problem are so delicate that none are applied in the construction of the cavity itself. The best method is still to place the whole radio system in an air-conditioned room, which is not always possible for several fairly obvious reasons.

4.4. Temperature stability

4.4.1. Compensation for ambient temperature

Let us suppose a coaxial cavity made entirely of copper, working at 500 MHz with a resonator of electric length 75°. Its real length (we are referring to the central conductor) is $l = \frac{75}{90} \times \frac{3 \times 10^4}{4 \times 500 \times 10^6} = 0.125$ m, that is, 12.5 cm. Assuming that for small variations the frequency is inversely proportional to $l$, we have:

$$\Delta F = -F \cdot \frac{\Delta l}{l}.$$  

The expansion coefficient of the copper being $16.2 \cdot 10^{-6}/^\circ C$, this means that for each $1^\circ$ increase in ambient temperature, there is a corresponding relative extension...
\[ \frac{\Delta f}{f} = 16.2 \times 10^{-6} \] and thus a decrease in frequency corresponding to \(5 \times 10^8 \times 16.2 \times 10^{-6} = 8,100 \) Hz.

To illustrate the idea, let us try to evaluate the shift in a concrete case. The frequency gap between two adjacent channels of a radio coupling system being of the order of 100 to 200 kHz, let us take a mean value of 150 kHz. Meanwhile we will consider that the normal value for the ambient temperature is 20°C with a possible range of variation (which is usually adopted as standard) of ± 40°C. We thus see a frequency shift of \(40 \times 8,100 = 324 \) kHz, which corresponds to a shift of two channels! If we focus on radio channels, and no longer on coupling channels; in this case it would no longer correspond to a drift of 2, but of 20 channels: this is catastrophic, because the radio system can no longer function. This numerical example shows that we cannot ignore the problem and that it is absolutely indispensable to treat it.

If we refer to Figure 4.1, we will recall that the central conductor is in fact composed of two parts, a fixed one with a length usually about two thirds of the electric length, and the other, mobile by means of rotation, possibly downshifted, of a central rod that allows fine and continuous frequency tuning. We fully understand that it is the material used to manufacture this rod that will define the apparent expansion of the piston as a whole, and we will be lead to turn towards reputed alloys, insensitive to temperature variations, such as Metalimphy products: ADR, N 42, dilver, invar. It is the latter, better industrialized and above all available in rods (which therefore allows it to be threaded) that is chosen by almost all manufacturers. Invar is an Fe/Ni alloy with a very large proportion of nickel (36%). It is broken down into four variants: standard, superior, geodesic and cryogenic, but only standard invar is commonly used in the application that is of interest to us. Its maximum expansion coefficient being 1.2 ppm/°C instead of the 16.2 for copper, the shift per °C will change from 8,100 Hz to 600 Hz for our 500 MHz cavity, which gives us a 24 kHz shift for a variation of 40°C. Does that suffice? This depends, of course, on the objectives we have set, and the selectivity to which we have tuned the cavities. But it is also necessary to take into account other parameters that are more or less controllable, which will add to the thermal shift, such as the method of fixing the cavity to its support, the nature of the support itself, the mechanical hysteresis due to friction, the differential expansion of different parts of the cavity in the event where it uses (which happens often) different materials, such as brass and copper, etc.

In fact, with respect to all of these second order actions, viewed as a whole they represent parametrical chaos that is rapidly inextricable. Above all with respect to the thousands of measurements taken in climatic chambers, we set, for low VHF (although is also more or less valid for the other frequency bands where cavities are...
used) a goal of ± 10 kHz for an ambient temperature variation of ± 40°C, all parameters considered. This is a shift value that is both satisfying and executable but, based on the numerical values mentioned above, we can readily see that simply using the invar is not theoretically sufficient to achieve it.

![Diagram showing expansion (Δθ > 0)](image)

**Figure 4.11.**

However, there are solutions. The most widely used is simple to understand if we take into consideration the parasitic capacitance that exists between the end of the piston and the back of the cavity. Let us consider Figure 4.11, which uses the standard architecture of the band-pass coaxial cavity. We have located, in relation to a reference plane marked O, corresponding to the base of the central conductor, certain noteworthy points A, B and C, which by themselves define the essential elements of the cavity. In all that will follow, we will, in addition, adopt the following conventions that will allow us, as we do in optics, to systematize the presentation of the phenomena related to temperature variations. First of all, we will always remain in the situation of a temperature elevation, supposing that the elongations become contractions when the temperature decreases. We will also suppose that variations in length are proportional to temperature variations, which is only true for a first approximation. Then, we will always draw the cavity as represented in Figure 4.4, the head to the left and the back to the right, with expansion towards the right evaluated based on reference plane O, considered set.

Let us suppose to begin with that the distance OA is small enough for us to ignore it, and that the central conductor is fixed on the plane O. When the temperature increases, planes B and C are displaced to the right, C further than B since it depends on the expansion of a piece made of copper, whereas B is moved by
the elongation of a piece made of invar, of smaller dilatation. As a consequence, the distance between planes B and C increases, and the back capacitance constituted by these two conducting planes decreases. If we call this capacitance $C_f$ and the capacitance of the coaxial section $C_l$, the capacitance $C$ that intervenes in the Thomson formula ($LC\omega^2 = 1$) is such that

$$\frac{1}{C} = \frac{1}{C_l} + \frac{1}{C_f}.$$ 

As a result, whenever $C_f$ decreases, $C$ decreases as well and $\omega$ tends to grow, contrary to the global effect of expansion, which usually causes a decrease in frequency. It is thus possible to ensure that one variation compensates the other exactly, adjusting $C_f$ by the BC dimension. It is a widely-used procedure, which allows the manufacturer to proudly declare that their cavity is “perfectly compensated in temperature”. In fact, perfection will require a supplementary study because there are two remarks to be made: first, the compensation is valid for one frequency alone, whereas the cavity systems are generally in bands. Second, and foremost, the value of $C_f$ – therefore the position of the back – must be compatible with the power that will cross the cavity, or in other words, the back must be at a sufficient distance from the end of the piston so that there is no arc between them.

Figure 4.4 gives a complete solution to the problem, concerning variations of the ambient temperature, and we insist on this precision, which will be justified in the following section. We have previously left to one side the modification of fastening OA by supposing it to be null. In reality this is not the case, and it is actually this that will enable a valid compensation to be performed over an entire band. The system is not complicated and it is of negligible cost, since it consists of simply moving the tube fastening on its plate: when the invar expands and pushes B to the right, the portion of copper or brass OA pushes A to the left and, according to its magnitude, to a lesser or greater degree opposes the action of the invar rod that acts in the other direction. We will choose the excess so that the length of the central conductor is constant in relation to the reference plane when the temperature varies. The capacitance $C_f$ still exists and will be added, as a parameter, to the preceding device. However, the latter is of first order, whereas $C_f$ itself is a corrective term of $C$, and so of second order, which pushes its own variation down to the third order. However, we will have to take it into account in the same way that all secondary parameters, at the point of global adjustments that accounts for the whole. The only disadvantage of the procedure is the slightly increase to the overall length of the cavity. This augmentation is, in addition, easy to calculate: if we use brass and invar, with expansion coefficients equal to 21 and 1.2 ppm/°C respectively, there is compensation when 21.OA = 1.2.OB, from where we derive OA when we know OB. We thus see that by using this system we could apply another system of alloys and replace the invar, which is therefore not indispensable, by a more ordinary material, less expensive and of surer supply, such as for example a martensitic stainless steel with a coefficient close to 12 ppm/°C. The tower, which is how we
refer to the OA excess, will of course be 10 times longer than with an invar rod, but it is simply a question of the available space, and one must consider this for high UHF or for laboratory cavities. Another advantage is that the expansion coefficient will be better defined than that of the standard invar, for which the indicated value is not a mean value, as for the other alloys, but a maximum guaranteed by the steelworker.

The procedure described above has in fact been known of for a long time in watch-making, in an application that carries the name of Leroy pendulum, by which we compensate variations in the length of the pendula in this manner (Boutaric, *La chaleur et le froid*, Flammarion 1927, p. 94). It has been widely tested and it is probably the simplest and most efficient method, but it is not the only one. On the other hand, all systems of this kind, based on thermal phenomena, have a common defect, which is the response time, from which we cannot free ourselves except by one method: automatic frequency correction.

4.4.2. Compensation for internal heating

The specifications relating to equipment involving cavities have for a long time been content with specifying the acceptable reversible degradations, in frequency and amplitude, as a function of the variations of ambient temperature. This was the case until someone, probably a bit more curious and patient than others, one day noticed an additional frequency shift, of the order of –40 kHz, on a UHF cavity submitted to its signal at nominal power for a long enough time, that is, largely greater than what we usually use to verify variations of loss, return loss and rejections, as well as intermodulation values. The phenomenon in fact starts to manifest itself after about a quarter of an hour, to reach stability after about one hour of activity, and its explanation is simple.

The zone of maximum current is situated at the foot of the central conductor, where the magnetic field is most intense, and it is consequently in this zone that the conductive parts, in particular the tube itself and the entry-exit coupling loops, will heat the most. The generated heat will propagate according to the principles described by Joseph Fourier in his "analytical theory of heat", and in particular it will prefer the paths of least thermal resistance, and drain off there by conduction and radiation. Starting off from the foot of the tube, there will therefore be a thermal flux towards the head plate and another towards the end of the piston. At this last point the conduction will stop, and the heat will accumulate, while it will escape relatively easily in the other direction. A very regular temperature gradient will then be established along the piston, with a maximum at the end, once thermal equilibrium is attained. *In situ* measurements are delicate and inaccurate, but we have nevertheless been able to estimate the temperature difference between the foot
of the piston and its end to be of the order of 30°, or maybe more. It is evident that the problem, which concerns all power cavities, must be treated.

Figure 4.12 shows a procedure successfully used in UHF cavities having the following original characteristics:
- volume: 6 dm³,
- frequency: 460 MHz,
- length of the tube: 17 cm,
- drift: –40 kHz,
- power: 50 W.

Instead of directly fixing the copper mobile piston (A) to the end of the invar rod, we do so by means of a manufactured aluminum part (B). The latter is rigidly connected to the invar rod, and the piston is extruded above, what implies a prior appropriate surface treatment, copper-plating, for example. The system then works by the same principle as the one used for compensation of the ambient temperature, that is, it uses a pair of metals with different expansion coefficients to create either an expansion or a contraction as a function of the elevation of the temperature. In this case it will be a contraction, since it is necessary to fight against a decrease in the resonance frequency, and thus reduce the electric length of the central conductor in compensation. Aluminum has a coefficient of 24 ppm/°C and copper 16 ppm/°C, and we understand very easily by looking at the sketch, that the aluminum will win and the head of the piston will recoil towards the left when the ensemble heats up.
We have previously remarked that this phenomenon needed a considerable time to set in, and that it was thus ignored for a long time. It is important to notice that the heat has only two ways to arrive at the end of the piston: the HF contacts and the invar rod. Both are poor conductive media, the former due to its weak support surface, and the latter due to its low coefficient of thermal conductivity: invar is a bad conductor of heat.

The procedure is efficient from the point of view of its realization, and it is this (along the same lines as the compensation for ambient temperature variations) that renders the whole industrially homogenous. But it is not any more unique than the latter, and the number of variations is limited only by the extent of creativity of the engineers. A spring, for example, could serve this purpose, as could a bi-metallic strip or any other thermally deformable mechanical assembly. In all cases, the system of internal compensation will add itself to that of external compensation and participate in it. There will thus be a different regulation of the different mechanical dimensions, an adjustment, but both systems are compatible and together offer excellent overall stability.

4.4.3. Various remarks. Humidity effects

The essential systems described above are all victims, if we can say so, of the laws of heat propagation. They have a reaction time that make then inefficient for the rapid variations of ambient temperature. It is a principle accepted by users, and this is why the measurements performed to verify specifications for temperature resistance are long: we measure in hours the increase and decrease times of the climatic chambers, and also in hours the stabilization times at the different levels defined in the specifications. But we also keep the advantages of an entirely passive device that can thus, theoretically, be installed permanently in places difficult to access and seldom visited. In fact, it is very much a relative advantage if we consider the fact that this equipment always accompanies a radio system that, even if much less sensitive to variations in temperature, needs on the other hand, an energy source in order to work. From the moment this source exists, we can use slave cavities. These cavities are equipped with an electric engine that activates the tuning rod through a reducer. They can work on open loop, in which case they are simply remote-controlled cavities that will be placed at a predetermined frequency thanks to a binary instruction, the engine being of the step-by-step kind. Or alternatively, they can work on closed loop, that is, truly under control. In this case it is necessary to insert into the normal operating program, sequences where we are going to introduce a signal to it, measure the reflected power at a specific frequency and control the cavity on the minimum return loss. We are dealing then with a particular class of products, where the price is entirely different to that of passive cavities, and that correspond to special applications where the additional cost is not a problem. On the
other hand, these cavities are almost entirely insensitive to any ambient variations that, whatever their nature may be, are instantly corrected.

The behavior of a standard cavity during an elevation in ambient temperature always proceeds in the same way. The envelope, subjected first to the perturbation, expands first, including the external part of the central tube which we have called the tower. The back capacity thus reduces at the same time as the invar rod is pulled towards the exterior and shortens the piston. As a consequence, the frequency increases rapidly, and then more slowly as the heat progresses to the interior and progressively reaches the end of the tube, to pass through a maximum and return afterwards, slowly, to its initial value. Once the cavity is built, its response time is determined. If the ambient temperature varies slowly, it will follow with a delay but this will be acceptable. If the variation is abrupt, there will be problems, but nothing can be done.

Another enemy of resonant cavities is humidity. A test of humid heat that is applied to equipment that has to work in tropical regions has brought to light a drift of the same magnitude as that produced by the heating itself, that is, about –40 kHz for an UHF cavity of 6 dm³ at 460 MHz, compensated in temperature, under the following climatic conditions:

– temperature 40°C,
– relative humidity 90%.

The relative humidity RH, or hygrometric state, is defined by the ratio \( \frac{p}{p_s} \) between the partial water pressure in the atmosphere at a certain temperature and the pressure of saturating water vapor at the same temperature. It is thus a quantity comprised between 0, for a perfectly dry atmosphere, and 1 for an atmosphere completely saturated with humidity.

Humid air contains water molecules in vapor phase. The mix has thus a refractive index whose value is comprised between 1, which is the index of the vacuum or, practically, that of the air, and 4/3, which is that of liquid water. The speed of propagation of the electromagnetic waves is thus theoretically comprised between \( c_0 = 300,000 \text{ km/s} \) and a value that has to be greater than \( 3/4 \) of \( c_0 \), that is 225,000 km/s, but in fact little different from \( c_0 \) if we regard humid air as dry air with the physical constants simply having been altered by a small amount due to an inflow of water in the vapor phase. From this point of view, we can equally say that the speed of light in humid air will be reduced by a small amount annotated \( \Delta c \), which we will endeavor to evaluate.
Moreover, it is possible, in complement to the definition given above, to quantify the humidity of the air by the mass of all the water molecules contained in a given volume of humid air at saturation point. We thus define a water mass per m³ of humid air at normal pressure (A), and a water mass per kg of dry air at normal pressure (B). The values as a function of the temperature, taken from the *Manuel de l’ingénieur* by Tideström (Dunod), are given in Table 4.2 and in Figure 4.13.

![Figure 4.13.](image)

<table>
<thead>
<tr>
<th>θ (°C)</th>
<th>-10</th>
<th>0</th>
<th>10</th>
<th>20</th>
<th>30</th>
<th>40</th>
<th>50</th>
</tr>
</thead>
<tbody>
<tr>
<td>A (g/m³)</td>
<td>2.14</td>
<td>4.84</td>
<td>9.39</td>
<td>17.34</td>
<td>30.37</td>
<td>51.18</td>
<td>83.0</td>
</tr>
<tr>
<td>B (g/kg)</td>
<td>1.62</td>
<td>3.83</td>
<td>7.74</td>
<td>14.88</td>
<td>27.55</td>
<td>49.52</td>
<td>87.5</td>
</tr>
</tbody>
</table>

*Table 4.2. Proportion of water vapor in humid air*
If we assume a proportionality law between the refractive index and the relative humidity, we can then establish a forecast variation curve of the speed of propagation of electromagnetic waves in the saturated air \((p/p_s = 1)\). The evaluation method is the following: let us consider, to clarify, an immaterial cylinder of 100% humid air (Figure 4.14) in which a plane EM wave propagates. For simplicity’s sake, we will suppose that the cylinder has a length of one meter and a cross-section of 1 m². We know that the quantity of water vapor contained in the cylinder is equivalent to the quantity of liquid water given by the curve above, and we will suppose it to be concentrated in a disk of thickness \(\Delta l\) and index of \(4/3\), the rest of the cylinder being filled with dry air of an index equal to 1, volumetric mass equal to 1.293 kg/m³, and of which a mass of 1 kg corresponds, as a result, to a volume of \(1/1.293 = 0.773\) m³ (Figure 4.14). This said, we can state that when the air is humid, the optical trajectory corresponding to the crossing of the cylinder is increased, in relation to that of dry air, by the thickness of the slice of liquid water that we defined above. There results a simultaneous variation in the time of propagation \(\Delta T\), the apparent index \(\Delta n\) and the velocity \(\Delta c\) such that:

\[
\frac{\Delta T}{T} = \frac{\Delta n}{n} = -\frac{\Delta c}{c} = \frac{4}{3} \times \frac{\Delta l}{l}
\]

The curve in Figure 4.13 indicates that, for saturated air at 40 °C, the water content per kg of dry air is of 49.52 g, that is, a volume of 49.52 cm³. Given a cross-section of 1 m², we have \(\Delta l = 49.52 \times 10^{-6}\) m and \(l = 0.773\) m.
From which we obtain the relative variation:

\[ \frac{\Delta n}{n} = \frac{49.52 \times 10^{-6} \times 4}{0.773} \frac{1}{3} = 8.54 \times 10^{-5} = -\frac{\Delta f}{f} \]

At a frequency of 460 MHz, a cavity should thus show, in saturated humid air at 40°C, a shift of \(-460 \times 10^3 \times 8.54 \times 10^{-5} = -39.3\) kHz. The recorded measurement of \(-40\) kHz in 90%-humid air, which we have already noted, would suggest about \(-45\) kHz with \(\text{RH} = 100\%\): we are thus not so far from the calculated value. It is enough to introduce a correction coefficient of 1.1 and we will obtain an excellent correspondence between measurement and calculation. Finally, we will write the following formula, which gives the shift of a cavity:

\[ \Delta F_0 = -1.1BF_0 \]

\[ B \text{ in g/kg; } \]
\[ \Delta F_0 \text{ in Hz; } \]
\[ F_0 \text{ in MHz} \]

[4.1]

For a humidity below 100%, it is enough to multiply the value found on the curve below, normalized in frequency to \(F_0 = 1\) MHz, by the RH index written in decimal form.

Example: \(F_0 = 220\) MHz, \(\text{RH} = 60\%\), \(\theta = 50\)°C: the shift will be \(220 \times 0.6 \times (-166.02) = -21,915\) Hz, that is \# 22 kHz.

It is evident that knowing how to describe the problem does not lead us automatically to the remedy. Either way, this problem, seen in common with the other types of frequency shift, is regulated globally by the use of slave cavities: these compensate regardless of the nature of the parameters. Only systems with large study funds can benefit from it, whereas for the others, with standard cavities, we do not know how to proceed, unless by completely climatizing the place where the equipment is located.

In the normal case, either the ambient humidity is permanent and we tune the cavities accordingly, or it is occasional and one or two decompression holes will allow it to disappear as quickly as it arrived.
Table 4.3.

<table>
<thead>
<tr>
<th>$\Theta$ (°C)</th>
<th>-10</th>
<th>0</th>
<th>10</th>
<th>20</th>
<th>30</th>
<th>40</th>
<th>50</th>
</tr>
</thead>
<tbody>
<tr>
<td>$\Delta F_0$ (Hz) for $F_0 = 1$ MHz</td>
<td>-3.07</td>
<td>-7.27</td>
<td>-14.69</td>
<td>-28.23</td>
<td>-52.27</td>
<td>-93.96</td>
<td>-166.02</td>
</tr>
</tbody>
</table>

Remark

There are *a priori* two pathways to follow in order to study the effect of atmospheric humidity on the velocity of propagation of EM waves in air: the one we have just used, which brings into play the refractive index, and another one, which should lead us to an identical result by recalling that the velocity we are looking for is given, according to Maxwell, by the formula:

$$
c = \frac{1}{\sqrt{\varepsilon \mu}} = \frac{c_0}{\sqrt{\varepsilon_r}}$$

putting $\mu_r = 1$. 
Starting from the value shown in the tables, \( \varepsilon_r = 81 \) for the water, we would arrive at a propagation velocity for light equal to \( c/9 \), that is, about 33,300 km/s, while the refractive index of 1.33 leads us to the experimentally verified value of 225,000 km/s. There is thus an awkward theoretical problem, mentioned by Yves Rocard in his work on electricity (Electricité, Masson, 1966, third edition, p. 379). The author notes the inconsistency of the numbers but concludes stating that when we measure \( \varepsilon_r \) not statically, as usual, but at the used frequency, a datum that the author does not provide, there is again concordance and we arrive at the law \( \varepsilon_r = n^2 \). This way of getting out of an embarrassing situation seems somewhat artificial and rather quick. It could, on the contrary, indicate that there is a real problem with the definition and above all that \( \varepsilon_r \), if it depends on the frequency, is not a characteristic of the material. This accident, and this continuity solution in the apparently unfailing structure of electromagnetism, is not unique and should remind us of the maxim that has the strength of a law in the scientific domain: prudence and modesty.

### 4.5. Cavity tuning

#### 4.5.1. Insertion loss and selectivity

![Figure 4.16.](image-url)
The response curve of a cavity of electric length $h$ is that of a resonant circuit of distributed constants. It represents an infinite number of maxima of transmission for frequencies such as $f = (2k + 1)\frac{c}{\lambda} = (2k + 1)\frac{c}{4h}$ separated by zones of large attenuation. In general, only the first maximum is desired and used (Figure 4.16), and we try to get rid of the others by means of a complementary low-pass device that only allows the frequency referred to as fundamental to pass. Two quantities are enough to entirely characterize the response curve of a cavity: the passing band and the insertion loss at $F_0$. For a given a passing band, this will be better the lower the loss; for a given loss its quality will be synonymous with a large selectivity, that is to say, of a narrow 3 dB band.

The passing band is defined as the frequency interval that separates the two points of the curve, on each side of the maximum $M$, for which the insertion loss is 3 dB greater than that of the maximum. We designate it by $BW_{3\text{dB}}$. The other important notations are:

- quality coefficient in vacuum $Q_0$ or $Q_u$ (unloaded $Q$),
- quality coefficient with load $Q$ or $Q_L$ (loaded $Q$).

If we return to the general representation of a resonator by an RLC circuit in series (see section 3.3.1, Figure 3.4), we know that its impedance is:

$$Z = R + j\left(L\omega + \frac{1}{C\omega}\right),$$

which we can rewrite in the form:

$$Z = R\left[1 + j\frac{L\omega}{R}\left(\frac{\omega - \omega_0}{LC\omega_0}\right)\right] = R\left[1 + jQ\left(\frac{\omega - \omega_0}{\omega_0}\right)\right],$$

as a function of the two equalities $LC\omega_0^2 = 1$ and $Q = \omega_0/R$. The quantity $\left(\frac{\omega - \omega_0}{\omega_0}\right)$ is called dissonance, a name that seems a little strange in the electricity domain but that is in fact derived from acoustics, which was the first branch of physics, relating to the general theory of the dynamics of vibrations, to be theorized. For short intervals between $\omega$ and $\omega_0$, we will put $|\omega - \omega_0| = \Delta \omega = 2\pi \Delta F$, which leads us to the expression:

$$Z = R \sqrt{1 + 4Q^2\left(\frac{\Delta F}{F_0}\right)^2} \text{ with } \frac{2\Delta F}{F_0} \approx |\frac{\omega - \omega_0}{\omega_0}| \quad [4.2]$$
If we refer to the current in the load at resonance as $I_0$, when the cavity impedance is reduced to $R$, and $I$ is the current at a neighboring frequency $F$, we have:

$$\left|\frac{I_0}{I}\right|^2 = 1 + 4Q^2 \left(\frac{\Delta F}{F}\right)^2$$  \[4.3\]

This quantity is equal to 2 for the two points of the curve which define the passing band at 3 dB, from which the following relationship is derived:

$$Q = \frac{F_0}{BW_{3\,\text{dB}}}$$  \[4.4\]

To be able to calculate the characteristics that interest us, we will now return to the equivalent circuit, by completing it with external elements, which gives the diagram of Figure 4.17.

Let us recall that the transformers $T_1$ and $T_2$ represent the entry and exit loop respectively, the in-out coupling values being represented by the transformation ratios $n_1$ and $n_2$, which generally have the same value $n$ if the two loops are tuned in identical ways, which is recommended. We can also use a diagram derived from the preceding one, where we have taken back to the interior of the cavity, in the high-impedance link, the external elements transformed by $T_1$ and $T_2$ (Figure 4.18).

These two diagrams, which we will use indiscriminately by choosing the best adapted to the problem in question, will allow us to completely characterize any given cavity. They each represent the real case of a loaded cavity driven by a source, as, for example, when it is connected to a network analyzer.
Let us now return to relationship \([4.3]\). If we no longer give \(I\) the particular value \(I_0/2\) but an arbitrary value corresponding to an arbitrary frequency \(F = F_0 \pm \Delta F\), this same formula will give us the attenuation \(A\) at this frequency. Putting

\[
\left\{ \frac{I_0}{I} \right\}^2 = x = 1 + Q^2 \left( \frac{2 \Delta F}{F_0} \right)^2,
\]

we have:

\[
A_{\text{dB}} = 10 \log x = 10 \log \left( 1 + \left( \frac{2 \Delta F Q}{F_0} \right)^2 \right) \quad [4.5]
\]

Numerical example: a cavity tuned to 460 MHz with a passing band of 170 kHz. What is the rejection at 465 MHz?

We have: \(Q = 460/0.17 = 2705.9\) and \(\Delta F = 5\) MHz

\[
A = 10 \log \left( 1 + \frac{10.2705.9}{460} \right)^2 = 35.4 \text{ dB}.
\]

We note that for the usual values of \(Q\) the quantity \(\left( \frac{2 \Delta F Q}{F_0} \right)^2\) is large compared to 1. We can thus simplify formula \([4.5]\) and finally write that which, applied to the previous example, gives the same result:

\[
\text{Attenuation at } \Delta F = A = 20 \log \left( \frac{2 \Delta F}{BW_{\text{sub}}} \right) \quad [4.6]
\]
The insertion loss is by definition the ratio $\alpha = P/P_0$ between the powers transmitted to the load with $(P)$ and without $(P_0)$ the cavity: $\alpha_{dB} = 10 \log \frac{P}{P_0} = 10 \log P - 10 \log P_0$. Let us now try to evaluate $\alpha$ with the aid of Figure 4.18 where we forget about $L$ and $C$ at resonance and we make, for the sake of simplicity, $n_1 = n_2 = n$. In the absence of a cavity, the current in the load $n^2Z_0$ is: $I_0 = \frac{nV}{2n^2Z_0}$. With the cavity in the circuit it instead: $I = \frac{nV}{R + 2n^2Z_0}$. We can thus write:

$$P = n^2Z_0 \left( \frac{nV}{R + 2n^2Z_0} \right)^2$$

and $P_0 = n^2Z_0 \left( \frac{nV}{2n^2Z_0} \right)^2$, from where we obtain:

$$\alpha = \left( \frac{2n^2Z_0}{R + 2n^2Z_0} \right)^2$$

In addition, it is possible for us to express $Q$ and $Q_0$ as a function of the components of the circuit: $Q_0 = \frac{L\omega_0}{R}$, $Q = \frac{L\omega_0}{R + 2n^2Z_0}$. Combining with the preceding values, we obtain:

$$\alpha = \left( \frac{1}{1 - Q_0/Q} \right)^2$$ [4.7]

In this relationship between $\alpha$, $Q$ and $Q_0$, the measurable quantities are $\alpha$ and $Q$. The usefulness of this formula is therefore, initially, to calculate $Q_0$ which is not, itself, directly accessible to the measurements. Rearranging the parameters, we arrive at the following practical identity:

$$Q_0 = \frac{Q}{1 - \frac{n}{10^{30}}} = \frac{F_0}{BW_{3dB}} \times \frac{1}{1 - \frac{n}{10^{30}}}$$ [4.8]

It is important to pay attention to the sign of $\alpha$: a loss of 1.2 dB is written $\alpha = -1.2$ dB.
Numerical example: a cavity is tuned with a band of 170 kHz and a loss of 1.2 dB at a frequency of 465 MHz. We must then find $Q_0 = 21,198$.

### 4.5.2. Loop tuning

The method used in the previous section, which consists of exploiting equivalent electric diagrams, is very prolific and allows us to solve, from a theoretical point of view, a considerable quantity of problems related to the cavity itself and to the grouping of cavities. We cannot tackle all of them, and the examples dealt with above should suffice for an initiation in how to proceed, and the potential complementary results will be given without demonstration, this no longer being essential.

Along these lines, we also show that the VSWR can be written as:

- **entry** $S_1 = \frac{R + nZ_0^2}{n^2 Z_0}$,  
- **exit** $S_2 = \frac{R + nZ_0^2}{nZ_0}$.

To clarify for $S_1$:

At resonance the R, L, C circuit of Figure 4.18 reduces to $R$. The total circuit can be considered as a generator of internal impedance $n^2 Z_0$ producing by means of a line of null length and characteristic impedance $nZ_0$ in a load impedance $R + nZ_0^2$. The reflection coefficient seen by the generator is thus: $\gamma = \frac{Z_t - Z_0}{Z_t + Z_0}$, and the VSWR is $S_1 = \frac{1 + \gamma}{1 - \gamma}$, that is:

$$S_1 = \frac{1 + \left(\frac{R + nZ_0^2 - nZ_0^2}{R + nZ_0^2 + nZ_0^2}\right)}{1 - \left(\frac{R + nZ_0^2 - nZ_0^2}{R + nZ_0^2 + nZ_0^2}\right)} = \frac{R + nZ_0^2 + nZ_0^2 + R + nZ_0^2 - nZ_0^2}{R + nZ_0^2 + nZ_0^2 - R - nZ_0^2 + nZ_0^2} = \frac{R + nZ_0^2}{nZ_0^2}$$

The same applies for $S_2$. If the two loops are symmetrically regulated, we have: $S_1 = S_2 = 1 + \frac{R}{nZ_0^2}$. It results that the VSWR are always greater than 1, that is to say, that there can be no perfect adaptation, but also that the larger the cavity, the better adapted it is, which experience confirms: a UHF cavity of 6 dm$^3$ has a $Q_0$ of about 10 to 11,000, a 3/4 wave cavity with 223 mm sides at the same
frequency has a $Q_0$ that surpasses 17,000. Moreover, once we have managed to regulate the return loss symmetrically, the fact of increasing one will diminish the other (and vice versa), which allows to anticipate the relationships above.

An important remark should be made about the respective position of the loops: we can tend, when we draw the cavity (and this is, moreover, a layout we encounter often), to place the two loops so that they are diametrically opposed, with the central tube blocking them from each other, so as to avoid any direct radiation. There is no use to this: one needs to know that in a cavity the loops do not radiate; they are not antennas, but transductors that allow the incident signal to transmit its energy to the dielectric volume, which resonates when the frequency is adequate. The consequence of that is that we can, from this point of view, put the loops wherever we want, and as close to each other as we want; and comparative measurements have established the invariance of $Q_0$ in this case.

Another remark concerns the loops themselves, and their dimensions. The loops being placed in principle where the current (in the electromagnetic sense of the term) is the strongest – that is, close to the base of the central tube – they must be made of a strap or of sufficiently-thick, silver-plated ribbon, without surface metallic discontinuity to avoid intermodulations. To clarify, we usually use silvered brass of 1 to 2 mm in diameter for powers from 50 to 100 W.

If we are dealing with deformation we will take smaller diameters (1 to 1.5 mm), whereas if they are rotation loops we can take them to be thicker. Figure 4.19 shows a conventional form, to the right the letters M and m indicate the orientations corresponding to the maximum and the minimum coupling respectively. It seems a priori that the rotation loops are easier to tune: with the help of a reference on the fixed connector we know at all times which the direction of the loop is in the interior of the cavity, but one must not forget that there is a cable attached to the connector and that its rigidity, even if it is of a type said to be flexible, requires us to proceed by small angular variations, constantly alternating the tightening of the loop and of the cable. The deformation loop, which is linked to a connector attached rigidly to the cavity, does not have this disadvantage, but on the other hand necessitates a special tool, that rather resembles a snail fork, as well as blind work which demands great ability and long experience. It is necessary in fact, alternatively, to monitor the network analyzer to know at each instant where we are concerned with the return loss and the passing band, and to give the little stroke that will make the tuning progress in the right direction, and this with full knowledge of the situation: we have understood, it is a craft!

In this type of operation, we will appreciate the true value of the type of analyzer that at the same time gives the entry and exit parameters $S$, allowing the symmetry of the return loss to be controlled, without having to worry about the accuracy
because above all it is a matter of a pre-tuning. That being so, the rules to achieve
the loss and band values that we have set are simple, from the moment we have fully
understood that the coupling is a direct function of the captured magnetic flux, or if
we prefer, of the number of Faraday lines that cross the loop, and that increasing the
coupling of a loop has the following effects:

– increasing the band by 3 dB,
– reducing the insertion loss,
– improving the return loss,
– degrading the return loss of the other access.

In order to increase the coupling we can:

– rotate the loop towards position M,
– deform it in order to bring it closer to the central tube,
– deform it so as to increase its surface.

To reduce it we can, conversely:

– rotate the loop towards position m,
– distance it from the central tube,
– reduce its surface.
It is very difficult, in the case of tuning by deformation, to make the position vary without changing the surface. It is again a question of needing the helping hand of a specialist, an aspect very characteristic of professions relating to high frequencies, and particularly relating to passive equipment. The manufacturer is rewarded for its efforts by the fact that the cavity, which is simpler and of slightly lower cost, can no longer be accidentally detuned by users. If despite this, the latter have a certain amount of skill and wish to tune the cavity themselves, at their own risk and danger, it is sensible to choose a cavity with graduated rotation loops.

4.5.3. Frequency tuning

Until now we have stated that to make the frequency of a coaxial cavity vary, we must proceed with the rotation of the invar rod on which the mobile piston is attached, this being the usual procedure. If we take the method explored in section 3.3.5 to determine the electric length and the effective length of the central conductor of a coaxial cavity, and if we take the case, for example, of a band-pass cavity with a ratio $b/a = 3.6$, with $b = 150$ mm, tuned to a frequency of 400 MHz, we can calculate that the total length $h$ of the piston is of:

$$h = \frac{\lambda}{4} \frac{b-a}{2} = \frac{3.10^{11}}{4.400.10^6} \frac{150-42}{2} = 133.5\text{ mm}.$$  

If we wish to vary the frequency by 10 kHz, it will be necessary to vary $h$ by:

$$133.5 \times \frac{10^4}{400.10^6} = 3.34.10^{-3}\text{ mm} = 3.34\text{ μ}$$

We usually use the invar rod of 10 mm diameter threaded at steps of 0.5 mm. An axial translation of 3.34 μ thus represents a rotation of:

$$360 \times \frac{3.34}{500} = 2.4°.$$  

Knowing that the nominal gap between two adjacent telephone channels in the same coupling band is of 150 kHz in average, we cannot envisage an adjustment precision worse than 1 kHz, that is, an angular uncertainty of a quarter of a degree. In order to better understand the magnitude of this value, it is enough to imagine what it means to tell a driver to turn their steering wheel by a quarter of a degree. But as it is unfortunately necessary to achieve this result, we must get around this difficulty, which will be done by making the system of containment of the invar rod slightly more sophisticated.
Beforehand, it is fundamental to recall certain purely mechanical notions, concerning the fitting in particular. Two pieces are never perfectly adjusted. When there is a male and a female part, in relation to a common theoretical dimension, the real value of the male part will have to be slightly larger and that of the female part slightly smaller, without which there can be no fitting: this is the slick, which corresponds a little to measurement precision in physics, and which results as a consequence in mechanical hysteresis. This is the case particularly for a rod threaded to its tap. In order to have no more play, it is necessary to exert a tightening, and then we are at the heart of a new problem: in order to regulate the cavity, the regulating rod must have some play, but for the adjustment to be stable it is necessary for it not to have. To get out of this delicate situation, there are four practical solutions, each giving rise to infinite variants.

The first method consists of blocking the rod with a lock-nut. It is the simplest procedure, the most evident, the most economical, and was the rule at a time when frequency plans were permanently set in the radio stations and were communicated to coupler manufacturers. The latter would deliver systems regulated in the factory and would be moved on the occasion of rare changes to the frequency plans. The tightening of the lock-nut, which also has the effect of shortening the rod by pushing the thread towards the exterior, and thus making the frequency vary, is part of the adjustment and must be anticipated, giving place to a rather long operation.

The second method consists of adding a “thread brake” to the rod fastening part. It is a second built-on tapped part, crossed by the rod, connected and as close as possible to the main tapping but fitted with spacing screws, which provoke a permanent mechanical tension on the threading by means of opposing actions and thus suppress the play (Figure 4.20a). The rod, terminated by a large adjustment...
button, thus turns with a controlled friction, sufficient to ensure a constant tightening but allowing nevertheless for a manual rotation. Another system, used on small diameters, is shown in Figure 4.20b. We find it, for example, on trimmers (on the mobile part), and in regulating systems of small cavities such as those which contain a ceramic resonator. The two slits represented on the drawing and used on the threaded fastening part are either enlarged or slightly compressed, so that the penetration of the rod can only be done with some effort. The play is then suppressed, there is a constant tightening, weak but enough to assure the keeping in position in the presence of impact and vibrations, and still allowing an adjustment by screwdriver.

The third method is an evolution of the second that consists in adding to it a gearing-down system. By increasing the torque on the rod, this construction allows us to increase at will the permanent tightening, while always keeping the option of manual adjustment. It is a solution adopted for coupling racks susceptible of being frequently re-tuned by operators in important networks, regional or national. It is also a much more efficient solution against vibrations and impacts when we are dealing with cavities of significant mass and volume.

Finally the fourth method makes use of perturbers: a conductive disk is perpendicularly fastened to the end of a threaded rod, forming a kind of mushroom. The rod is itself fastened onto the envelope, either on the head or the back, or even close to the end of the piston. The progressive penetration of the disk in the dielectric modifies the geometrical parameters and makes the resonance frequency vary. We have represented in Figure 4.21, three different positions of the perturber, which correspond to plausible installations. The one marked (a) is the most practical, since it is on the same side as the frequency adjustment. Position (b) forbids the juxtaposition of cavities, and position (c), whilst compatible with this, obliges the person performing the adjustment to apply a somewhat gymnastic action that can be prohibitive. It is necessary to note, besides these technological considerations, that
these placements are not all electrically equivalent. When we screw (a), located in what is said to be the current zone, the disc moves away from the mass and makes the mass, which is globally considered with the disc, act as if it was moving closer to the end of the piston, whose apparent length decreases: the frequency thus increases. By screwing (b) or (c), situated in what is referred to as the voltage zone, the frequency decreases, since the back capacitance increases.

In summary, we have at our disposal a substantial choice of solutions for the adjustment of the frequency, each having its advantages and disadvantages. The first three have the common advantage of not introducing any variations, neither in the adjustment range of the cavity, nor in its quality coefficient. The third has a cost that is about 20% greater than the second but on the other hand it permits a tuning with better resolution and can be easily automated. The fourth solution, which uses a screw disturber which we equip with a brake thread to suppress the play, degrades the quality coefficient, but in a proportion that is usually acceptable. On the other hand, the adjustment, limited to about 1% of the pre-tuned frequency (of the order of 5 MHz for F₀ = 500 MHz), becomes extraordinarily easy and of an incomparable precision.
The ideal band-pass filter allows only a determined frequency band to pass, completely reflecting all the others; also has no losses and a constant group propagation time $d\phi/d\omega$ (phase is a linear function of the frequency). Real filters try to execute in the best possible way one or more of these characteristics, but never all of them at the same time, because the latter, at least, is achieved at the expense of
the others and necessarily leads to a compromise. In order to help solve this problem, mathematical physicists have defined approximating functions that, as indicated by their name, allow a theoretical approach to the practical solution by means of modeling.

The most used approximating functions are those of Butterworth, Cauer (elliptical), Legendre, Gauss, Bessel, and above all Tchebycheff, a name well known in multiple spellings, which in fact covers a family of functions comprising a considerable number of variants. It is necessary in particular to distinguish the Tchebycheff polynomials from the rational functions of the same name, of which the polynomial is a particular case. Here we enter into details that are not really of interest to application engineers, but we will find, in case of necessity or mere curiosity, all the developments in the works of reference such as Zverev (see bibliography).

The polynomial Tchebycheff filters of the first kind are the most used in radio because they possess the greatest out-of-band selectivity, after elliptical filters. Another characteristic is that the poles are rejected to infinity, that is, the decrease of the response in amplitude is continuous on both sides of the passing band, which is not the case for the Cauer filters, which have fluctuations on the rejection band. In the interior of the passing band, the Tchebycheff presents a constant amplitude ripple (equiripple), whose number of maxima and minima corresponds to the order of the function. From now on we will use this approximating function exclusively, which is the best adapted with respect to selectivity and which, as a consequence, is systematically used in the industry of HF filtering, for the design and production of pass-bands. The fact that the group propagation time is not constant in the passing band is not a big problem in the domain of telecommunications, because the width occupied by a carrier is sufficiently narrow for us to consider the phase to be linear as a function of the frequency.

Figure 5.1 shows the form of the response of a Tchebycheff band-pass of the third order. We note that the order of the function corresponds to the number of extremes in the equiripple band of the prototype low-pass and to the number of maxima in the band of the transformed band-pass, and this number will be also that of the resonators used to physically produce the filter. The three loss levels a, b, c indicated in the diagram correspond to the resistive loss, the equiripple band and the return loss, respectively. Let us look at these more closely.

The top of the model of the ideal low-pass represents zero loss. We thus have increasing attenuations which go towards the bottom in a conventional system of amplitude/frequency coordinates resembling what we see on the analyzer. The real filter has a non-null resistance which is translated by a thermal dissipation, and thus by a certain loss: it is the abscissa level a, to which the maxima of the equiripple
band are tangent. The minima of this band are located on level b. This level is chosen, in the filter calculation procedure, as a function of the minimum adaptation required, and more generally as a function of the constraints that we impose on it and that are globally translated, on the books of specifications, by a model. This choice is oriented by conflicting properties of the approximating function: the selectivity, the ripple and the adaptation. The greater the selectivity, for a given order, the more significant the ripple and the worse the adaptation. There is thus the necessity, for each new band-pass project, to define before anything else a minimum adaptation, represented by the level c. The value retained for this leads in fact to the maximum value for the ripple, whose minima correspond to the maxima of the “return loss” curve. Let us take for example a ripple of 0.1 dB = b-a. We deduce:

\[ \frac{P_t}{P_i} = 0.977 \]
\[ \frac{P_r}{P_i} = 1 - 0.977 = 0.023 \]

return loss = 10 log 0.023 = -16 dB

This value is clearly insufficient in relation to the ordinary specifications of the adaptations, which are in principle greater than 20 dB. If we re-perform the calculation starting from a ripple of 0.01 dB, we will find a minimum PSWR of 26 dB, a value which this time is acceptable. But we also know that the price to pay for this improvement is a decrease in the attenuation outside the band, as shown by the different families of curves provided by the reference works, each corresponding to a round value for the ripple.

Remark: we must not confuse the return loss and the resistive loss, which corresponds to an internal calorimetric dissipation and translates by an elevation in temperature of the filter. This has no effect on the value of the adaptation. This means, among other things, that a “bad filter”, that is, a filter with high losses, can be perfectly matched.

5.2. Calculation of a Tchebycheff band-pass

The study of a filter always starts by a drawing of its model (template). This is supposed to graphically represent the amplitude limits between which the response function in frequency must be situated, from 0 to \( \infty \). That is to say that it includes not only the intrinsic curve of the filter, but also all the imposed constraints, whether made by the client, or by the institutional organizations managing telecommunications, or even the designer themselves, if they want, for example, to allow for extra security margins. Figure 5.2 gives the typical appearance of a complete model.
The useful band is comprised between \( F_2 \) and \( F_3 \), the other frequencies specified being of variable quantity depending on the spectral, physical or administrative constraints. Once the model is established, the following step consists of determining the curve of the smallest order which fits into the main model, that is, the one that goes from \( F_1 \) to \( F_4 \). For this, we need to refer to the graphical resources of specialized works (Zverev, Matthei, etc.), where they are usually presented with the attenuation axis inverted, the reference of 0 dB being at the bottom and the increasing attenuations towards the top, which gives the appearance of Figure 5.3. In this example, it is the curve of the 4th order that we will choose: curves 2 and 3 hit against the selectivity template, and the 5th against that of the losses. It is evident that the filter alone will not be able to do everything, due to the periodicity of the systems of distributed constants, which will translate by passing bands with odd-numbered multiples of \( F_0 \). It is thus necessary to complete it with additional devices such as low-pass, which we can either integrate into the filter, or place elsewhere in the emission or reception networks. It is necessary to recall at this point that it is not always worthwhile, during the conception of a complete telecommunications system, to systematically remove the totality of the filtering problems towards the filter manufacturer: it is precisely an element of the project that is almost always worth distributing along the whole emission or reception network, where it is the simplest to integrate it. It is not a question of the filter itself, but complementary filtering elements which allow it to be made compliant with the off-band specifications.
Once the order of the function, that is, the number of resonators, is established, what remains is to choose the technology, which will determine the dimensions of the filter and its performance.

5.3. Technologies

There are as many ways of designing a filter as there are types of resonators, of which we can define, a priori, five large families: discrete elements, planar circuits, ceramics, coaxials and waveguides. The first and last ones mentioned are devoted to the HF and low VHF, and to the SHF, respectively, where the guiding section is reasonable, that is, in practice less than a square decimeter. We are thus left with, for the VHF, the planar circuits, ceramics and air coaxials.

5.3.1. Planar circuits

As their name suggests, these are plane capacitors in the form of thin plates for which one side constitutes the electrical mass and the other carries the resonators constituted of strips, the base of which is joined to the mass. The dielectric is made
either of air or a substrate of insulating material: epoxy glass, Teflon glass, ceramics, etc. to mention the conventional ones, regardless of those the industry will provide in the future, on a permanent effort of evolution towards the perfect product, with very high $\varepsilon_r$ and very low losses. In this technology the filter is made as a printed circuit and will have the same industrial characteristics: reproducibility, low cost, small footprint, but also average quality, even with the ceramic support, and relatively limited power. However, the rapidity of progress is such that we can foresee excellent band-pass products, knowing beforehand that the battle is already won for low and high-passes.

On the other hand, there is an enormous advantage in the usage of metallic substrates, which is the fact that we can just as easily make $\lambda/4$ or $\lambda/2$ resonators, or more generally, line sections of identical length whose properties are well known. Figure 5.4 shows three classical dispositions for coupling by proximity, all three corresponding to a band-pass structure: (a) represents half-wave resonators coupled by proximity, (b) is a disposition referred to as interdigital, with quarter-waves, (c) is a combline structure, always with quarter-waves. The views are on the track side, the other face being entirely metal-plated and constituting the electric mass, to which the mass bands of (b) and (c) are connected by metallic holes not represented in the figure.

![Figure 5.4.](image)

The calculation of this type of filter has for a long time constituted the substance of specialized works (of which Matthei), but this should now change with the appearance and diffusion on the market of electromagnetic simulation programs, such as *Sonnet Suite* or an equivalent, which allow us to obtain, in increasingly shorter times, the forecast response curve of a circuit analogous to that of Figure 5.4.
The use of such programs, which in a way are “false experiments”, gives way to a considerable widening of the fields of investigation. In fact, it is enough to produce a drawing where logic is not broken in order to obtain on the screen a faithful projection of the output of the circuit once it is built, which allows us to come off the beaten track and better explore imagination and intuition.

We spoke above about a “medium” quality of this technology, yet it is compensated by an excellent stability in temperature if we limit ourselves to wide-band filters, or let us say, not too narrow for the band-pass.

5.3.2. Ceramic filters

This technology, launched in the industry by the Japanese in the 1980s, experienced great success before being countered in Europe by the Finn, Lauri Kuokannen, who succeeded in miniaturizing and manufacturing mobile air duplexers on a large scale, slashing prices and taking hold of the market, until the arrival of digital technology, which signaled the end for this type of equipment and the disappearance of the industry. A second emergence came about with the arrival of GSM, with another generation of materials and other industrial forms, to return to the benefits of air models, much more flexible to employ and manufacture. In fact, ceramic resonators had a lot of difficulty setting up in the UHF domain, because air cavity filters offered, for a larger, but generally non-prohibitive footprint, qualities that ceramics will never have. However the situation is reciprocal, because ceramics propose $\varepsilon_r$ around 35 and 90, which permit reductions to dimensions that are of interest. Figure 5.5 shows the two conventional representations that we have available for frequencies comprised between 300 and 3,000 MHz. The stick shape to the left is convenient to filters of small dimensions, such as reception filters, with a dimension $a$ that roughly varies by a tenth to a half-inch. The dimension $L$ depends on the frequency: $L = \lambda/4$ or $\lambda/2$. The resonators are kept grouped side-by-side by a small metallic structure designed to be easily fastened almost anywhere, including
on a printed circuit. We thus obtain a 3-pole pass band filter with sides of about 3 cm for a thickness of 15 mm, whereas an equivalent air filter would be of the order of a 10 cm with a thickness of 25 mm. This being so, the choice is not that evident and it is worth drawing up a comparative assessment of the qualities and the flaws before deciding on the procedure.

Advantages:

– with the materials that we find on the market, the quarter-wave is 5 to 10 times smaller than with the air dielectric. The footprint saving is thus evident in theoretical terms;
– the dilation coefficient is low and fairly stable, to ± a few ppm; it can even be null.

Disadvantages:

– the available sections are small. It is difficult to find sizes larger than half an inch in diameter or in the side measurement;
– the dispersion of the dilation coefficient is not always sufficiently well-defined to suit the narrowest band-pass;
– the decrease in the length of a resonator can cause problems when, in order to have an acceptable quality coefficient, we are led to choose the largest sections: we arrive more quickly, having chosen a TEM mode, at forms where higher modes develop, which means that we do not automatically benefit from the footprint saving that \( \varepsilon_r \) allowed us to expect;
– we depend on the supplier, on their durability and on the choice they propose. It is an industrial risk which is even more real as the specialists are of limited number and are usually very dispersed geographically;
– the tuning devices are more restricted and less precise than with air filters, since we cannot penetrate the interior of ceramics. As regards the adaptation, for example, a return loss of 15 dB constitutes the norm for large series. We can do better if the constructor has planned for zones accessible from the exterior, in which we can make the surface vary to adjust the capacity, but there is nevertheless a lack of flexibility related to the kind of manufacturing;
– the prices of the materials are still too high.

Conclusion

When saving space is primordial, filters with prismatic ceramic resonators are of interest (on UHF) for wide band-pass, very large series and low powers. This, of course, is under the condition of accepting the principle of buying the filters
regulated in advance, ready to be integrated into the sub-assembly, and of not being too demanding regarding the adaptation. Nevertheless, with ceramic sticks it is possible to produce a bandpass, the performance of which is of interest, but which is very small in size compared with an air filter, by installing good quality (that is, air or ceramic) trimmers, to at the same time perform the tuning of the in-out coupling and of the coupling between resonators, and the frequency tuning (Figure 5.6). It is absolutely necessary that these three types of tuning exist on the filter so that we can achieve the optimal global tuning. The cost of such a filter is around three times that of ready-made market filters, because it is necessary for it to have a more elaborate body so that the trimmers can be lodged; these are as expensive as they are small, those devoted to the tuning of couplings often being less than a picofarad as soon as we reach 400 MHz.

Figure 5.6

Figure 5.7
The shape on the right of Figure 5.5 is mostly used in bandpasses of fairly large dimensions, for emission or for reception filters with particularly steep slopes. The ceramic is enclosed in a silvered metallic case and resonates according to a mode, for which we show the magnetic field lines. Figure 5.5 indicates a conventional set-up where the fastening system is not explicit and where the couplings are not represented, in order not to overcomplicate the drawing: we in fact use loops placed perpendicularly to the magnetic field (as for air cavities) in the aerial part of its course. The dimensions of the case are very important and must not be much greater than those of the ceramic, fixed to an insulating stand. A screwed-on disc allows the back capacity of the case to be varied and fine-tuning of the frequency to be assured.

5.3.3. Dielectric air filters

We now come to the oldest, most widespread and most conventional family of UHF filters: dielectric air filters, a name that avoids mention of the word “vacuum”, the ambiguous character of which when it comes to the propagation of EM waves we have already highlighted. Such a filter can be constituted either by isolated cavities connected by coaxial cables, or be presented in the form of a set of compact elementary cavities of squared section, coupled together by openings, or windows in the common walls; it can also consist of an alignment of resonant tubes coupled by proximity: this is the combline, upon which we will particularly insist.

![Figure 5.8](image)

Figure 5.8

The first category mentioned (Figure 5.8) is chosen when there is the need for filters of very high performance and when we have all the space necessary, which
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fairly frequently presents a major problem, with space now being so restricted in technical places. Besides this practical consideration, there is no theoretical limit to the selectivity of these filters, given that it is enough to increase their volume to obtain the necessary $Q_0$. However, there is a practical limit which we have already dealt with, regarding coaxial cavities: that of the quality of the temperature stability. It is evident that there is in fact no point installing an extremely selective bandpass around a frequency to be protected if the slightest variation in ambient temperature causes it to drift beyond the predicted band: the higher the $Q_0$, the more the thermal stability must be looked after and put at the same level as the electrical performance. The coupling is done with the aid of coaxial cables of an electric length close to $k\lambda/2$, which put two consecutive cavities in phase opposition, the final tuning being made by rotation or deformation of the loops, whether it be by entry-exit coupling or inter-cavity coupling.

![Figure 5.9](image)

The compact filters are less bulky, and thus more easily accommodated, and are constituted by the juxtaposition of sections of aluminum profiles, screwed over a tube-carrier plate and capped with a second common plate bearing the frequency tuning screws. The profile (Figure 5.9) is obtained by extrusion from a pattern. This usually remains the property of the extruder; it is generally free and reserved exclusively to the ordering party, who must however place an initial order of around 500 kg to a ton, in the form of sections of length as is convenient, which limits the number of companies who desire to invest in this type of product. On the other hand, once we have several profiles with well-staggered dimensions, we acquire a production flexibility which allows us to respond rapidly to demand. It is enough in fact to group the resonators as shown in Figure 5.9b, screwing them with the aid of self-threaded screws, thanks to the screwing holes planned for in each corner. We note that in the illustration these holes are not closed, and this is a technological necessity that comes from the fact that we cannot extrude otherwise. It is strongly recommended, once the profile has been sectioned, to slightly mill these holes, so as
to avoid the aluminum rolls inevitably caused by the penetration of self-threaded screws that would compromise the direct contact of the profiles up against the plates: interstices in the HF assemblies always provoke an increase in losses and induce non-linearity, and thus intermodulations. For the same reasons, in the coupling windows we have silver-plated pliers which ensure electrical continuity, and thus equipotentiality, between two adjacent profiles (Figure 5.9b).

Figure 5.10 shows a 3-pole band-pass filter, of the compact type with open tubes, provided with three types of tuning that is indispensable to attaining an optimized global tuning:

- the frequency tunings are constituted of blocking screws by lock-nuts which penetrate more or less into the tubes, the whole of which we can consider as cylindrical capacitors. The further the screw penetrates, the more the resonant frequency decreases;
- the coupling tunings between resonators are also performed by lock-nut screws placed in the windows, perpendicularly to the screens. The more the screw penetrates, the more the coupling decreases, this only being valid in this scenario. With helical resonators notably, the effect is the inverse. The depth of the windows is experimentally determined on the prototype;
- the entry-exit couplings are constituted by silver-plated straps called “taps”, which connect the pin of the connectors to a point on the end of the tube. What matters, and really characterizes this coupling, is the quantity of the field captured, that is, the surface comprised between the tap and the tube. The tuning will thus
consist of modifying the geometry of this loop, or of this tap, either by deforming it
or by fastening it at a lesser or greater height on the resonant tube.

We thus come to the best performing model of the band-pass filter: the
“combline”. The importance and interest of this type of filter are such that it is not
superfluous to dedicate a whole chapter to it.
Chapter 6

The Combline Filter

6.1. Architecture

6.1.1. General structure

The word combline is formed by the combination of the terms “comb” and “line”. This pleonastic expression designates a filter where the resonators are aligned in a common plane-parallel space. It is thus no longer a question of juxtaposed profiles: we can say that it is the evolution of a compact filter where we progressively increased the coupling windows, by moving the tubes slightly more apart each time, in order to compensate for the increasing of the coupling, doing so until the walls were completely suppressed. In comparison to the compact filter constituted by the same tubes, it will thus be a little longer, but on the other hand the suppression of the walls implies a decrease of material that will translate into a reduction in losses and weight.

The combline is necessarily a bandpass. Figure 6.1 shows the typical structure of a 4-pole. The connectors are placed on the tube-holder plate, but they could just as easily be found on the sides, to the right and to the left (Figure 6.2): what is essential, from an electrical point of view, is to be able to make a closed loop with the tap, the resonator and the mass. We also note that the taps are not linked to the tubes directly, but by means of a silver-plated strap that runs along the tubes over approximately the lower third of their height, parallel to their axis and very close to it so that it will be under the same potential. The fastening is done easily using a traditional iron, while a small welding torch is needed if we want to fix it directly to the tube. However, in the case of mass production, from the moment the geometry of the tap is well determined and fixed, we can make a direct welding, which would
have the advantage of being silver-plated with the rest: it is a question of a choice to be made by the manufacturing committee.

![Figure 6.1](image1)

![Figure 6.2](image2)

The combline is the best filter we can produce in UHF. The suppression of the walls, in relation to the compact filter, translates into a reduction of the losses that, even if it is not extraordinary, can help to meet specifications. We can count a gain, around 500 MHz, of about 5 hundredths of a dB by resonator, attributable to the suppression of the walls, to be multiplied by the number of intervals (n-1). On the other hand, the direct coupling between the tubes will lead to a larger gap, in such a
way that a combline will always be longer than a compact, the other dimensions being kept the same. By means of this, the combline will be the best and the lightest, because it is the simplest, that is, it presents the least amount of conducting surface.

6.1.2. Fastening the tubes

![Diagram](image)

**Figure 6.3.**

The first practical condition of the quality of a resonant cavity is the electrical continuity, especially in the zones of strong current. The foot of the resonators is thus the most sensitive place, where in return for the slightest discontinuity, the smallest fissure or slightest lack of surface treatment, there is immediately a general degradation, of lesser or greater significance. We have long considered that only welding or brazing could offer a sufficient guarantee on this point. Yet a welding ages, and if not well done (even if this is not apparent) could break in the event of a shock, mainly due to diurnal and nocturnal variations in temperature. In addition, if we want the welding to be silver-plated, which is strongly recommended and may be imperative, this means that it will have to be done in advance, and thus on good welding materials: among the industrial metals, we find ourselves almost confined to using copper and above all brass, which results in heavy filters. It is therefore not the best solution. Other procedures are described below, the first two of which were tested at the time of important productions in series. It is above all necessary to retain the basic principle: pinning the tube base against the tube-carrier plate with an elastic device (spring), which permanently maintains a force pressing one against the other. This principle, if well-executed, assures the durability of the attachment and its quality, not to mention the facilitated set-up and modification that depend selectively on the variants used, which are very numerous.
Figure 6.3 illustrates four examples of possible procedures on which we can comment as follows:

- 6.3a: this is the traditional, or indeed, old, procedure of tin welding or brazing on a brass plate where beforehand we have applied positioning bores of low depth (0.5 mm). The resonator is generally sectioned in a robinet copper tube 1 mm thickness. With the available diameters ranging from 6/8 to 58/60 mm or more, we can usually find one of a convenient size. The ensemble is then silver-plated.

- 6.3b: a tapped brass disc is inserted into the tube, at about 20 mm from the base. It is fixed by welding or setting. In the latter case, it has the form of a pulley, in the throat of which we push back the material of the tube, which this time is made of aluminum or an aluminum alloy, providing, with respect to brass, an appreciable weight saving. The central tapping is of a small diameter, such that the fixing screw must be thin and long in order to offer a certain elasticity, in combination with the portion of the plate where it is supported.

- 6.3c: this is a variant of the preceding device that needs no screwing. A helical spring is slipped by one of its ends into a small hole formed in the plate. We then present the tube equipped with a pin over which, with the aid of special pliers we hook the other end of the spring, which will ensure the stock of elasticity and the permanence of a pulling force.

- 6.3d: this version is the most interesting, but demands special equipment, for obvious reasons regarding investment profitably, with which only companies capable of producing in large series will be able to equip themselves. We continue to use, as in the two previous examples, tubes and plates made of aluminum alloy, but the tube is bored for the most part of its length, in order to leave no more than a shoulder of about 2 mm at its base. On this shoulder a thin calibrated spring will be positioned, pierced in the middle and forcibly connected by means of a pop rivet to an identical disk which is supported by a symmetric shoulder performed by pressing on the tube-holding plate. The assembly is self-centering; but in addition, the fact of reaming the tube in order to make the interior diameter compliant with that of the disc-spring takes away almost half its mass, which will lead in the end to a filter that is the lightest, most robust and most stable possible, to which we also add that its assembly time will be shorter.

Versions b and c present the disadvantage of not assuring the precise positioning of the tube. We correct this defect by the addition of a card or light plastic disc, of a diameter only slightly greater than the interior diameter of the tube, which we slip onto the screw or the spring beforehand and which we force into the interior, assuring a perfect centering.
6.1.3. Housing, tuning, stability in temperature

The times of mass-milled brass housing are in the past. Computerized machines have dramatically changed the technologies of mechanical fabrication, and amongst them, crossfold machines today enable us to make aluminum alloy cases which are close to the optimum regarding lightness, robustness, cost and deadline. The thick threaded walls are replaced by bordered thin metal sheets, which look similar, on paper, to children’s cardboard at the beginning of the 20th Century, which they assembled with glue to make constructions or toys. In the modern industry, glue is replaced either by pop rivets, when we have important series or we do not expect frequent disassembly, or bolts in the opposite case. The basic diagram, the simplest and most sure and efficient is shown in Figure 6.4. The metal sheet of 15/10 mm constitutes in general a good compromise between rigidity and lightness. The tube-holder plate is part of the body of the filter, the tuning plate fits into the upper part and is fastened by pops or by bolts. The small sides are of similar conception and fastening and are found at least 1.5 times’ the width of the filter in relation to the end resonators, a condition imposed by experimental findings and by Dishal’s method, which will be explored later, so that they have no electric influence over the main design.
As we could suspect, the variants in the making of the parallelepiped that constitutes the body of the filter are infinite in their details, and the choice of the cutting of the case usually depends on secondary arguments, related to the immediate environment of the filter, in particular in the way we have predicted its fastening in the sub-assembly that it will form a part of. Ultimately there is only one, extremely important, constraint, which is illustrated by Figure 6.5: it is absolutely essential to avoid that the assembly slits, especially if they are located at the foot of the tubes, be on the same plane as that which contains the field lines. The non-observance of this instruction regularly leads to instabilities and non-linearities, and thus to intermodulations if we are dealing with a power filter.

What now remains is to define the technology of the tuning elements, which are split into three groups:

- frequency tuning,
- tuning of the coupling between resonators,
- entry/exit coupling taps.

The height of the tubes is generally calculated for a frequency slightly higher (about 1%) than the working frequency, or, put in another way, they are slightly shortened. It is then easy to make the resonant decrease to the desired frequency by adding to the upper end a capacity supplied by an adjustable capacitor, plane or coaxial. Figure 6.6 shows three devices that are commonly used, extremely simple and corresponding, each one, to a particular necessity. For each one of them we have drawn a part of the upper plate where the most common system for fastening the tuning rod is represented: an opened-tap, applied in an aluminum sheet of 15/10 mm, which allows us to block with a nut a M4 screw over about two threads, which is enough to ensure a good fastening. These numbers are valid for filters of 1 to 4 dm³ in low UHF (400 to 900 MHz). These precisions lead us, without a break,
to evaluate the fineness of a coupling or frequency tuning, both being of the same order: in the thread SI standard, the height of the standardized thread is a tenth of the diameter. When we tune a filter and are close to the result, the angular variation which we subject one screw for the last time is of the order of 3 degrees. With an M4 thread, this corresponds to a horizontal translation of $\frac{3}{360} \times \frac{4}{10} = \frac{12}{3600} = \frac{1}{300}$ mm, that is about 30 μ. This is very little, but is nevertheless all the precision required for all the frequency tuning, the coupling tuning being a little bit more tolerant. But it is important to observe that this small flick of the wrist, given by the technician on no matter which frequency screw, can make the return loss vary by several dB. We thus see through these numbers that this is an extremely sensitive operation, precise and delicate, that right away gives the exact measurement of what must be the stability in temperature of a filter that must potentially work between -20 and +70°C, and leads us to explain the principle. To do that we follow the similar study developed in section 4.4 relating to the coaxial cavity, recalling that we place ourselves systematically in the case of an increase in the ambient temperature, knowing that everything is inverted when it decreases.

One of the resonators is represented in Figure 6.7, the h-L dimension representing the difference between the height of the tube and the tuning plate being systematically equal to its diameter D. The materials used are brass for the tube, aluminum for the housing and iron for the frequency-tuning screws. The origin of the expansions is the tube-holding plate, to the left. When the temperature increases, the tube elongates towards the right and its own resonance frequency decreases. But
the dimension h increases more than L, and pulls towards the exterior of the tube the screw, whose contrary elongation is much weaker. This results in the capacity of the cylindrical capacitor formed by the screw/tube assembly decreasing, having the tendency to make the frequency grow. The whole system is thus naturally self-compensating, and we can adjust the diameter and the length of the screw for the compensation to be exact. If we have understood the phenomenon correctly, we can note that it works even with an aluminum tube, it being enough for the contrary action of the screw to be weaker, but in any case there is always a solution to the problem.

![Diagram](image)

**Figure 6.7.**

The linear expansion of the housing in the direction of the length also means that the gaps between tubes increase when the temperature increases. The couplings then decrease, but cause no frequency drift: they simply have the tendency to reduce the band. However, there will be a corresponding variation of the return loss that must be monitored. If it is very significant, to the point of taking the filter away from the specifications, the design or sizing of the tuning screws will have to be corrected in order to vary their compensating effect in the direction desired, knowing that the coupling decreases when the screw penetrates towards the interior (the opposite is true for helical resonators).

We can now return to Figure 6.6 and detail the models b and c. b is a variant of a, which we use when we need a greater tuning capacity, which happens in two cases: either we have shortened the tubes for reasons of bulk in terms of height, and a large capacity is necessary to return to the right frequency, or the standard screw is not appropriate to the temperature stability. It is evident that such a configuration limits the acceptable power, which depends directly on the breakdown voltage, and thus on the distance between the height of the tube, where the voltage is maximum, and the mass. If it is necessary to pass on an appreciable power, we will have to turn towards version c. This is then especially suitable for power filters, by limiting the proximity to the mass and increasing as much as possible the radius of curvature.
Matthaei provides very complete data on the profiles of cavity resonators of very high power. Without going that far we can satisfy ourselves (in the case of radiotelephony UHF filters, which most particularly interests us) with welding, before silvering, a round-off disc as drawn in 6.6c, with a curvature radius equal to one quarter of the diameter. The tuning screw will comprise a disc of the same diameter, which will be located in the same place that the mass is located in a reception filter, that is, at a distance of the order of the diameter. It is obvious that this is a supplementary constraint that implies a reduction of the tuning range. We will also have to ensure that it has an impeccable finishing, with mandatory polishing, with a hard cloth, of all the zone with a strong electric field, since the smallest spike could reduce the efficiency of the device to nothing.

**Remark: other technologies**

The construction methods described above are very suitable for prototypes and to average series in low UHF. In high UHF (1.5 to 3 GHz), and for large series, numerical machines of the “HSM” (high-speed machining) type provide an interesting alternative. They allow the complete body of a filter, comprising the housing with the resonators, to be produced from a block of aluminum, in a matter of a few minutes – these being either pierced or full, as desired. We have also produced, for the GSM 1 800, considerable quantities of double duplexers for fixed relays, with an incomparable reproducibility.

### 6.2. Dimension calculations. Dishal’s Method

We can ask ourselves, in view of the constant progress of systems of electromagnetic simulation, what interest there can be in developing algorithms specifically dedicated to a particular type of product. The answer to this question is in another one, for which we excuse the triviality: “why use a sledgehammer to kill a fly?”. Especially since heavy tools such as the HFSS reference or all other 3D latest-generation programs, alone capable of leading to a solution, imply the use of a sufficiently powerful computer and demand a non-negligible calculation time, although this is being constantly reduced as clock frequencies increase. Finally, what is more convenient than to have at our disposal on a programmable calculator that accompanies us everywhere, a dozen kilobytes that will instantly provide us with all the dimensions of a band pass filter for which we know only the template?

#### 6.2.1. Parameterization

We have previously seen (section 5.2) how to determine the order of a Tchebycheff from its response curve. Once we know the number of resonators necessary, it is then necessary to determine the dimensions of all the elements of the...
filter, but there is an important datum that is lacking, and even thought it is well described in the specifications, it depends on the choices of some parameters that we must fix \textit{a priori}: we are talking about the insertion loss. It depends on the size of the filter according to laws that render it theoretically determinable. But it is important to notice that this is a number that, firstly, is only crucial in cavities of very large dimensions where powers of several dozens of kW circulate, and secondly, that, in the majority of cases, it is less than 0.5 dB. In other words, to make a calculation error of 0.1 dB on such a value can have very bad consequences, whereas it would be without importance on a out-of-band rejection of 60 dB. The designers prefer to base themselves on the results acquired and to choose the size of the filter in the table that they possess on the characteristics of already-existing devices: this is not really gratifying, but is of an infallible sureness! The width $h$ of the filter will thus be, in general, empirically chosen by scaling from an already existing filter.

However, there are certain situations, for example the liberation of a new frequency band, where we do not have a close enough reference point. It is thus desirable to be in possession of a few formulas that allow us to estimate, even if approximately, the width parameter. We have previously defined (section 3.3.8) a quality coefficient in the vacuum, $Q_0$ for an isolated resonator. A filter is never just a grouping of resonators and, if we recall that in a combline the resonators are all identical, it seems natural to consider that the quality coefficient of the filter in the vacuum is the same as that of one of the resonators. If $n$ is the order of the filter, that is to say, the number of resonators, and $\alpha_0$ is the loss associated with a resonator, the insertion loss of the filter can then be written (see equation [4.7]):

$$\alpha = (\alpha_0)^n = \left(\frac{1}{1 - \frac{Q}{Q_0}}\right)^{2n}$$

with $Q = \frac{F_0}{\Delta F_{\text{int}}}$ and $Q_0 = \frac{4\pi R_0}{F \delta \mu_0 \left(\frac{\lambda}{d} + \frac{\lambda}{h} + \frac{8}{3} \log \frac{h}{d}\right)}$ (see equation [3.8]),

(6.1)

It is a case now of knowing what we will choose as the external diameter to define a coaxial cavity equivalent to the elementary resonator, which is itself bordered by two conductor planes. Figure 6.8 shows in dotted lines two circles supposed to represent the external conductor of the equivalent cavity, one too small and the other too large, without us being able to specify where the correct layout is between the two. But we can always consider that the smallest corresponds to the upper limit of losses, and it is that which we will therefore take into consideration, assuming that we find a correction coefficient as a function of the measured results. The ratio we called $b/a$ in the study of the cavity becomes $h/d$ with the annotations of the drawing, and if we fix $h$ arbitrarily and if we take, as usual, $d = h/3.6$, it allows us to calculate $Q_0$. We thus have all the necessary tools to
calculate the corresponding loss and see where it is situated in relation to the specifications. Afterwards we exploit the result as we want, perhaps with the aid of a small iterative program that will allow us to make a given value of $h$ correspond to a value of losses by means of successive approximations.

\[ h \]

\[ d \]

**Figure 6.8**

**Example**

We wish to build a 6-pole Tchebycheff at 940 MHz with a 3 dB band of 25 MHz, and an insertion loss of less than 0.5 dB. We choose \textit{a priori} a width of 50 mm and a tube diameter of 18 mm. We have $Q = 37.6$ and $Q_0 = 4762$, if we take the width of the filter as the diameter of the equivalent cavity. Having done all the calculations, we find $\alpha = 6 \times 0.0688 = 0.41$ dB, which is suitable. If the maximum insertion loss was of 0.4 dB, we could have taken a width of 60 mm instead of 50, which would have lead to a $Q_0$ of 5680 and to a loss of 0.35 dB. These values are very close to the effective results.

Remark: We can see, by means of the preceding example, that the quantity $Q/Q_0$ is very small compared to the integer: 0.008 in the case of a 3 dB band of 25 MHz, and 0.004 if the band is doubled. We thus find ourselves in the condition of being able to use the series expansion of the function $\log (1+x)$ which is equivalent to $x$, for small $x$, and write, putting $Q/Q_0 = x$:

\[ \alpha_{\text{db}} = n \times 20 \log (1+x) = 20n \times 0.434 \log (1+x) \]

that is:

\[ \alpha_{\text{db}} = 8.48n \frac{Q}{Q_0} \]  \hspace{1cm} [6.2]
We will verify that to a negligibly close difference, we find the proceeding result. Additionally, going back to the definition of Q, this approximation allows us to state a very simple and practical scaling law: for the same filter, and at a given frequency, the product of the 3 dB band by the losses is constant: if we have 0.4 dB of losses for a band of 25 MHz, we will have 0.2 dB for a band of twice the value. We can summarize this in a very simple formula:

\[ \alpha \times BW_{\text{3dB}} = \text{Const} \]  

[6.3]

To finish this section, let us recall that in the estimation of the losses we overlooked adaptation losses, which is allowed when we have chosen a ripple of 0.01 dB, which allows us to tune the return loss to 26 dB. It is very evident that if the ripple becomes of the same order of magnitude as the insertion loss, it will be necessary to add it. If, for example, in the preceding case, we had decided to sacrifice the return loss to increase the selectivity and for that we had chosen a ripple of 0.1 dB, the total losses would have increased to 0.51 dB and would surpass the specifications.

### 6.2.2. The modified method

Dishal is part, along with Cristal, Cohn, Matthaei, Zverev and others, of the group of specialists in HF filtering from the mid-20th Century, that bequeathed to us the essence of knowledge on the subject. The method of calculation of the dimensions of a combline that we propose and that we have widely used is extrapolated from the one he developed for interdigital filters, presented in the specialized magazine “IEEE Transactions on MTT”, September 1965, p. 696. We will refer to this article for details of the calculation. It will also be of use to those who wish to go deeper into the study to consult another article cited in reference by Dishal and entitled Coupled Circular Cylindrical Rods Between Parallel Ground Planes, by Cristal, “IEEE Transactions on MTT”, July 1964, p. 428 onwards. The original method was modified (very little in its principles) to adapt it to the combline model, and to include some principles that we introduced previously, as well as to take into account certain parameters that we fixed \textit{a priori}:

- the tubes are open at their upper end,
- the frequency tuning is made by fine screws,
- the distance between the height of the tube and the upper plate is equal to the diameter of the tube,
- the filter is completely silver-plated,
The approximating function is a Tchebycheff polynomial,

all tubes are identical: same height, same diameter.

\[
\begin{array}{ccccccc}
n & q_0 & I.L. & q_1 & q_0 & k_{12} & k_{23} \\
\hline
\text{inf} & 0.000 & 1.1811 & 1.1811 & 0.6818 & 0.6818 \\
23.622 & 0.937 & 0.9796 & 1.5869 & 0.7027 & 0.6630 \\
11.811 & 2.006 & 0.9649 & 1.7472 & 0.6785 & 0.6759 \\
7.874 & 3.227 & 0.9686 & 1.8730 & 0.6465 & 0.6917 \\
5.905 & 4.642 & 0.9810 & 1.9817 & 0.6091 & 0.7093 \\
4.724 & 6.313 & 0.9994 & 2.0787 & 0.5667 & 0.7283 \\
3.937 & 8.351 & 1.0230 & 2.1653 & 0.5180 & 0.7491 \\
3.375 & 10.963 & 1.0518 & 2.2408 & 0.4607 & 0.7721 \\
\end{array}
\]

Table 6.1.

This being agreed upon, the goal to achieve is to calculate the distances between tubes from the coupling coefficients as given in the tables. To achieve that there are several steps, of which the first is to explain the notations. We will take as a basis Table 6.1, relating to the third order, p. 344 of Zverev, where \( k_{12} \) and \( k_{23} \) represent the coupling coefficients between resonators and \( q_0, q_1 \) and \( q_n \) are the standardized values of the quality coefficients such that \( q_0 = \frac{\Delta f}{f_w} Q_0 \) and allocated to the elementary resonator, the entry circuit and the exit circuit, respectively. One must remember that the values entered into the table are calculated from an approximation function, and are not related to any particular manufacturing technology. In particular, they are valid for discrete element filters, for which the quality coefficients are incomparably worse than for resonators, and where, additionally, we have the right to insert resistances. In the case of comblines we are only interested for the moment in the couplings between resonators that are supposed perfect, and we will thus deal neither with \( q_1 \), nor \( q_n \), and we will always place ourselves, even if the assumption is not exact, in the hypothesis \( q_0 \approx \infty \). From this we derive Table 6.2 which will be the sole basis of the calculations in what follows.

We note that these values lead to filters with a geometric center of symmetry, which is not the case for those composed of discrete elements, and for which the central coefficients vary very little from rows 7 or 8: with the habitual screw tuning,
we will be able to build a filter of a higher order by considering that $k_{ij} \cong \text{Const}$ for $i$ and $j \geq 8$.

It is necessary now to convert the values of $k$ into distances between the axes of the tube, as a function of the parameters that we will choose beforehand and that depend on the predicted performance of the filter: the width $h$, the diameter of the tubes $d$ and their electric length $\theta$. These distances will be annotated $C_{12}$, $C_{23}$, etc. to correspond with the coefficients $k_{12}$, $k_{23}$, etc. We have thus implicitly decided that the couplings are represented by the distance between the tubes, which has lead many theoreticians to identify them with respect to the capacities of the capacitors they form. This notion of coupling is in fact very subtle in the sense that we can make multiple representations of it, either electrical, magnetic or even geometrical or energetic, without necessarily dissipating the mystery and unveiling its true nature in an incontestable way. In fact, in electromagnetism, two resonating systems that are not isolated from each other by a perfectly conducting obstacle are always coupled: if they are distant from each other the coupling is weak, if they are close to each other the coupling is strong, and with respect to our immediate goal we will stop here.

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<th>$k_{45}$</th>
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<td>0.5673</td>
<td>0.8430</td>
</tr>
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</table>

Table 6.2.

Dishal reminds us in the quoted article that the coupling coefficient $K$ between two circular rods can be expressed, amongst others, by the approximations of Honey or Cristal:
The Combline Filter

C₀ is the capacity of the bar itself relative to the inner walls, Cₘ is the mutual capacity relative to the two neighboring resonators. We note that the quantities of the form C/ε depend only on geometrical parameters: for a plane capacitor, for example, we have C/ε = surface/thickness. There is thus, implicitly, a link between the coupling and the dimensions, by ways of capacity, and it is not surprising that the development of one or the other of these relations leads, via very complex calculations, plus empirical touch-ups that will not be detailed here, to a less hermetic expression which we will call Dishal’s formula, in homage to its discoverer:

\[
\log K = \left\{ -1.37 \frac{C}{h} + 0.91 \frac{d}{h} - 0.048 \right\} \quad \text{[6.4]}
\]

We then possess a fairly simple relationship between K, coupling coefficient, and c, the corresponding distance between the two consecutive resonators concerned, that we seek to determine. It is necessary now to evaluate K as a function of the accessible parameters. It is possible to transform Cristal’s relationship by introducing the electric length θ to arrive at the practical formula which we will ultimately use:

\[
K_{i,i+1} = \frac{4}{\pi} \left| \frac{k_{i,i+1} F(\theta) \tan \theta}{1 + (k_{i,i+1} - k_{i-1,i}) F(\theta) \tan \theta} \right| \quad \text{[6.5]}
\]

where \( F(\theta) = \frac{1}{2 \sin \theta} (\cos \theta + \frac{\theta}{\sin \theta}) \), \( w = \frac{\Delta F_{\text{norm}}}{F} \), and where ki,j are the coefficient of Table 7, representing the normalized couplings between the resonators of rows i and j. The use of [6.4] and [6.5] together allows us to extract the desired value:

\[
C_{0} = \frac{h}{1.37} \left( 0.91 \frac{d}{h} - \log K - 0.0048 \right) \quad \text{[6.6]}
\]
In formula [6.5] we find the quantity $w$, which is the normalized band at 3 dB. Now, in the specifications, we usually give the usage band, which is a band at 0 dB, we could say. This has of course no physical signification, but it is absolutely necessary to define a band-pass that is realistic and at the same time does not betray the spirit of the specification by implying null losses. We can, as suggested by Zverev, use an equiripple band, but we prefer a 0.1 dB band which reveals itself to be more efficient in usage. The different correspondences are indicated in the table above, valid for a ripple of 0.01 dB.

The calculations performed above, which in the past were performed “by hand”, are now presented in the form of more or less elaborate programs, that can range, as regards the presentation of the results, from a simple listing of the numbers to a final picture in 3D, or even direct instructions for computer-controlled machinery. We now have all the data to calculate the coupling distances, except for one extremely important remark: it is essential, in daily practice, to make, in the first place, a filter as we have planned for, satisfying all specifications, with an optimal adaptation and a compliant off-band rejection. This is all the more so as in general it is planned that the usage band may at some point be modified. This is the reason why we have planned for tunings. We will thus introduce before any parameterization, in the project of a combline, a band correction, in such a way as to aim for a filter that, with all tuning screws unscrewed, will find itself pre-tuned to a band at the same time narrower and slightly shifted towards higher frequencies. To do so, we will multiply the specified band by 0.7 to obtain below [6.7] the value of $w$ to introduce in formula [6.5]. $F_1$ and $F_2$ are the lower and upper limits respectively, $F_0$ the central frequency and $c$ the coefficient extracted from the last row of Table 6.3 (0.1 dB band):

$$w = 0.7c \times \frac{(F_2 - F_1)}{F_0} \quad [6.7]$$

What remains is to specify a last parameter that has not yet been mentioned: it is the distance $e$ between the axes of the first or the last resonator and the end walls.
Dishal points out that the influence of these walls is negligible as long as the distance we are referring to is at least equal to 0.75 h. We will systematically respect this condition, experimentally verified, while remarking that in case of imperative necessity it is possible to reduce it even further. We will thus write the last formula to be used:

$$e \geq 0.75 \, h$$  \[6.8\]

**Example of application**

We want to make a combline for a central frequency of 460 MHz, with a 3 MHz band. The specifications of insertion losses and selectivity have lead us to take the 4th order with \(h = 40\) mm. What are the dimensions?

The optimal diameter of the tubes is 0.37h, that is, 14.8 mm. We take the standard diameter that is closest, that is, 14 mm. The construction of Figure 3.9 indicates to us that \(\theta_{\text{max}} = 86.4^\circ\). We choose 60° to reduce the height dimension. The corresponding height is \(l = 108.7\) mm. The internal height of the filter will thus be:

\[
H = l + d = 108.7 + 14 = 122.7 \text{ mm}
\]

We thus find: \(C_{12} = C_{34} = 68.6\) mm
\(C_{23} = 72.4\) mm
\(e = 0.75h = 30\) mm

The filter is entirely determined (Figure 6.9).

Had we chosen to keep \(\theta_{\text{max}}\) instead of shortening the tubes, we would have found:

\(l = 156.5\) mm
\(C_{12} = C_{34} = 43.8\) mm
\(C_{23} = 47\) mm

From which follows a total internal length of 194.6 mm.

Comment: considering these numbers relating to two electric lengths, we conclude that shortening the tubes immediately leads to enlarging the filter in the sense of its length. In addition, the shortening must be compensated by a corresponding increase in the tuning capacity, given that it is necessary to reach an effective path of \(\lambda/4\), which means increasing the dimensions of this capacity, or alternatively, a reduction of the depth of the dielectric, and thus a decrease in the
stand-off distance to the mass and in consequence, a weaker breakdown voltage: this will not be suitable for a power filter. If, as shown by the calculations, we find ourselves obliged to increase the initial distance between the tubes when we reduce their height, this means that, without that, the coupling would have increased. There is something surprising there when we have chosen to define the coupling as a capacity, which would normally increase linearly with the dimensions of the conductors. In addition, the development of this way of seeing has lead to formulas that were previously justified and which are, in fact, the basis of methods of widely-proven efficiency. That is to say, as we have already mentioned, this notion of coupling is delicate and subtle, when we try to analyze its implications from a physical point of view. In its energetic aspect, in any case, it seems that it fails to bring into play, not the energy itself, but the energy “density”.

![Figure 6.9.](image)

6.3. Tuning of filters

6.3.1. General rules

Before moving on to the tuning of comblines, it is indispensable to detail a bit more the working mechanism of the taps, for which we have given only a short technological description until now. The term tapping appears in the articles of IEEE Transactions that we have already cited, and in particular that of Dishal. In section 5 of that publication, Dishal gives, following the description of the method of calculating couplings between resonators, which we have explored, a complementary theory intended to determine, with precision according to the author, the height of the connection of the taps. We have the right not to agree with this concept, based on a notion of impedance which is arguable and which, while it may be capable of producing results, can only do so in the particular geometrical configuration presupposed by the author. We also have the right to contest its theoretical validity, because it has been experimentally verified that it is possible to correctly couple the entry and the exit by connecting the linking strap at any height,
from the base to the top, even forming a simple loop at the mass close to the resonator, without a direct connection to it, as in a cavity. What is important, which we have been lead to take into account to be able to parameterize, is the magnetic flux generated in the loop of the entry tap by the signal, or created in the loop of the exit tap to restore it to the external circuit.

Figure 6.10 shows the two most frequent layouts with regard to the position of the connectors on the housing. Shown as a full line is the initial form of the tap, as a dotted line is an arbitrary form after tuning by deformation. The deformation of a geometry is a very normal practice in the tuning of filters and cavities, because in the most cases it is the only possible way, given that we have not find another of the same order of simplicity: we introduce a small tuning hole into one of the large walls of the filter, approximately at the height of the horizontal part of the tap, and we lean against the tap by levering it. From the original configuration drawn as a full line, we cannot in principle do anything but reduce the surface of the tap/mass loop, and thus the coupling. It is thus necessary to build this hole a little bigger than necessary, which demands a certain number of preliminary trials. It is necessary to assimilate the idea that a filter has a unique tuning. If we gave two identical filters to be tuned to two people located in two different labs, and supposing each one is an expert and arrives at the optimal tuning, we would see that all the screws would be pushed in by the same length, respective to their location on the housing. It would also be necessary that these filters be rigorously identical from a mechanical point of view, which we never find. Even in this case, what is said regarding tuning by screws would be false for taps, as far as their exact forms are concerned, because there is an infinity of geometrical configurations that can give place to an adequate magnetic flux, which determines the correct entry/exit coupling and whose value will be perfectly identical in both cases.

Figure 6.10.
We have already insisted upon the extreme fineness of the frequency tunings (section 6.1.3). When we connect a filter that has just been manufactured to the analyzer, it is not rare that we see nothing emerging from the background noise on the screen. The first thing to do is to align the screws in the upper position. This done, we normally observe an increase in the transmission curve in the vicinity of the central frequency and a concurrent decrease of the return loss curve, which will allow us to recognize the presence of the filter (6.11a). This line will be above the correct frequency and will be under-coupled, thus giving rise to a band that is too narrow, and we will thus fix the band and the central frequency by progressively and methodically pushing the screws in according to the following rules:

- the aim to attain is to have a reflection curve having the greatest possible surface, with slight maxima at the same level and the lowest possible on the trail (6.11d),
– we turn all the frequency screws, one after the other, blocking them at each time. We stop for each of them when we can no longer make the return loss curve decrease, and then we go on to the following one. Afterwards, we repeat the process with the coupling screws, bearing in mind that pushing a coupling screw has the effect of increasing the coupling (this is the inverse of what happens with the helical resonators). We then restart the process with the frequency screw, as many times as necessary, until we obtain a curve of identifiable appearance (6.11b),

– with the aid of frequency screws, we support the high frequency of the return loss curve on the high frequency of the specified band F₂. The increase in the band will in fact be made towards the low frequencies, from the high band (here still, and this is a systematic phenomenon, things will be reversed for the helical resonators, but we will come back to this),

– the coupling tuning has an influence on the frequency tuning, but the reverse is not true,

– we slowly and progressively push to insert the coupling screws until we obtain the desired band, and we put the finishing touches to the tuning, possibly allowing for a little band supplement for the variations in temperature,

– close to the final result, we act on the taps to correct any under- or over-coupling, increasing or decreasing the effective surface respectively, in the sense of Faraday lines, that is to say, if we prefer a practical point of view, by distancing or approximating the strap from the base of the tube. Figure 6.12 shows the physiognomy of the response curves when the access coupling is not harmonized with that of the resonators, the correct aspect to obtain being that of 6.11d.

![Figure 6.12.](image_url)
In summary, we realize that the tuning of a combline, when we expect a high-performance from it (for which it was in fact conceived), requires extraordinary care. Besides the meticulous aspect of the process, it is equally indispensable, to minimize the tuning times, to correctly master the effects of the variations of the parameters, that is to say, knowing what to do to modify the response curve in the way we want: change the central frequency, change the band width, smooth the return loss curve, etc. The different descriptions and remarks above must provide the essential weaponry to reach it, whereas the rest is a question of practice and self-will.

To finish, let us add that a well-tuned combline filter is not symmetric: when the resonators are linear (tubes or bars), the left side is less steep than the right one. When they are helices, as we will perhaps have guessed, it will be the right slope that is less steep. We will be able to possibly use this property to make duplexers with small band intervals, by combining the two technologies.

6.3.2. Elliptization of the Tchebycheffs

It is possible to modify the response curve of a Tchebycheff filter in the lower part of the lateral slopes to give it the appearance of a Cauer. We sometimes call this modification an “elliptization”, which is not fully justified because the appearance of the curve of the group delay indicates that it is always definitely a Tchebycheff. Figure 6.13 shows in dotted lines the track that can be obtained from a standard curve, the latter presented as a full line.
We thus manage, keeping all the advantages of a Tchebycheff, to create two poles tangent to the passing band that allow a gain of the order of a dozen dB at a crucial place in the making of duplexers. The transformation of the curve, which then resembles a Zolotarev function, is paid for in fact in an out-band increase that compensates the off-band selectivity gain obtained in this way. But a gain of 10 dB may be a decisive advantage when we are close to the limit of the specifications, and today it is indispensable to know this much used procedure.

The basis of the technique of elliptization is to physically build on the filter, unexpected links of weak interaction between distant resonators (by distant we mean resonators that are not consecutive). It is not a question, of course, of satisfying ourselves with connecting zones: as in all devices intended to create attenuation, it is necessary to respect a double condition of phase and amplitude which will be realized with the combination of two signals, in which one will have a precise progression of electric length in relation to the other. It consists in yields that, initially, were discovered by chance, helped by healthy curiosity; nevertheless, the mechanical model of the combline that followed made it possible to produce an extremely simple and productive representation of the phenomenon. The positioning of the system depends on the configuration of the filter. Until now, we have only described filters that are geometrically linear, that is, which present themselves under the appearance of aligned resonators, with or without separating walls, because it is the simplest model to imagine and also the most common.

But, as long as we do not compromise with the general laws of coupling, we can give filters many varied forms: folded comblines are relatively widespread, but if necessary we can stretch our imagination further, as shown in Figure 6.14, which is by no means limiting. According to the layout adopted, and according to the result expected, the elliptization will thus be made in many ways, all having a common principle of removing a small part of the signal in the resonant space of a tube, and
reintroducing it into another non-adjacent one, in phase opposition. From this point of view, the folded filters offer an undeniable advantage: the additional coupling can be made very simply by loops passing through small windows in the central wall (or the internal walls when the filter is folded several times). It is thus not indispensable to use a semi-rigid cable to perform the counter-reaction, even if this designation is usually used in the domain of amplificators. On the other hand, the solution by cable allows, by its absence of specificity, all the configurations to be performed, including, in the case that it is indispensable or easier, the coupling of the base of a tube with the top of another, with the aid of a mixed loop/probe system at the ends of the coaxial.

The radio band-pass always has very weak relative bands, say of a few % around the central frequency $F_0$. It is at this frequency that we will calculate the step differences of the elliptization circuit. As regards the amplitudes of the re-injected signals, we know *a priori* that they will be weak because they are of the same order as the attenuation we want to improve. We will thus install either small deformable loops close to the bottom of the tubes, or small probes, also deformable, close to the top, the deductions and reinjections of energy being so weak that they will not be sensed except on the places where we expect their effects to occur. The diagram of Figure 6.15 shows the two procedures that are most widely used to execute the elliptization by secondary coupling of resonators 2 and 5 of a 6-pole band-pass, a coupling that works particularly well. We immediately see the advantage of having chosen a folded combline, where all these mechanical operations are simplified, all the more so as the pair 1 and 6 also works well and can be accumulated with the preceding association.

Figure 6.15.
The Combline Filter

It now remains to be known, once we have defined the technological procedures of the secondary coupling of eliptization, which resonators to link in order to obtain the attenuation poles on both sides of the band, and above all how to do it. The answer can seem strange, but it will find justification in the mechanical analogy: we can use whichever couple, on the sole condition that the electric length L of the path of the deducted signal be such that it is re-injected in phase opposition with the main wave at the exact place where the summation of both occurs.

The condition can then be written:

\[ L = \frac{k}{4} \frac{c_0}{\varepsilon_r F_0} \]  

[6.9]

where \( c_0 \) is the velocity of light in the vacuum, \( \varepsilon_r \) the relative permittivity of the insulator of the coaxial (or the integrated line) and \( k \) an integer depending on the relative phase of the vibrational state at the two ends of the coaxial, which phase will by necessity be a multiple of \( \pi/4 \). The amplitude tuning, which will simultaneously be the tuning of positioning of the attenuation poles on the response curve, will be done by deformation of loops or probes, according to the layout chosen.

![Diagram](image)

**Figure 6.16.**

It is evident, after what has been said, that the higher the order of the filter, the more possibilities we have. In particular, if the practical combinations start to be sufficiently numerous, we can have the idea of installing several retroactions on the
same filter, if for example we want to reinforce the effect of a first system at the same place on the response curve. In fact, we will never exceed the number of two because there is the possibility of interaction, in such way that the tuning, already delicate by nature, can become problematic. In any case, it is a case for extremely qualified specialists, but which is important to know because in this way we can gain more than 10dB of rejection at the foot of the band. As an example, Figure 6.16 shows the addition of two counter-reactors on a 6-pole filter. The two diagrams on the left of Figure 6.14 are particularly well adapted to this installation.

6.3.3. Mechanical model

The aim of this section is to construct a completely mechanical equivalent of a filter to try to give a representation of its working mechanism without reference to electromagnetic concepts.

The analogies between vibratory mechanical and electromagnetic systems are almost as old as the domains referred to. In the 1950s and 1960s, in lectures ranging from basic to specialized mathematics, the study of the correspondence between these systems were taken relatively far, in order to make the students capable of easily passing from one domain to the other, and then solve more easily a problem that could prove arduous from a certain point of view but was more evident from another. It is partially the memory of this philosophy, today abandoned, that motivated us to try this methodology in certain applications with microwaves, and in particular on combline filters, with impressive results.

![EM Model](image1.png) ![Mechanical model](image2.png)

*Figure 6.17.*

We cannot proceed with this genre of modeling except by giving life to the dielectric, the vacuum, by considering it as a perfect fluid with a volumic mass. In order not to shake up habits too much, we will refer to it as an “equivalent dynamic...
mass”, which will avoid us, for the moment, having to deal with all the risky theories about the true nature of the medium of propagation of electromagnetic waves; and in case the analogy opens up to fruitful applications, we will also be spared directing towards the relativists the inevitable difficult questions that would follow.

From this moment then, the current notions of fields and potentials will be abandoned to only consider, in the space called filter, the vibration modes of a volume of fluid of very low compressibility, a sort of powder or “jelly”, as Wheeler qualified it, whose volumic mass is very large, the latter point being in accordance with the extremely small magnitudes of the amplitudes of the vibrations when we choose to calculate them.

The physics of incompressible, or very slightly compressible, fluids is necessarily swirling, and the vortex holds a dominant place: it is the only commonly observable phenomenon that we can relate in analogy with a number of electromagnetic manifestations, where we establish the simultaneity of presence and interaction between a current, in the electric meaning of the term, and a magnetic field. The other key notion is precisely this systematic interaction between the dielectric fluid, supposedly present everywhere including in the interior of the material, and the free electrons moving in the conductive parts. Figure 6.17 establishes the parallel between the conventional vision of an end resonator of a combline, which is equipped with a tap, and its mechanical model, where the tap is replaced by a vortex. The vibrating part of the “vacuum” dielectric, is considered to be located between the two right prisms, of which the smallest, the central tube, is of circular section, and the largest, constituted partially by the external envelope, of a more or less flat section, or elongated in such a way that is at a tangent to the neighbor, also having a part in common as suggested by the diagram. The form is not very important, once there is interaction. The vortex model that we will use is that of Rankine, which is unilateral, the simplest possible because it is, amongst other things, a modeled representation of the drain whirlpools, that anyone can observe in their bathtub. In our usage, this model is simply pushed to its extreme limits of perpendicularity and of thinness between the layer and the funnel. We refer to as a funnel here that which is habitually the flow, or if we prefer, the vorticity filament, that is to say, the axial part where the movement of the fluid is orthogonal to the layer.

From these definitions and preliminary descriptions, we can now draw (Figure 6.18) a combline filter in the form of resonant prisms, vibrating in torsion, coupled by the interpenetration of their dielectrics and excited by vortexes. Let us describe the mechanism, not losing sight of Figure 6.17.
The current that crosses the entry tap sets in movement the layer of the associated vortex, which instantly gives rise to an axial displacement (filament). If this current is alternating, the filamentary flow changes direction at each period and becomes a vibration that excites the first dielectric volume at its base. Given the direction of the mechanical excitation, horizontally and perpendicularly to the radius of the prism, the latter can only vibrate in torsion and in the vertical direction. At resonance, the upper section is in phase quadrature with the lower section, which translates by a little angular displacement. The prism \( n_0 \) \( 1 \) leads the prism \( n_0 \) \( 2 \) which is, as Figure 6.18 shows, in phase opposition with the previous 1, labeled 0, and so on. As a consequence, the different resonators have relative alternated phases 0 and \( \pi \), consecutively. The latter activates, by its filamentary axis, the second vortex, which carry the electrons of the tap, giving rise to the exit current.

We then have at our disposal a consistent mechanical representation of the mechanism of a band-pass filter, which can be easily extended to other types of filters, and which will allow us to use another logic, concurrently to “conventional” electromagnetic theory.

6.3.4. Abundances of the mechanical model

A section was previously consecrated to the eliptization of Tchebychefs, using a mechanical model which is now explained. It is tempting to put this model to the test, for a first evaluation of its interest, and this first test will in fact be a way of validating it.

The principle of elliptization, let us remind ourselves, is to install on the filter a circuit or a feedback system whose effect, invisible in the passing band due to the low amount of power that is diverted, will be to generate two attenuation poles.
below the sides of the response curve: with the aid of a loop or a probe, we remove a small quantity of the signal from a given place in the filter and we reinsert it at another place, in phase opposition with the main signal as it is in this other place of reinjection. Let us apply the procedure to the 5-pole filter of Figure 6.19, making the link 2 to 5 which is a conventional configuration. The phases, referenced this time to the entry tap connected to tube 1, are indicated in the boxes below and above each resonator. It should be noted that the filter is, as a whole, a multi-resonant volume where there is no propagation, in the usual sense of the term, between the entry and the exit; two filters of different lengths, tuned in the same way, present the same group time and phase characteristics. This said, if we decide to create a feedback 2 to 5 by means of loops at the bottom of the tubes, we see that between these points there is already the correct value of phase difference, that is, $\pi$. If a 6-pole filter is folded once, and resonator 2 is then close to 5, it will thus be enough to have one common loop, of length close to zero to make both their volumes communicate. If the filter is of conventional length, it will be necessary to use a coaxial loop-to-loop link, of electric length $n$ times 360° which will assure the phase opposition, a unique and indispensable criterion.

![Figure 6.19](image-url)

But it seems that once we respect this criteria, we can push the investigation a bit further: the layout of the diagram suggests that there are multiple solutions, considering, amongst others, that we can also use the zone located at the top of the tubes. There is thus the possibility of putting secondary coaxial links terminating on one side by a loop and on the other by a probe, so as to connect the bottom of a tube with the top of another. This will enable us, in particular, to perform the summation of two feedbacks, even with a filter comprising few poles, by simply respecting a
rule of intuitive logic which forbids two systems from overlapping geometrically, and not to cross, which would make interaction and tuning almost impossible. We could for example make a link “2 to 5” coexist with a “1 to 6”, which is what is effectively done and allows a supplementary gain in rejection of several dB, which in some cases is not to be disregarded. We thus show that a problem such as the elliptization of band-pass can be understood in another way, which speaks for itself, incomparably more intuitive than the traditional one that appeals to approximation functions and the theory of circuits. It is never bad to have at your disposal more than one method to tackle a problem, but the decisive argument in favor of the mechanical model will come from the representation it proposes for the working of the taps, and above all the consequences one can take from this representation.

The frequency and inter-resonator coupling tunings lead to a certain level of return loss, which is the best we can obtain with respect to the pre-tuning of entry-exit loops, for which we are obliged to fix a certain dimension a priori. But, since this is only a pre-tuning, it is always possible to improve it by optimizing the taps, which is in any case indispensable if we do not respect the specifications to begin with. The inspection of the adaptation curve allows us to know if we are over or under-coupled (see section 6.3.1 and Figure 6.12). If we are overcoupled, it is necessary to reduce the effective surface of the taps, which is easily done by deformation (Figure 6.20a). If on the contrary there is undercoupling, it is necessary to increase this surface, and we can find ourselves in a situation where this is not possible, because we are already at the maximum configuration for one reason or another. Within the framework of the classical electromagnetic model, there is no solution to that. However, in the mechanical model, where the tap is an oscillating vortex in the dielectric fluid, we are rapidly lead to wonder whether it would be possible to create a second vortex by welding a second parallel loop to the first one, in between the connector and tube n0.1. Now, it would seem that that works, and that it works well: once the second tap is installed (Figure 6.20b), it is sufficient in
general to alter the frequency screws of the two end tubes (we treat the two ends of
the filter symmetrically) as well as the couplings with their immediate neighbors, in
order to see the return loss curve decrease all the more neatly as we are already
relatively well adapted. If that is not enough, despite everything, we can, provided
we have enough space to do it, add a third loop, etc. The procedure of course has its
limits: if we suppose that the second loop doubles the action of the first, the third
will increase that of two loops by 50%, the fourth by 33% of that of the three
preceding ones, and so forth. In practice, we rarely surpass the number four. But this
is not all, the most surprising is still to come: if we need to gain only a few dB in
adaptation, we could content ourselves with adding a part of a tap in parallel with
main one (Figure 6.20c), by taking care to install it in a plane at a tangent to the
displacement lines, which are usually considered as lines of magnetic field, thus
carefully avoiding the installation represented by a dotted line, placed in the same
plane as the loop, which has no effect. In the case of the partial loop, it is difficult to
interpret and to say if we increase the thickness of the vorticial layer or if we create a
second vortex, secondary and smaller, it always being that in conventional
electromagnetism we do not encounter the phenomenon; the amount of loop that we
add is not crossed by any magnetic flux, and also creates none since it is only a
current diversion. We can even take the experience a bit further and choose
diameters of conductors such that the HF resistance of the new installation is
identical to the original one: the gain in return loss shows itself to be indifferent to
these parameters. This will probably be the subject of in-depth consideration for
some, but as regards what concerns, us we will be satisfied to apply with pleasure a
procedure that has many times helped us out of apparently inextricable situations.

To finish with the taps, and still with an eye towards the mechanical model, the
fact that we see the loop as a vortex frees us from the condition of having to attach it
to the resonator. If the entry coupling must be weak, which corresponds to a band-
pass of narrow band, we can effectively, as in Figure 6.21, close the loop to the mass
without any contact with the tube, provided only that it is on a plane containing the axis of the tube. This reduces to nothing, but it is thus, Dishal’s theory on the connection dimension of the tap as well as its way of calculation.
Chapter 7

Channel Multiplexing

7.1. Definitions

Having finished the individual study of filtering elements, it is now possible to tackle that of the association of filters or cavities. The generic term which covers this technique is that of multicoupling. We have seen previously that the word coupling entered the definition of different concepts: there was the coupling of resonators, entry/exit coupling, undercoupling, overcoupling, and there is now multicoupling, which is not a concept but a technique. It consists of making arrangements of several filters, in the most general meaning of the term, in order to realize in the first place a multiplexing of radio channels. The problem is always the same: on one side there is a single antenna and on the other several radio channels that must reach it, which means several emitters or receptors, or both at the same time. It is thus necessary to conceive equipment ad hoc, as passive as possible, and with minimal losses, that allows each channel to use the antenna as if the other channels did not exist. To do this it is necessary to dispose of two types of components: firstly, selective filters, cavities or band-pass, which attenuate the neighboring frequencies sufficiently, and afterwards a system of interconnection by means of lines, very often coaxial cables, that by transformation of the impedance will present, at a common central point, the other channels as open circuits. This will be the harness.

7.2. The duplexer

The simplest of the associations is that of two filters. We will call it a duplexer by extension, this designating in fact, at the origin, a device allowing vocal
radiotelecommunications in duplex, that is to say, using two frequencies, in opposition to the alternate that uses only one: the first frequency serves for the emission of interlocutor 1 and reception of interlocutor 2, the second for the opposite, and this is all done with a single antenna on each side. Although the receptors are all equipped with quartz filters of very high selectivity, these cannot support a very large power and must be protected by a supplementary filtering to avoid the local emitter deteriorating the local receptor. This is the first role of the duplexer. Its second role, as important as the first, is to assure the separate adaptation of the emitter and the receptor to the common antenna. Possibly, if we are dealing with a band-pass duplexer, we can also ask it to participate in the suppression of the emission harmonics as well as in the suppression of noise and off-band lines. Figure 7.1 shows the functional diagram of the two most commonly-used types of duplexers: the band-pass and the pass-reject. We never find, in practice, purely rejecting duplexers, even though there is no particular problem with conceiving them, simply because their symmetrical response does not allow a small inter-band separation and is not in general adapted to the practical problems.

![Diagram of Band-pass and Pass-reject Duplexers](7.1)

**7.2.1. The pass-reject duplexer**

The pass-reject duplexer is formed by two associated groups of resonators: low-pass reject and high-pass reject. The dissymmetry of their individual response curves makes it possible, in this association, to be compatible with the small duplex separations (see section 4.1.3), while assuring important rejections. This is one of the reasons for their antiquity, this term not to be taken pejoratively but with the consideration that we owe to the precursors, or if we prefer, of their durability. Figure 7.2 shows the typical appearance of a global response curve (high-pass F1F2 and low-pass F3F4). The transmission curves are shown in full lines and those of the return loss in dotted lines. As a function of the habitual relative bands of the radio,
the number of cells generally varies between 2 to 4 for each band, and from 4 to 8 for the complete device.

![Diagram of coupling system](image)

Figure 7.2.

The coupling system, be it the coupling between cells or blocks of cells, is perfectly defined: it is exclusively a matter of quarter-waves (Figure 7.3), that present themselves usually in the form of semi-rigid coaxials, most of the time curled up to save space and arranged on the exterior, between each elementary cell (Figure 7.4). When, for very low frequencies, the quarter-wave coaxials become too cumbersome, it is then necessary to remember that a line is equivalent to a concatenation of serial inductance/parallel capacity cells, and that we can effectively produce it in this form, in discrete elements. We would be advised, in the event that we choose this solution, to anticipate an additional compartment to lodge them, so as to assure a total separation between them and the resonators. But we can also, when the dimensions allow, produce the $\lambda/4$ in strip lines, air or substrate, integrated into the very interior of the housing (Figure 7.5). With respect to the latter, in the large majority of cases we use the compact structure, each resonator being completely isolated from the others, from the point of view of the radiation, at the interior of a square-section profile which allows the most rational usage of the available space. We can nevertheless, precisely if we do not have this problem of bulk, put all the tubes inside the same parallelepiped, as for the combline, but keeping them sufficiently distant so that the direct coupling is negligible. Referring again to
Dishal, who indicates that in a combline the end walls have no influence as long as they are placed at more than 1.75\(h\) from the axis of the end tubes, \(h\) being the width of the housing; we thus take this value as being equally applicable to the minimal distance between the resonators. This technology has the consequence of elongating the filter, but it allows us to simplify it, to make it lighter and do without the profiles.

![Figure 7.3. Low-pass structure](image)

The rejections we can obtain with a pass-reject duplexer obviously depend on the number of resonators used. In UHF, with tube diameters from 12 to 18 mm, which is more or less the norm, one should count on 30 dB of rejection per resonator. The link by quarter-waves completely isolates them and allows us to juxtapose the response curves to obtain a rejection band of lesser or greater significance, or to add
them to a common frequency to favor a strong rejection. In this latter case we can effectively obtain as many times 30 dB as there are cells. Moreover, it is curious to note that we can hardly surpass this value, even with large cavities, but that conversely, it is possible to attain it with resonators of a mediocre quality. The difference in the quality coefficients will be noticed in this case on the passing loss and the selectivity.

Besides its aptitude to small band separations, the pass-reject duplexer has, relative to the band-pass, a second large advantage which concerns the acceptable power. In fact, resonators have as a role to provide the rejection in the neighboring band and consequently are not subject to the overvoltage they would have to support if they were made in band-pass. In the passing band, the power passes through the line, the resonators connected in derivation on the line are supposed to present, at the corresponding frequencies, an infinite impedance. The behavior in power will thus depend on the quality of the line and of the connection elements, selfs or capacitors, between the line and the tubes. In the case of selfs (low-pass side) there is no problem given their robustness, whereas in the case of capacitors (high-pass side) it will be necessary to ensure we choose components with low ESR (Equivalent Serial Resistance). In a common fashion, a pass-reject duplexer can support a power 5 to 10 times greater than that of a band-pass with the same dimensions.

The tuning of the duplex separation is done by adjusting the impedance of the link between the line and the resonators: its nature, inductive or capacitive,
determines its function, low-pass or high-pass respectively, and its value the separation itself. If we take the example of the low-pass, the larger the self, the greater the duplex separation. On a large pass-reject cavity we have the habit of installing, in series with the self, a variable capacitor whose minimum and maximum values are such that, by tuning it, we can obtain, as we wish, a globally inductive or a globally capacitive coupling, that is to say, either a low-pass or a high-pass function. Once the duplex separation is tuned, it is more or less conserved when we vary the anti-resonance frequency of the tube by its tuning screw, that is to say that we can displace curve 81 in block. But we notice, in the case of the duplexer, a very big difference of behavior between the two constitutive parts, which limit its operation to a few % (2 or 3) of the nominal frequency. We understand clearly, in fact, that if the frequency varies, the duplex separation cannot be conserved unless the coupling impedance changes parity. Now, whilst this is effectively the case for an inductance, which varies like the frequency, it is no longer valid for a capacitance that varies in inverse function. There will thus be a specific problem on the high-pass side and in fact, when we want to build a duplexer of the pass-reject type that is tunable in a large enough band, of the order, for example, of ±20%, it is absolutely necessary to anticipate variable capacitors, with tuning holes, on the high-pass side. A pass-reject duplex of 2n tunable cells in wide band will necessarily comprise 3n tunings: 2n frequency tunings and n capacitance tunings.

**7.2.2. The band-pass duplexer**

![Diagram](image-url)
This other type of duplexer, also widely used, will be made up, as its name allows us to anticipate, of two passband filters, almost always combiners, linked to a common point called nodal point by a harness formed by two adapted lines whose length will no longer systematically be \( \lambda/4 \), but will have a value at times arbitrary but nevertheless precise which we will determine with the aid of the Smith chart, according to a general principle that will afterwards be applied to more complicated cases where the number of filters and functions will also be arbitrary.

![Figure 7.7. Band-pass duplexer](image)

Before starting the study itself, it is important to specify a certain number of notions, conventions, principles, which will facilitate comprehension and efficiency. In the first place, it is absolutely indispensable to have, otherwise it must be acquired, a good experience with the handling of the Smith chart, and to have a good understanding of the correspondences with the conventional Cartesian display. It is necessary to assimilate, in fact, that the chart contains the phase information and that it perfectly takes into account – it is moreover conceived for this – of the periodicity of the impedances along a line: the length of a line calculated in the chart is always modulo \( \lambda/2 \), which corresponds to a full turn on the external circle. Figure 7.6 shows both associated traces which the analyzer gives us, the response curve in amplitude is faded because it is useless in the process of determining the harness. When we move from left to right, that is, in the sense of increasing frequencies, on the amplitude response, we turn in the clockwise direction on the chart. The undulation of the return loss, or that of the losses curve, which is linked to it and consists of \( n \) maxima, translates in the chart by as many convolutions around the center, of impedance \( Z_0 = R_0 \) (in principle 50 \( \Omega \)). The frequencies \( F_1 \) and \( F_2 \) are those of the 3 dB band of the low filter, \( F_3 \) and \( F_4 \) those of the high filter that will be coupled to it (Figure 7.7) according to Figure 7.7, where \( l_1 \) and \( l_2 \) are the lengths, to be determined.

Figure 7.8 shows a possible appearance of the practical curves that we can see on an analyzer screen, relating to the theoretical trace, as taught in microwave courses, of an isolated resonant cavity (Figure 7.8a). The full circle that passes by the points \( Z = Z_0 \) and \( Z = 0 \) is that of the perfectly coupled cavity, that presents a perfect adaptation to the central frequency; the dotted circles, meanwhile, correspond, for the largest, to an entry/exit overcoupling, and for the smallest to an undercoupling.
A real band-pass cavity has rather the appearance of the central drawing (Figure 7.8b), relative to which we can make the following remarks: the symmetry axis is no longer the horizontal axis of the chart, the curve is no longer a circle and does not pass through the centre. The angular discrepancy $\theta_0$ of the axes, which we have represented as negative, because this is generally the case, has as a cause the inevitable connectorization of the cavity (or the filter), which introduces a supplementary trajectory and thus a phase difference. The other visible differences, in particular the degeneration of the circle, stem from losses and are all the more enhanced the more the cavity has smaller dimensions. As for the curve of the filter, to the right, it can be even narrower, to the point of not showing the loop anymore, but always present, around the center of the chart or to the side and contained in a small circle corresponding to a certain value of the adaptation in the band; an inflorescence of $n$ petals, $n$ being the number of poles. With the bridges now down, let us move on to the method.

![Figure 7.8.](image)

To be determined are the lengths of the lines $l_1$ and $l_2$ of Figure 7.7, which connect filter 1 and filter 2 respectively to the nodal point. Both filters are initially tuned separately and, if the coupling is correctly performed once the two filters are joined, it should not be necessary, in theory, to alter it. In practice, this depends on the quality of the filters, but it is very rare for there to be no small adjustments to be made, at the end, on the frequency of the last resonator. It is then necessary to draw the two charts, which are in a sense the identity card of the filters, and a first precaution must be taken concerning the calibration: the phase reference, at the exit of the filter, is that of the internal plane of the housing where the exit connector is fixed (Figure 7.9). Now, the calibration is usually performed with the “standards” of the calibration kit, which each possess a connector, which has a certain length, in general different of that of the fixed connector of the filter which will take their place afterwards. From this comes a cause of error that must be completely suppressed, all the more meticulously when the frequency is high and the dimension
of the connectors is no longer negligible with respect to the wavelength. But it is a fact that the manufacturers have not predicted this particular application, and technicians have no other resource than to manufacture as best they can other sets of adapted calibration standards, which are done relatively simply by using sockets similar to those of the filter and completely equalizing the female pin. This will be the standard of the open circuit, that of the short-circuit will be the same fixed connector over which we will support with the thumb, for the corresponding calibration time, a small conductive plate. There is no problem for the impedance calibration (alignment of the chart), the standard being a pure resistance. It will be necessary, on the other hand, for the other scenarios, to prepare a complete set of fixed connectors with respect to the types of connectors we are likely to find: N, SMA, Subclic, RIM, 7/16, etc.

![Figure 7.9](image)

We are thus capable, from now on, to arrange traces on the chart of the two filters to couple (Figure 7.10). Adding a portion of line to the exit of a filter, means extending the course of the signal, which is translated visually, on the chart, by a clockwise rotation of the trace. Conversely, if we need to make the trace turn in the trigonometric direction, it will be necessary to shorten the path, which is not possible unless the filter is already equipped with an exit coaxial. This scenario exists, if only when a badly calculated harness must be rectified. But in any case, there is always a solution to orientate the trace as we desire, even if it entails performing a complete lap of the chart. The action will consist, here, of covering the point of infinite impedance, to the right on the horizontal axis, by the sector divided on the exterior circle by the frequency band of the other filter. But this implies two theoretical conditions. The first is that the band must be narrow, so that we can assimilate it to a single point ($Z_{\infty}$), the second is that each filter must be entirely unmatched in the band of its future associate, or in other words, the trace for these frequencies must be on the external circle. In practice, we never find these two conditions, and Figure 7.10 is realistic from this point of view: the markers are never on the external circle, and the band is as it is. Study engineers will have to accommodate this fact and do the best they can.
Concretely, there is still a certain flexibility in the application of these principles, and one must not forget that we can intelligently cheat (we say: perfect the tuning) by altering the tuning of the filters once they are joined. But it is certain that if the initial conditions are degraded (filters with very significant losses, bands that are too large, separations that are too short), the final tuning will also be degraded, at the same pace, and we end up, as result of approximations, having an unusable result. However, in order to arrive at this point, it is really necessary to transgress the logic and go beyond certain limits which it is, besides, rather difficult to define precisely, but that we can try to quantify approximately.
As regards the band, in a project we could reasonably take as limits those which are indicated in Figure 7.11: the shaded part is forbidden, and it is not recommended to overstep, or even to reach, the sector limited by the radii at ±30°. Regarding the tolerance on the reflection coefficient, a module of $\sqrt{2}$ will be a limit not to cross, with a good tolerance for $\gamma \geq 0.9$.

This being specified, when the coupling is made according to the rules, it should not be necessary to alter the tuning of the coupled elements. In all cases the line lengths $l_1$ and $l_2$ are determined from the relationship $\frac{\phi}{2\pi} = \frac{\delta}{\lambda}$, which is the simplest and the most fundamental of the theory of lines, and which is read “in old style”: $\phi$ is to $2\pi$ that which $\delta$ is to $\lambda$. It is simply necessary to watch out that the phase differences determined on the chart are relative to a return path and that it is necessary as a consequence to affect the raw results to a factor of $\frac{1}{2}$. If $\varepsilon_r$ is the relative permittivity of the insulator of the line, the speed of propagation is of $c_0/\sqrt{\varepsilon}$, and we have, finally:

$$l_1 = \frac{\theta_1}{4\pi\sqrt{\varepsilon}} \frac{c_0}{\sqrt{F_3 F_4}}, \quad l_2 = \frac{\theta_2}{4\pi\sqrt{\varepsilon}} \frac{c_0}{\sqrt{F_1 F_2}}$$

[7.1]

In these two formulas, the terms with the form $\sqrt{F_1 F_4}$ represent the average frequency, in the physical sense, of the band $F_3$, $F_4$. But often, considering the narrowness of the radio bands, we take the arithmetic mean, out of habit.

**Remark 1**

From the progressive insufficiency of the allocated bands, the separations between bands in radiotelephony are usually of the same order of magnitude, and sometimes narrower than the bands themselves. It becomes difficult, in these conditions, even with very good filters, to ensure that one is completely unmatched in the passing band of the other. We can thus wonder what the effect will be of a non-infinite impedance on the selectivity of the filters after coupling: we find that the variation due to coupling, which can be significant, of the order of 10 dB, occurs in the correct direction, and for both filters at the same time. We will not look here for a theoretical justification; it exists but is complicated, and of no use for our purposes. Let us note, however, that a possible elliptization works just as well when the filters are coupled as before their pairing up.
Remark 2

In the same way, we have already noted (see section 6.3.1) that the combline, especially for orders below 6, are not completely symmetrical, with a slope to the right greater than that to the left, which penalizes the high filter in its inter-band rejection function. More than a tendency, this is a systematical phenomenon, very difficult to explain but where the configuration of the taps probably plays a certain role. Moreover, we can partially correct this dissymmetry by placing the taps in the interior side of the end resonators, that is, between the first and the second and between the penultimate and the last, but staying well inside their zone of influence, where we have placed the part of the dielectric attached to this or that resonator. The correction is only partial, but uncontestable. The installation of connectors that follows can allow us, besides, by freeing them from the ends, to solve access and bulk problems.

Once the procedure has been assimilated in its principles, the door is open towards more complicated applications where the quantity of filters and frequency bands will grow.

7.3. The combiner

This device plays an essential role in the base stations of all the mobile radiotelephony networks. Being a central part of the transmitter multicoupler, it allows, as the duplexer of which it is a direct extension, the simultaneous use of \( n \) radio channels on a single antenna, assuring its two key parameters that constitute the conditions for the correct functioning of the system: insulation and adaptation, which are nothing more than the two conditions of the goal set, that is, the separation of channels.

The typical combiner is that with a nodal point (star combiner), in which a certain number of cavities or band-pass filters are linked to a common point, that will be connected to the antenna, by coaxial cables of the same length modulo \( \lambda/2 \), the assembly of which is called a harness. It is conceived from standard hypothesis of spectral occupancy, that in mobile radiotelephony are the following:

- the channels to be coupled are grouped into consecutive “packets”,
- the frequency separation between channels is the smallest possible,
- the channels are permutable (agility).

With respect to what has previously been said about the duplexers, we understand that the conditions above are limiting, and in fact we very quickly ascertain that the more flexibility is required from the device, in terms of frequency
band, tuning band, bulk or the number of channels, the more difficult it is to construct. We find ourselves in the position where commercial reason ends up opposing common sense, which largely motivated, in our introduction, the disenchanted comments on “decibel manipulators”. It is thus appropriate to recall, once again, that the quality of a cavities coupling depends on the selectivity of the elements, and thus on the quality coefficients, and that the latter are higher the more the cavity is voluminous: this is an inviolable and definitive law.

This being established, let us move on to the technology. We have the habit, but this is not entirely justified, of classifying cavity combiners into two broad categories, corresponding to series or parallel groupings. The parallel mode is the nodal-point combiner that have mentioned above. The series mode will designate a combiner where the cavities are connected in sequence to each other in such a way that we can add one or more additional cavities to the end of an already constituted group, if there is the need. It is a presentation that has always interested the military, who see in this conception of the coupling a flexibility in the usage that the first would not have, due to the apparent possibility of being able to add (or remove) a new radio channel as desired, according to the necessities of the moment, without changing what already exists and without interrupting the traffic. We will show that this advantage is contestable, and that it is limited by the choices made at the start of the dimensioning of the components.

7.3.1. The nodal-point combiner

We will equally find the expression “star combiner”, because this is how the Americans or the Germans often refer to this model. It is, from the point of view we are interested in, a direct extension of the band-pass duplexer, and the method of
determining the harness is strictly the same. The actual graph resembles that of Figure 7.12, where we find as a dotted line the ideal presentation and as a full line the actual trace, about which we can make two comments: firstly that the angle $\theta_0$ which makes an axis of symmetry of the trace with the horizontal, is always small, and secondly that the degree of degeneration of the circle depends on the volume of the cavity. We note as well that the trace is underneath the axis, which means a slight capacitive predominance, that can almost entirely disappear if we calibrate the analyzer with the corrected standards. In all cases, we find ourselves, in principle, in the zone of tolerance where we can consider that the impedance is, if not infinity, at least sufficiently large so that we can consider it as such. The primary harness (Figure 7.13a) of a group of cavities will thus be constituted of cables of electric length $k\lambda/2$, where $k$ integer can be null: it is in fact common, and practical, to build up integrated modules of four cavities (Figure 7.13b) where the four central holes are directly linked to each other by large silvered straps welded onto the pin of a fixed connector. The latter should be able to support an effective power quadrupole that applied to the entry connectors, and will thus not necessarily be of the same model. The straps being of inductive nature, we sometimes, in parallel, add a variable capacitance to the nodal point of a few pF (1/14 for example), which allows us to perfectly balance the return loss out. It will be necessary, as for the connector, to choose a capacitor capable of supporting the peak voltage of the four modules together.

![Diagram](image.png)

**Figure 7.13.**

With what has previously been said, we can envisage that the number of channels to couple will very quickly be limited by two parameters already mentioned: the frequency band of the packet, which must not leave the authorized zone of the chart, and the total peak power which, little by little as it approaches the
disruptive limit of the weakest component, will generate increasing intermodulations before achieving the critical value synonymous with destruction. In the first French mobile network, Radiocom 2000, we coupled up to 64 channels of nominal power equal to 50 W each, so as to draw the maximum from the enormous candle-antennas sized to cover zones of 30 km in radius. We have seen filters exploding, coaxials melting, and we have quickly gone back to much smaller cells and a maximum of 12 channels to couple per antenna at the same base station. This shows that the peak power, which corresponds to the breakdown effect, has as much importance in the study of a coupling as the effective power, related to thermal effects.

Another limitation, this time purely technological, comes from the growing difficulty in producing a nodal point connectorized to a great number of accesses. We have no knowledge, in conventional industrial productions, of numbers greater than 8 in this domain, and even this is an architecture that is little pleasing to the eye and very inconvenient, given the number of cables that go along with this. The layout that is the most used is that of modules of 4, coupled among themselves by a secondary harness whose elements have a length multiple of \( \lambda/2 \).

7.3.2. The series combiner

![Diagram](attachment:image.png)

Figure 7.14.

The line lengths being defined to about \( \lambda/2 \), a nodal point of zero dimension can be replaced by a concatenation of half-wave sections connected by T-shaped links. This allows us to immediately transform a parallel coupling into series coupling (Figure 7.14), using the same cavities. This procedure, which is electrically equivalent to the preceding one, is more esthetically-pleasing for the multicoopers with a large number of cavities, it necessitates no nodal point, and it has got its supporters. Prima facie, we conceive that it could be interesting when we do not
know exactly the number of filters to couple, and especially when the latter is susceptible to varying in sizeable proportions. This hypothetical uncertainty is rare in radio, however, the majority of the networks are planned with precision and always with respect to a programmed extension. Despite all, some constructors have chosen this technology as standard and integrate a portion of line in the cavities to delete the T-junction. Let us repeat, the performances are identical and do not depend on anything else but the volume and the way in which the cavities are manufactured.

There is, however, another type of series multicoupler which leads to a system that is truly different, and it is the one using guiding filters: each filter of channel is constituted by two cavities (or two band-pass filters) arranged between two directional couplers that provide the outlet, in the common path, with characteristics which this time are effectively large band. These will be studied later in the chapter on transmission multicouplers.

### 7.3.3. Harness adjustment

Whatever the kind of coupling chosen, it is subjected to the necessity of making connections of a well-defined length, most often exterior (non-integrated) and coaxial, and this is maybe the most compelling aspect of high-frequency roles. An HF laboratory has to have a set, at least, of stripping machinery, not only because the fabrication of cables is long and tedious, but above all because at times it is necessary to have good precision and good reproducibility. At 500 MHz, a sector of 10° on the chart corresponds to

\[
\frac{1}{36} \times \frac{3.1 \times 10^5}{500 \times 10^3} \times \frac{1}{2} = 0.0083 \text{ m},
\]

that is, 8.3 mm with an air dielectric coaxial, and 5.5 mm with a polyethylene dielectric coaxial where c = 0.66c₀. If we set a cutting precision of ± 1 mm, which is reasonable, we will have, over the large circle of the chart, an uncertainty of ± 2°; that is reasonable, without excess. One must not forget, in fact, that added to the uncertainty error of the cutting will be that of the setting or the welding, of the braid and the central conductor, to be multiplied by 2 because there are two ends, plus the errors intrinsic to the dispersion of the characteristics of the cable itself, and those of the connector. In fact, after execution we establish that at this frequency, the cables must not differ by more than 1 mm.

Added to this is a second constraint: the spacing of the radio frequencies is of the order of 10 kHz, and a sufficient separation to be able to couple with such a separation cannot be made with cavities. Only quartz filters will be able to achieve this, and it is this type of filter that, for this reason, we find in receptors, where they allow channel-to-channel protections, with a large number of poles, greater than 60 dB. But they only support a few Watts, and the cavities cannot be avoided when
it comes to the transmission of power. Thus, in order to make things possible in a relay- or base station, we will couple radio channels with a minimum separation of the order of 100 to 200 kHz, in return for an adequate distribution of frequencies in a zone of given coverage. It is a common compromise between the selectivity of the cavities, which from experience must be of at least 6 dB from one channel to the other, for us to have a coupling worthy of this name, and the width of the packet of frequencies which must be the smallest possible in order to minimize the dispersion of the markers on the chart (see Figure 7.11). This having been said, let us move on to the tunings.

We will reason over a module of four cavities, each of which will be tuned with an insertion loss slightly below the specifications of the ensemble, in order to allow for the coupling loss; more importantly, the entry and exit matching will be identical. We will not alter these matchings until the harness is finalized.

First step: the approach. The four cavities, being tuned in the same way, give rise to equal charts; so we can pick any of them as a reference, and we will center the corresponding cavity at the median frequency $F_c$ of the packet of frequencies limited by $F_0$ and $F_n$ (Figure 7.15). We can see straight away appearing on the chart, in angular form, all the factors that will allow us to evaluate the probabilities of success of the coupling: size of the cut-out sector on the great circle by the packet, value of the coefficient of reflection at adjacent frequencies, selectivity, etc. Afterwards the

![Figure 7.15.](image-url)
theoretical length \( L \) of the connection to the nodal point is immediately calculated as a function of the angle of the exit phase \( \theta_0 \):

\[
L = \frac{1}{2} \left( \frac{c_0}{F_e} \right) \left( \frac{360 - \theta_0}{360} \right)
\]

[7.2]

But in general this connection has three physical parts: the exit connector of the cavity, the cable itself and the trajectory at the interior of the housing of the nodal point. The connector part does not matter if we have calibrated the analyzer with the corrected standards, there remains the two other lengths. It is there that the chore starts for the technician which, whatever the variant chosen, will consist of cutting four cables of electric length slightly greater than what is necessary, with connectors simply placed, not fastened, and then progressively reducing the common length by removing, strictly, at each operation, the same small length from each of the four coaxials. When must we stop, knowing that it is forbidden to exceed the correct value, at the risk of having to restart? It is at this stage that the choice of the method intervenes, systematically making use of the network analyzer.

First method: return loss out at the nodal point. The four cavities are loaded at the entrance and connected to the nodal point with four identical cables of a certain length. The analyzer is connected at the exit of the nodal point in Cartesian mode (it is thus a method for which we can use a scalar analyzer). According to the appearance of the ensemble of the return loss out, we can know if the cables are too short or too long, on the condition of being always relatively close to the correct length. Figure 7.16 summarizes the operation: if the base of the return loss curve decreases towards the right, the cables are too long. If it decreases towards the left, they are too short. As a dotted line is the hypothetical curve for one cavity alone, identical to those we are in the process of measuring and tuned like them. The return loss minima are thus found necessarily on this curve, and we see, on the middle curve, that it is normal that the central return losses are better than the end ones when the length of the cable is correct. It is only at the end that, more for esthetic than logical homogeneity reasons, we will be able to adjust the exit loops of the cavities to balance the levels of reflected power. This is an operation that is not physically justified, from the moment the return losses are according to the specifications, but it is a habit.
Second method: direct measurement of the losses. The cavities are connected to the nodal point, three are loaded at the entry, and we inject the signal of the analyzer at the entry of the fourth and watch the tracing at the exit of the nodal point: as the cables, initially too long, shorten, the apparent loss of the cavity reduces so as to pass through a minimum and then regain the value measured for the isolated cavity, notwithstanding approximate coupling losses. Simultaneously, the return loss passes equally by a minimum which is more easily noticeable than the minimum of the losses, due to its sharpness.

Third method: the stubs. The stub is a portion of line, open or short-circuited, connected in derivation at a point of the main line. The coupling cables being supposed to return an infinite impedance to the nodal point means that simply their connection to the nodal point with the other open end constitutes an equivalent system, and this allows an approach of the result using one cavity alone. This is the main advantage of the method, that in addition is more precise than the preceding ones if the cavities do not have a good quality coefficient.

We can imagine other variants, once we have assimilated the simple principle of determining a harness, but none will let us avoid the tedious and obligatory part of the process. What was said for a group of four cavities (or more generally four band-passes) is also valid for a larger number, but it is evident that the difficulty will grow with the spectral width of the packet to couple. This difficulty is even more accentuated by the fact that operators now demand, contrary to the early days of the mobile, that the system be “agile”, that is to say, that we be able to exchange frequencies at will in the channels of the combiner. All that while using, evidently, because competition demands it, material that is ever smaller, and thus less performing!
7.3.4. Secondary harness

When we have chosen the mode of nodal coupling with, as a base, the module of 4, integrated or not, we are lead, to surpass this number, to couple the modules among themselves. We do that with a harness said to be “secondary” which will be systematically composed of sections of length \( kn \lambda / 2 \) adjusted to the central frequency. But we are not condemned to construct couplings at a number of channels that is a multiple of 4: a module can be very incomplete, under the condition of replacing the missing cavities by equivalent impedances. Figure 7.17 illustrates an example of coupling to six channels, extensible to eight, where we have completed the second nodal point by two short-circuited stubs whose electric length, taking into account the path interior to the nodal point, is of \( \lambda / 4 \). The secondary harness comprises in its effective length the number of times the half-wave that physically allows us to create the connection of the ensemble, and its exact length can be adjusted with the help of the analyzer, independently of the cavities. In the event we have to add a channel, we remove a stub and replace it by a cavity, pre-tuned at the mid-band with symmetrical return losses, and then translated to the frequency of the additional channel. We do the reverse if we remove a channel: we notice that there is as much modification flexibility here as on a series setting.

![Figure 7.17.](image)

If there are more than eight channels to couple, we will have the choice of making a tertiary harness and using a nodal point identical to that of the modules, with the exception that it must be verified, in this second hypothesis, that the exit connector will be correctly dimensioned to resist the resultant peak voltage.

7.3.5. Projected losses

It is interesting to be able to predict the losses brought about by a coupling as a function of its parameters: frequency, number of channels, separation between the
channels, quality coefficient of the cavities. The different modelings of Chapter 4
give, for that purpose, useful relationships to characterize the cavities taken
individually, that then must be completed by others reflecting the coupling itself. Let
us first recall the expression of the matching given in section 4.5.2:

\[
S_1 = \frac{R + n_0^2 Z_0}{n_1^2 Z_0} \quad \text{entry} \quad S_2 = \frac{R + n_0^2 Z_0}{n_2^2 Z_0} \quad \text{exit}
\]

And \( S_1 = S_2 = S = 1 + \frac{R}{n^2 Z_0} \) when the loops are symmetrical, which we will
suppose from now on.

Given the expressions of \( Q_0 \) and \( Q \) which appear in 4.5.1, \( Q_0 = \frac{L \omega_0}{R} \),
\[
Q = \frac{L \omega_0}{R + 2 n^2 Z_0}, \quad \text{and taking the first into the second, we have:}
\]

\[
\frac{Q}{Q_0} = \frac{L \omega_0}{R + 2 n^2 Z_0} \times \frac{R}{L \omega_0} = \frac{R}{R + 2 n^2 Z_0} = \frac{1}{1 + \frac{2 n^2 Z_0}{R}} = \frac{1}{1 + \frac{2}{S - 1}} = \frac{S - 1}{S + 1} = \gamma
\]

\[
\frac{Q}{Q_0} = \gamma \quad [7.3]
\]

This simple formula is to be memorized; it is of interest because in a way it
creates the link between the interior of the cavity and the exterior. It leads in
particular to another expression for the loss:

\[
\alpha = 20 \log \frac{1}{1 - \gamma} \quad [7.4]
\]
It will also allow us to know the value of the equivalent resistance of the cavity. Let us consider Figures 7.18 and 7.19 which draw a diagram of a resonant cavity, loaded by \( R_0 \) and attacked by an internal generator of impedance \( R'_0 \). Let \( R'_e \) be its equivalent resistance and let us put \( R_e = R_0 + R'_e \). From what was said before, follows:

\[
\gamma = \frac{Q}{Q_0} \quad \text{and} \quad R_e = \frac{1+\gamma}{1-\gamma} R_0
\]

Let us take \( R_0 = 50 \, \Omega \) and let us consider a quarter-wave cavity, tuned to 426 MHz, over which we have a loss of 1.32 dB and a band at 3 dB of 339 kHz. We have:
\[ Q = \frac{426}{0.339} = 1257, \quad Q_0 = 8913 \text{ and } \gamma = 0.141. \]

\[ R_s = \frac{1 + 0.141}{1 - 0.141} \times 50 = 66.4 \text{ } \Omega. \]

We can deduct on the way \( R'_s = 16.4 \text{ } \Omega \), a value which, individually, is not useful for our calculations, but gives an order of magnitude to this parameter, which is always interesting. Let us examine now what happens, with the help of Figure 7.20, in a group of cavities coupled to a nodal point, and let us suggest that we calculate the total loss dissipated in the central channel. Let \( n \) be the row of a central cavity and \( \Delta F \) the frequency separation between two adjacent cavities. The pattern of the response curves clearly shows that the two neighboring cavities and even maybe the following two will have an influence on the real loss of the cavity \( n \), presenting at this frequency impedances that we must take into account, because in general they are not negligible, and they come to place themselves in parallel with \( R_s \). If the intrinsic losses of the cavity \( n \) are represented by the point A, the point B designates the supplementary losses due to cavities at \( \pm \Delta F \) and point C those due to cavities at \( \pm 2\Delta F \). Even if, theoretically, the others also intersect on the median axis, we do not need to take into account those beyond the second row, whose influence is in general of only a few tenths of dB, whereas that of the closest ones can be of the order of a dB. Regarding the upper ranks, it will be necessary, for them to be taken into account, that the cavities be so poor that we will also have other problems, such that the coupling will not work. The real losses of cavity \( n \) are highlighted in the electrical diagram presented in Figure 7.20. We will suppose (and this is a condition that is always met) that the separation between the channels \( \Delta F \) is substantially greater than the half-band at 3 dB. It results from this hypothesis that the neighboring cavities present, at frequency \( F_n \), a complex impedance for which the nature of the imaginary part is almost totally capacitive or inductive, and depends on their relative positions in relation to the central cavity. For the cavity \( n-1 \), for example, which works above its resonance, we will have at \( F_n \) an inductive predominance with an impedance of the form \( Z_{-1s} = R_s + jX \), for \( n-2 \) we will have \( Z_{-2s} = R_s + 2jX \), etc. These impedances put in parallel with that of the central cavity, constitute leakages and absorb a certain power that will add to the nominal power, and we have now to evaluate the global loss in \( n \) taking these neighboring effects into account. If \( \alpha \) represents the intrinsic loss of the cavity \( n \), \( P_{1n} \) the additional losses to the cavities \( n-1 \) and \( n+1 \), \( P_{2n} \) those of the following two cavities, we write:

\[ P_n = \alpha + P_{1n} + P_{2n} + \ldots \text{ in dB.} \]
Let us focus for the moment on $P_{in}$ and let us try to evaluate the different currents and the different impedances in play. Let us remark to begin with that the two impedances which absorb $P_{in}$ are conjugated, what incites us to treat them globally. On the other hand we can deal with the central cavity from the nodal point, as suggested by Figure 7.19, or from the other side, and depending on the case the electrical scheme will differ. We should then expect two different values of the losses depending on the position we give to the generator. If we put ourselves in the usual case where the generator is at the entry of the cavity and the nodal point is loaded by $R_0$, we can adopt the model of Figure 7.20 as an equivalent. In fact we would say that, in each case, $P_{in}$ stems from a supplementary current $I_1$, due to the two lateral circuits, that adds to the current $I_0$ that normally crosses the load $R_0$ when the cavity is alone.

![Figure 7.20.]

In these conditions, the calculation depends on the way of evaluating this current, and there are several leading to different, but close enough results, according to the reasoning adopted. If we refer to as $P_1$ the power dissipated in the cavities $n-1$ and $n+1$, $Z_1$ their global impedance and $I_1$ the current that crosses it, we can say here, supposing a constant applied tension, that:

$$P_{in} = \frac{P_1 + P_0}{P_0} = \left(1 + \frac{I_1}{I_0}\right)^2 = \left(1 + \frac{R_0}{R_0 + R_i}\right)^2,$$

and the value of the loss in dB will be given by the relationship:

$$P_{in} = 20 \log \left(1 + \frac{R_0}{R_0 + R_i}\right)$$

Where $R_i$ is the resistance equivalent to the perturbors that we must now calculate. If we refer to as $Z_i$ the impedance equivalent to the ensemble of the two
circuits n-1 and n+1, we have: \[
\frac{1}{Z_i} = \frac{1}{R_i + jX} + \frac{1}{R_i - jX} = \frac{2R_i}{R_i^2 + X^2},
\]
from where we obtain \[
Z_i = \frac{R_i^2 + X^2}{2R_i} = \text{Re} \text{ since it is a real quantity. We know } R_s, \text{ so what is left is to determine } X \text{ for which we will apply the relationships and procedures explored in 4.5.1.}
\]

The impedance of an R, L, C series circuit can be written:
\[
Z = R + j(\omega L - \frac{1}{\omega C}) = R \left[ 1 + jQ \left( \frac{2\Delta F}{F_0} \right) \right] \text{ and } |Z| = R \sqrt{1 + 4Q^2 \left( \frac{\Delta F}{F_0} \right)^2}
\]

It results, once all calculations are done, that the imaginary part of \( Z_i \), the impedance equivalent to the two cavities n-1 and n+1, will assume the value:
\[
X_i = \frac{2\sqrt{2} \cdot \Delta F \cdot Q \cdot R_s}{F_n}
\]

The complex impedance, equivalent to the two immediate neighbors is thus:
\[
Z_i = R_s + j \frac{2\sqrt{2} \cdot \Delta F \cdot Q \cdot R_s}{F_n}
\]

In order to have the impedance equivalent to the cavities n-2 and n+2, it will be enough in this formula to change \( \Delta F \) to 2\( \Delta F \). If we take the last example, we obtain:
\[
Z_i = 66.4 + j \frac{2\sqrt{2} \times 0.3 \times 1257 \times 66.4}{426} = 166.25 \quad \text{R}_i = 482.65 \ \Omega \text{ and } P_{1n} = 0.78 \ \text{dB (losses due to cavities n-1 and n+1)}.
\]

We will also have \( P_{2n} = 0.24 \ \text{dB (losses due to cavities n-2 and n+2)} \).

The effective loss of the central cavity, whose value in isolation was of 1.32 dB, will thus be 1.32 + 0.78 + 0.24 = 2.34 dB. This value takes into account nothing but the presence of the neighboring cavities, without prejudging the effect of other parameters. We will see in fact, when it comes to predicting the total losses, when the combiner will be integrated into an transmission multicoupler, that it will still be
necessary to add to the preceding quantity the losses in the cables as well as in the auxiliary components such as circulators and others. The complete assessment will be made in the chapter on multicouplers.

In what has been said, we have considered that the cavity \( n \) was surrounded by at least two other cavities on each side. This is not the case for the end cavities that only have neighbors at one side. To also make a reasonable evaluation of the losses, we will remind ourselves that \( \log(1 + x) \) is equivalent to \( x \) when \( x \) is small, and we will say that the loss due to a cavity is the half, in dB, of that of a pair. In the preceding example, if we suppose that the cavity \( n \) is the second of the group, its effective predicted loss will be of \( 1.32 + 0.78 + 0.12 = 2.22 \) dB. If it is the first or the last, it will be of \( 1.32 + 0.39 + 0.12 = 1.83 \) dB. These values are, however, not very important in the sense that, for reasons of homogeneity with regard to the specifications, we will return to the tuning of these cavities in order to put them at the same level as the others.

7.3.6. Optimal \( Q \)

If we look again at the curves of response of Figure 7.20, we can wonder what is going on when, with the data of the ensemble kept the same, we make the passing band of the cavities vary, it being well understood that they remain absolutely identical. If the cavities have their band increased, their intrinsic loss decreases but it is compensated by an increase of the additional losses that end up becoming predominant, and the adaption deteriorates. If they are too selective, the coupling works better but the intrinsic loss increases when the band decreases, without limitation. We thus sense that there is a compromise to be found for the effective loss of each cavity to be minimal. There exists a relationship between \( Q, \Delta F \) which we are looking to determine and \( Q_0 \), which expresses this compromise with a good accordance with the experiment; it is the following:

\[
Q^2 + 2KQ = KQ_0 \quad \text{with} \quad K = \left( \frac{F}{2\Delta F} \right)^2
\]  \[7.6\]

The solution of this equation is easily found on a computer by successive approximations; afterwards we apply a correction coefficient of 0.9 to finally obtain:

\[
Q_{\text{opt}} = 0.9 Q
\]  \[7.7\]
We then deduce the value of the band at 3 dB to which the cavities must be tuned in order to have the best coupling. It is still necessary that this result be compatible with the specifications of the return loss, which is a priority, and this is why the range of this calculation is limited. It is nevertheless a part of the theoretical tools.
8.1. Introduction

HF multicoupling and filtering rely to a great extent on theoretical and technological knowledge of cavities. However, we are obliged in practice to use supplementary devices to achieve the integrality of the specifications of a system. For example, it is impossible to assure a channel-to-channel isolation, which is measured in tenths of dB, with the cavities alone: it is necessary to add circulators. If we have difficulties with the adaptation, it will be necessary to add impedance adaptors. If we need to cancel an off-band increase of the response curve, it will be necessary to use a secondary filtering, and if there is no space to do so by means of cavities, we will have to find another technique, etc.

It is thus indispensable that technicians dispose of a complete arsenal that allows them to solve all kinds of situations, and we will call these objects auxiliary.

8.2. Circulators

8.2.1. Operating principle

These are passive components, and thus linear, at least in a certain dynamic, but non-reciprocal. Their specific properties are based on those of ferrites subject to a strong magnetostatic field. In informal language the ferrite can designate both the material, and the product, or object. Ferrites are sintered ceramics (pulverized mixture, compression, cooking, fabrication) based on iron oxide. From the chemist’s point of view, we can refer to as ferrites all the components of the form Fe$_2$O$_3$, XO,
where X can be either Fe (this is the magnetite), or one of the elements of the YBaCuO family (yttrium, baryum, copper, oxygen), which we find for example in superconductive materials and that are always at the basis of particular properties. The crystallographic structure of ferrites is cubic of the spinel type (MgAl₂O₄), the metallic ions being positioned in the interstices.

As oxides, ferrites are insulators, which grants them two major qualities: firstly, they are transparent to sub-luminous electromagnetic waves, and secondly, they have no free electrons and so there is no possibility of Foucault currents being formed, and so no thermal losses.

![Diagram](image-url)

Figure 8.1.
When we apply a magnetic field of a suitable value, the elementary magnetic moments, seen from the perspective of associated kinetic moments, start to precess, which in itself is not an intrinsically detectable phenomenon, because it is quickly dampened. But we can cancel this damping if we send into the ferrite an HF wave possessing a component perpendicular to the magnetic field, and by maintaining the phenomenon we can provoke an electromagnetic resonance of the volume at a certain frequency, and set off what we call gyromagnetic phenomenon. In particular, if the polarization of the HF wave is rectilinear, we can consider it as a combination of two circularly polarized waves, one in the inverse direction to the other. We can then easily see that there can be, according to the direction considered, either a favoring of the propagation or the contrary. In fact we can push these two tendencies to the extreme and obtain a gate effect: in one direction the wave propagates without attenuation, whereas in the other it is totally reflected. We can then produce a device that may be integrated into a wave guide. In the UHF domain we use no guides, their dimensions being prohibitive. But, still using the same principle of gyromagnetism, we can make localized-constant systems called circulators, whose basic model comprises three accesses or ports.

There are multiple technologies which allow us to create the interaction between the HF signal and the gyromagnetic volume ("fillet" circulators, in Y, strip), but only the latter has survived the cost war. The common characteristic is the trajectory of the signal in the interior: connector → track of impedance Z0 → adaptation circuit → interaction zone → mass. Figure 8.1 shows the simplified assembly of a 3-port strip, today the standard model in UHF, with the three tracks at 120°: the conductive circuit is packed in between the two gyromagnetic ferrites (soft); the magnets are also made of ferrites (hard). The field is adjusted with the aid of a superposition of metallic discs of diverse materials and thicknesses thanks to which we tune the permeability of the magnetostatic circuit, which closes by the mass of the housing.

We have also produced models with tunable fields, disposing the magnets in threaded bowls with a very thin step. The printed circuit is not indispensable: there are models where the central conductor is a simple circular metallic strip (silver-plated copper or bronze or silver-plated beryllium) with three bands linked to the connectors. But the printed board is best adapted to series production and more practical, for example, for the installation of adaptation components comprising a series self, and thus an interruption of the track going from the center to the connection points. The adaptation circuits, located by small gray circles on the diagram, are composed, in general, of a series capacity, a series self and a second capacity in diversion (Figure 8.2), these three elements having to be of a superior quality (silver-plated strap for the self, ceramic tiles with low ESR for the capacitors).
8.2.2. Usage

Even if it is never useless to know a little about the physical operating principles of a component or a sub-group, we must recognize that in the particular case of the circulators, we could, if need be, do without it, since its usage is simple. In terms of the functioning, in fact, we can compare the object with what a one-way roundabout is to road traffic: if we take the example of the 3-ports, the signal can be injected by any of them, numbered 1, but can only go to that which is numbered as 2, which corresponds to the predefined rotation direction marked on the housing, and the power reflected in 2 can only go to the remaining port, which is numbered 3 and connected to an adapted load. When this load is integrated, the circulator becomes an isolator. The main function of the circulator is then to channel the reflected power towards a resistance and to suppress it, transforming it into heat. Figure 8.3 illustrates the conventional presentation in a connectorized housing and the associated block diagrams. There are also miniaturized models without housing or
connector (drop-in) which can be directly soldered to the printed circuit as any other component.

A circulator is a device that we can build in short or medium band according to the performance needed. A 3-ports displays a similar response to that of a hybrid coupler, with, in particular, a sharp adaptation curve (Figure 8.4) where the accepted value will in fact determine the specified band width. This adaptation curve ($S_{11}$) depends largely on the internal circuit, which is very delicate to adjust, and is not tunable during production, and whose inevitable dispersion translates into a rather awkward uncertainty on the return loss. We must add to that that the ambient temperature and the internal dissipation are two factors that are extremely difficult to compensate, such that the specifications guaranteed by the manufacturers are often surprising by their apparently excessive prudence, the measures of the entry control being much better in general. But we find ourselves faced with the problem typical of specifications and their philosophy: should we or should we not allow degradations in performance when the conditions of usage are no longer the nominal ones, and if so, of what importance? To illustrate this idea, Table 8.1 gives an outline of the main characteristics of the connectorized circulators which we find on the market in the VHF and UHF bands, for a nominal power of 50 W, making it clear that the table only gives orders of magnitude, the values indicated depending largely on the definition adopted for the usage band.
<table>
<thead>
<tr>
<th>Frequency (MHz)</th>
<th>Dimensions (mm)</th>
<th>Amb. return loss (dB)</th>
<th>Extr. return loss (dB)</th>
<th>Amb. losses (dB)</th>
<th>Extr. losses (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>40/70</td>
<td>100×100 thick 35</td>
<td>20</td>
<td>15</td>
<td>0.3</td>
<td>0.6</td>
</tr>
<tr>
<td>40/70</td>
<td>100×100 thick 35</td>
<td>21</td>
<td>16</td>
<td>0.6</td>
<td>1.2</td>
</tr>
<tr>
<td>70/110</td>
<td>70×70 thick 25</td>
<td>20</td>
<td>15</td>
<td>0.2</td>
<td>0.4</td>
</tr>
<tr>
<td>70/110</td>
<td>140×70 thick 25</td>
<td>21/22</td>
<td>15</td>
<td>0.2</td>
<td>0.4</td>
</tr>
<tr>
<td>220/470</td>
<td>50×50 thick 25</td>
<td>21/23</td>
<td>16</td>
<td>0.1</td>
<td>0.2</td>
</tr>
<tr>
<td>220/470</td>
<td>100×50 thick 25</td>
<td>21/24</td>
<td>17</td>
<td>0.2</td>
<td>0.3</td>
</tr>
<tr>
<td>&gt;900</td>
<td>35×35 thick 25</td>
<td>22</td>
<td>17</td>
<td>0.1</td>
<td>0.2</td>
</tr>
</tbody>
</table>

Table 8.1.

In a circulator, the precision of the tuning of the magnetostatic field is of the order of 10^{-3}. We can then easily imagine the difficulties that will exist to obtain the performance stability, knowing that the final user usually demands that the system works at minimum power as well as at maximum, since radio networks nowadays are agile in power, and possibly in a large range of temperatures planned for external equipment. In the case of double circulators, real specialists know how to lay out the response curves to the ambient so that the variations of two levels are more or less compensated, as a function of the variations of power and temperature, but the problem remains in its entirety for simple circulators.

8.2.3. Characterization

The choice of a circulator is thus something extremely delicate and demands an in-depth investigation with regard to the manufacturer, should we want to avoid certain disappointments. For example, besides the predictable drifts mentioned above, it is also necessary to be attentive to the mechanical design, where two important details are to be systematically verified: the correct sealing of the magnetic circuit and the fastening system. The magnetic circuit must not be perturbed by the environment, given the fineness of its tuning. The central part
schematized in Figure 8.1 closes normally by the housing, which must then be made of a ferromagnetic material whose sealing must be perfect. The first thing to do when we evaluate a circulator is to place a scrap metal directly above the ferrites: a good product should not attract it, or, shall we say, barely do so. Along the same lines, the tightening of the fixing screw on the support must not induce deformations on the housing, which would be another cause of variation in the internal static field. All this can be seen and verified on the network analyzer, on which we should not note any noticeable deformation of the response curve as we proceed with these manipulations.

![Diagram](image)

**Figure 8.5.**

We can, if we take certain precautions, use the analyzer to visualize the frequency response of a circulator subjected to its nominal or maximum power. Figure 8.5 proposes an arrangement that allows this operation, and it is necessary only to make sure that the total attenuation of the measuring coupler and the attenuator is such that the return power is less than the maximum authorized (in general of the order of a Watt, that is, 30 dBm). This being so, it is the only way of knowing what will happen under the real conditions of usage, and it is thus absolutely necessary to use it at the moment when we choose the model to be integrated into the ensemble. It will also be necessary, with the aid of the set-up used in section 2.2.2, to verify that the levels of intermodulation are compatible with the rest of the equipment, that is to say, that they be at least 10 dB below the most restrictive specifications. The conventional method consists of adding the two frequencies with the aid of a 3 dB coupler before injecting them into the circulator. The disadvantage of the coupler is in making the system lose half of the available power, and this is the flipside of the total agility in frequency, that is, the possibility of using two arbitrary frequencies, even if they are very close. If we do not have amplifiers capable of delivering double the normal power specified, we can also couple by cavities or filters, by producing a mini-combiner, or also by duplexer, but it is necessary then that the two frequencies be different by a certain minimum separation, which can pose problems if, for example, these are imposed in the specifications.
But there also exists a second method, which is the direct translation of the real case where the system receives a perturber by the antenna: it consists of creating an intermodulation of the third order by injecting the second frequency at the exit, as indicated by the bottom diagram of Figure 8.6, and then deducting the intercept point. For the precise definition and properties of the latter, we will refer to the chapter on low-noise amps.

First set-up: we send two carriers of equal amplitude N with a separation ΔF, the IM₃ are equal in amplitude and to ΔF of the carriers. If we refer to as Δ the difference in level between the carriers and IM₃, in dB, (c for carrier), we have:

\[ IP₃ = N + \Delta / 2 \]  

[8.1]

Second set-up: we have a sole carrier at nominal power N₁, and at the exit we inject a small signal said to be a “perturber” of level N₂. We can still apply the preceding formula, considering that Δ now is the difference between N₂ and the highest intermodulation IM₃, and then putting N = N₁.
<table>
<thead>
<tr>
<th>Parameter to verify (in the specified band)</th>
<th>qualification</th>
<th>Unit inspection</th>
<th>Sampling control</th>
</tr>
</thead>
<tbody>
<tr>
<td>Insertion loss $\Theta_{amb}(S_{21})$</td>
<td>*</td>
<td>*</td>
<td></td>
</tr>
<tr>
<td>Entry adaptation ($S_{11}$)</td>
<td>*</td>
<td>*</td>
<td></td>
</tr>
<tr>
<td>Insulation ($S_{12}$)</td>
<td>*</td>
<td>*</td>
<td></td>
</tr>
<tr>
<td>Exit adaptation ($S_{22}$)</td>
<td>*</td>
<td>*</td>
<td></td>
</tr>
<tr>
<td>Temperature drift</td>
<td>*</td>
<td>*</td>
<td></td>
</tr>
<tr>
<td>Power drift</td>
<td>*</td>
<td>*</td>
<td></td>
</tr>
<tr>
<td>$S_{11}, S_{12}, S_{21}, S_{22}$ in power</td>
<td>*</td>
<td>*</td>
<td></td>
</tr>
<tr>
<td>Size rating</td>
<td>*</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Exterior magnetization</td>
<td>*</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

Table 8.2.

In conclusion, the circulator is a particular product for which it is necessary to proceed, during manufacturing, with systematic individual verifications, and sometimes repeatedly, because the object is susceptible, more than others, of varying in time. The standard list is given in Table 8.2.

8.3. The antenna alarms

The infrastructure of a radiocommunications network generally consists of a meshing of stations disposed in such a way as to geographically cover a certain territory. There are base stations, sometimes multifunctional, where permanent personnel work, and relay-stations, some of which are located in places difficult to access that are therefore not often visited and so have been automated. Even if the reliability of the equipment has seen enormous progress, nature takes care of supplying the technique for the random element, with which all human activity is confronted, particularly in the form of storms which strike antennas placed in high locations. The necessity to have monitoring and alarm systems was thus rapidly felt, after a heroic period where operators only knew about a malfunctioning through the complaints of customers. It was thus decided to equip the exits of multicouplers with what we call an “antenna alarm”, an electronic surveillance device that activates a dry loop, normally closed, which opens when it detects a misadaptation of the aerial.

Detecting a misadaptation means first of all characterizing it, which is only possible with knowledge of the reflected signal in relation to the incident signal. This implies two conditions: the first is that this evaluation is not possible unless
there is transmission, and the second is that we must install on the feeder of the antenna a bidirectional coupler. We will explore the theory and the functioning of this component in more detail later, but for the moment we will satisfy ourselves with a simplified approach: we will define it as a passive device of negligible loss, linear, directional, and capable of simultaneously providing a sample of the incident power (that is, towards the antenna), and a sample of the reflected power. We will subsequently deal with these two quantities in an electronic circuit which will analyze them and compare them with a predetermined value.

An antenna alarm is not a power meter. The latter is a costly laboratory or equipment device, which evaluates the return loss with precision and which for this reason often uses thermal, perfectly linear procedures. The antenna alarm is an auxiliary system and must be as cheap as possible; it does nothing but compare the direct and reflected powers without trying to determine their exact value. It must be able to work with a certain level dynamic whose lower limit corresponds to a carrier at the minimum power (of the order of a Watt or a few Watts), and at the upper limit for the totality of the carriers (often 8 or 16) at maximum power (of the order of 100 W per channel). As a function of the ambient parameters, the phases and what has been said, we have often fixed and it has become a standard, the alarm to be triggered for a reflection coefficient between 2 and 4, all the variables together.

8.3.1. Detection alarm

Figure 8.7.
The most widespread system uses the block diagram above (Figure 8.7). A bidirectional coupler provides two proportional signals in direct power (D) and reflected power (R), which are sent to the two channels of the alarm. They are first of all attenuated in such a way that they present themselves at the same level to amps 1 and 2. Each signal is then rectified in a diode detector, and the two continuous tensions proportional to D and R are presented to the two entries of a comparator, whose triggering triggers the opening of the relay when R surpasses a predetermined value. The two channels are identical, at values close to the attenuations $A_1$ and $A_2$, which are determined as a function of two criteria:

- the two voltages detected are equal for a reflection coefficient comprised between 2 and 4 and arbitrarily fixed to 3,
- it is indispensable, in order to avoid the return of intermodulations on the main line, that the insulation between this and the detectors be of at least 60 dB for a power per channel of 50 W.

We note on the block diagram the presence, on the comparator, of an inhibition function, which is active when there is no carrier emitted. In effect the alarm, being always in a state of functioning and being as the antenna at large band in relation to the utilized frequencies, is susceptible to receiving by it all the carriers emitted by other channels. The alarm can thus be triggered when the coupler receives through the aerial a parasite frequency of arbitrary origin. The inhibitor allows, to a large extent, protection against this eventuality by not authorizing the opening of the loop except during a transmission, but it is very evident that, even within this sequence, a parasite signal captured by the antenna can perturb the system.

8.3.2. Thermal alarm

The detection alarm is a common and largely used device. It has nevertheless some disadvantages, due to its working principle. First, the detectors are sources of non-linearities and require, as said before, a certain separation between the diodes and the antenna path: we are sometimes lead to add in the amplification chain an attenuation stage and a supplementary transistor, only to form this separation. Then, when there are several carriers, the system is sensitive to the phase, unless we use quadratic detectors instead of simple diodes, and this is a complicating factor. The conventional HF power measurement devices, such as the bolometer or the more modern “power meters”, use linear thermal sensors assembled in bridge, that transform a signal into imbalance current and allow us to take a proportional measurement.

*A priori*, the idea of transforming an HF signal into heat is attractive because it seems to eliminate immediately, by its very principle, all the potential causes of non-linearity that can inevitably be present in a purely electronic device. On the other
hand, and this is a decisive advantage when we are dealing with multiple simultaneous signals, we are certain to eliminate the problem of phases. The application of this principle first involves choosing the sensor which will perform the transformation of electromagnetic energy into thermal energy. If the final objective was to perform a real measurement of power, it would be necessary to conceive a linear device. For a long time, we deferred this constraint to the sensor itself, that was made in the form of metallic filaments in vacuum, for which we measured the variation in the resistance, considered proportional to the increase in temperature thanks to a manufacture using judiciously-chosen metallic alloys. This was the case until the arrival of the microprocessor, which allows us to linearize any sensor through the use of memorized corrective tables. This property authorizes the use of new technologies of sensors such as, in particular, NTC (Negative Temperature Coefficient) thermistors, whose value decreases when the temperature increases. In the case of antenna monitoring, as we are no longer dealing with a measurement but with the comparison of two signals, it is not necessary then to linearize and the microprocessor is consequently of no use.

In comparison to old sensors, NTC resistances present the advantage of having a greater sensibility (of the order of 5% per degree) for a much lower cost. These are semi-conductive sintered ceramics that we find on the market in the form of small discs or solid drops with two connection wires, sometimes in chip version. The initial product is a polycrystalline powder of metallic oxides (mainly Mn, Fe, Co, Ni, Cu, Zn), to which we add stabilizing additives. As a reminder, and without any importance for this application, the law of variation of the resistance as a function of the temperature is of the form:

$$ R = R_0 \cdot e^{B \frac{(T - T_0)}{T_0}} $$

$R_0$ being the value at a reference temperature $T_0$ (in general 25°) and $B$ a constant characteristic of the material. PTC thermistors also exist in this technology, but their non-uniform curve of variation directs them towards other applications. Either way, all the sensors that the industry will provide and whose resistance will vary in a continuous manner with the temperature, in one direction or the other, will be susceptible to being used, as long as their sensitivity is suitable.

Figure 8.8 shows as a block diagram, a simple circuit using two thermistors assembled in differential bridge. A bidirectional coupler provides two signals proportional to the directed and reflected powers. These HF signals are directed uniquely towards the thermistors thanks to the blocking selfs L and cause a heating up, and thus a variation of the resistance, as a function of the injected power. The continuous tensions $V_D$ and $V_R$ are thus functions of the two powers and allow the comparator to lock in one state or another according to their relative value. $R$ must be of the order of 5.6 to 10 kΩ for the polarization current not to be too important (for a 12 V bias voltage), and we will choose thermistors of about 1.5 kΩ at 25°. We will remark that the secondary outputs of the coupler are loaded with resistances of
56 Ω, which on their own give an already excellent adaptation of 25 dB. The placing of the thermistor in parallel will further approach the impedance load to the ideal value of 50 Ω. The symmetry of the set-up assures the compensation of all the perturbing factors: ambient temperature, variations in the bias voltage, warming-up of the thermistors themselves by polarization current. The installation of thermistors on the printed circuit is very important. First of all, they must be close to each other in order to be at the same ambient temperature, and also be away from all sources of excessive heat and be protected by a common cover.

![Diagram](image)

**Figure 8.8.**

*Remark*

We used for the thermal alarm a two-way coupler, unlike the transistor alarm which, in the diagram of Figure 8.7, is connected to a one-way coupler giving at the same time the direct and reflected powers. The latter is evidently simpler than the former, but does not work properly unless the loads are well-defined on the two sides and both are equal to the nominal value. This is normally achieved in the case of the transistor alarm but not in that of the thermal alarm where the thermistors constitute a variable parameter as regards the load impedance. The effect of a misadaptation is to degrade the directivity of the coupler, and so introduce a possibility of error, especially in the evaluation of the reflected power, the direct power being supposed larger and therefore less affected. However, it is necessary to recall once again that the antenna monitor is not a precision device and that only qualification tests will say if the two-way coupler is indispensable.
The following table will help with the technological choice by comparing the main characteristics.

<table>
<thead>
<tr>
<th>Comparison criterion</th>
<th>Detection alarm</th>
<th>Thermal alarm</th>
</tr>
</thead>
<tbody>
<tr>
<td>Complexity</td>
<td>simple</td>
<td>Very simple</td>
</tr>
<tr>
<td>Reliability</td>
<td>good (of the order of 40,000 h)</td>
<td>excellent (no transistor)</td>
</tr>
<tr>
<td>Possibility of IM3</td>
<td>yes</td>
<td>no</td>
</tr>
<tr>
<td>Phase sensitivity</td>
<td>yes</td>
<td>no</td>
</tr>
<tr>
<td>Temperature stability</td>
<td>very good</td>
<td>sufficient</td>
</tr>
</tbody>
</table>

Table 8.3.

From simple block diagrams, the two types of alarms are always updatable, from the technological point of view, as a function of the progress made on their specific components, transistors for one and thermal sensors for the other.

8.4. Loads and attenuators

8.4.1. Loads

A load (termination) is an adapted connectorized component intended to totally transform an electric signal into heat. It is constituted by an absorbing transductive part, where the transformation happens, which we can obtain separately in the form of a "chip", and of a radiator where the chip is fixed. It exists in a great number of models distinguished by the following characteristics:

– dimensions,
– min. PSWR,
– maximum acceptable power,
– connectorization,
– maximum frequency.

The definition of maximum power of a load is a subject of eternal discussion. Each manufacturer indicates a number, but not all specify under which conditions it is valid. It is necessary in particular to know, since the dissipated heat creates an increase in temperature, which equilibrium temperature must be taken into account.
If the equilibrium is attained by convection in ambient air, it is also necessary to specify the ambient temperature on the one hand, and the position of the radiator on the other hand. We must also not forget that its life cycle strongly depends on the temperature of use.

We often buy the full load, ready for application with its radiator and its connector, but it is also possible, if we want to enrich the catalog with this genre of product, to fabricate it starting from the chip. The latter usually presents itself in the form of a brazed ceramic patch over a fixing platform. Figure 8.9 shows two usual examples of this presentation. The part where the energetic transformation happens was initially constituted by a carbon-based “ink” and specific additives, which were replaced afterwards by a technology of a thin layer of beryllium oxide or aluminum nitride. The heart of the chip is dimensioned so that it presents the desired characteristic impedance, valid up to a certain cut-off frequency that we want to be as high as possible. For this reason, it is necessarily small, and the energy density is consequently considerable. The main difficulty, in the conception of the power load, will thus be the evacuation of the thermal energy towards the exterior.

Figure 8.10 shows the diagram of the device viewed generally and in the sense of its physical principle: the signal heats the chip from where the heat cannot escape except via the platform, of thermal resistance $R_{th1}$, which transmits it integrally to the radiator, of thermal resistance $R_{th2}$, which in turn is cooled down by convection in a fluid, generally ambient air, or by internal circulation of a heat-transfer liquid for loads of strong power. It is thus necessary to use materials whose thermal resistance is the smallest possible, whilst remaining compatible with the mechanical constraints, the cost price demands and the facility of supply. We will find there, without much surprise given certain similarities that exist between electromagnetism and thermodynamics, from the moment we consider electricity and heat according to
their fluid aspects, metals having the best electric conductivities, and in the same order: silver, copper, gold and aluminum, except for the fact that all are preceded by a metal forgotten by the mechanical industry, despite a promising start at the beginning of the 20th Century. We are talking about calcium, an even lighter metal than aluminum (density 1.54), whose thermal conductivity is one and a half times better than that of copper and silver, but that is not available on the market in an exploitable form from the point of view of fabrication. In the remaining group, the price of silver and gold means that these are reserved for the treatment of surfaces by electrolytic means, for the improvement of conductivity or protection against corrosion respectively. Finally, from a practical perspective, there are only two materials left for manufacturers of loads: copper and aluminum. Table 8.4 specifies the orders of magnitude of their conductivities.

![Figure 8.10](image)

<table>
<thead>
<tr>
<th>metal</th>
<th>calcium</th>
<th>silver</th>
<th>copper</th>
<th>gold</th>
<th>aluminum</th>
</tr>
</thead>
<tbody>
<tr>
<td>thermal conductivity (W/m·°C)</td>
<td>600</td>
<td>420</td>
<td>395</td>
<td>300</td>
<td>218</td>
</tr>
<tr>
<td>resistivity (Ωm)</td>
<td>4.5.10⁻⁸</td>
<td>1.58.10⁻⁸</td>
<td>1.72.10⁻⁸</td>
<td>2.21.10⁻⁸</td>
<td>2.72.10⁻⁸</td>
</tr>
</tbody>
</table>

Table 8.4.
Copper conducts heat almost two times better than aluminum, and some consider it necessary to interpose between the platform and the radiator an intermediary piece of this metal in the hope of reducing the global thermal resistance. But by adding a mechanical part, we also add an interface, and thus the possibility of trapping air during the set-up. With air being a good thermal insulator, this is something to avoid at all costs, which implies a mechanical finishing of the type “polished mirror” of the parts in contact, and if necessary the placing of a silicone-based flexible seal: we are not certain, ultimately, of reaching the desired goal. In fact, we note that a carefully manufactured load, with a platform fixed directly onto the aluminum radiator with a small amount of silicone grease to keep air out, behaves just as well and does not heat up, for the same dimensions as a load including an intermediary copper part. Figure 8.11 gives the usual appearance of a load of about 30 W, open, with a radiator of dimensions approximately 100 x 80 x 25.4 mm, darkened by anodic oxidation or another similar procedure to reduce the surface resistance and thus optimize the convective exchange with the ambient air. The small additional drawing below highlights a very important fabrication detail that is nevertheless not always respected by manufacturers, and this is why it is recommended to verify its execution: it is absolutely necessary to plan for an elasticity, in other words, a possibility of deformation without rupture, at the link between the core of the connector and the chip. This link is in principle made by high-temperature tin soldering, and the temptation is evidently strong to make it the simplest way, with a sole direct soldering. It is forgotten that in this way we closely join several rigid mechanical parts having different dilatation coefficients, and that every temperature
variation will create mechanical strains, the strength of which is always surprising. We thus noted, and this is only one example, that 30% of a contingent of loads stored for several months, waiting to be fitted onto their definitive place, revealed themselves outside of usage to have broken connections, during the verification inspection, uniquely due to the continuously and prolonged effect of diurnal and nocturnal variations in temperature of the place where they were kept. It is therefore preferable to perform not only one but two solderings, with an intermediary, thin and flat part, generally made of beryllium bronze and comprising an elasticity crease, not too accentuated so as not to reduce the adaptation band by the introduction of a reactive part, but enough to absorb these variations.

8.4.2. Reminders concerning the transmission of heat

The propagation of thermal energy, both in general and at the interior of a load, is a complex phenomenon with its laws described by Fourier in his *Traité analytique de la chaleur* (analytical treatise on heat), which can be considered in this domain to be an equivalent of the *Treatise of Electricity* by Maxwell with respect to electromagnetism. It is not necessary, however, for a projects engineer to go this far to be able to determine the dimensions of a load radiator or attenuator, as long as the profiles’ suppliers provide all the formulas and charts necessary. However, a reminder of certain definitions is probably useful.

![Figure 8.12.](image)

The transmission coefficient or thermal conductivity, annotated $K$, is a number indicating the amount of heat, expressed in Joules or calories, passing during the unit of time from a warmer body to a cooler one, with a difference of temperature of 1°C, and through an isothermal surface equal to the unit. $K$ will thus be expressed in W/m.°K or W/cm.°C. If we consider only two mediums to be crossed (Figure 8.12),
delimited by infinite parallel planes, we can define a thermal flux going from the
heat source, also supposed infinite in the upwards and downwards direction, to the
cooling medium where it is completely dissipated. This flux, relative to a surface S,
is equal to the thermal energy crossing it by time unit:

\[ \Phi = \frac{Q}{t} \]  

[8.2]

On the other hand, the value of the flux is given by Fourier’s law, written as:

\[ \Phi = \frac{K S}{e} (T_1 - T_2) \]  

[8.3]

If we rewrite this relationship in the form \( T_1 - T_2 = \left(\frac{e}{KS}\right)\Phi \), we cannot miss the
striking analogy with Ohm’s law \( V_1 - V_2 = R.I \), and this is why we refer to as
thermal resistance the quantity \( e/KS \), expressed in °/W:

\[ R_{th} = \frac{e}{KS} \]  

[8.4]

We remark that on the drawing of Figure 8.12 we have written down, for four
mediums, six temperatures. In the theoretical case, where the interfaces are simply
geometrical, we have \( T_1 = T_2, T_3 = T_4 \) and \( T_5 = T_6 \). In practice though, this is not
true: the entities which we have called mediums, become mechanical parts that we
are trying to connect as best we can, but that all present a certain surface state, and
between which we can trap air or be obliged to insert a seal: there is always
something. It is such that, with the thermal resistances adding in series in the same
way as the ohmic resistances, we risk finding ourselves in the presence of an
unaccounted-for resistance, susceptible of causing, by an unpredicted increase of the
global value, a degradation of the cooling, with all the consequences this entails.
This justifies the remark, already made previously, that it is better to fix the chip of a
load directly onto the aluminum radiator, not forgetting the little gram of silicone
grease that will dispel the air, rather than adding an intermediary copper part, if we
are not certain of the good quality of the contacts.

Besides this traditional technique of manufacturing power loads, it is necessary
to mention the usage of simple resistances, inserted in cases and connectorized, for
powers less than 5 W, and in addition, the loads constituted by coaxial cables. There
are, in special execution, lossy cables specially devoted to this application, but any
cable can be used in principle: all that is necessary is to wrap a certain length of it in
order to obtain the attenuation which will allow the dissipation of the heat to be
better distributed through the space, a low-power load to be used at the termination,
and, most importantly, a termination with very weak intermodulations to be obtained.

8.4.3. Attenuators

All the generalities valid for the loads are evidently also valid for the attenuators. The need to have the adaptation at the entry and exit leads to a basic set-up which is a bit more complicated, where we go from one to three resistances. From this fact, we already know that an attenuator will perform less in terms of frequency than a load of the same power, because the chip will be larger. The set-up is necessarily symmetrical in its scheme, in T or in π, but not necessarily in the construction: the first resistance absorbs more than the last, the two can then be dimensioned in a different way if there is a benefit in doing so. This is the reason for which certain power attenuators bear a mark for the connection direction, which is not arbitrary and must in fact be respected at the risk of unacceptable incidents. Aside from that, the laws and rules mentioned before apply, and it is only necessary to keep in mind that in this case only a part of the power is converted into heat. Thus, when we inject an HF power of 50 W into a 3 dB attenuator, this must respond to a nominal power of 25 W. An attenuator of 30 dB connected in place of the previous one will have to support 50 W.

Figure 8.13.

The same final remarks of the last section regarding loads are equally valid for the attenuators: below 5 W, we can use chip resistances in T or π inside a connectorized case, adding close by, to the interior of the main housing, a deformable intermediary cover to add and tune the parallel capacity necessary to obtain the adaptation. On the other hand we can conceive high-power models with weak intermodulations, distributed thermal dissipation, using a coaxial cable whose external insulating protection will have been removed and that we can wrap over a
connectorized support in order to create high-performance laboratory equipment (Figure 8.13) that is extremely hardwearing, avoiding the concentration of heat at one point. If we want to use the principle without possessing or wishing to build the product, for an isolated experiment, we can satisfy ourselves with diverting, for several instants, a reel of coaxial cable taken from a stock and equip it with two plugs; this is an operation that has been performed many times.

8.5. Reception amplifiers

8.5.1. Introduction

![Figure 8.14. Kurokawa set-up](image)

We may be surprised to find in the chapter on auxiliary devices an electronic sub-assembly of such high technicality such as a low noise amplifier, that constantly demands the best-performing components in terms of transistors. It is simply because we wanted to leave it to real specialists in the physics of semi-conductors to do the fundamental study, while we are happy to simply apply the results of their research where it must be placed within the architecture of a system. The real work we expect from the radio engineer is to know how to use the best possible way the last transistor made available on the market, incorporating it into the plan of an amplifier of average gain, generally comprised between 3 and 15 dB, of excellent reliability, and which, for these two reasons will be the simplest possible. It is almost only in the collective distribution of TV antennas that the number of channels to be split leads to set-ups with two or three transistors. In telephony the standard is to have one transistor per reception system, the aim being to compensate the attenuations and assure a final gain of a few dB: when, in 1999, the Radiocom 2000 network was dismantled, having been in service for about 15 years, almost the totality of the monotransistor amps of the reception multicouplers were still working. There are nevertheless those who support a so-called Kurokawa set-up (Figure 8.14), where we have two amplifiers between two directive couplers, as in the filters of same name, and whose justification was initially to provide, in the case...
that one of the amps fails, an element that is degraded but still exploitable. This is, in our view, a disputable reason because the extent of the degradation depends on the destruction mode, but on the other hand it is certain, by virtue of the reliability calculation mode, that the latter is almost two times lower than that of an amp with a single transistor. On the other hand, the intercept point theoretically increases by 3 dB, and this is the only real advantage that we retain, but the noise factor is no better than that of a single amplifier.

It is always good to recall that, in the amplifier, the source of the power is not the transistor, but the power supply. The study of the latter thus needs to be given particular attention. It must first be capable of providing the necessary current, and anticipating a safety margin is not a luxury: we must not hesitate to over-dimension. It must also be perfectly stable, that is, exempt from residual ondulations and instabilities. It is necessary in particular to beware of low-frequency oscillations that are most of the time undetectable by HF measurements focusing on the useful signal, aside from noise measurements, which will trigger the alarm.

This said, we have at our disposal, to assure the amplification in reception, two families of transistors: the bipolars, silicium-based, and field-effect transistors, referred to as III/V, in reference to the number of the columns of the Mendeleiev table where we find the basic components. Between 1990 and 2000, we witnessed an undecided battle between two technologies, one alternatively taking the lead in relation to the other in an ephemeral way. Then the AsGa FET broke away to rejoin, then surpass, bipolars in the domain which constituted their initial weakness, exit power, it being made very clear that we are dealing only with low noise transistors here. To say that the fight is over would probably be unwise because, in the same way that there is today no point in increasing the power of car engines, the performance achieved by HF transistors is such that we can no longer expect any technological upheaval: below 0.5 dB of noise factor and simultaneously with an intercept point of several Watts, we have what is needed to resolve all current problems in terms of radio reception, and it is possible, and probable, that the future of these components depends on economical and reliability considerations, rather than purely technical.

Either way, our proposal is uniquely, without wishing to enter into the details of the domain of fundamental physics, for which we refer to the specialized monographies, to know how to use these transistors as they become available to us. Figure 8.15 shows the two simplest basic set-ups, with a positive feeding tension, one for an NPN bipolar transistor, the other for a field-effect P-channel transistor, used in the function of a reception multicoupler. These two set-ups served as basis for several generations of amplifiers which have perfectly fulfilled their mission. They are cascadable, since by principle adapted to entry and exit, that is, they can constitute multi-level amplifiers capable of a high gain in power. In this case the
first transistor will be chosen for its noise factor and the latter for its intercept point, which evidently reduces simplicity and reliability, but gives more possibilities than with a one-transistor model, which must achieve excellent performance at the same time in both respects.

In all that follows, we will suppose that we are in the most common conditions, that is to say that the bias voltage is positive, that the bipolar transistor is of the NPN type and that the field-effect transistor is channel N.

1) Bipolar transistor (8.15a)

We will not make reference here to a particular transistor, the aim being instead to set out the practical rules to be employed, if possible with permanent validity, that is, adaptable with minimum modifications to all new products. We will exclusively use the set-up of a common emitter, scrupulously applying the installation instructions of the manufacturer. The latter provides with each type of transistor a complete documentation where one can find the test plan, from which the measurements are made, noise circles, the intercept point or the maximum power,
the entry and exit impedances as a function of the frequency, the max current not to be surpassed, as well as all the useful details which allow us to install the transistor on its printed circuit in the same way the manufacturers did it. It is important to know, in fact, that the first question posed by them, if we do not obtain the expected performance, will be whether we have followed their recommendations properly, and that is what logic and caution also lead to. These precautions of a general order being taken, we realize a first prototype circuit according to the model in diagram 8.15a. The self-transformer $T_s$, which usually comprises between two and six turns over a low-$\mu$ ferrite, in order to avoid saturation, is not vital in itself, but constitutes a change of impedance at wide band that can facilitate the exit adaptation if in the beginning we are too distant from $Z_0$, while assuring the self-blockage function which, from the point of view of the signal, isolates the transistor from the power supply, and that is itself necessary. If the transformer is reduced to a self, we connect the exit of course on the collector. The resistance $R_4$, of the order of 33 k$\Omega$, transmits the emitter voltage to a surveillance system which triggers a dry-loop alarm in the event of a failure of the transistor, a complement that now has become conventional. The entry and exit lines, if necessary equipped with stubs, but also made of discrete components which take less space, allow us to transform the impedances according to well-known rules, for which we will refer to the specialized University courses on microwaves. No resistive component should be found upstream of the transistor, so as not to degrade the noise factor: the attenuator of the gain adjustment, for example, will necessarily be installed in the vicinity of the out port.

2) Field-effect transistor (8.15b)

In the same way as the bipolar transistor is installed in a common-emitter configuration, we will use a set-up in a common-source configuration for the FET. All the general remarks made before apply here, the specific feature of the FET simply being that the gate voltage must be negative relative to that of the source. The entry will thus be connected to the mass by a resistance that can be of a fairly large value, since the entry impedance is much larger than that of the bipolar, and the source will be carried to a positive potential, if we take an N-channelFET, via resistance $R_2$. The value of this is usually chosen so that the drain current is fixed at $I_D = 0.6 I_{D\text{max}}$, a value corresponding to the best yield, and under the condition that the achievement of the specified intercept point does not demand a higher value, in which case the drain current will have to be increased. Whatever the type of transistor chosen, the capacitor $C_4$ will have to be of excellent quality: we will choose in practice metal-coated ceramic chips of very low ESR (Equivalent Serial Resistor), that is to say, very low $\text{tg}\delta$. If the source has got two connections, we will symmetrically split $R_2$ and $C_4$ to each side of the transistor.
The self L1 and the capacity C5 have no functional utility in the amplification of the signal, but they allow us to protect the field-effect transistor from electrostatics discharges, to which it is very sensitive and that could instantly destroy it during the manipulations of tuning and measuring. This accidental phenomenon has a considerable importance in production. When the first large serial production chains of the first generations of mobiles was released, a failure rate was noted that was considerably higher than that forecast, and it was quickly realized that these failures, which affected field-effect transistors, were correlated to the phases of production. The transistors were not necessarily destroyed, but at the very least their noise factor was strongly degraded and they were in any case weakened. Today all cabling staff are equipped with ground straps, and they work over graphite carpets, and things have fallen back into place.

![Figure 8.16. Architecture of low-noise amp](image)

Finally, and this is a useful complement that we add to the malfunction alarm we mentioned before, we often insert into the exit line a strip coupler, with a coupling of the order of 30 dB, that does not disturb the set-up and allows a global view of the received frequencies by connecting a spectrum analyzer to it.

Figure 8.16 shows the typical appearance of the installation of a low-noise amplifier for a standard reception multicoupler, where the essential characteristics of the product are summarized: entry and exit distanced from each other, adaptation lines, alarm, exit coupler, global simplicity, low cost.
8.5.2. Intercept point

For a long time we characterized transistors by their saturation power, the maximum value we can obtain at the exit and to which it did not necessarily survive, and then by the compression point at 1 dB, measurable without risk of destruction, which is the value of the exit level situated 1 dB below that which we would have been able to predict by extrapolation of the linear part of the power response curve.

The growing importance of the levels of intermodulation parasite rays quickly asked for a new approach, providing a possibility of predicting the level of the intermodulations generated by the transistor, particularly of the IM3. 

In fact, in multi-channel reception, when these are regularly spaced as in mobile telephony, the IM3 created by the carriers of two adjacent channels automatically fall in the two lateral channels, which they disturb to a lesser or greater extent according to their level. In fact, all intermodulations of odd-numbered order will be found in channels and will be potential perturbers. TRT, which produced the radiocommunications system of Jeux d’Albertville in 1992, developed a program allowing us to calculate the frequency combinations susceptible to fall in the channels, up to order 17. Laboratory measurements have effectively shown that a part from the IM3, the strongest intermodulation was of the order 15! The maximum powers of the carriers were of 120 watts, that is, 50.8 dBm, and the minimum level of reception at –175 dBc, that is, an absolute value of –124.2 dBm, which brought the level not to be exceeded for parasite rays to about –130 dBm. The network worked.

Figure 8.17 shows the geometrical construction of the intercept point and its correspondence with the signals. We have two neighboring carriers of frequencies $F_1$ and $F_2 = F_1 + \Delta F$, of the same level $P_s$, which give IM3 at $F_1 - \Delta F$ and $F_2 + \Delta F$. When the levels of the two carriers increase simultaneously by 1 dB, the level of IM3 increases by 3 dB, that of IM5 by 5 dB, etc. If we continue to increase, we arrive at a value for which all IM products are at the same level as the two main ones: this is the intercept point. This is therefore the center of the beam of lines representative of all the odd-numbered intermodulations generated by the amplifier. On the graph to the left, in log/log coordinates, the first bisector angle represents a unitary gain, and the response curve of the amplifier is, in its linear part, a parallel line offset by the value of the gain G. The line of IM3 has a slope equal to 3, that of IM5 equal to 5, etc. and in particular we see that the length of the segment BC is double that of AB. It is therefore extremely easy, with this representation, and this is why it exists, to predict a relative intermodulation level of arbitrary order, once we know the intercept point. This is generally situated at 10 or 15 dB above the
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compression point at 1 dB and it will have as a value, if $\Delta$ is the difference between the level of the carriers and that of IM$_3$:

$$IP_{3,\text{out}} = P_s + \frac{\Delta}{2}$$

see [8.1]  [8.5]

And in a more general fashion:

$$IP_{n,\text{out}} = P_s + \frac{\Delta}{n-1}$$

[8.6]

We use this relationship in the way we want, be it for calculating the value of the intercept point, or to predict the level of IM$_3$. Some have the habit of taking this value to the entry:

$$IP_{3,\text{in}} = IP_{3,\text{out}} - G$$

[8.7]

Figure 8.17.
There is another method to determine the intercept point, which we have already described before, that uses a sole carrier and a perturber which we inject at the exit. It is less physical but is considered more legitimate as regards complete equipment, because they better reflect the real problems. It is still necessary to define the level of the perturber (see section 8.2.3).

8.5.3. Noise factor

The noise factor expresses the degradation of the signal/noise ratio in the crossing of a quadrupole or of a chain of quadrupoles. It is equal to the signal/noise ratio at the entry divided by the signal/noise ratio at the exit:

\[
F = \frac{(S/B)_{in}}{(S/B)_{out}}
\]  

[8.8]

In fact, we measure it in absence of signal, with the help of a noise meter which injects at the entry of the quadrupole to be tested, a white noise generated by a wide-band noise diode, whose emission curve is recorded in the form of a table of level/frequency relations, and to which we compare the noise at the exit of the device under test. This must imperatively be adapted at the entry and the exit so that the measurement is valid. The measuring device directly gives the value of the noise factor, but taking into account all that is detected: if there is some instability in the power voltage, for example, or a parasite signal is received in a way or another, the additional untimely power will make the result increase by a value that could be very significant. It will be up to the operator to uncover the "spurious" by making good use of their intuition.

Along these lines, we should remark that the real noise is not limited to a thermal white noise, but also comes from different origins: shot noise, partition noise, generation/recombination noise, scintillation noise. However, we will suppose, for the sake of simplicity, that all are reduced to a phenomenon of thermal agitation, and this is, moreover, the sense of the measurement with the noise analyzer, which implies that the noise depends on the temperature. We must therefore specify, at the time of a noise measurement, at which temperature it was done. If this specification is not mentioned, we can conclude that it was made at the reference temperature of 290 K, that is, 17°C. At this temperature, the noise spectral density is:

\[
kT_0 = 3.98 \text{ W/Hz}
\]  

[8.9]
That is $-174$ dBm in a band of 1 Hz or $-114$ dBm in a band of 1 MHz. If we make the calculation for a telephone channel of 10 kHz band, we obtain a noise level of $-134$ dBm: we are practically at the level at which we measure the intermodulations, which leads radio technicians to say that they are starting to “see” the noise. This means as well that the performance of mobiles is getting closer to the physical limits in terms of sensitivity.

What is now interesting to know is the relationship existing between the total noise of a network of quadrupoles and that of each of its constituents. This relationship is called the Friiss formula, and we refer to the works of reference for its demonstration, which uses the concept of noise temperature.

\[
F = F_1 + \frac{F_2 - 1}{G_1} + \frac{F_3 - 1}{G_1 G_2} + \ldots + \frac{F_n - 1}{G_1 G_2 \ldots G_n}
\]

[8.10]

This very important and frequently used formula requires some comment. First of all, it is always necessary to keep in mind that the different noise factors are expressed in real numbers: we can only pass to logarithms and decibels when all the algebraic operations have been performed. Afterwards, the relationship shows that $F$ depends directly on $F_1$, which must therefore be minimized as much as possible, and that the noise of all the successive quadrupoles depends on the gain of the first stage. It is therefore important that it be as large as possible, so that it can “mask” well, from the point of view of the noise, the one which is found behind it. Finally, if we are dealing with a passive quadrupole, such as an attenuator, the noise factor of this quadrupole is equal to its attenuation in dB: a filter of 1.2 dB of losses will have a noise factor of 1.2 dB, an attenuator of 10 dB will have a noise factor of 10 dB.

Example: a reception multicoupler is composed of a pre-selective filter with a loss of 1.4 dB, followed by an amplifier of 12 dB gain and noise factor of 0.8 dB, and a 4-channel distributor with a loss of 7 dB. What is the global noise factor?

\[
F_1 = 1.4 \text{ dB} \rightarrow 1.38 \\
G_1 = 1/1.38 \\
F_2 = 0.8 \text{ dB} \rightarrow 1.20 \quad G_2 = 12 \text{ dB} \rightarrow 15.85 \\
F_3 = 7 \text{ dB} \rightarrow 5.0
\]

\[
F = 1.38 + \frac{1.20 - 1}{1/1.38} + \frac{5 - 1}{15.85 / 1.38} = 1.38 + 0.28 + 0.35 = 2.01, \text{ that is, } 3 \text{ dB.}
\]
We can see very well through this numerical example what we mean when we say that the amplifier “masks” what is found behind it, in this case a passive distributor whose apparent contribution to the noise factor passes from 5, when it is isolated, to 0.35. We could thus conclude that it is always beneficial to have a high gain: we will see later on (reception multicouplers, Chapter 11), that this is not an absolute rule, though.

8.6. The impedance adaptor

![Figure 8.18.](image)

The precision of an adaptation can have very important consequences on certain characteristics that directly depend on it, such as, for example, the directivity of the coupler. It is now necessary to be able to go beyond 20 or 23 dB, which constitutes the norm of the “correct” adaptation, in order to go down to the suitable level of reflected power, which can be as small as necessary. There exists a very simple device which thus allows us to improve an adaptation as we desire, only subject to the condition that at the start it is better than 15 dB. It is a simple \( \pi \) circuit composed of a small self and two adjustable capacitors, that we can easily fit in a connectorized housing of small dimensions and low cost (Figure 8.18). The capacitors are air trimmers of 1 to 10 or 14 pF, the self is composed of a few turns of silver-plated strap wire of diameter 1 mm on a chuck of 5 or 6 mm; three turns in VHF, only one in UHF. The tuning is performed with the aid of a network analyzer, by alternately varying the two capacities. If we use the Smith chart, we always manage to arrive at the center, starting off from a point located at the interior of the 15 dB circle. The direct loss, of a few hundredths of dB, is negligible, and we can obtain directivities greater than 40 dB on an octave with any 3 dB coupler.

8.7. The 2\(^{nd}\) harmonic rejecter

In the same way as the previous point, we find in certain catalogs another small case with the same appearance, but with only one trimmer, which can be of great use in a radio station, as part of the fight against the parasite rays due to power
amplifiers. The values of the components are the same as before and the appearance very similar (Figure 8.19). We will easily obtain rejections of 40 dB of the 2nd harmonic, which is an economic and rapid way to solve certain problems regarding compliance with the norms and the hertzian cohabitation in the battle-like space of radiocommunications. The direct loss is as low as that of the adaptor, one of two hundredths of dB, the adaptation depending on the quality of the line section linking the two connectors, and it is very easy to attain 30 dB.

Figure 8.19.
Chapter 9

Directive Couplers

9.1. Introduction

We use the adjectives “directional” and “directive” indiscriminately to qualify the couplers that present this property. A directive coupler is formed by two parallel lines, one said to be the main and the other the coupled, the second being close to the first and consequently being influenced by its radiative electromagnetic field. The ensemble is therefore an octopole since it comprises four accesses. The lines can be of any type, microstrip on a single circuit, microstrips in opposition, wired, hybrids, or even wave-guides. In the last case, it will simply be necessary to create one or more openings in the common wall to guarantee the coupling. When the parallelism of the lines and the matching of the four accesses are done, the coupler presents a directional function which makes the coupled line deliver a signal that is proportional, either only to the direct power crossing the main line (exit 3), or only to the reflected power (exit 4). The theoretical study is based on the principle that the main signal induces in the coupled line two signals propagating in opposite directions, called even mode (for that propagating in the same direction as the main
signal), and odd mode, for the second. Each one gives rise to a different impedance: $Z_{0e}$ for even mode and $Z_{0o}$ for odd mode, these two quantities depending on the type of the line and the procedure used. We thus show that there is matching when the condition $Z_0^2 = Z_{0e} \cdot Z_{0o}$ is satisfied. The multiplicity of possible technological solutions implies just as many different formulas, too numerous to be listed here, and for which we refer to the works of reference (Gupta, Wadell, university courses, etc.). Figure 9.1 shows a conventional representation, in microstrip, because this is the simplest to imagine, but equally valid for the other types, and for which we will establish the essential characteristics and definitions:

– the coupling coefficient indicates the relative level of the available power on one of the two coupled ports, in relation to the power of the main signal: if this ratio is of 1/100, that is, – 20 dB, we will say that the coupling is of 20 dB, or that the coupler is a 20 dB coupler. The coupling depends on the length of the part in question. When this length reaches $\lambda/4$, the response is symmetric in relation to the central frequency. When, in addition, the power available on the coupled line is equal to the transmitted power on the main channel, we have a 3 dB coupler;

– a coupler is theoretically reciprocal, or in other words, reversible, which means we can invert the role of the two lines and decide that the coupled line will be the main line. In practice, this is not always true, from a purely technological point of view: we can be lead to under-size the coupled line, if the coupling is very strong (say, greater than 10 dB), for economic or lodging reasons, and in this case the accesses are consequently marked. A 3 dB coupler, on the other hand, is always permutable and we can enter via any access. In its splitter function, taking into account the phases of the two exit signals, we distinguish between the 90° and 180° couplers;

– the directivity is expressed, in practice, by the ratio between the power of the exit signal on one of the accesses of the coupled line and the power on the same access when we invert the direction of propagation on the main way. It is therefore a quantity, generally expressed in dB, which represents the effect on the accuracy of the measurement of the direct power made with the aid of the coupled line, of a reflected power in the main way and above all the reverse, that is to say the error made on the measurement of the reflected one, because it is evident that it is mainly the direct one that will disturb the measurement of the reflected one, rather than the reverse. It is therefore a very important datum, which reflects very well the quality of the coupler and that will by necessity have to be taken into account if we make an estimation of the error. It must be greater than 20 dB, but we very easily reach 40 dB on an octave by proceeding with judicious impedance matchings, on one or more accesses, what can be done either externally with the adapters, which we dealt with in the last chapter, or internally with procedures that will be detailed later on.
9.2. Technologies

The way in which the two lines are coupled depends on the use we want to make of the product: strong or weak coupling, high or low power, etc. The technological solutions are many, notably in microstrip, and we will limit ourselves to the most common ones. Figure 9.2 shows a few of the main geometries, the most used, whose advantages and disadvantages we are going to explain.

![Figure 9.2: microstrip coupler](image)

This is the basic coupler, which serves as a model for almost all theoretical studies, and consists of two parallel strips brought closer over the same substrate. The sides under consideration having as height the thickness of the strips, either 35 or 70 μ, the coupling is by necessity very weak. This is a model that is suitable for coupling coefficients that are greater than 20 dB (one should say: less than –20 dB, but this is not usual usage), and it is out of the question to produce a “3 dB” with this technology. To illustrate the orders of magnitude, the separation between the strips for a coupling of 20 to 30 dB is a few tenths of a mm for a coupling length of λ/4. Nevertheless we use it fairly often, notably as equipment complementing the low-noise amplifiers on a printed circuit, because by means of a few additional cm2 on an already existing support, we can consider it is of zero cost. In addition, given the low value of the coupling, the resulting perturbation of one line by the other is negligible and their impedance remains practically equal to Z0. On the other hand, it is clear that the acceptable power will itself also be limited, but this is of little importance under the usual conditions of use. If we want to be more precise, a spacing of 0.3 mm corresponds to a breakdown voltage of 375 V in air (value measured between two needles). The line will then be able to allow a signal having this value of peak voltage, that is, an effective voltage of 265 V, or further still, an effective power of about 1,400 W over a load of 50 Ω. This can seem a lot, but one should remember that such a weak interstice between two conductive bands constitutes a very efficient receptacle for all the dust, fillings and other micro-debris of which certain are conductors and can be joined by gluing vapors, in such a way that, even in the case of a glazed printed circuit, this device is not made for power because it is not reliable.
Figure 9.2b: layered hybrid coupler

We call a layered hybrid coupler a coupler “3dB”. This name comes from the previous usage of the “hybrid transformer”, so named because it in fact involves two transformers, a splitter and an impedance transformer (see later on in this chapter), and which was the first directive device capable of separating a signal into two equal parts or, reciprocally, to combine two independent signals. The basic idea is of putting face-to-face the tracks of two identical quarter-wave microstrip circuits, fixing the separation between the two by inserting an insulating HF layer of relative permittivity $\varepsilon_r$, the whole being compressed inside a housing. The latter will guarantee at the same time the mechanical hold, the precision of the separation, on which the coupling coefficient depends, and a dielectric rigidity much greater than that of air. This is the most conventional model and the most widespread in the industry. We usually find it in the form of a plain aluminum case with four fixed connectors matched to the frequency band. The other procedure that applies the same principle is to dispose the two tracks in opposition over a double-face printed circuit placed in central position, and to make the separation with the ground by means of a piling-up of insulating layers whose permittivity and thickness will assure the correct impedance. When the fixed connectors are of the N type, we usually trim two angles in such a way that they do not oversize the housing (Figure 9.3), elegance which is moreover pointless because the plugs that will be connected above will do so in their place. We have here one of the basic products of radiotelephony, well mastered, of large diffusion, robust and allowing relatively high powers. There are multilayered miniaturized versions for direct set-up on a printed circuit, with degraded performance but which even so fulfill the desired function. The fact that we use a double-face circuit allows us in addition to fold the tracks in a reduced space and to be able to descend fairly low in frequency: a hybrid 70 MHz takes up little more place than a 400 MHz model.

Figure 9.3.

Figure 9.2c: air coupler

When the power in transit increases, the printed circuit ends up presenting more disadvantages than advantages, notably from the point of view of losses, and we pass now to a technology where the dielectric is the air and where the tracks are thick silver-plated metallic strips. But above all, there is the possibility here to build a coupler of very low losses, and consequently of minimum heating, which is very
Directive Couplers

important when the power is measured in tens of kW. In this kind of coupler the
tacks are fixed directly onto the central conductors of the fixed connectors. These
must then be chosen sufficiently solid to assure the mechanical hold of the internal
equipment, upon which the stability of the electric performance directly depends, in
particular of the adaptations. This is a structure that is electrically equivalent to that
of Figure 9.2a, with the difference that the insulator is all air. We will therefore tune
the line impedance, or in other words, the capacitance between the line and the
ground, by offsetting the latter to a lesser or greater extent within the interior of the
housing, with the aid of a supplementary part as shown in Figure 9.4.

In these two types of coupler, and in fact in all those which are connectorized,
the link between the tracks and the connectors introduces an impedance solution of
continuity which degrades the matching to the accesses by virtue of the
predominantly inductive nature of a direct strap link. We remedy this disadvantage
by adding at this place a parallel capacitance that will create the equivalent of a line
section and that will be either a discrete component in the case of the coupler with
strips, or a patch in the case of a printed-circuit coupler. We then transform the
perturbative connection into an equivalent of 50 or 75 Ω line.

Figure 9.4: coupler for measuring device

This is a very specific directive coupler we use in the measurement of direct and
reflected powers and that is associated with a monitoring system that can be a simple
galvanometer graduated in Watts, as we can see in the popular power measurer
“BIRD”, that we insert in series into the circuit where the signal to be evaluated is
passing. The latter passes through the interior of the measuring device by a
cylindrical bar of sufficiently large diameter so that the resistance is negligible, and
the low part of the coupled line is parallel to this main line matched by the proximity of a ground plane. The coupled line is fixed onto a rotating device with two positions at 180°, in the form of a plug that we position, after choosing it as a function of the frequency band, in a housing that will mechanically guide it until the correct position and will then allow the technician, by a simple operation, to switch the measurement to direct or reflected.

9.3. The hybrid transformer

A hybrid transformer is in fact an association of two transformers whose ensemble is intended to fulfill a function of 3 dB coupler/splitter, hence its name. It is a device used mainly in reception circuits because the ferrites over which we place the windings have not got the necessary size to support high powers. Their aim is moreover that of making dividers by two of small dimensions, able to be integrated into reception cases where they will often be associated with low-noise amplifiers. The first transformer T1 is in reality a self-transformer of ratio $\sqrt{2}/2$, whose role is to transform the impedance $Z_0$ into $Z_0/2$ (Figure 9.5). The second T2 is a symmetric splitter giving two signals of equal amplitude and of opposite phase, from the signal present at the middle point M, the three accesses 1, 2 and 3 being adapted thanks to
the impedance transformation. Theoretically, this is a high-pass device whose passing band is not limited at the top end except by that of ferrite (Figure 9.6c). It is here, moreover, that the problem is, and this is why we have great difficulty, starting from a few tenths of MHz, in arriving at 1 GHz. The equivalent microstrip represented in 9.6b has better performance in frequency, but reaches a much greater size. The ferrite model needs, to compensate for the low band of the core material, the addition of a capacitor C whose value is comprised between 1 and 10 pF in VHF and in UHF. The resistance $R$ that links the two symmetric accesses absorbs the imbalance currents due to the imperfections of the windings. It should theoretically be of $100\,\Omega$, but it is almost always greater and serves, by its proper value as well as by its parasitic inductance and capacitance, in association with the central capacitor, for the optimization of performance, both in adaptation and in passing band. The windings are generally made of enameled copper wire and have as few turns as possible: often three turns with two for the impedance transformer, which gives at the point $M$ $0.44\,Z_0$ for 0.5, or better 3.5 with the possibility of 2.5, still closer to $\sqrt{2}$, which gives $0.51\,Z_0$, and two times one turn for the symmetric transformer. Certain constructors develop models for TV reception where the winding is a printed circuit track on which we thread a two-hole ferrite, thus obtaining unbeatable production costs and reproducibility, well suited for production in large series, the winding being suppressed, and a single passing band covering all the television sub-bands. The choice of ferrites is extremely delicate, and demands an enormous amount of tests and manipulations to open onto a narrow selection. A good combination, since the cores can be very different for each type of transformer, is the following:

- impedance transformer: $6 \times 7.1 \times 4$ mm Siemens two-hole ferrite, ref. B62152 A0007 017;
- crossover transformer: $4.3 \times 2 \times 3.1$ mm LCC tube, ref C3BT0410H.

We introduce the whole onto a double-faced circuit as shown by Figure 9.7 as an example of production. There are also ready-made products, in chips or connectorized housing, with guaranteed performance, which are more expensive but will suit specific applications where the user has no time or means to perform a study by themselves.

The hybrid transformer is evidently reversible and above all it is cascadable, that is to say, we can make successive divisions by 2, in power, of a received signal. This is the most conventional disposition, called “Wilkinson”, of the distribution over several channels of reception multicouplers, whose detailed description will be detailed further ahead.
9.4. The 180° hybrid ring

This type of coupler – because it is, functionally, in fact, a 3 dB coupler – has in common with the preceding one the fact that it does not bring into play the coupling between two lines, but it uses, as that one, simple phase-difference devices. We mean by this a phase difference caused uniquely by differences of electrical lengths in separate lines, in this case by coaxials on the chosen model, but that can also be microstrips. The comparison stops here: the hybrid transformer is intrinsically a wide-band system, whereas the hybrid ring is by conception narrow-band. It is nevertheless a commonly used object, robust, very stable and capable of supporting high powers, limited only by the breakdown voltage of the lines used; it therefore
has its fans. Figure 9.8 reminds us of its structure and the type of its response in frequency: four line sections, three of them of length $\lambda/4$ and the fourth of length $3\lambda/4$, are welded to each other to create a bridge. We suppose, for the sake of clarity, that the system works as a splitter and that the signal to be split enters at 1 and exits at 2 and 3, with half the power, but it is clear that it will work just the same as an adder, with the entries at 2 and 3 and the coupled exit at 1. If we move to 4, we see that at this point there is the addition of two signals of equal amplitudes and phase difference of 180°, and that therefore cancel each other out. In practice, imperfections will always result in the existence of a small residual signal, which will be absorbed at a matched low-power load. On the other hand, in the adder function with entries at 2 and 3, half of the power will be dissipated in the load placed at 4, and this will have to be consequently dimensioned. We also see, if we progress along the left and the right tracks, that the electrical length difference is always of $\lambda/2$. As in all other types of couplers, it is implied that the four ports are matched to $Z_0$; 1 therefore sees two loads in parallel whose ensemble is equivalent to an impedance of $Z_0/2$ to which it is linked by two lines in parallel, each one with a characteristic impedance $Z$. For the impedance transformation to be correctly done, it is necessary that the two quarter-waves (to the nearest $\lambda/2$ ) present a global impedance $Z/2$ such that $Z_0 \times Z_0/2 = Z^2/4$, or further still, $2Z_0^2 = Z^2$, which gives:

- for $Z_0 = 50 \, \Omega$ \quad $Z = 70.7 \, \Omega$;
- for $Z_0 = 75 \, \Omega$ \quad $Z = 106 \, \Omega$.

We find without much problem, in the specialized catalogs, coaxial cables having these characteristic impedances, in semi-rigid or flexible braid versions, and which are therefore intended for this particular application.

Remark: even though this type of coupler is used in the same way as others, its functioning is fundamentally different from that of a radiating coupler. As we have already specified before, we are dealing here uniquely with phase compositions, without the notion of coupling intervening in the sense of the proximity phenomenon that we studied until now. In particular, the terms over- and undercoupling no longer more sense, and there result forms of response that, even if they lead in the end to the same category of applications, they are not at all of the same nature. Figure 9.9 will help us to make the distinction: we have represented the response curves of two models used in signal adders, and we have traced a line at just over 3 dB attenuation to take into consideration resistive losses. It is on this line that the response curves of a perfectly balanced coupler should be found. In the case of the hybrid ring, a sole point is found at the same time on the two curves, and we distance ourselves from the dotted line as we move away from $F_0$. In the case of the coupler, which was regulated with a slight overcoupling, the curves cross twice at half-distance between $F_1$ and $F_0$ and between $F_0$ and $F_2$, what allows us to optimize
the passing band as a function of the loss criteria defined in the specifications that were fixed at the start.

![Figure 9.9.](image)

Certain manufacturers seem to like to have this coupler in their range of products, which is certainly original but ultimately does not present any advantages in relation to other models, except if the fact it is narrow-band contributes to a better selectivity, which is an appreciable asset in interband couplings.

9.5. The wireline

![Figure 9.10.](image)

It is a special cable, intended for high-frequency applications, of the family of shielded pairs, where we have paid particularly attention to the material and the dimensions of the insulators so that the ensemble presents a given characteristic impedance, in this case that of the 50 or 75 Ω standards, with a well-determined coupling coefficient (Figure 9.10). To manufacture a hybrid coupler with this cable, it is enough to cut out a section of length \( \lambda/4 \) and to install it on a printed circuit or in a connectorized housing. As with all other types of housed couplers, it will be
necessary simply to compensate the inductance of the internal links with the connectors by little grounded capacitors, which we will do favorably by cutting them in printed circuit made of Teflon glass because of the modesty of their value (fractions of pF). If we use the flexible version of the cable, it is very easy to roll it up to store an appreciable length in a housing of the type shown in Figure 9.10. We could then use the procedure in low VHF while keeping dimensions of interest.

The original Wireline, a registered trademark, whose precedence earns it the right to use a capital letter, owed its coupling and adaptation characteristics to an asymmetric structure, with a supplementary insulator (kapton) around one of the conductors. This is not mandatory, and manufacturers such as Alcatel-Câbles know how to make symmetric shielded pairs having the same characteristics – it is nevertheless a special product that does not appear in the catalog. In fact, since it is very hard to produce with sufficient accuracy the parallelism of the two internal conductors, we calculate the thickness of the insulators in such a way that they are in contact and we twist them (twisted pair), a procedure that simplifies the manufacturing a lot. The insulator marked $\varepsilon_{r1}$ on the drawing is often Teflon in the form of very thin bandages rolled up one over the other to form a multilayered ensemble. It is necessary to take care, at the moment of the welding, to choose the correct conductor, corresponding to the ohmic continuity between the entry and the exit of each of the two lines, and that we will identify with the aid of an ohmmeter.

9.6. The “groundless” coupler

![Diagram of a groundless coupler]

Figure 9.11.

The appearance, in the field of high frequencies of electromagnetic simulation programs has completely changed things and habits concerning free laboratory studies. In particular, they allow us to produce in record time, and in general with a very good accordance with reality, virtual experiments whose number and duration were previously discouraging for engineers. Thanks to them, it is now possible to extend the field of investigations to the infinite and to test, with an approximation
that is more and more convincing, new ideas that would previously have had to pass by necessity via the stage of a real-scale model. It was in applying this new way of proceeding that the groundless coupler was born.

Originally, starting from the working of the hybrid coupler described above in Figure 9.2b, and that we have called a layer coupler, the question was posed regarding what would happen if we removed the ground, and in particular if we could make up for the alteration in impedance that would probably result from it, by an increase in the width of the microstrip lines in opposition. Now, it turns out that this does not only work, but that we can push the experience further, to distance the exterior ground planes, which constitute the electrical limits of the product, from the central circuit to such an extent that they become useless to it after a certain distance (Figure 9.11). This distance is of the order of twice the width of the coupled tracks. The hybrid coupler is then reduced physically to a double-face printed circuit with two exactly opposed tracks (paired strips), without the need for ground planes on either side, hence the name we give to it. We can now put it in a housing on the condition that the walls are sufficiently distant from the central substrate so that they no longer have any influence on its characteristic values. With the thicknesses of the usual circuits of 0.8 or 1.6 mm, this leads to envelopes that have a dimension slightly greater than that of an N socket. We have here a hybrid coupler that is amongst the simplest to manufacture, of excellent reproducibility, of a bandwidth of the order of an octave, and a thickness that does not exceed, with an N-type connection, the diameter of male plugs that will be fixed above, of small overall dimensions since we can draw the tracks in zigzag, and being able to support fairly significant crossing powers according to the adopted substrates. In addition, in return for the precautions indicated above, with this concept we can make – receiver splitter-dividers integrated into a common printed circuit comprising the amplification and the monitoring, the whole in a drawer measuring an inch in height (2.54 cm).

The name “groundless” makes reference to the ground planes external to the central printed circuit, but it can comprise a ground plane of its own over the same substrate as the tracks, which will allow the compensating capacitances to be welded at the connectors, as well as possible conduction points towards the housing: these are details which are revealed at the time of the set-up of the scale model, and there is no theoretical obligation regarding the presence or not of this ground. If, on the other hand, we decide it is of use, it will be always beneficial to ensure the continuity by connecting the two faces by regularly spaced metallic through-holes. The ground planes will moreover have to be found at a certain distance (of the order of 2 mm) from the central tracks so as not to interfere in their impedance.
9.7. The “catnose” coupler

This coupler, as the previous one, was developed by the radio department of “Etablissements Normand” in France. Conceived originally to be integrated into a antenna monitoring, within the framework of the “Acropol” system (450 MHz), the fixed objective of its study had the aim of producing an economic bidirective coupler, presenting a very good directivity and capable of conveying at least eight 50 W carriers, all with the most reduced dimensions possible. The main line is a silver-plated annealed copper band with two right angles, welded on the central conductor of the N bases and kept fixed in relation to the housing by two anti-rotating Teflon slugs that prevent any rotation during the screwing of the plugs onto the fixed connectors. The width is of 1 cm for a distance to the wall of 2 mm, which implies a characteristic impedance of 50 Ω. The coupled line is a quarter-wave drawn over a double-face epoxy of thickness 16/10, welded over the pins of two SMA fixed connectors with an adjustable dimension that gives the desired coupling and that we tune to a certain value, ordinarily comprised between 20 and 30 dB. Usually, we do not use epoxy glass in UHF, but in this particular case the ohmic losses have no importance: they simply come into consideration in the adjustment of the magnitude of the coupling, without influencing the directivity, which depends only on the matching. It is precisely with regard to this aim that the specific feature becomes involved, which has given its name to the coupler and consists of two straps welded to the ends of the secondary line as indicated in Figure 9.12; technicians quickly named these “whiskers”, leading to their name being given.
The case is closed by a welded lid presenting three tuning holes. The central one allows the main line to be deformed by distancing it from or approaching it slightly to the wall in order to fine-tune its matching. The two others allow, the lid being fixed, the whiskers to be deformed in order to optimize the directivity. The modeling of these attributes is so delicate that for the moment it escapes simulation programs, and we will simply say that they present the impedance needed to lead that of the coupled line to the right value, that is to say, that which will imply maximum directivity. It is better to admit that this procedure was found by empirical means, but also by following a certain logic regarding the chosen location to install them. Regarding the dimensions, the system reveals itself to be very flexible, and the length of the straps, comprised between 1 and 2 cm, seems not to be critical. The tuning is performed, evidently, with the aid of a network analyzer and consists of alternately passing from one strap to the other, deforming them slightly each time, and controlling on-screen the decrease of the trace of isolation until it is located entirely below the maximum level specified. This will depend on the working band, and we can take as reference the following values:

- BW = 1 octave: directivity > 30dB,
- BW = 1/2 octave: directivity > 40 dB,
- BW = F0/10: directivity > 50 dB.

If we add to this the fact that the dimensions of the main line are such that the losses are negligible, we see that the performance of this coupler is quite exceptional, for the homogeneity of its ensemble as well as for its very level. Nevertheless, for some it oversteps certain strongly-held principles in the domain of high frequencies, in particular that which recommends, especially in the case of directive couplers, taking care that the propagation velocities are indeed identical from one line to the next: here, with the coupled line on epoxy glass, the velocity of the waves is two times less than in the main air line, but despite that we find no problems in the exploitation of the product, even regarding intermodulations.

9.8. Discrete-elements coupler

The hybrid ring described in 9.4 becomes an increasingly dominant creation as we decrease in frequency. At 70 MHz, for example, with a semi-rigid Teflon insulator (c = 0.69 c0), the quarter-wave has a length of 1.55 m, which, even rolled up, starts to become cumbersome, heavy and expensive. It is then in our interests to pass to an electrically-equivalent discrete-elements production, consisting of replacing a section of the line by a self and two capacitances disposed in π. We can now build an economic hybrid coupler as drawn in Figure 9.13, insertable on a printed circuit but that can equally well be put in a connectorized housing. The model
is cascadable, the right-hand side of the illustration represents a 4-way adder/divider composed of three 2-way Wilkinson-grouped elementary cells. For an adder in the band 72-79 MHz, the selfs are 6-turn reels made of 60/100 silver-plated strap formed on a 5 mm chuck, and the capacitance C is of 33 pF. The desired value 3C will thus be 100 pF. This device has been well implemented with the following performance:

- channel/channel isolation > 21 dB,
- adaptation 18 to 22 dB,
- losses < 6.6 dB,
- crossing power: 4 carriers at 120 W,

- $R_0 = 50 \Omega$. The resistances of 100 \( \Omega \), which in each cell must dissipate half the injected power, were offset over a 19” front surface 8 units high (one unit = 1.75") and fitted with a common radiator.

This procedure is of particular interest in low VHF, but can also be used with some benefit in high VHF (up to 300 MHz) with 3-turn selfs, or even in low UHF (up to 500 MHz) with single-turn selfs. Beyond that, the suitable self would be a short strap whose inductance magnitude would be of the same order as the fixing elements, which is not possible and so it will be necessary to move on to other models with distributed constants.
9.9. Numerical data

We find in the specialized works a great number of formulas which give the geometrical parameters of lines or components as a function of their impedance. They allow us to have a response to all conceivable configurations, but are sometimes of an unjustifiable complexity in relation to the services they can provide, all the more so as their predictive efficiency is not always certain. We are not obliged, for example, to make use of the Bessel functions in the calculation of reels: that borders on the ridiculous, but it is done. In fact, when we are determining the dimensions or geometrical characteristics of lines as part of a pre-study, a few basic data are largely enough for a rapid approach, and one must not forget that adjustments (always empirical) are part of all prototype studies and, generally speaking, of the work itself. It is above all necessary, once we have a few reference points, to know the laws of similarity in order to be able to deduce the variations of parameters implied by the change of one amongst them.

\[ 1 \quad \quad \quad \quad 2 \]
\[ 3 \quad \quad \quad \quad 4 \]
Main line

Coupled line

Figure 9.14.

Figure 9.14 will serve as a support, by means of a schematization that is the simplest possible, to specify or recall certain definitions and conventions relating to couplers. Despite the theoretical symmetry we will distinguish the central line, that we will choose as such, and the coupled line, even for 3 dB hybrids.

The measurements made with the analyzer are of two types, if we refer to the S parameters:
- \( S_{xx} \) = access impedance \( x \),
- \( S_{yx} \) = gain of \( x \) towards \( y \).

They are the following:
- entry 1, exit 2, loads in 3 and 4: impedance 1 (\( S_{11} \)) and main channel loss (\( S_{21} \)),

\[ \]
– entry 2, exit 1, loads in 3 and 4: impedance 2 ($S_{22}$) and main channel inverse loss ($S_{12}$) = ($S_{21}$),
– entry 3, exit 4, loads in 1 and 2: impedance 3 ($S_{33}$) and secondary channel loss ($S_{43}$),
– entry 4, exit 3, loads in 1 and 2: impedance 4 ($S_{44}$) and secondary channel inverse loss ($S_{34}$) = ($S_{43}$),
– entry 1, exit 3, loads in 2 and 4: coupling $k$ ($S_{31}$),
– entry 1, exit 4, loads in 2 and 3: insulation $i$ ($S_{41}$).

The directivity $D$ is deduced by subtraction of the measurements of $S_{31}$ and of $S_{41}:

$$D = S_{41} - S_{31}$$

[9.1]

It is of interest to be able to know the common impedance to give to both quarter-waves, as a function of the value of the coupling. Let us call this one $k = P_3/P_1$ and let us reason only on the main line. If we look at it alone, we can consider that the decrease of power registered in 2, whose real cause is the absorption of power by the coupled channel, can also be seen as the effect of a terminal impedance $Z_t$ different from $Z_0$ and which will thus give rise to a reflected power equal to $P_3$, precisely with the essential difference that there is no reflected power. It is thus not a case of applying the law of reflections, but that of circuits, with the conserved relationship $P_1 = P_2 + P_3$. The impedance $Z$ that we are looking for will then be that of an matching quarter-wave and will be related to $Z_0$ and to $Z_t$ by the relationship $Z^2 = Z_0Z_t$ (Figure 9.15). It remains therefore to evaluate $Z_t$, which is done in a simple manner and without calculating, remarking that at constant tension the impedance is inversely proportional to the power, what we will write $Z_t/Z_0 = P_1/P_2$, that, is $Z_t = Z_0/(1-k)$, from which we ultimately have:
We deduce from this the following numerical values.

<table>
<thead>
<tr>
<th>k = P_3/P_1</th>
<th>( \frac{1}{\sqrt{1-k}} )</th>
<th>coupling (- dB)</th>
<th>( Z_\Omega (Z_0=50,\Omega) )</th>
<th>( Z_\Omega (Z_0=75,\Omega) )</th>
</tr>
</thead>
<tbody>
<tr>
<td>1/2</td>
<td>1.414</td>
<td>3.01</td>
<td>70.71</td>
<td>106.06</td>
</tr>
<tr>
<td>1/3</td>
<td>1.225</td>
<td>4.77</td>
<td>61.24</td>
<td>91.86</td>
</tr>
<tr>
<td>1/4</td>
<td>1.155</td>
<td>6.02</td>
<td>57.74</td>
<td>86.60</td>
</tr>
<tr>
<td>1/5</td>
<td>1.118</td>
<td>6.99</td>
<td>55.90</td>
<td>83.85</td>
</tr>
<tr>
<td>1/6</td>
<td>1.095</td>
<td>7.78</td>
<td>54.77</td>
<td>82.16</td>
</tr>
<tr>
<td>1/7</td>
<td>1.080</td>
<td>8.45</td>
<td>54.00</td>
<td>81.01</td>
</tr>
<tr>
<td>1/8</td>
<td>1.069</td>
<td>9.03</td>
<td>53.45</td>
<td>80.18</td>
</tr>
<tr>
<td>1/9</td>
<td>1.060</td>
<td>9.54</td>
<td>53.03</td>
<td>79.55</td>
</tr>
<tr>
<td>1/10</td>
<td>1.054</td>
<td>10.00</td>
<td>52.70</td>
<td>79.06</td>
</tr>
<tr>
<td>1/100</td>
<td>1.005</td>
<td>20.00</td>
<td>50.25</td>
<td>75.38</td>
</tr>
<tr>
<td>1/1000</td>
<td>1.000</td>
<td>30.00</td>
<td>50.03</td>
<td>75.04</td>
</tr>
</tbody>
</table>

Table 9.1.

If we calculate the adaptation corresponding to \( k = 1/6 \), that is to say, of a line of 54.77 \( \Omega \), we find 26.8 dB. This will also be the value of the adaptation if we use a line of 50 \( \Omega \) in a 6 dB coupler. It is safe to say that already for this value of coupling, and all the more so for the lowest values, there is no point calculating a special line, and a characteristic impedance equal to \( Z_0 \) could suffice, avoiding complications in the case of microstrip circuits.

We will find in the works dedicated to planar lines, of which the most important are mentioned in the bibliography, an extraordinary quantity of formulas that are supposed to allow us to calculate the dimensions of lines, components and discontinuities covering all possible configurations. Some are very complicated, and additionally a non-negligible number amongst them are not at all verified in the
measurements and are, as such, questionable with regard to their well-foundedness. However, we must recognize that it is a considerable work, the fruit of long years of specialization, that solidifies the efforts and the competence of their authors. However, the day-to-day exercising of the profession of HF engineer does not require permanent recourse to these complicated and exhaustive sources, sometimes redundant. On the other hand, to possess a limited number of simple references, capable of immediately providing, without calculations, precise orders of magnitude, is particularly appreciable. We can also add that only the relatively simple formulas are deduced, and that the more complicated they are, the more they contain empirical terms of statistical origin, hence their less rigorous foundations.

<table>
<thead>
<tr>
<th>line</th>
<th>dielectric</th>
<th>w/h</th>
<th>w (mm)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>air</td>
<td># 5</td>
<td></td>
</tr>
<tr>
<td>2</td>
<td>Epoxy glass</td>
<td># 1.9</td>
<td>3.1 mm (h = 16/10)</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td>1.6 mm (h = 8/10)</td>
</tr>
<tr>
<td>3</td>
<td>Teflon glass</td>
<td># 3.1</td>
<td>5 mm (h = 16/10)</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td>2.4 mm (h = 8/10)</td>
</tr>
<tr>
<td>4</td>
<td>Epoxy glass</td>
<td># 3.0</td>
<td>4.9 mm (h = 16/10)</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td>2.4 mm (h = 8/10)</td>
</tr>
<tr>
<td>5</td>
<td>Teflon glass</td>
<td># 4.8</td>
<td>7.8 mm (h = 16/10)</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td>3.8 mm (h = 8/10)</td>
</tr>
</tbody>
</table>

Table 9.2.
It was in this frame of mind that Table 9.2 below was composed, concerning the planar lines defined, for a given dielectric, by the ratio $w/h$ (for width and height), and where some practical values are given in tabulated form, from which we can deduce others. It is still necessary to have a good understanding of the rules to be applied when we make a parameter vary.

The five structures shown above, with, for the last four, standard printed circuit thicknesses of 8 and 16/10 of a mm, represent more than 90% of the cases found in the daily activities of the radio industry. This table will thus constitute a practical basis, and that is its aim, for drawing as a first project the planar circuits in question. For the other cases, we will refer as usually to the reference works. We now have to adjust these numerical data in order to conserve the impedance when we need to change one of the three basic parameters: the relative permittivity $\varepsilon_r$, the thickness $h$ of the dielectric and the width $w$ of the strip. Now, if we want to keep the same impedance $Z = \sqrt{L/C}$, or in other words, the same ratio between the capacitance and the linear self, we notice that it is impossible to modify only one parameter, without also modifying the two others. Changing $\varepsilon_r$ results in changing only the capacitance, but not the self. The line impedance is thus modified, and to reestablish it and keep the ratio $L/C$ constant, it will be necessary to change both $w$ and $h$. If we change the width of the track $w$, increasing it, for example, the capacitance increases, but also, simultaneously, the self decreases: the impedance will not be able to be retrieved unless we change $\varepsilon_r$ and $h$ at the same time. If $h$ is modified, we find ourselves in the same situation as with $\varepsilon_r$. The homothety does not conserve the impedance, which could lead to the idea that the electromagnetic universe is not fractal and has a scale. Whilst this remark opens up metaphysical perspectives that are probably very interesting, it is not the business of the engineer, who will not be able to use analogous similarity laws, as we do, for example, in the Careen test-pools to deduce the dimensions of a ship from those of its x-scale model. In fact, specialists are used to considering the ratio $w/h$ as the variable on which $Z_0$ depends, and they give approximations that, because of this, do not have general scope. Wheeler, for example, proposes two series of curves corresponding to two different approximations, according to $w/h$ being large or small, and which do not intersect: at the middle, for $w/h$ around the unit, we have, according to the approximation, chosen two different values, which is embarrassing to say the least. We can nevertheless trace a family of curves giving $Z_0$ as a function of $w/h$ (Figure 9.16), which proves very practical and reliable from the moment it includes experimentally-verified reference points, but that are not valid except for $w/h > 1$, in a region where the relative variation of $C$ (linear) reflect very markedly on that of $L$ (logarithmic) and where we will allow a quasi-linearity of $Z_0$ as a function of $w/h$. Here again, simulation programs provide an alternative solution, of much greater interest, but also heavier. Table 9.3 shows a modeling on Sonnet Suite of the same kind of circuits, with different permittivities.
Figure 9.16.

<table>
<thead>
<tr>
<th>Substrate</th>
<th>$\varepsilon_r$</th>
<th>$\tan \delta$</th>
<th>No ground</th>
<th>Ground at 1 mm</th>
</tr>
</thead>
<tbody>
<tr>
<td>Teflon 8/10</td>
<td>2.2</td>
<td>0.002</td>
<td>2.5 mm</td>
<td>2.4 mm</td>
</tr>
<tr>
<td>Teflon 16/10</td>
<td>2.2</td>
<td>0.002</td>
<td>4.9 mm</td>
<td>4.3 mm</td>
</tr>
<tr>
<td>Epoxy 8/10</td>
<td>4.3</td>
<td>0.02</td>
<td>1.6 mm</td>
<td>1.5 mm</td>
</tr>
<tr>
<td>Epoxy 16/10</td>
<td>4.3</td>
<td>0.02</td>
<td>3.1 mm</td>
<td>2.7 mm</td>
</tr>
</tbody>
</table>

Table 9.3.
Important remark

When we develop a directive coupler, whatever the principle chosen, we are obliged to complete it with additional capacitances placed at the accesses. This necessary completion is often presented, as we have done in this book, as an obligation associated with the presence of connections that will regularly constitute a transition of inductive predominance and in that way a solution of continuity in the impedance. It is almost always true, and particularly evident in the case of fixed connectors, but this is not the only case: pre-studies with the aid of simulation programs, where no connections are present, except as limits of the drawing, themselves lead, independently of the reasons mentioned before, to capacitances systematically being added at the ends of quarter-waves in order to reach the optimal characteristics in band symmetry, adaptation and directivity.

Microstrip productions are particularly practical from this point of view, because the additional capacitances, evaluated in picofarads, are a little cumbersome and transparent in terms of capital cost. In other models, such as the wireline, we will use discrete components, in air couplers we will be able to add small columns placed as reinforcements between two carefully-chosen walls for which we will adjust the value by the diameter and the distance to the ends of the lines: there is always a solution.

9.10. Applications

There are two large categories of application of the directive coupler, directly related to multicoupling techniques: summing/splitting and the pairing of quadrupoles. It is also necessary to add a particular association of four hybrids that allow us to couple antennas on a single channel and that we often use in confined environments, such as, for example, the superstructure of a ship or an emission tower.

The first category was already mentioned and corresponds to the actual function of the device: we can either group onto a single channel two independent signals, or, conversely, split a given signal into two identical signals with similar phases, each with half the power of the original signal. Moreover, it provides the adaptation of all the accesses and an insulation between the channels. The system is in principle infinitely cascadable (but beware the peak voltage!) and allows successive additions or divisions in powers of 2: 1, 2, 4, 8, etc. The set-up is in parallel or Wilkinson. But there is also the possibility of a set-up in series that makes it possible, by constituting a chain of asymmetric couplers, to achieve any order.
The second category demands the simultaneous usage of two couplers and allows us to give particular properties to a pair of identical quadrupoles. The general diagram is that of Figure 9.17. When the quadrupoles $Q_1$ and $Q_2$ are amplifiers, we have a set-up called Kurokawa, already mentioned in the section on low-noise amps (section 8.5.1). When $Q_1$ and $Q_2$ are bandpass filters, the ensemble constitutes what we call a directive filter and becomes the basic element of the emission multicoupler of the “series” type, in contrast with the parallel model said to be “star-shaped”. In this configuration, the relative phase of the two channels is extremely important: if we fix the entry in 1, it must be zero between ports 3 and 4. We often choose the layout of Figure 9.17, which allows us to compact the system into one single-bloc product, with the only condition that the in-out phase difference be the same from 1 to 3 and from 1 to 4. We can choose it to be equal to zero, which reduces the dimensions of the whole, but we can also produce other variants, with couplers at 180° instead of 90°, or with long connections, as long as the phase condition is respected. In multicoupling applications, the entry of channel 1 is qualified as “narrow band” because it is filtered by the band-pass filters, accesses 3 and 4 are qualified as “wide band” as they are, in one sense or another, the first being the entry of signals coming from a certain number of identical filters located upstream and the second the exit towards other quadrupoles in line. The functioning is detailed later on in Chapter 11.
Chapter 10

Helical Resonators

10.1. Introduction

The radio-communications industry, under an impressive pace of development, is at the same time the creator and the requestor of new technologies. Now, those discussed in this work, besides the fact that they are not new, are not taught in their technological details in traditional courses, and the acquisition of this knowledge in the profession demands such a considerable time that little is left for research. Consequently, this is not a field where we expect to see shattering discoveries, to the point that we often have the feeling that everything that was to be found has already been so. This is an additional reason to try to become familiar with it, if not completely, then at least in the fundamental aspects of the tools it makes available to us.

Straight quarter- or half-wave resonators constitute the basis of high-performance filtering products. Their geometrical simplicity leads to simplicity in the modeling that has allowed an exhaustive theoretical study and made it a reference in the study of resonant TEM-mode systems. We can say the same thing about the planar circuits or microstrips that are added to them, without creating competition, but that benefit with respect to them from particular perspectives essentially related to the discovery and development of new substrates. Besides these two essential families, there exist other less conventional objects that could nevertheless prove to be of great utility, such as for example helical resonators, and it is indispensable to know them.
10.2. Functioning

When we go down in frequency, the size of the quarter-wave cavities progressively grows until it reaches dimensions that end up posing problems of transport and accommodation. We generally consider that the low VHF (30 to 100 MHz) is the frequency zone where the question is usually posed as to whether one should continue to use distributed constants or move to localized constants, that is to say, to use discrete elements. These most often constitute a radical solution to problems of size, but at the expense of a quality that we see to be excessively lowered in relation to linear resonators. Helical resonators offer a transitory solution, widely used between 50 and 150 MHz, where their quality/dimensions ratio is unequalled, and even at higher frequencies, towards 500 MHz, for solving drastic dimension problems.

It is a technology that very much suits the production of duplexers of the pass-reject type, for which the passing loss depends above all on the quality of the common line constituted, let us recall, by the series quarter-waves that connect the resonators, whereas the rejection is almost not affected by the reduction of the Q₀ of the resonators and remains of the order of 30 dB per cell. In relation to the model with prismatic resonators, the visible quality difference essentially concerns the selectivity, which is a general rule already mentioned in the case of coaxial models. Besides that, the basic principles remain the same and we are happy to simply replace the tubular resonator of a cavity by a solenoid. The exposition is simple, the functioning and the production much less so: it is not at all evident, a priori but above all in relation to the usual point of view and the methods used for the study of coaxial resonators, that this geometrical form could be suitable. First of all, it was necessary to wonder in which mode the waves could propagate in such a structure, so distant from the conventional criteria.

From this particular point of view of the propagation mode, it is difficult to associate that of the helical resonator with anything known. It is not a TEM mode if we consider the ensemble of the helix and its envelope, but this could at a pinch be associated with it if we consider the helix alone, momentarily ignoring the presence of the external conductor, because now the propagation can be compared to that of a wave guided by a linear conductor. But, even if the external envelope plays a less important role here than in the case of the coaxial cavity, it is not at all negligible: when we remove this envelope, the resonance collapses. We use in certain works, such as the Zverev, the term “quasi-TEM” mode, which is not entirely clear, but we will neither contest nor approve it, in order to devote ourselves, as we have done up to now, to the practical problems of producing the system. For those who nevertheless want to increase their knowledge on the subject, we recommend following the research paths indicated in H. Aberdam, l’Aide-mémoire d’électronique et radio-électronique, Dunod, p. 142, and more particularly the
article: “Surface waves and their application to transmission lines”, *Journal of Applied Physics*, Volume 21, November 1950, p. 1119-1128, Georg Goubau. There are also several cross-linked Internet sites that allow more in-depth investigation to be performed and the article in question to be reached: do an ISP search on Georg Goubau, also see Sommerfeld, Fromageot, Louis, Zenneck. All these specialists in high frequencies studied the same phenomenon, consisting of stretching a conductive line (Goubau line) between an emitter and a HF receptor, directly connecting it to the core wire of the access cables, their external conductor being linked to the propagation medium by means of two cones, one at the departure and the other at the arrival; to briefly sum it up, as it is of great interest. We thus note a guide along the line, with a transmission loss which is less than that of a coaxial having the same central conductor, provided there are no poor conductors nearby.

The criteria which will determine the quality of the resonator are the same as for coaxial models, with an important difference though: this time, the participation of the envelope in the resistive losses is much smaller, to the extent that we will not take it into account in the evaluation of the quality coefficient. This is confirmed by the fact that we can use a raw surface of non-silver-plated aluminum as an external conductor, without noticing any appreciable decrease in the selectivity or any other value representative of the quality of the application in progress. It is also a supplementary connection with the specific features of the Goubau line, which shows lower losses than other lines of guided propagation because they are localized in the central conductor, the closest external conductor not intervening because its limits are pushed away to the position of the nearest conductor, and therefore infinitely if there is no conductor nearby.

**10.3. Structures**

Figure 10.1 shows the two most common aspects of helical resonators: in (a) the helix self-fixed thanks to a thread welded to the lower part, and in (b) a helix wound over a Rexolite (or any other good insulator) chuck, with both a low $\delta$ (0.00066 at 10 GHz for the Rexolite 1422) and a good mechanical quality, that is, allowing at the same time a not overly delicate manufacturing, solid fastening and acceptable behavior in temperature. It is clear that, the insulating chuck never being perfect, it will only degrade the quality coefficient of the helix. We will thus use the layout in Figure 10.1a, every time it is possible, that is to say, whenever the total permitted dimensions allow us to arrive at a freestanding geometry of sufficient robustness. In a practical way, this criterion comes true by an absence of vibrations and deformations due to impacts, which will be controlled by the durability of the response curve after these tests.
We can then, according to their apparent shape, separate the helices into two categories: the short, self-fixed and the long, which need a support to acquire the necessary rigidity. Let us first look into the first type. The geometric parameters are summarized in Figure 10.2. They are theoretically independent, but practice has
imposed justified habits. First of all, the width or the diameter of the conductive interior where the helix is found will taken to be about the double of the average diameter of the turns. Zverev chose a mode of calculation of $Q_0$ which gives a decisive importance to this dimension, from which we cannot distance ourselves too much, under the risk of seeing the quality factor decrease rapidly. The built-up length of the solenoid part of the helix, excluding the part that serves for the fastening, is equal to a quarter-wave:

$$n \pi D = \frac{\lambda}{4}$$  \hspace{1cm} (10.1)  

We notice that there are no end corrections here as for coaxial cavities. Afterwards comes pitch $p$. It is a free choice, \textit{a priori}. However, the main interest of the helical resonator being the space gain, we will of course take it as small as possible. The relationship $p = 2d$ is often chosen, and it offers a practical advantage as well, which is to allow us to produce at the same time and on the same chuck two tubes kept side-by-side, whose common diameter assures the correct spacing between the turns. We thus make two helices at the same time, but this is a manual procedure that is suitable only for making prototypes, using an annealed copper tube presenting the necessary pliability. This will not be suitable for a production in series. For this, we use a solid brass rod and we turn to spring manufacturers, whose computationally controlled bending machines are perfectly adapted to the conformation of helices as required by HF specialists.

There are, however, restrictive conditions for the material: given the power of the machines, made for piano cords or solid circular profiles made of sufficiently hard material, the slim tube is excluded because readily broken. Now, it is regrettable to be obliged, uniquely for mechanical reasons and all the more so if the diameter is large, to use solid material when only the exterior surface is of interest to us. A helix made from a tube of 0.5 mm thickness and 5 mm diameter, made of brass (density 8.4) weighs 2.78 times less than in a solid stem. If we take the example of a band-pass FM filter (78.5-108 MHz), requiring 10 helices of deployed length 0.694 m each, we gain 733g on a product that initially weighs little more than 2 kg: it is a lot. It can therefore be beneficial to construct special machinery to reel the tubes intended to make helical resonators in series.

Those who possess this machinery are also subject to an additional constraint, coming from the fact that the brass or copper tube of small diameter is generally supplied in cold-drawn sections, which are impossible to bend as they are. It is necessary to anneal them at 600°C and let them cool slowly before working on them. But the result is able to rise to the difficulties.
For the second category of helices, those whose rod diameter and length make it indispensable to use a permanent chuck, all the difficulty resides practically in the conception of the latter. The material is essential, because the electrical and mechanical qualities of the resonator will depend on it. We will choose an insulator with a guaranteed tgδ and that is the weakest possible, a softening temperature that is the highest possible and a good mechanical memory: Teflon, for example, allows functioning at relatively high temperatures, but it is not advisable if we have to take it to more than 180°C during one of the production phases, because it would keep any shape acquired during softening. Rexolite is commonly used below 90°C, despite a delicate fabrication. The only valid mechanical solution to use long helices at high temperatures is not to use a chuck but to support the coil by welding it in places on contact-points made of HF ceramics, regularly distributed and fixed to the envelope. It is thus necessary to design the cells in such a way that we can make the welds once the helix is installed in its housing. The German constructor Kathrein solved in a particularly clever way all the practical problems regarding this kind of product all at once by editing a special profile that we saw to manufacture completely equipped cells that are afterwards juxtaposed alternately with added-on thin walls, the whole being kept together by external constraining rods to form a compact ensemble that is perfectly modular. Moreover, the dimensions were chosen so that we could equally house the helices on high or low VHF and tubes for medium and low UHF. This does not condemn chuck helices which, retained at their base by a fixing screw and at the top by the tuning screw (Figure 10.3), offer a
configuration that assures a very good stability against vibrations and allowed us, for example, to perform mobile duplexers for the vehicles of the French national police.

The procedures presented above cover most of the conventional practical applications of helical resonators, but not all of them. The intrinsic form of the helix constitutes an unbalanced object that necessarily needs at least one supplementary fixing point in addition to the foot, given that the inertia of its mass can lead to a possible deformation following a shock. We realize, when we consult the catalogs of filter manufacturers, that the helices are very little employed, not to say not employed at all, in the making of bandpass power filters in low VHF (under 80 MHz) and in HF. There are at least two reasons for this: firstly there is not a large market to support its production, and secondly, the technology is complicated because it demands specific tools difficult to pay off financially. We can also add that the absence of simple modeling of coaxial resonators has not encouraged the theoretical developments, and the domain remains somewhat mysterious for the cautious manufacturer. Despite all that, it is perfectly possible to use procedures of artisanal manufacturing, such as manually making the helices from plumbing tubes, with the aid of simple conforming equipment: the profession is use to such procedures and is not ashamed of them.

10.4. Tapping and coupling

In relation to the tubular resonators, the helices radiate less and work in a different way that does not allow us to conceive entry/exit couplings and inter-resonator couplings in the same way, with the exception of the pass-reject function where everything remains the same. There is no place, in the space between the helix and its envelope, where the field is sufficiently strong for us to use loops or probes. We proceed therefore, for the link with the connectors, by tapping, that is to say, passing via a strap welded directly between the core of the connector and a point of the helix situated somewhere on the lower turns, near to the earth, to a dimension we usually define by a fraction of a turn with an accuracy going up to 1/16th of a turn and which is determined experimentally. We generally do not exceed the lower 1/6th of the deployed length of the helix, which corresponds roughly to the third turn for a total of 15 to 20 turns. In low-pass filters, the length of the strap is not important. It becomes so in the case of pass-reject duplexers where the nature and magnitude of the impedance determine the function in the same way as with tubular resonators: an inductive tap gives place to a low-pass cell, a capacitive tap to a high-pass cell (see section 4.1.3), and their value determines the duplex gap. The tuning procedure is identical.

Concerning the bandpass function, the helix combline may be produced just as well as with tubes. If we prefer a compact structure with profiles, the couplings
between resonators are generally made by proximity, with windows added into the
common walls and tuning performed by screws, exactly as we would do for tubes. It
is necessary only to recall that the effect of the screw is inverted: when it penetrates
further into the window, the coupling decreases. Moreover, the left-hand slope of the
response curve will be more abrupt than the right-hand-side one, which is the
inverse of what happens with prismatic resonators.

In the work of A.I. Zverev cited in the bibliography, we can read the following
on page 508: “coupling between helical resonators is considered as the most arduous
problem in carrying out a filter project. This problem finds its origin in the difficulty
we have in proceeding with its mathematical analysis”. We notice, in the view taken
by the author, a particular aspect, a supplementary illustration of the opposition that
can exist, in the practical domain, between theoretical physics and rational physics.
While it is true that the modeling of the coupling is considered extremely
complicated in general, even impossible in the case of helices, it is equally true that
fine-tuning a filter is always the same, whatever the type of resonators used. From
the moment mathematics, instead of helping the engineer, confuses things by means
of an overly systematic attempt at modeling, it is better to refer to notions that are
less precise, but more physical and easier to grasp, and above all simpler, such as
considering that the coupling is an electromagnetic interaction associated with the
proximity and whose magnitude is in an inverse function to the separation. This
definition does not allow us to calculate the dimensions of a prototype, but it has the
merit of setting out the limits of our real knowledge and is enough for the
technicians, in their laboratory, to be able, by means of successive experimental
approximations, to determine the correct distances between tubes or helices,
indifferently. We can moreover add to what Zverev hastens to give, following his
remark, a set-up of experimental devices permitting the indirect measurement of the
coupling between two helices and the tracing of a curve of variations, which is a
commendable and honest concession to the experimental method.

10.5. Quality coefficient

In order to calculate the quality coefficient $Q_0$ in vacuum, we cannot operate
exactly in the same way we did for coaxial cavities, given that we no longer have the
equivalent of a simple line section. We will proceed by applying the defining
relation $Q_0 = \frac{L}{\omega R}$, evaluating $L$ and $R$ separately according to a modeling that we
will choose amongst several possibilities we have, both for $R$ and $L$. Referring to
section 1.1.1, we have, for the value of the skin depth:

$$\delta = \sqrt{\frac{2}{\omega \mu \sigma}} = \sqrt{\frac{\rho}{\pi F \mu}}$$

$\rho$ resistivity

[10.2]
The helix resistance can now be calculated in a conventional way, like with direct current, supposing that the HF current is located in a peripheral ring of width \( \delta \). We thus write: 

\[
R = \rho \frac{l}{S}
\]

where \( l \) is the length of the helix, that is, \( \lambda/4 + D/2 \), supposing that the fixation point has a length of \( D/2 \), which is the usual case. On the other hand we will suppose that the ensemble of the resonator can be modeled by a hypothetical symmetric line whose helix constitutes the central conductor and envelope, by means of a wave path that we suppose also to be helical, the external conductor having the same global resistance that adds to the first. If \( R \) is the resistance of the helix itself, the quality coefficient in this hypothesis will be: 

\[
Q_0 = \frac{L}{\omega R}/2
\]

with a total effective resistance \( R' = 2R \) whose value will be given by the formula:

\[
R' = 2\rho \frac{\lambda}{4 + D/2} / \pi d \delta
\]

What remains is to evaluate the self \( L \). We find ourselves faced here with an impressive number of formulas varying in complexity, some even involving elliptical integrals of the first and second kinds (Wadell), and none of which, in fact, can claim to be valid for all cases. It is also important not to lose sight of what we want to find, that is to say, an order of magnitude only, an approximate estimate of the quality coefficient which allows us to compare the helical resonator to a linear one with the same envelope section, so that we may know what to expect if we choose this solution. In summary, we do not need any large precision. On the other hand, we are looking for an expression that is related only to the inductance itself, understood as the seat of the stocking of electromagnetic energy, without taking into account the corrections associated with parasitic capacitances which modify the impedance when the frequency increases. For this reason, and maybe paradoxically, we will use one of the simplest formulas, proposed in the short summary by the French publisher Dunod, cited in the bibliography, developed for low frequencies and valid for a ratio \( l/D > 0.3 \), a condition representing the general case:

\[
L (\text{Henry}) = \frac{10^{-6} n^2 D}{0.46 + l/D}
\]

1 and \( D \) in m.

Numerical application: we wish to make a band-pass filter at 75 MHz with the aid of resonators contained in a squared profile with 50 mm sides, and we are limited to 20 cm in height. What are the possible solutions?

At 75 MHz we have \( \lambda/4 = 1 \) m. A coaxial cavity of the “Rubis” type, of circular section of diameter 15 cm, will have, according to the program “Qo coaxial cavity” (see the chapter “Utilities”), a \( Q_0 \) of about 4600 (4662 exactly, with a silver-plated
The helix which will be in the 50 mm squared profile will have a diameter of 25 mm, a step of 10 mm and will therefore comprise 12.7 turns, with a length of 127 mm and a foot of 12.5 mm. Relationship [10.3] gives based on these numbers a self of $7.28 \times 10^{-7}$ Henry. The skin thickness at 75 MHz is $\delta = 7.3 \times 10^{-6}$ m, the effective resistance of the helical cavity will then be, according to relationship [10.3]:

$$R' = 2 \times 1.58 \times 10^{-8} \times 1.0125/ \left( \pi \times 5 \times 10^{-3} \times 7.3 \times 10^{-6} \right) = 0.279 \ \Omega$$

$$D' \text{ where } Q_0 = 7.28 \times 10^{-7} \times 2 \pi \times 75 \times 10^6 / 0.279 = 1230$$

It is necessary to compare this value to two others: firstly that which a coaxial cavity of the same external dimensions would have with a central tube of 50/3.6 # 18 mm, and then that of the same cavity when we have replaced the central tube by the helix completely straightened. The program to calculate $Q_0$ gives 1552 and 1205 respectively. These numbers are of really interest and deserve a comparative table and some comments.

<table>
<thead>
<tr>
<th>Type of resonator</th>
<th>Coaxial cavity 150/40 mm</th>
<th>Coaxial cavity 50/18 mm</th>
<th>Coaxial cavity 50/5 mm</th>
<th>Helix cavity 12.7 winds Φ5</th>
</tr>
</thead>
<tbody>
<tr>
<td>method of calculation</td>
<td>$Q_0$ program (27) see 3.3.8</td>
<td>$Q_0$ program (27) see 3.3.8</td>
<td>$Q_0$ program (27) see 3.3.8</td>
<td>$L_0/R'$</td>
</tr>
<tr>
<td>$Q_0$</td>
<td>4662</td>
<td>1552</td>
<td>1205</td>
<td>1230</td>
</tr>
<tr>
<td>$F_0$</td>
<td>75 MHz</td>
<td>75 MHz</td>
<td>75 MHz</td>
<td>75 MHz</td>
</tr>
</tbody>
</table>

**Table 10.1.**

The first evident point is the importance of the degradation of $Q_0$ when we go from 150 to 50 mm of external dimension, but this is nothing but confirmation and a supplementary example of what we have already learned about the coaxial cavities, in particular the strong dependence of electrical performance on its dimensions. What strikes us the most, and that which relates more to this chapter, is the little difference that exists between the results of the three columns to the right, that is to say, regarding the 50 mm shape, under three different ways of usage, the reference
value being the column relative to the ratio 50/18 (external/internal diameter), expected to lead to the maximum value of the quality coefficient. But most surprising is the remarkable concordance between the results of the two last columns, which come from completely different models. It would certainly be careless to draw immediate theoretical conclusions from this fact, but it is necessary to recognize that to find two results that are so close having followed two very different paths, not only gives us confidence as to the validity of the results, but also gives us elements to reflect on.

We cannot avoid, in particular, thinking about the guided propagation of Goubau and Sommerfeld, for which the devices we describe may be considered as an extension or an individual evolution. There is perhaps material for investigation in this, but for now we will focus on the fact that we can have a good estimation of the predicted $Q_0$ of a helical cavity by likening it to a coaxial cavity using a central conductor of the same diameter as that constituting the helix, and applying the ordinary method of calculation. We see then, and maybe this is the most important point in practice, that the quality coefficient of a helical cavity is very similar to that of an optimal cavity of the same section, for a considerable gain in size (ratio 7 to 8 in height).

The industrial usage of this technology is therefore of great interest given the compactness of the products to which it leads, and it is also widely applied in the fabrication of duplexers and VHF filters of moderate power. The arguments exposed above should incite designers to extend it to devices of average and strong power, for which the gains in space should be even more impressive.

10.6. Set-up rules

We have previously seen the different ways of constructing a given model of a helical resonator, as well as the greatest problems of its fixing and rigidity. Also mentioned was the similarity of organization, when we have to group them with coaxial resonators in a housing in order to make the filters. But there are also differences, related to the structure of the helix, that are noticeable in particular when, from necessity, we are lead to use taps. In effect, when we apply this kind of connection, such as for example between the end resonators and the housing connectors, the disappearance of symmetry planes that exist in the tube band-pass combline filters somewhat changes the set-up possibilities and introduces a supplementary degree of freedom.

It is normal, when we draw the plan of the ensemble of a helical filter or duplexer, to pose the question of how we will order their alignment. Is it necessary for them to be laid out in the same way, or should we, on the contrary, alternate their
fixation points, putting them in the direction of the length or perpendicular to it? etc. In fact, when we have enough space, when the construction is spaced out, we can lay them out as we want, taking into consideration only what is the easiest set-up to adopt. However, whenever both the filter and the duplexer decrease in size and the specifications become more difficult to fulfill, we realize that the details of the installation of the helices have an effect on the behavior of the whole.

We regularly notice an improvement, and notably a greater facility of tuning, when we adopt the layout of Figure 10.4 where we have highlighted a centered axis of vertical symmetry. The illustration is explicit enough that there is no need to add further comments. If the number of resonators is odd we will place the central one indiscriminately in one position or the other.

**Figure 10.4.**

10.7. Applications

Helical resonators constitute a technological transition between devices with distributed constants and discrete elements. The border is fairly blurred on both sides, and the choice depends on particular circumstances. But most of the time, it involves solving a problem of size, of limit dimensions, in exchange for a consequent but generally tolerable decrease of the quality coefficients. It is in the production of pass-reject duplexers in low VHF that there is an industrial market for helices, at a level of more than 90%. Nevertheless, applications at moderate or high power, in coupling racks working below 100 MHz, could be envisaged, but a certain reticence seems to exist on the part of designers to take this direction, and we do not really know why. This is a pathway we will try nevertheless to promote in the following chapters.
Chapter 11

Multicouplers

11.1. Transmitter multicouplers (TX)

11.1.1. Choice of the technology

We now have, after the product-by-product analysis and descriptions made in the preceding chapters, all the elements necessary to define and produce complete coupling systems. Therefore, the question now is to decide what type of technology will be the best suited to the project in question, and this is probably the most important point, by its definitive nature, since once the system is chosen, it is practically impossible to change it after production has started. It is thus absolutely necessary to have a precise view, not only of the functionalities that will be implemented, but of the foreseeable or possible evolutions of the system, this last point being often forgotten by beginners. In fact, no radio network is fixed, because in addition to its intrinsic demands and specifications, it is subject to an eminently variable environment, due, for one thing, to the provisional and constantly revised character of the frequency allocations: it is always necessary to expect modifications, and thus strive to conceive a system that is as flexible and modifiable as possible.

No system is perfect, none can enjoy all the advantages, and from this results a great diversity of possible technical solutions whose number does not necessarily facilitate the task of the author of the specifications. Because it is absolutely necessary to have one set of specifications: while the major ordering parties are familiar with these procedures and know its unavoidable character, there also exists a large number of occasional or ill-informed clients with whom it is better, when we are manufacturers or project managers, to spend the necessary time to elaborate
precise specifications, under the danger of finding ourselves later in a conflicting situation over overlooked or ill-considered points.

It is necessary therefore to come to a decision about the exclusive choice to be made between agility in frequency – an expression dedicated in the radio medium to designating the possibility of changing the frequencies of channels without retuning the coupling – and a system with low losses, given that we cannot have both at the same time. Agility unavoidably leads, in fact, to hybrid couplers set-up in cascade, whereas low loss implies filters or cavities. The number of channels to be coupled is also important as soon as it surpasses 2, but the frequency is also to be taken into consideration in the measurement, on which the dimensions of the objects used depend. In fact the parameters to take into account are so numerous that the best solution is to list the technologies by examining for each one the advantages and the disadvantages, as well as their limits.

### 11.1.2. Cascading hybrid TX multicouplers

The unitary hybrid coupler has two equivalent decoupled entries, a coupled exit and a load dimensioned to dissipate half of the sum of the incident powers. The current models allow agility over a band of an octave, which places them in the of “wide-band” coupling, contrary to the cavity couplings which we also call “narrow-band”. When we have to couple 3 or 4 channels, it is necessary to pass to three
couplers and we lose 3/4 of the power on each signal. We can see here the two main
disadvantages of the system: at each stage crossed, the power of each signal is
divided by 2, and there are one or more channels that are not used when the number
of entries is not a power of 2. Figure 11.1 shows the case of the 5-channel coupling
where we are forced to lose an additional 3 dB, be it with a single supplementary
channel. The advantage is that of creating a simple construction, without tuning,
made from a single product that is, in addition, relatively cheap and in general
available off the shelf.

We remark, and this is specifically related to the cascade set-up, that whatever
the number of stages the total effective power at exit is equal to the nominal power
per channel, supposing that all carriers have the same power, which is usually the
case in radiotelephony. We could conclude that we can thus couple as many
channels as we want, which one should be careful not to believe: apart from the fact
that we lose 3 dB at each stage, which ends up being penalizing at the antenna and
the radiated power, the evaluation of the peak power shows that the latter increases
proportionally to the number of channels (see following chapter on limiting power).
The growth of the latter thus leads us, at a given moment, to reach the breaking
tension of one of the constitutive elements. This question of maximum acceptable
power is of the greatest importance in a coupling project, and it is imperative to
always evaluate the peak power.

11.1.3. **TX multicoupler with degressive coupling**

Among the disadvantages of the preceding set-up, that related to the obligation
of using a structure governed by the power of 2 can be by-passed using special
couplers set-up in series, and conserving the wide-band characteristics of the
coupling. The couplers in question are no longer 3 dB hybrids, apart from a single
one situated at the head of the group. The others each have a coupling coefficient
that depends on its rank within the set-up. Figure 11.2 shows the set-up of n
directive couplers in series, where we can analyze the operating principles of the
system by assessing the power at each intermediary exit: each signal to be coupled is
marked by its voltage $V_1, \ldots, V_n$, with $V_1 = V_2 = \ldots = V_n$. At the exit of the first coupler,
of coefficient 1/2, we have two coupled and attenuated 3 dB signals. At the exit of
the second coupler of coefficient 1/3 the two preceding signals undergo an
attenuation of 1.77 dB, which makes in total 4.77 dB and therefore exactly the same
value as for $V_3$ which arrives on the coupled channel $C_2$. We thus have, at the exit of
the second coupler, three carriers of a same level, which will, all three, undergo an
attenuation of 1.23 dB when crossing the main channel of $C_3$, which has a coupling
coefficient of 1/4, which will take their global attenuation to 6 dB, whereas $V_4$ will
see its level reduced by 6 dB, which will put it in equality with the three others, etc.
We thus see that it is possible to produce in this way an optimized wide-band
coupling where there are no unused entries and where the attenuation is always minimal. The system, of great interest for strong powers since we lose only the strictly necessary, may be extended at will within the limits imposed by the breakdown voltage of the last element. It is necessary, however, to have special couplers, but there is no particular difficulty involved in their construction.

11.1.4. Cavity TX multicouplers

The essential principles were explored in Chapter 7, where you can find all the necessary tools necessary for determining the dimensions of a cavity as a function of the expected performance. Now we want to look into the practical details of its creation.

Let us firstly recall an essential principle: a coupling bay of cavities of the nodal-point type can only be optimized, that is, present the best characteristics we could expect, when the frequencies are fixed for all of them at once, in such a way that we can fine-tune the cavities definitively. This was the case at the beginning of mobile radiotelephony, when a coupling bay coming out of a workshop was transported to the site with a definitive frequency marking and harnesses built precisely and in a definitive way. From the moment we need flexibility in the usage which implies that this condition will not be fulfilled, there will necessarily be a degradation of the performances which will depend on the degree of flexibility demanded. This consequence of the properties of lines is always very difficult to make future users understand, even if it is perfectly evident to those who construct the systems based on impedance transformations by line sections. The commercial escalation and competition have generated an increase in the sophistication of technical characteristics of PMR (Private Mobile Radio) networks, which sometimes end up being a challenge to common sense. Each time creators of networks consult coupling specialists before writing the technical specifications, things come out much better. The exchange of technical views is always fruitful and everybody gains from it.
The functionalities added over time are the following:

– a cavity multicoupler must be able to work at a certain frequency band, the tuning range is increased;

– the mixing of channels must be possible, that is to say that a given cavity must be able to accept an arbitrary and instantaneous change in frequency in the group;

– the final user must be able to tune the cavity by simply turning a button, which possibly implies a mechanical automatic tightening system;

– the coupling bay must be installable elsewhere than in the base station, in a non-protected environment, and must be subjected to shock and vibration tests (the first being regulated on site);

– the dimensions must be reduced, whereas to compensate for the new demands they would need, on the contrary, to be increased.

Despite all these unnatural modifications, we manage nevertheless to make a product that ensures the totality of the functions, but on the condition of agreeing to make a sacrifice on the insertion losses (we go, for example, from 3 to 5 dB) and other parameters if this alone is not enough.

The typical composition of a channel of a cavity TX multicoupler is that indicated in Figure 11.3. In general, the amplifiers are found in another bay, we thus enter on the circulators that are placed here to assure sufficient isolation between two adjacent channels, so as to maintain the intermodulations below the levels specified by international norms. At the time of the appearance of “drop-in” circulators on the market, which may be directly installed on a printed circuit, certain power amplifier manufacturers integrated one of these components at the exit, in the same housing, but the cost and performance analyses done in retreat lead to a return to the conventional layout, the fashion trends having exhausted its arguments. We have represented a double circulator, and this is in fact the habitude and the norm in UHF, but the number of stages depends on the specific channel-to-channel isolation and can pass to 3 in VHF and in HF, whereas one alone is no longer sufficient in general.

![Figure 11.3](image-url)
The channel-to-channel isolation specification is thus essential in the specifications. It is completed by a specification of the antenna-channel isolation, corresponding to a protection of the same physical nature, but coming no longer from adjacent channels, but from parasite frequencies captured by the antenna in the working band. These parasitic frequencies can themselves combine within an amp and intermodulate in the reception band at very low levels. We will thus put as many stages of circulators as necessary to attain the most severe isolation possible between the two.

The theoretical calculation demands knowledge, in addition to the maximum levels of emission and reception intermodulations, of a supplementary datum that characterizes the behavior of the amplifiers from this particular point of view. It is a quantity $K$, expressed in dB, given by the manufacturer and called the “intermodulation coefficient” or “protection coefficient”, or further still “conversion factor”. This is of the order of 10 to 20 dB and is added to the inverse losses of a channel (circulators + cavity) with regard to the calculation of the antenna-channel isolation, as well as to the selectivity of a channel between channel and the other of the cavity when we are dealing with channel-to-channel isolation.

The inverse isolation of circulators being difficult to master in relation to the ambient temperature and traversing power variations, we generally perform verifications via a series of measurements taking into account all these parameters. The network analyzer ambient measurements are not enough, and it is necessary to use a power bench (see section 2.2.2).

Figure 11.4 proposes a multipurpose, but inevitably incomplete, basic set-up, which, according to the position of the measurement components such as the couplers or circulators, allows measurements to be conducted at conditions very close to reality, with nominal power.

The layout of the diagram as it is drawn corresponds to a measurement of direct power losses, but simple permutations of the connections can make it compatible with other tests. The dual power meter represented is a very practical device but has a very wide band, which in most cases leads to its limitation by filtering cavities not represented here. On the other hand, we often use a spectral analyzer for the isolation measurements, in conjunction with a supplementary circulator inserted just before the load or the power attenuator, that allows us to inject a second signal towards the entry at a different, but close enough frequency, in order for this one not to suffer from supplementary attenuation in crossing the coupling cavity.
A particular remark is to be made concerning the output return loss of the complete multicoupler and its measurement. This parameter, associated with the quality coefficient, is often specified at about 15 dB (say, 15 to 18), and a cavity of reduced size, pre-tuned symmetrically at the entry and exit, cannot necessarily attain this value naturally. To satisfy, despite that, the specifications, the usual practice is to destabilize the cavities with the aid of their loops, favoring the exit at the expense of the entry, which is not noticeable because in order to reveal it, it would be necessary to open link B of Figure 11.3 and connect the network analyzer to it, which is not done in order not to further extend an already substantial measuring time.

We manage then to bring down to the desired level, without limitation, the exit of all coupling cavities, in such a way that the measurement of direct losses of any channel is apparently not affected. If we look more closely, this procedure introduces a measurement error first of all coming physically from the attenuation of the reflection in the crossing of the cavity and perhaps from the choice of algorithms of the software packages of the measuring devices and their data processing, when the set-up (when it contains circulators) makes the reflected disappear at an arbitrary location.

Even if the error is not usually high, it is always positive and can mask a real loss that would be impossible to detect except through a measurement of the power radiated by the antenna, which could possibly surpass the specifications by a few tenths of dB. It is therefore not only pointless, but in fact detrimental to specify the exit PSWR of a cavity transmitter multicoupler in these conditions, because it is nothing more than the reflection of the quality coefficient.

If we absolutely wish to improve it, it is necessary to do so using one or more impedance adaptors (see section 8.6), be it two per cavity, or only one placed at the...
exit (beware the dimensioning of the capacitors!), whilst ensuring, in this latter case, that the in-out return loss of the cavities are equal, a condition for having the best yield, that is, the best loss/selectivity compromise. In support of this important remark, a series of fine measurements were performed with a network analyzer HP-8753-C on a cavity operating at 485 MHz, pre-tuned normally and afterwards destabilized in return loss so as to present a correct exit adaptation, starting from an initial value of about 13 to 14 dB.

![Figure 11.5.](image)

The results are summarized in Figure 11.5. Let us add, to remove any doubt regarding the accuracy of the manipulation, that the calibration was verified after the measurements with a difference of two hundredths of dB over the zero of transmission with the aid of the “short” of the calibration kit. The difference observed is not insignificant when, during a test, we know it is necessary to verify the losses that always come within a tenth of a dB of the specifications.

The following complementary measurements allow us to better define the order of magnitude of the uncertainties:

- IL sole cavity A to B: 2.84 dB,
- IL sole cavity B to A: 2.93 dB,
- IL simple circulator: 0.17 dB,
- IL double circulator: 0.35 dB.

When we subtract the losses of the circulators from the results of Figure 11.5, we find very similar measurements to those made on the cavity alone, which seems to exonerate data treatment in the analyzer. On the other hand, the increase in losses due to the imbalance of the tuning is clearly evident.
11.1.5. TX multicoupler with directive filters

The principle of the directive filter was briefly introduced in section 9.10 as an application of directive couplers. It is appropriate now to study it more in detail with a view to using it as a coupling element. Let us recall for this purpose the operational diagram of Figure 9.17, slightly modified to give it a more functional appearance (Figure 11.6). We suppose the tuning has been performed and we consider the directive filter as a component, in its entirety. A radio channel entering in 1 will be split in two by the first coupler, filtered and then recomposed in the second coupler to re-exit in 3. We will use 1 to refer to the “narrow-band” entry. Another radio channel, whose frequency will be situated outside the passing band of two identical filters, will be injected in 2, called the “wide-band” entry, and will re-exit in 3 without attenuation (a part from the negligible one that is due to the directive coupler), since the two filters present an almost infinite impedance towards it. If we desire the channel 2 to be selective, it will be necessary to add, upstream, an analog filter to the other two.

We have thus created a low-loss channel adder with selectivity, but that intrinsically is not better performing than a two-cavity coupling with harness.
The interest of the device appears once we set-up several of them in cascade (Figure 11.7). If the exit 3 of a first directive filter is connected to the wide-band entry 2’ of a second one, wherein the two filters are tuned at the preceding misadaptation band, we can inject in 1’ a third frequency which will be added to the first two at the exit 3’, and so forth. In this way we build an transmitter multicoupler that has as low losses as a model with cavities, without harnesses to adjust and being capable of accepting frequencies randomly located in a band of about an octave. We often find this type of material in FM and TV emission, as well as in radiocommunications systems of military and civil aviation. The frequency F1 can be provided, as is often the case, by a spare transmitter capable of supplying any of the following at all times, in the eventuality of a failure.

The main disadvantage of the system is its cost price and its size, in relation to a system of simple cavities, but it has its supporters in the operation sectors where breakdowns are dramatic and must be repaired in a very short time. From this perspective, the fact of not having to take up the tuning of the channels, when we add or remove it, is decisive in the choice of the procedure. We can thus fine-tune a directive filter taking all the necessary time, and then benefit immediately from a lull in traffic to insert or remove a channel with the assurance that there will be nothing to alter. Operators subject to this kind of activity do not find this guarantee of efficiency a high price to pay. Yet some have the tendency to see in it a system that can be extended at will, and it is necessary at this point to remember that when we have coupled n carriers, each of an average power P at the common point, the total peak power is equal to 2nP, and the corresponding voltage must be less than the disruptive voltage of the weakest element.
11.2. Receiver multicouplers (RX)

An RX multicoupler is composed of a preselector, a low-noise amplifier and a splitter (Figure 11.8), each of these elements having already been studied before. Its role is to distribute to a given number of receivers the signals in the reception band, coming either from the reception antenna, or the antenna duplexer when this is unique.

The preselector filter is a band-pass, that can moreover coincide with the RX filter of the antenna duplexer if this is of the bandpass type. Its role is to protect the amplifier from saturation and inter-modulations by eliminating the undesirable frequencies situated outside of the reception band. It is here that we find the major application of the ceramic filters, in the sector of radiotelephony, when the size of a network allows large series manufacturing with a stable frequency plan. But comblines, in VHF and low UHF, keep all their appeal with their flexibility of production. The preselector is by definition a low power filter, which could lead to very low dimensions, but its noise factor, in other words its loss, interferes such that in the total noise factor of the reception chain, it will then possibly be necessary to concede it a certain volume for this reason. We do not conceive that a pre-selector could have a loss largely exceeding a decibel.

The low-noise amplifier, located immediately behind the pre-selector, is responsible for bringing up, if necessary, the level of the received carriers to make them compatible with the threshold of the receptors. The corresponding gain is a function of the number of exits and can vary by 3 dB for a 2-channel PMR at more than 30 dB, as, for example for a CATV (collective antenna) distribution system. In this last case, multiple stages will be necessary, three in general. The basic diagram
proposed in section 8.5 remains viable because it is cascadable, the transistors will simply not be the same according to the stage considered: the first will have a better noise factor, the last the best intercept point.

The power-splitter, as the name indicates, distributes the multifrequency receiving signal over as many channels as there are receptors in the station. In the great majority of cases, it sits next to the amp on the same printed circuit, which strongly incites the usage of a strip technology, on the condition of the central frequency not leading to quarter-waves that are too large, even when folded several times. If this is the case, we will then have to turn to the hybrid transformers described in section 9.3, but we must note that the cost price jumps due to the manufacturing time of the windings. It is necessary meanwhile, if there are strict environmental standards regarding shocks and vibrations, to fix the ferrites to the printed circuit, which complicates the set-up even more. But this is also the solution that takes up the least place, and this is the reason it is always used in low VHF. Let us note, however, a particularly well regarded production by Kathrein in CATV, which consists of using fairly thick ferrites slipped onto single-turn windings in a printed circuit, which allows us, with the aid of a simple mechanical set-up, to reconcile the two technologies; this is evidently only an interesting solution for very large series.

11.3. TX/RX multicouplers

In radio we can use one or two antennas at the base stations. One layout or another is chosen as a function of criteria that are in general quite vague, or where practical, technical and economic, as well as political, views can mix, that moreover can sometimes be modified shortly after their establishment. From a technical point of view, the only thing that is of interest to us, is that each of the two options must guarantee minimum isolation between the transmission and receiving, defined by the system’s specifications. This obligation leads us, for the two-antenna configuration, to distance them enough from each other to attain the necessary value of attenuation. This isolation between antennas is nothing but a fraction of the total isolation, which is a function of the emitted power and sensitivity of the receptors, and that brings into play the whole of the network. It thus also depends on the inverse attenuation of the circulators, on the selectivity of the cavities, the directivity of the couplers (if there are any) and the different filterings put in place. We cannot be more precise without knowing the specifications and the options chosen, but it is generally situated in a range of from 30 to 60 dB in PMR radiotelephony. We can perform it in two ways: vertically and horizontally. Vertically, this means on a same tower and, though the usual radiation diagrams lead to feasible heights, it is still necessary to have authorization or simply the material possibility of erecting it. In addition, the construction must be sufficiently rigid so that the wind does not put the two antennas
out of true, which could greatly deteriorate the isolation. Horizontally, it is necessary to have a distance between the two antennas that can surpass by a large amount the extent of the site, for values greater than 50 dB. As an example, Table 11.1 gives some figures for two half-wave dipoles at an exploitation frequency of 150 MHz, corresponding to a wavelength of 2 m. For complementary values and further details, consult the curves in Chapter 13 (utilities).

<table>
<thead>
<tr>
<th>Shielding (dB)</th>
<th>30</th>
<th>40</th>
<th>50</th>
<th>60</th>
</tr>
</thead>
<tbody>
<tr>
<td>Vertical distance (m)</td>
<td>2</td>
<td>2.8</td>
<td>4.2</td>
<td>6.2</td>
</tr>
<tr>
<td>Horizontal distance (m)</td>
<td>8</td>
<td>26</td>
<td>84</td>
<td>260</td>
</tr>
</tbody>
</table>

Table 11.1.

The solution with a single antenna thus offers an alternative and attractive perspective, where these problems disappear, but only to be replaced by other ones. It is necessary in fact to use a duplexer to couple the transmission and reception parts, and this device must assure a cohabitation without damage between a certain number of power carriers (medium or strong – dozens of Watts), and an access where the signals will be directed at very low level (fraction of picowatts). The selectivity of the two constitutive filters will pose no problems unless we impose a maximum size, the delicate point residing essentially in their linearity. This could a priori come as a surprise, given that the passive filters are theoretically considered as perfectly linear, but the investigations conducted when we note the creation of intermodulations of a forbidden level in the receiving band lead invariably to the duplexer, whose antenna exit, common to the two high- and low-pass filters, is in general the critical place where the non-linearity happens. Its accomplishment thus demands particular care, which will be dealt with in greater detail in section 11.5.

Figures 11.9 and 11.10 propose, amidst an infinite number of possibilities, two architectures of TX/RX multicouplers where we have tried to integrate the totality of the elements that we have at our disposal to create the ensembles, with the exception of progressive couplers which we can easily imagine in similar devices. The first of them shows a hybrid 4-channel wide-band coupling with a double-reject duplexer; the second a cavity coupling, therefore narrow-band, where the duplexer was chosen to be double passband, which allows us to suppress the pre-selector from the reception part. It is evident that we can add, switch, suppress, or multiply each type of constitutive element to make the diagram the most suitable to a given project, as a function of all the technical, commercial and administrative data, among others. The influence and the wealth of these factors make it impossible to set up a universal
model capable of resolving all possible scenarios, and moreover, for this reason they would be non-optimal.

This being so, the establishment of the operating diagram of a coupling system is not too lengthy to perform. The choice between one or two antennas, and so of the usage or not of a duplexer, is practically always imposed from the start by considerations relating to the site more than to the network. Afterwards comes the apparent radiated power (ARP) necessary for the coverage of the zone and that, as a function of the power of the transmission amps, will lead to a maximum loss value that can itself signify the elimination of a certain type of material. Then, depending on us being in a cellular network with total coverage or in a PMR with a limited number of stations, or in an isolated station, the possibility of guaranteeing at each station a sufficient separation, through interleaving of the frequencies, usually of the order of 10 to 20 channels between adjacent ones, will not be always possible and will exclude the cavities. In brief, and proceeding mostly by elimination, we arrive very quickly indeed at determining the structure that is suitable and that ultimately imposes itself. It is necessary afterwards to calculate all the parameters that will figure in the specifications.
One of the reference points is given by the sensitivity of a mobile. We will adopt for this, only as an order of magnitude because it is constantly evolving, a value of 0.25 $\mu$V with a protection of 12 dB SINAD (Signal to Noise And Distortion, the signal-to-noise ratio where the noise includes not only the thermal but also the spurious signals), that is, about $-120$ dBm. This places the maximum level of parasite rays in the receiving band to a value that we will round to $-130$ dBm, which corresponds to a power of $10^{-13}$ mW, that is a ten-thousandth of a picoWatt. The coupling bay should not create any signal exceeding this level in the receiving band, and this will be one of the specifications that will condition the value to be given to the total attenuation between transmitting and receiving bands.

The values to be provided in the specifications, to completely define the coupling, are the following:
Table 11.2.

We find in this summary table (Table 11.2) two sorts of specifications: firstly those we can call structural, such as the passing loss, because they are a consequence of the physical and geometric characteristics of the constitutive elements, and afterwards those which come from the radioelectric requirements of the radio system, as the various protective losses and intermodulation levels. The two families share the common point that miniaturization makes them more difficult to be achieved.

11.4. TMA

This abbreviation stands for “Tower Mounted Amplifiers”. This material appeared in the wake of microunit networks, to solve problems of balance of channels going up and down, as well as to complete coverage comprising “blind zones”. There exist simple TMAs, consisting of a bandpass filter and a low-noise amplifier, and duplicates, working in both senses, with a transmission and a reception channel, that are composed of two duplexers and a low-noise amp (Figure 11.11). The base stations being more powerful in transmission than the mobiles, we find it advantageous to re-amplify the latter to the level of the reception antenna of
the station. We thus decrease the Rayleigh fading by increasing the reception sensitivity of the station, allowing mobiles to demand less from their batteries. The gain is never very high because it is necessary to avoid saturation and intermodulations; it is generally of the order of 10 to 12 dB. The two duplexers use a common filter for the transmission channel; its performance as bandpass is not critical and it is often a 3-poles whose main function is assuring duplexing the adaptation. The two filters of the reception channel (of the base or relay station), are, on the other hand, very selective.

In order of appearance, we have seen the following accessories appear close to the base station antennas: first, simple spectrum-cleaner filters, without amplification, and next single, then double, TMAs. These are now products that are part of the normal catalog of radio equipment, their industrial take-off having been such that studies were conducted on a supraconductive solution, in order to reduce the noise factor to a minimum, i.e. theoretically to 1. The administration financed an experimental network, that has in fact worked, but allowed us to note the appearance of technical problems that made the noise factor of 0 dB unattainable. Meanwhile, by itself, the constant progress of transistors made the total noise factor of a traditional TMA decrease to less than 1.5 dB, for a price 20 times less than the cryogenic model, equipment and maintenance cost included. With a gain of just 1.5 dB or less at such a cost, this solution is no longer considered, but it is known that it was tried and that it worked.

The problem common to all these small pieces of secondary equipment that may be added to the infrastructure of a network is that they are subjected to the worst climatic conditions, being susceptible of being installed anywhere and in all types of
locations: cold, heat (and sometimes both alternately), humidity, salty mist, nothing is spared them, and their environmental mechanical specifications are very strict. The range of working temperatures is generally from \(-40\) to \(+70\)°C, and they must be totally impermeable, and this leads to a technical parenthesis. At first sight, it would seem that we are distancing ourselves from the actual HF here, but this sealing problem is noteworthy in that it is precisely of great generality in this domain, because it concerns all housed equipment. It is also the cause of regular controversy about the way of solving it, in particular regarding the basic principles to be adopted: is it better to have an absolutely impermeable housing system, resistant to water coming even from below, or is it better to keep a small aperture at a low point for the humidity that might have entered despite all precautions taken to drain? One should be convinced of the fact that water always ends up entering the system, and this is evident for constructors. A housing necessarily possesses an entry and an exit for the “signal”, and so two fixed connectors on which two coaxial cables are connected, plus a power plug if this is not done via the core of one of the coaxials. A simple lesion of an insulator and rain water will seep into the braid, which will lead it precisely where we do not want it to go. But even without an accident of this kind, whatever we do, water will enter because no connector is perfectly sealed over time. Knowing that, the reasonable and realistic position is to renounce the concept of absolute sealing and to plan an evacuation passage below the system, protected from vertically ascending sprinkling by a simple device of which an example is schematized in Figure 11.12.

![fibers](image)

Figure 11.12.

The conception of new materials for base stations is evolving continuously, towards greater sophistication in the treatment of the signal, so that this kind of material, restrictive by its installation demands and additional costs, has the tendency of disappearing as a result of the recognition of the problem it was supposed to solve at the level of data management of the network.

Radio-range enhancers are devices that we can group in the same family but which are a bit more significant in size and in power, because they are intended to extend the coverage zones where an insurmountable obstacle for radio waves creates a zone of total shadow (mountains, car parks, tunnels, etc.) and requires the use of a
bidirectional relay. The structure is schematized in Figure 11.13: two bandpass
duplexers, which assure all of the filtering functions, allow the head-to-tail
arrangement of two amplifiers with each other, one for the mobile located in the
shadow zone and captured by one of the two antennas, the other for the signal
coming from the transmitting station and collected by the other antenna. The aerials
can be directional or omnidirectional antennas, or further still, radiating cables in the
case of sites of a certain length, such as tunnels.

![Figure 11.13.](image)

11.5. Power and intermodulations

In the definition of signal-to-noise ratio, we can consider noise as being
composed of all that is not the useful signal. This being so, it is at the same time
convenient and evident to distinguish two large components in it: wide-band noise,
of which the best know is thermal noise, and spurious radiation. Thermal noise is a
well-known and predictable physical phenomenon, whereas spurious radiation is on
the contrary of multiple natures, sometimes predictable in their frequency location
but not in amplitude, except in particularly simple cases. Intermodulation radiation
constitutes one of the families that is most particularly of interest to us, given that
they can be generated by the equipment itself. The notions of power and
intermodulation are linked to the extent that they vary together and in the same
direction. Yet, there is no direct causality between one and the other: the appearance
of parasitic frequencies is due to the non-linearity of one or more elements, a
common and convenient term to use, but that hides a reality not as simple as it
seems, and that deserves for this reason a more in-depth study.

11.5.1 Limit power

When we progressively increase the HF power applied to a component, we see
the development of two kinds of very distinct manifestations: firstly there are
thermal effects due to the transformation of EM energy into heat, in the places
where we have identified passing losses, and that manifest themselves as an increase in the temperature; and secondly, we also identify distortion effects associated with the creation of harmonics that, combined, generate undesirable radiation as an increasing function of the intensity of the power applied. Then, if we continue to increase the power, a disruptive discharge is produced in the circuit. It is most often destructive by means of an elevation of temperature that is no longer distributed and gradual as in the thermal effects, but which is instead sudden and of great intensity, and that leads most of the time to combustion or evaporation of an essential element at the point of creation or arrival of the discharge. One of the roles of the project engineer will thus be that of evaluating, to begin with, and then harmonizing the breakdown voltages of all constituents of the structure, striving to suppress all abnormally weak points.

It is not possible to lead these kinds of studies if we do not know the characteristic values of breakdown voltage of the materials we use, and first of all of atmospheric air, which is taken as a reference for the other substances, in the same way as the vacuum serves as reference for relative permittivities. Unfortunately, these characteristics depend on several parameters that absolutely must be specified when a value is foretold. In fact, the breakdown voltages are measured at the moment of the appearance of a spark between two electrodes subject to an electric field that is progressively increased until the discharge, and the analysis of the measurement result readings shows the existence of variations as a function of the separation of the electrodes, their shape, that we associate with the curvature radius of the element, as well as their composition and the meteorological conditions (humidity, pressure, temperature). For this reason, the experimental results, of which the first serious publications date back to the end of the 19th Century, show fairly large discrepancies, enough to create a certain difficulty when the moment comes to fix the value based on which we will perform the dimensional calculations. We will find in Figure 11.14 average curves, derived from more complete tables reproduced in the utilities chapter, compiled from several databases dating from the beginning of the 20th Century, but which are not challenged these days.

The sources are the following:

– De Laharpe, *Notes et formules de l’ingénieur*, Albin Michel 1927,

– Desarces, *Nouvelle encyclopédie pratique d’électricité*, Quillet 1936,


The dielectric strength of the insulators grows as the thickness decreases. It is a general rule that is difficult to interpret.
For air, we have for example, according to Gray, 29.8 kV/cm for 1 cm, 33.6 for 5 mm, 51 for 0.5 mm, and we can thus achieve 200 kV/cm for very small thicknesses. It seems that the records for dry air, between two plates distanced 1 cm from each other, are of 32 kV, which is very close to the value indicated by Gray. At the electrostatic workshop of the Palais de la découverte, every day the general public can admire a discharge between an sphere of about 50 cm diameter taken to 150,000 V and a ground plate situated 15 cm below it, which gives a disruptive field of approximately 10 kV/cm. We can certainly object that in this particular case the conditions of the experiment are unfavorable (excessive assistance, high humidity level, ionization of the air) and that the equipment, used daily for several years, probably presents acute microfissures at the starting of the arcs, but we are still able to grasp through these few numbers the complexity of an apparently simple phenomenon, and the necessity to always specify all the parameters when we propose a numerical value.

In practice, and especially in the domains that are of interest to us, we often use a value of 29 kV/cm proposed in Matthei, whichever the distance we are lead to, a value which seems convenient to everyone, and that has always lead to satisfactory predictions. However it is worth noting a formula taken from the compendium of constants from Abraham/Sacerdote which gives a simple law of variation for a
dielectric strength in dry air at 15°C between two infinite planes, annotated \( V_d \) (disruptive voltage), valid for a length of first spark \( d > 2 \text{cm} \):

\[
V_d \frac{(\text{V/cm})}{d(\text{cm})} = 25.9 + \frac{5}{d(\text{cm})}
\]  

When a band-pass combline filter is intended to support a certain peak power, a very different parameter from the mean power when we are in multicarrier mode, it is necessary to give a particular form to the upper part of the resonators and to the tuning elements, already mentioned in section 6.1.3 and developed in Matthei (p. 900 onwards), in order to protect the filter from an internal overvoltage due to a radius of curvature that is too small. On the other hand the silver-plating will have to be polished with a rough cloth in order to suppress all micro-needle-points. The evaluation of the effective peak voltage at the interior of the filter ultimately depends on three parameters: the number of carriers, the high voltage due to the passing band and that due to the geometric form, or more precisely, to the radii of curvature. To simplify the calculation of this last category we can, for crossing powers of the order of a few hundred Watts, limit ourselves to standardized proportions illustrated by
Figure 11.15. For the very selective filters and powers measured in tens of kW, with a frequency tuning no longer done by the mass but by the central conductor, it will be necessary to lead a more elaborate study, based on the indications given by Matthaei, but the proportions of the drawing in Figure 11.15 allow us to save time by offering a tested solution for common cases. The curvature radius of the added part welded at the top of the tube is thus chosen to be of a quarter of the radius $R$ of the tube, and the distance $b$ with the mass disc varies between $R$ and $R/2$, which leads to a high voltage coefficient due to the curvature between 1.45 and 1.75. In this configuration, for a signal of effective power $P_0$ entering the filter, the voltage to take into account for the calculation of the separation between the resonators in relation to the ground is:

$$V_{\text{peak}} = n \sqrt{2P_0 Z_0 \frac{F_0}{\Delta F}} \times 1.75 \times 1.1$$

(11.2)

$n$ being the number of carriers. 1.1 (or 1.2) is a coefficient corresponding to a security margin that we choose according to habit. Once the maximum peak voltage has been determined, we use the upper curve of Figure 11.14 or the fixed value of 29 kV/cm to deduce the minimum corresponding distance.

Example: a bandpass of 5MHz band at 460 MHz receives 8 carriers of nominal power 50 W each. The internal crest voltage is:

$$V_c = 8 \times \sqrt{2 \times 50 \times 50} \times 460/5 \times 1.75 \times 1.1 = 100,183 \text{ V}$$

If we choose the value of 29 kV/cm for the disruptive voltage, the minimum separation to the ground will be: $100,183/29,000 = 3.45 \text{ cm}$.

There is nothing better than a numerical application to enable us to understand the effect of one parameter change or another. In the preceding example, we could see that reducing the band by half will lead to having to double the security distance, and so increase the thickness of the filter by an equivalent amount; we can also see that going from 8 to 16 carriers will produce the same result. We also find an order of magnitude that often surprises the uninformed user. In 1986, on the site of Lyon La Sarra, France-Télécom made an attempt to regroup 64 standard radio channels on the same antenna, via an experimental coupling comprising cavities and hybrid couplers, the whole being “covered” by a two-pole anti-harmonics filter, of a volume of about 20 dm$^3$ and comprising in the interior additional centering pieces made of Teflon. The experiment lasted no more than 30 seconds, after which time traces of black smoke started to leave the system by the tightening screws, indicating to the technicians present there that they had only just avoided the filter exploding. The calculation which should have been done beforehand revealed after the fact that
in the interior there were peak voltages greater than a million volts. The autopsy showed an extraordinary sight of silvered aluminum sheets presenting practically all the colors of the spectrum, a viscous surface for the parts mechanically intact and engravings of several centimeters at the starting points of the discharges. It is thus necessary not to forget to systematically carry out these calculations, be it when we are dealing with very high powers or multicarriers, and even with relatively modest nominal powers, and to never insert Teflon parts in the strategic places of the power filters.

![Diagram](image)

**Figure 11.16.**

This problem of high-voltages is not the only one to be posed when we have to deal with power. In a notch filter, the resonators work in the rejection band and are not subjected to the same phenomena as in a bandpass. A pass-reject duplexer with the same dimensions as a bandpass can hold a power usually 5 to 10 times greater, which is in fact not limited except by the quality of foot impedances on the one hand, and of the quarter-wave line on the other hand. The selfs of the low-pass side being practically indestructible, the vulnerable point is usually (we should even say practically always) constituted by the capacitors of the high-pass part. These components, especially when they are tunable, have two weak points: their breakdown voltage, which is quite easy to reach in multicarrier mode even if there is no high voltage coefficient to take into account, and their internal resistance (ESR), which can cause a destructive heating-up and a piercing of the dielectric. This last point is in common with power devices with printed circuits, such as certain models of hybrid couplers. It will thus be necessary to carefully take into consideration all the characteristic parameters of the materials used, to choose them well and to determine for each usage the thermal energy dissipated by internal losses and the critical arc path. When, for example, we must use capacities less than a picoFarad in a pass-reject, it is convenient to do it in cut-out patches on a printed circuit plate. It will then be necessary, as in the stripline couplers, to introduce onto the edge a non-
metallized zone that will sufficiently extend the spark path to bring the disruptive tension in the air to the same level as that of the substrate (Figure 11.16).

When we add on the same channel two signals of effective nominal power $P_0$ via a hybrid coupler, we do it by sacrificing half of the power, which leaves in the form of heat in an matched load. This is the price to pay in order to benefit from a channel-to-channel isolation which allows total agility within a relatively large band. We find ourselves then, at the exit, with two carriers of an effective power of $P_0/2$ each. But when the power is divided by 2, the peak voltage $V_{c0}$ is divided only by $\sqrt{2}$, and the two signals combined generate a peak voltage $V_{c1}$ which is in the end $\sqrt{2}$ times higher than the original peak voltage of a single non-coupled signal. If we are dealing with a 4-channel coupling, we will have 4 carriers at the exit, each with an effective power of $P_0/4$ but a peak voltage $V_{c2} = 2V_{c0}$. Generalizing now to a Wilkinson coupling of order $n$, the peak voltage at the exit will be $V_{cn} = V_{c0} \times \sqrt{n}$. Such a progression does not generally pose many problems in terms of the couplers themselves, unless we have very high power signals: if we take, for example, a 50 W hybrid coupling of 16 radio channels, the maximum voltage at the exit will be 283 V instead of 70.7 V for a single entry. On the other hand, if this coupling is followed by a band-pass filter, the latter will see its maximum internal voltage increase by the same proportion and could suffer permanent damage if we have not taken this initial high-voltage into account in the calculations.

11.5.2. Intermodulations

The non-linearity of a reportedly linear quadrupole is a complex phenomenon that manifests itself by parasitic signals at such low power levels that it was practically ignored up until 1980. But mobile radiotelephony quickly resulted in receptors whose sensitivity is such that the weakest useful carrier levels used are now of the same order of magnitude as parasite rays, which consequently constitute a sort of technological barrier against a new lowering of the reception thresholds.

The origin of this evil was firstly discovered in duplexer, by successive elimination. It is clear that in the few cm$^3$ where the transmission and reception channels are assembled, the cohabitation of signals having levels of 170 dB of ratio cannot occur without a few difficulties arising. But is it not only here, and we notice in fact that from a certain level of observation no element is linear. Hence the relentless pursuit of parasite signals, which is now the subject of specifications concerning all the subgroups, of transmission or reception.

When two frequency carriers $F_1$ and $F_2$ pass through a non-linear element, this causes a deformation of the sinusoidal signal that results in the appearance of
harmonics and creates an infinite number of signals of the form \( mF_1 \pm nF_2 \), by multiplication. The sum \( m+n \) is the order of the intermodulation. Whilst the harmonics are located above the fundamental frequency, the intermodulations can be anywhere in the spectrum. Those of order 3 are of the form \( 2F_1 \pm F_2 \) or \( 2F_2 \pm F_1 \), with \( F_2 > F_1 \) and \( \Delta F = F_2 - F_1 \), and are of more particular interest for radiotelephony, because those corresponding to the sign “–” in the relationships above, are found at \( \pm \Delta F \) of \( F_1 \) and \( F_2 \), that is to say, in the adjacent channels. They are therefore potential perturbers whose level must absolutely be kept under the sensitivity of the receivers, which is by definition the minimum allowed level of the carriers received.

The greatest problem in the fight against low-level intermodulations is that we do not know how to compute their probable level, given that we do not know their causes with certainty: the low-level intermodulations are not calculated, they are measured. Those who must master them find themselves before a parametric chaos somewhat similar to that of meteorologists, and they have as their only weapon an ensemble of rules, principles, techniques and precautions that together make up an tooling of a certain efficiency and that is the state of the art, at the same time revealing to us that there are limits to what it is possible to do.

Simple vibratory phenomena, relaxation oscillations apart, are represented by a sinusoid: a pendulum, a vibrating slat, a spiral spring, a resonant circuit, all manifest themselves as oscillating systems having sinusoidal variations around an equilibrium point. But as soon as an external parameter is added to the mathematical model, the sinusoid is deformed and harmonics appear. This phenomenon can be desired, as for example in the harmonics generator of a hyperfrequency “source”, or fought against for the reasons previously mentioned. In the first case, let us say, to illustrate the idea, that of a multiplicative diode, the non-linear action, that has as a result the disappearance of half the representative curve, is so evident and of the first order that we can model it and break it down into a Fourier series, having a complete representation of the harmonics spectrum, including the amplitudes. This is one of the rare cases where we can make predictions. Regarding the secondary phenomena of non-linearity, we are completely unarmed from a theoretical point of view. There remains a principle, which is the breadcrumb trail of the intermodulation hunter: all that introduces a variation into the propagation conditions of a wave is a potential source of non-linearity. We can thus sketch out a strategy based partly on an attempt at classification of the noxiousness level, and partly on an inventory of probable causes.

From a practical point of view, we can distinguish three broad types of non-linearity:

a) induced non-linearities that we will call of the first kind. These are the ones intended to generate strong harmonics of which one or more will be chosen for a
specific usage (frequency multipliers, frequency converters, hyper sources, detectors, etc.);

b) non-linearities of the second kind. We will refer to in this way those which give intermodulations commensurate with the fundamentals, that is to say, visible on the same spectrum analyzer screen without the need for us to change the sensitivity, and whose cause is particularly evident, such as the diode effect of a badly-performed welding or an intermetallic couple. The remedy is generally found after the identification of the problem, and is in most cases a repairing operation. The preventive action consists of silvering all the welded parts, so as to make all the electric couples and potential diodes disappear. We recall here that silver is the best conductor, and that it is non-oxidizing and the blackness that appears after a certain amount of time exposed to air is due to sulfuration, which remains conductive. The simultaneous use of golden connectors with silvered coaxial cables is strongly inadvisable;

c) non-linearities of the third kind are those which cause very low-level intermodulations, invisible to the analyzer in normal observation conditions, and which we cannot detect except after an in-depth and focused investigation.

Only categories (b) and (c) are of interest to us. The non-linearities of the second kind (b) are primarily the effect of devices we know to present this particular feature by their very nature, at least under certain operating conditions: an amplifier, for example, is studied as a linear element, but when the power increases it is subject to a saturation phenomenon that causes it to become increasingly non-linear. The same happens for a circulator, which nevertheless is a passive component. There are also involuntary non-linearities that link themselves to a known pattern by their structure or their symptoms: a badly performed welding, an intermittent contact, a crack, a couple between two non-silvered metals after welding, are all comparable to a diode or another semi-conductor. The latter are actually failures that are often repaired by a simple visual examination, and their repair is accompanied by the disappearance of the parasite signals that triggered the alarm. The others are characteristics of the components, multipoles or subgroups that cause them, and are the subject of specifications: an amplifier manufacturer, for example, always provides the value of the intercept point of the order 3, which allows us to anticipate the level of the intermodulations of the same order. If these indications are not provided, they must be measured.

Non-linearities of the third kind (c) are those that create parasite signals at very low levels, but enough to be detectable by receivers of operating equipment, which consider them as ordinary carriers and inform the computerized management system that the corresponding channel is busy. It is therefore very important in a project for a mobile network or another multi-carrier radiocommunications system, to know beforehand the recurrent risks of "spurious" effects in the allocated channels. For
this reason we have an interest in entrusting a computer with the task of calculating all the combinations of the form $mF_x \pm nF_y$ that fall within the reception band. For the 1992 Albertville Olympic Winter Games, for example, the company responsible for the mobile radio security system decided to develop a program for locating the risky combinations up to the order 17 (i.e. $m+n = 17$). Such a program made it possible to establish a list of “double” frequencies corresponding to false signals susceptible of taking the place of the correct frequencies, regardless of their amplitudes since we are incapable of justifying a spectral variation law. In fact, while it was reasonably predicted that a globally regular decrease and a progressive cancellation would take place with the order of the intermodulation, the measurements showed a particular strength of signals of the orders 13, 15 and 17, which were found to be of the same order of magnitude as those of order 5.

The search for parasite signals is further complicated by the fact that the measuring bench itself produces them, independently of the equipment to be tested. The internal clocks of the synthesizers, for example, are often made from standard 5 or 10 MHz oscillators, whose signature we find throughout the spectrum in multiple forms. A systematic preventive action will consequently be to identify, before the measurements, all the frequencies coming from the bench. During the measurement, on the other hand, it is necessary to identify the measured signal, verifying, when it is present in the list of frequencies to be monitored, that it follows with exactitude the amplitude variation laws that its order implies: if the two carriers creating intermodulations vary by 1 dB, the parasite signal of the order $m+n$ must vary by $(m+n)$ dB. If this is not the case, we do not take it into account, subject to verifying, of course, that it is not generated by another subgroup of the equipment instead, but the essential, from our perspective, is to eliminate the coupling from being a possible cause.

Predicting the position of the undesirable intermodulations allows us to verify their existence but does not provide a solution to eliminate those that surpass the specified level. For that, there is still all the work to be done and this can be an operation that is difficult, unpredictable and tedious, in the completion of coupling material. The above in fact illustrates well and justifies the term “parametric chaos” which we used before when comparing the “hunters” of intermodulations with meteorology forecasters, in relation to their working conditions and their theoretical equipment. In fact, the main weapon against parasite intermodulations is still the quality of manufacturing, resting first of all on the elaboration of a system of manufacturing rules resting on a search of the possible causes. Yet these possible causes cannot be more than hypotheses, whose validity depends on the level of dependence of the non-linearities on the physical phenomena that we believe to be responsible for them. Seen from this angle, the task seems complicated, or even impossible, but there is another way of looking at it, simpler and more realistic, based on the principle that if the energy propagates in pure TEM mode through...
perfectly adapted ideal quadrupoles, there are no intermodulations. The action comes down to searching for and identifying over the path of the signal, from the channel entry until the exit of the antenna, all the possible causes of accidents along the propagation, in the general sense of the term, and of imperfections in relation to the theoretical models. It will suffice afterwards, if we may say so, to check the suspected locations and to try, as far as possible, to vary the parameter considered the most representative to see if the intermodulation varies. The list that follows, which we have divided in three large categories, gives an example of classification that could serve as the basis for finer analysis.

**Adaptation faults**

These are synonyms of return power, or in other words, of the creation of a signal that propagates in the opposite direction and is susceptible of combining with the direct signal in a non-linear element. It therefore not a primary cause, but a parameter that can amplify the effect of a nonlinearity located at a different place. All the line transitions are to be watched from this point of view, including those that are at the interior of the connectors and are therefore invisible: a plug or a socket which are ill-designed have the same effect as a damaged connection.

**Modes of propagation**

This hypothesis appeared quite late on in articles dealing with the causes of intermodulation, simply because the technological progression of receivers called for more extensive thinking on the origins of intermodulations of very low level due to the non-linearities we have referred to as the third kind. The presence of several modes of propagation at any place in the path of a signal typically constitutes a non-linearity of the third kind. This may be associated with a cable where the coaxial is not homogeneous, in which case it will be necessary to consider changing the kind of product. This is an additional reason why it is recommended to have a good stock of cables of all models in a HF laboratory. Double-braid cables are just as preferable as simple-braid ones.

**Energy density**

Here we are within a little-explored domain due to the permanent equivocation about the existence or non-existence of the propagation medium called vacuum, as a physical object. We have observed, thanks in particular to the multiplication of radio-mobile networks and the different models of multicoupler they have required, that the size of the elements, their dimensions, were more or less connected to the minimum level to which the intermodulations could decrease. As a function of the dimensions imposed on transmitter couplers, it seems impossible, in fact, to have better values than those obtained after every attention has been directed to the manufacturing of the equipment. With Acropol-type cavities, for example, which have a volume of about 5 dm³ and operate at 450 MHz with 50 W carriers, we have
never succeeded, in a regular manner, in decreasing the transmission IM3 below the
–100 dBm mark. Attempts at miniaturization of the filters have always lead to this
kind of consequence. It is possible then, to come back to the Maxwellian
conceptions of the vacuum, that there exists in free-space a limit to the density of the
EM field, of which the progressive approach leads to a distortion and the progressive
creation of intermodulations.

11.6. Multiband coupling

In zones of radio obscurity where mobiles can be found (tunnels, car parks), the
operators must assure coverage by extending the transmission range of the signal by
bidirectional TMAs or similar (we must be able to reach the telephone, and make
calls). Yet the communications of subscribers are not the only ones to take into
account: the security services should benefit from the same advantage, and this is
particularly evident for fire services, ambulances and the police; and ordinary users
expect to receive at least FM and long-wave radio.

We thus find ourselves faced with a certain quantity of frequency bands, going
from low VHF to high UHF, which must be transported in parallel from the
coverage zone to the non-covered zone. The necessary devices comprise amplifiers
and a coupling system. There are now two strategies, whose choice depends both on
the technical possibilities and on the number of bands to pass. The first consists,
when we have high-power, wideband amplifiers, of using a hybrid coupling, with a
large loss compensated by a strong amplification, but a band that can include all
those that are to pass. The second, selective coupling allowing an adapted
amplification at each band, is more complicated from a filtering point of view but
has no limitation on the number of bands.

![Diagram](image_url)
The principle of multiband coupling is the same as that of a cavity multicoupler (see section 7.3.3): a harness allows us to adjust the relative phases so that at the common point a channel to be coupled sees the other as an infinite impedance. It is thus the same problem, with the simple difference that the monofrequency channels turn into bands. In the multicouplers of base radio stations, the frequency separations between channels are very small and allow us to define an average length for the harness that suits the ensemble. In the case of multiband coupling, we deal with arbitrary frequency bands that, seem from a given filter, are distributed anywhere on the external circle of the Smith chart, so that we can calculate one harness for two groups, but only two, unless we are in a particularly favorable situation. Yet inside a tunnel, it is usual to have to assure the operational continuity of a dozen bands going from 30 MHz to 2 GHz, just for the mobile radio and broadcasting, not to say beyond, in both directions for other bands.

Before generalizing the method of resolution, we will take a simple example with two coupled bands and a third incompatible one. Let these then be the three bands B_1, B_2 and B_3 of frequencies F_1, F_1' for the first, F_2, F_2' for the second and F_3, F_3' for the third. The first two bands pass through filters A_1 and A_2 which are coupled by a first harness and are seen by filter A_3 in the layout shown on the chart associated with it (Figure 11.17).

We readily see that there exists no phase rotation that could simultaneously bring the representative points of A_1 and A_2 to the point $\infty$, which seems to be an extremely awkward limitation of the technique of coupling by harness. It is thus necessary to find a palliative method that can work in all cases. This method consists of using phase equalizers, which are wideband filters of the high-pass or low-pass type.

In the example considered, if we have the condition F_1< F_1' < F_2< F_2' < F_3< F_3', it will be a low-pass filter that will cover bands B_1 and B_2 and it will be inserted downstream of the nodal point of A_1 and A_2. The effect of this supplementary filter, very easy to build since it is a concatenation of LC cells, and also easy to tune if the capacitors are trimmers, of low loss and little cost, is to draw all the characteristic points close to the point of infinite impedance, as Figure 11.18 shows. The coupling of the three initial filters boils down to a coupling with A_3 of the group A_1 + A_2 + equalizer considered as a sole filter, an operation which is performed by setting-up the harness l_3 + l_4. If bands B_1 and B_2 are above B_3 instead of below, the equalizing filter becomes a high-pass and produces the same effect of regrouping the frequencies. When the number of bands increases, it is always possible to add equalizers to, each time, bring the coupling to that of two filters, starting by the two lowest and the two highest ones in terms of frequency to arrive at a central nodal point where all channels are coupled. Figure 11.19 shows a practical example made for the tunnel of the urban motorway A14, in the suburbs of Paris.
Figure 11.18.

Figure 11.19.
Chapter 12

Utilities

In this chapter we will find first a series of small BASIC programs which, coded in a pocket calculator, constitute a very practical support that every HF specialist should always have with them and which the author has used daily himself for the 20 years he has worked in the field. Afterwards come tables of values whose frequent consultation justifies their presence within this work. The programs were developed in 1985 on a pocket calculator, Tandy PC-6, that no longer exists, and they consume no more than 10 kilobytes of memory. They enable the main dimensions of a combline filter to be given instantly according to Dishal’s formula, for example, or the Q₀ of a cavity to be obtained as a function of its dimensions, as well as enabling us to solve the little conversions that we have to do dozens of time every day, such as passing from VSWR to adaptation in dB, etc. Anyone can add to them whatever they want according to their habits and needs.

12.1. BASIC programs

Program P₀: Return Loss.

Gives the adaptation (or reflected power or return loss), as a function of the VSWR, as well as the reflection coefficient. \( RL = 20 \log \frac{1 + R}{1 - R} \) with \( R = \frac{V_{\text{max}}}{V_{\text{min}}} \).
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10 PRINT "Adaptation"
20 INPUT "VSWR=",R
30 X=20*LOG(1+R)/(1-R)
40 Y=.01*INT(100*X)
50 PRINT "Return Loss =";Y
60 Z=(R-1)/(R+1)
70 A=.01*INT(100*Z)
80 PRINT "COEF REFLEXION=":A
90 GOTO 20
100 END

Test: VSWR = 1.8  return loss = 10.88  reflection coef. = 0.28

Program P: unloaded Q.

Gives the $Q_0$ of a coaxial cavity as a function of the measurements of loss and passing band at 3 dB.

$$Q_0 = Q_u = \frac{F}{\Delta F} \frac{1}{1 - 10^{-L/20}}$$

10 PRINT “unloaded Q”
20 INPUT “F(MHz)”,F
30 INPUT “BW(kHz)”,B
40 INPUT “IL(dB)”,L
50 X=10*(-L/20)
60 Y=1000*F/B/(1-X)
70 PRINT “Q_u=“;INT Y
80 GOTO 20
90 END
Test: F=410 MHz   BW=160 kHz   IL=2.2 dB   Q₀= 12459

Program P2: Qₖ optimal.

Gives the optimal value of the band at 3 dB for cavities, as a function of their Q₀, for a given separation between the channels, as well as the projected loss.

\[ Qₖ^3 + 2KQₖ = KQ₀ \quad \text{with} \quad K = \left( \frac{F}{2\Delta F} \right)^2 \]

10 PRINT “Qₖ optimal”
20 INPUT “Q₀”,A
30 INPUT “F(MHz)”,F
40 INPUT “kHz separation”,B
41 SET F 0
50 K=(1000*F/(2*B))^2
60 Y=K*A
70 FOR N=0 TO 10000 STEP 1000
80 X=(N^3)+(2*K*N)
90 IF X≥Y THEN 110
100 NEXT N
110 FOR O=N TO Ø STEP –100
120 X=(O^3)+(2*K*O)
130 IF X≤Y THEN 150
140 NEXT O
150 FOR P=O TO 10000 STEP 10
160 X=(P^3)+(2*K*P)
170 IF X≥Y THEN 190
180 NEXT P
190 FOR Q=P TO Ø STEP –1
200 X=(Q^3)+(2*K*Q)
210 IF X≤Y THEN 230
220 NEXT Q
230 C=0.9*Q
240 D=INT(1000*F/C)
250 PRINT “QL=”;C;”B3dB=”;D;”kHz”
251 SET F1
260 E=C/A
270 G=50*(1+E)/(1-E)
280 R=10*LOG(G/50)
290 H=G*C*SQR2/SQRK
300 I=(G↑2+(2*H)↑2)/G
310 J=20*LOG(((100+I)/(50+I))
320 L=(G↑2+(2*H)↑2)/G
330 M=20*LOG(((100+L)/(50+L))
340 S=R+J+M
350 PRINT”IL=”;S;”dB”
360 GOTO 10
370 END

Test: Q0 =14510  F =425 MHz  Separation 150 kHz  Q=2025  B 3dB=209 kHz  IL=2.13 dB

Program P3: Dishal 2 to 7 poles.

See section 6.2. Degree mode. Take, for a given bandwidth, the operational width specified (that of the specifications), without correction. It is considered as being defined at 0.1 dB. The parameters C_{xy} are the interaxial distances between the resonators.

This gives the dimensions of a combline filter as a function of the central frequency F_c, of the operating band and the chosen width, h. e is the distance between the axis of an end resonator and the end wall. L is the total internal length.

10 PRINT”DISHAL 2 to 7 POLES”
11 PRINT"RUN DEG"
20 INPUT"Fc(MHz)"",F
30 INPUT"passing band at 0.1 dB (MHz)"",S
40 INPUT"h(mm)"",H
41 SET F 1
50 E=.37*H
60 PRINT"d opt=";E;"mm"
70 INPUT"d(mm)"",D
71 I=3*10↑5/F
72 R=90*(I-H+D)/I
73 PRINT"θ max=";R;"deg"
80 INPUT"θ₀ deg"",X
90 G=X*3*10↑5/360/F
91 J=.75*H
100 PRINT"l=";G;"mm"
110 A=(.5+X/57.29578/SIN(2*X))*B:F
120 INPUT"number of poles N="",N
130 ON N GOTO 120,140,210,280,380,480,700
140 K=1.2732*A*2.1*.7075/(1+A*2.1*.7075)
150 C=H*((.91*D/H)-LOGK-.048)/1.37
160 PRINT"C₁₂=";C;"mm"
170 PRINT"e=";J;"mm"  \text{2 poles}
180 L=C+(2*J)
190 PRINT"L=";L;"mm"
200 GOTO 20
210 K=1.2732*A*1.43*.6818/(1+A*1.43*.6818)
220 C=H*((.91*D/H)-LOGK-.048)/1.37
230 PRINT"C₁₂=C₂₃=";C;"mm"
240 PRINT"e=";J;"mm"
250 L=2*(C+J)
260 PRINT"L=";L;"mm"
270 GOTO 20
K = 1.2732*A*1.25*0.737/(1+A*1.25*0.737)
C = H*((0.91*D/H)-LOGK-0.048)/1.37
M = 1.2732*A*1.25*0.541/(1+A*1.25*(0.541-0.737))
O = H*((0.91*D/H)-LOGM-0.048)/1.37
PRINT “C12=C34=”;C; “mm”
PRINT “C23=”;O; “mm”
PRINT “e=”;J; “mm”
L = O+2*(C+J)
PRINT “L=”;L; “mm”
GOTO 20

K = 1.2732*A*1.15*.78/(1+A*1.15*.78)
C = H*((.91*D/H)-LOGK-.048)/1.37
M = 1.2732*A*1.15*.54/(1+A*1.15*.54-.78)
O = H*((.91*D/H)-LOGM-.048)/1.37
PRINT “C12=C45=”;C; “mm”
PRINT “C23=C34=”;O; “mm”
PRINT “e=”;J; “mm”
L = 2*(J+C+O)
PRINT “L=”;L; “mm”
GOTO 20

K = 1.2732*A*1.11*.809/(1+A*1.11*.809)
C = H*((.91*D/H)-LOGK-.048)/1.37
M = 1.2732*A*1.11*.55/(1+A*1.11*.55-.809)
P = 1.2732*A*1.11*.518/(1+A*1.11*.518-.55)
Q = H*((0.91*D/H)-LOGP-.048)/1.37
PRINT “C12=C56=”;C; “mm”
PRINT “C23=C45=”;O; “mm”
PRINT “C34=”;Q; “mm”
PRINT “e=”;J; “mm”
L = Q+2*(J+C+O)
PRINT “L=”;L; “mm”
GOTO 20
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600 GOTO 20
700 K=1.2732*A*1.07*.829/(1+A*1.07*.829)
710 C=H((0.91*D/H)-LOGK-.048)/1.37
720 M=1.2732*A*1.07*.56/(1+A*1.07*(.56-.829))
730 O=H*((0.91*D/H)-LOGM-.048)/1.37
740 P=1.2732*A*1.07*.517/(1+A*1.07*(.517-.56)) 7 poles
750 Q=H*((0.91*D/H)-LOGP-.048)/1.37
760 PRINT “C12=67=”;C; “mm”
770 PRINT “C23=56=”;O; “mm”
780 PRINT “C34=C45=”;Q; “mm”
781 PRINT “e=”;J; “mm”
790 L=2*(J+C+O+Q)
800 PRINT “L=”;L; “mm”
810 GOTO 20
820 END

Test: Fc=465 MHz   BW 0.1dB=3.2 MHz   h=40 mm   d opt=14.8 mm   d=14 mm
max=86.4°   θ0=86°   l=154.1 mm   no. of poles=5   C12=C45=44.7 mm   C23=C34=49.4 mm   e=30mm   L=248.3 mm. We can complete the program up to 8 poles, as an exercise, if we have sufficient variables.

Program P4: Q₀ coaxial cavity.

Gives Q₀ as a function of the dimensions: external diameter D, internal diameter d, number of quarter-waves.

10 PRINT “Q₀ coaxial cavity”
11 SET F 0
20 INPUT “F(MHz)=”,F
30 INPUT “D(mm)=”,D
40 INPUT “d(mm)=”,E
41 INPUT “No. 1/4 wave=”,N
50 A=LN(D/E)
60 B=600/F
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```
70  Q=(9.07916*10^6)*.85*A/((B*1000/D)+(B*1000/E)+(8*A/N))/SQRF
80  PRINT “Q_0 with copper=”;Q
90  R=Q*1.0445
100 PRINT “Q_0 with silver=”;R
110 GOTO 10
120 END

Test: F = 418 MHz   D = 180 mm   d = 52 mm   3 \lambda/4   Q_{0Cu} = 13172   Q_{0Ag} = 13759

Program P5: Rejection of cavity.

Gives the value of the loss at x MHz of the central frequency of a cavity.

10  PRINT “Rejection of cavity”
20  SET F 0
30  INPUT “Separation/F_0(MHz)”,A
40  INPUT “BW at 3 dB(kHz)”,B
50  C=10*LOG(1+(2*A/B*1000)²)
60  PRINT “A_s=”;C;”dB”
70  GOTO 10
80  END

Test: separation = 15MHz   BW_{3dB} = 0.19 MHz   A_s = 44dB

12.2. Varia

Matched attenuators (R_0 = Z_0 = 50 \Omega). For another characteristic impedance R, multiply the values by R/R_0. For 75 \Omega, for example, multiply by 1.5.
```
### Table 1: Attenuation (dB)

<table>
<thead>
<tr>
<th>Attenuation (dB)</th>
<th>$R_1$($\Omega$)</th>
<th>$R_2$($\Omega$)</th>
<th>$R_3$($\Omega$)</th>
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### Table 2: Velocity of propagation of EM waves. ($c_0 = 300,000$ km/s)

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## Conversions of line parameters

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<th>IL(dB) (loss)</th>
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<th>$P_R$ (%) (refl. power)</th>
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Coaxial cables 50 Ω most commonly used. DB = double braids. SR = semi-rigid. Doc Filotex.

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<th>damping (dB/100m)</th>
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Plastic materials usable in HF

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Inductance and capacitors out of market:

1 cm² Teflon glass 16/10 ($\varepsilon_r$=2.1) = 1.16 pF
1 cm² epoxy 16/10 ($\varepsilon_r$= 4.3) = 2.34 pF
Coaxial line 50 Ω: $C = 96.4$ pF/m
$L = 242$ nH/m
Linear inductances (*straps*):

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<th>φ (mm)</th>
<th>L (nH)</th>
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<tr>
<td>10</td>
<td>0.1 mm</td>
<td>150</td>
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Conventional inductances (Henry):

Reel

\[ L = \frac{\pi^2 N^2 D^2}{l} \times 10^{-7} \]

1 circular turn

\[ L = \frac{2\pi D}{10} \left( 2.303 \log \frac{8D}{d} - 2 \right) \]

1 squared turn

\[ L = 0.8a \left( 2.303 \log \frac{2a}{d} - 0.774 \right) \]

*Strap*

\[ L = 0.2l \left( 2.303 \log \frac{4l}{d} - 1 \right) \times 10^{-6} \]

Dielectric rigidity in air (\(T=15^\circ\text{C}, \ p=76\text{cm Hg}\)). Abraham and Villard, 1911.
### Vd (V)

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### Spark length (mm)

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<th>5 cm</th>
<th>points</th>
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<td>6.0</td>
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<td>5.8</td>
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<td>9.3</td>
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Chapter 13

Various Questions and Exploratory Ways

Each profession, each activity, each branch of physics, has its specific problems, its techniques and ideas. Depending on the field of application, innovations come more or less quickly, discoveries are more or less easy, but each engineer has to take care, when daily obligations allow them the time, to improve their knowledge and technologies, to contribute to the progress made. In spite of this being a particular domain where the essential already seems to be known, the small world of high-frequency passive devices also possess its margin of progression, related particularly to the ever new demands of radiocommunications. We will find in what follows a certain number of ideas not implemented yet, but that merely await the opportunity to be so, as well as certain questions with or without answers, but that constantly inspire existential thinking among technicians.

13.1. The coupler without intrinsic loss

A myth, or an eternal hope, the coupler without internal loss has become a recurrent theme for free discussion that sometimes takes place during breaks in the most diverse mediums of HF applications. We can ask ourselves if there is a precise justification for the fact we are condemned to having to sacrifice half of the power of two signals, from the time we have decided to combine them in the same propagation channel. In fact, there are many devices where we can group two carriers without having to pay the price these fatidic 3 dB losses, which are necessarily associated, so it seems, with the couplers of the same name. In the course of the development of microwave links, for example, there was, at a given moment, a saturation of the network, for which it was necessary to find a remedy. We thus conceived a process allowing us to instantly double their capacity without
increasing the number of frequencies. This procedure consists, in a link between two microwave link towers, of passing from one to two sources in each parabolic antenna and feeding each one of them by a polarized plane wave, orthogonal to the other, at the same frequency. For this reason we call it frequency reuse, but it is in fact nothing but a coupling of two signals. The propagation check-up of the link shows that there is no supplementary transmission loss, beyond those we expect in the conventional way: there effectively has been a coupling of two channels without energy dissipation in a load. Another example is given by wave guides, which, we recall, are high-pass devices that do not propagate the TEM mode. We can create a connection between a guide and two others of same section by means of a T, of which one of the characteristics is, as in the preceding case, to bring an orthogonality condition between the separated channels, to the place where this division happens, it being well understood that, on the other hand, the link is perfectly reciprocal, that is to say that it fulfils the double function of adding or division according to the way it is used. The two preceding examples have in common the fact that they involve a guided propagation system (the parabolic antennas constitute a sort of guide because the propagation is made in a beam and consequently there is no dispersion), as well as an orthogonality condition, understood as a condition of isolation between the two signals. Finally, we find ourselves exactly at the same point when we endeavor to analyze the functioning of the hybrid coupler, and we highlight the necessary orthogonality condition of the vectors representing the two signals in order to have insulation between them. The difference with the previous examples therefore seems to be more technological than theoretical: in the case of the hybrid coupler and the related devices, the orthogonality of the vectors representing the two signals comes from the capacitive nature of the coupling between the two lines. The propagation difference between an entry and an exit, which gives rise to a phase difference totally independent of the preceding one, is produced by a certain line length which is equal to a quarter-wave when we desire a symmetrical system. In order for this line to work normally, it must be closed on an adapted load, and this load absorbs, in this case, half of the incident power.

In fact, the more or less great difficulty that we have when we try to understand how a 3 dB coupler works in its adding function comes from the way it is represented. Precisely, it is one that simplifies reasoning and consists, as suggested by the architecture of the coupler, of considering it as a superposition of two dividers. This could seem paradoxical in a first analysis, given it is an adder, but it readily becomes evident that, if each signal undergoes an equipartition as soon as it enters the coupler, there is effectively a division of the power by 2. Figure 13.1 proposes an equivalent diagram based on this observation, that takes into account its functioning in another way than the usual one, and that is well in concordance with the facts. As a function of this interpretation, it thus seems that all ambiguity is
dissipated concerning the intrinsic loss of 3 dB, and that seems inevitable and due to the properties of the lines. Things are not this simple, however.

At the beginning of the 20th Century, a French engineering graduate of the Ecole Polytechnique named Maurice Gandillot, convinced etherist and passionate anti-relativist, author of several works on this subject, proposed the following thought experiment, which we will call “Gandillot’s paradox”: on an electrically isolated laboratory table there are two identical capacitors $C_1$ and $C_2$, of capacity $C$, connected as indicated in Figure 13.2 by a switch B. Another switch A allows $C_1$ to be loaded by means of an external static generator, and then the ensemble $C_1 + C_2$ to be isolated. We load $C_1$ under a voltage $V$, the switch B being open, and then we open A. $C_1$ as well as the ensemble $C_1 + C_2$ now possess a load $Q = CV$ and a potential energy $W_1 = Q^2/2C$. We close B: the load $C_1$ is equally split between $C_1$ and $C_2$, each one now possessing a load $Q/2$. The total energy has become $W_2 = Q^2/4C = W_1/2$. The fact that we split the load of a capacitor into two identical ones has divided by two the associated energy, which constituted Gandillot’s argument to suggest that the missing energy had gone into what could be none other than ether. Even if we are not in agreement with this personal conclusion of his, it is...
necessary to recognize that there is material for explanation there regarding the subject, and it would be desirable to trace the lost energy. We can find in Maxwell a number of theorems that, in simply being interpretations adapted from the principle of energy conservation, give mathematical justifications to the curious result of the experiment, but a mechanical analogy already used by Henri Poincaré in the volume Électricité et optique of his course on mathematical physics will allow us to update the missing part of the manipulation in a much more evident and explicit way, and to clarify that which is invisible in Gandillot’s experiment.

Let us replace the two capacitors, considering that electricity behaves in our problem as a fluid, by two identical containers $R_1$ and $R_2$ (Figure 13.3), open at the top and whose lower openings are connected by a flexible tube fitted with a gate $V$ in its middle. In the same way that we loaded $C_1$, we fill $R_1$ with a liquid up to the level $a$, $V$ being closed. Afterwards we abruptly open $V$: the liquid flows out and, carried by its inertia, will be found completely inside $R_2$ with the same level $a$, then the phenomenon will be reversed again and, if we consider the experiment theoretically, with total absence of viscosity and friction, the fluid will see its mass oscillating indefinitely between the two containers. This analogy leads us to suppose that in the case of the two preceding capacitors, what happens is something similar: supposing there are no resistances between them: the system oscillates. To reach the final state represented by 2 in Figure 13.3, it is necessary that the oscillations be damped by the viscosity of the fluid, or by the presence of a resistance in series with switch $B$ in the case of the electrical experiment. In both cases, the equipartition of potential energy is reached through a damped oscillatory regime, whose damping is made by the transformation of half of the initial energy into heat. We can see very clearly now, thanks to the mechanical analogy, how the loss of 3 dB happens in the electric device described. To transpose this directly to the case of the hybrid coupler is not possible, but it is a line of reasoning that has the advantage of providing simple and intuitive reading, under the condition of constructing an adapted analogy. To go from the static regime to the alternating one is an act of gymnastics that is common in electromagnetism.
What we have just seen is thus not a demonstration: it is an attempt to connect common phenomena at different configurations, multiple set-ups and multiple devices, leading to a better view of a principle that little-by-little comes to light under the combination of electrical phenomena seen from the aspect of the movements of loads that accompany them, and particularly seen from the energetics aspect. All that, despite the temptation, does not permit us to affirm that the loss of the 3 dB is inevitable in the coupling of two signals, but if we keep in mind the structure of an adder seen from the aspect of two superposed dividers, we can better understand, from what has been said, how the separation of an electrical quantity is accompanied by the transformation of half of the associated energy into heat. Where this associated heat goes after is a question of free interpretation.

13.2. Infinite rejection band-pass

![Figure 13.4.](image)

The continual hardening of the specifications relative to the duplexers leads us to search for ways to improve the off-band selectivity of the bandpass filters. Elliptization is one of them, the increasing of the number of poles is another more conventional one, but we can ask ourselves if there might be a theoretical method to rigorously nullify the response at least in a part of the rejection band. Let us examine with this aim in mind the set-up proposed in Figure 13.4. Associated with the bandpass filter PB₁, whose off-band rejection on one side or the other of the passing band we want to improve, are two directive couplers C₁ and C₂ which have the same coupling coefficient of the order of 20 dB. A part of the incident signal is deducted from point A and directed in a circuit that comprises a tunable attenuator, an adjustable phase-difference line by means of an LC circuit, and a two-pole filter that can be helical in order to save space. The small part of the entry signal that passes
through this derivation circuit will have a response curve like that of the two-pole bandpass, but with an amplitude and a phase tunable at will. The principle is to select from the slope of the two-pole filter a portion superposable to the part of the slope of PB₁ (Figure 13.5), which has an arbitrary number of poles, that we want to annihilate and which must already present, for the operation to be possible, a loss largely greater than the 40 dB that represent the sum of the coupling coefficients of the two directive couplers. Theoretically, this amplitude/phase-type compensation works for a single frequency only, but the phase rotation that takes place as a function of the frequency in the slope of one of the filters more or less follows that of the other filter, and the number of tuning parameters (filter, line length, amplitude), must allow an over-attenuation band largely greater than that which is needed, for example, for an elliptization or any method susceptible of creating a simple attenuation pole. The coupler C₂ can just as well be found downstream or upstream of the main filter, and this is a question of what is convenient. The directivities of the couplers must be excellent. The rejection band said to be “infinite” cannot be very large, but the need for exceptional off-band attenuations a priori only concerns low-band duplexers and those with a small duplex separation. The complexity of the arrangement is relative and does not give rise either to a significant cost overrun, or to exaggerated dimensions. We can foresee, on the other hand, a particularly long and delicate tuning. With the action of the corrective signal affecting all frequencies, in places it will increase the level instead of the nullifications. In the passing band, this will translate as an almost invisible ondulation, and elsewhere in the spectrum the increases will not be able to exceed 3 dB.

![Figure 13.5.](image)

13.3. **Helix TX multicoupler**

The production of multicouplers working below 100 MHz requires cavities of dimensions such that it is often necessary to reserve a full rack to the coupling system. Helical resonators can solve this problem of size and thus allow the
integration of the multicoupler in another bay, but at the cost of a decrease in the quality coefficient, which can be prohibitive. However, the gain in space is such, at the elementary resonator, that we can foresee the usage of two per channel in the form of a two-pole filter. What remains is to determine if the performance of such a filter is comparable to that of a large model coaxial cavity.

The methods of projected calculation explored in Chapter 10 will allow us to answer this question. Let us take as reference, to start with, a coaxial cavity of an external diameter of 150 mm, with a 40 mm inner silver-plated tube, working at 75 MHz, which corresponds to a quarter-wave of one meter. Dishal's program gives a predicted $Q_0$ of 4,662. The volume of the cavity is comprised between 12 and 13 dm$^3$. Let us consider now using helices constituted of a tube 10 mm in diameter wound up in around 4 turns with an average diameter of 8 cm. The calculation now gives a $Q_0$ of 2,881 for a single resonator. If we form a narrow-band filter with two of these resonators, the off-band rejection at any given frequency will go from A to 2A, that is to say that the loaded Q doubles. In addition, the flatness in the passing band, although narrow, allows us to obtain a passing loss less than that of a single resonator. According to the tuning value chosen, we see that it will be possible to ultimately obtain a better quality coefficient, be it loaded or unloaded, than with the
large coaxial cavity. We can house the two helices in a parallelepiped of 150 x 300 mm with a depth of 100 mm, which corresponds to a volume of about 4.5 dm³ (Figure 13.6). The numbers speak for themselves: a size three times smaller, a better quality coefficient, a manufacturing cost that is \textit{a priori} lower, and a disposition favorable to a rack installation with access to the connectors by the front face; there are only advantages to the system at first sight, and certain existing networks could have already benefited from this equipment. Yet, we do not find it in any catalogs or practical applications, which shows that there are still things to be done in the field, and this is fine. Let us add, regarding this example, that we can very easily deport the frequency tuning to the same face as the connectors. For this it is necessary to replace the internal coupling tuning (passing-band tuning) with a return rod terminated by a cogwheel that leads two other wheels fixed onto the end of the frequency tuning axes. We then proceed with the band tuning by means of another device, and the chapters dedicated to filters and cavities should provide the central concepts well enough so that this will not constitute a problem for creative designers.

All of these advantages must in principle be moderated by experimental evidence related to the propagation mode of the helix: the latter radiates little, or in any case less than the coaxial quarter-wave central conductor, and it results that the frequency tunings, on the one hand, and the coupling tuning between resonators from the other are markedly less efficient. In a wide-band filter, as for the FM band, the helices almost touch and two coupling tunings are necessary, whereas in an ordinary combline there is only one. We thus know that the tuning range will be limited, which is circumvented by the creation of sub-bands.

On the other hand, the two-pole helical filter offers the possibility of producing it with half-waves, which cannot be done in coaxial technology unless we abandon quarter-wave frequency tuning. The helical half-wave is two times longer than the quarter-wave, but the quality coefficient remains the same, and the fact of being kept directly at the mass at its two ends grants it a much greater robustness and mechanical hold, whilst allowing frequency adjustment by plunger.

13.4. Conclusion

There are still so many technological pathways to explore that we can only give a few examples here. Those mentioned above are for some the extension of actions effectively taken, but the uncertainty of industrial life has prevented from being pursued and brought to fruition. The links that exist between extended HF and “true” microwaves allow interesting adaptations, and we should not hesitate, when within one of the specialties, to consult our neighbors: there are always, somewhere, good ideas to be taken up.
The mechanical models that were proposed in the text are another pathway where investigations are particularly fruitful. The most interesting are those where the propagation medium is taken into account, because these are the only ones which are complete and could lead to a truly efficient mechanical analogy. This is a formidable means of investigation which allows, by continuously moving back-and-forth between electromagnetism and the general theory of vibrations, to be able to comprehend a phenomenon from two very different points of view but actually submitted to the same logic. A diversion of results between the two is a sign of error in the reasoning, which is then easier to correct. Moreover, and this is maybe the greatest advantage, a problem of apparent complexity in a certain formulation can be revealed very simply when seen by another angle. Unfortunately, the systematic application of the analog tools is left to personal initiative, it is not at all evident, and its usage will only really be efficient when an education decision-maker decides to make it a subject to be included in a physics course.

Precisely in relation to physics, we will have perhaps remarked the sum of knowledge that the conception of a simple resonant cavity requires: besides the indispensable purely electromagnetic bases, which are already of a certain level, let us say at least that of a specialized second-year university cycle, a truly complete study makes use of practically all branches of physics: mechanics, heat, chemistry, etc. Nevertheless, of the 800 or so pages of Feynman’s Electromagnetism, part of his physics course, he dedicates only 18 of them to it. Moreover, we only find the general principles in it, which would be a poor resource to a technician in charge of a study. This is very little, but it is necessary to recognize that the industrial applications are almost exclusively linked to the development of radio-communications systems, and still it is necessary to emphasize that we find it nowhere other than in the infrastructures of networks, that is to say, in aspects completely ignored by the general public, which may explain in part the neglecting of a specialty that deserves more attention. Perhaps this work will contribute to making it better known.
Bibliography

The works proposed are presented in three categories: firstly those considered as indispensable in the considered domain and called “basics”, next the monographs exhaustively dealing with a well-defined related subject, and finally the documents regarded as general culture in relation to the text of the book. The titles preceded by an asterisk are reputed to be out-of-print. Another important source of documentation is the collection of specialized reviews on microwaves, and particularly the IEEE Transactions on MTT, of which some articles are explicitly cited in the text.

1. The basics

2. Monographs
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