SECTION I—GENERAL INFORMATION

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SECTION I—GENERAL INFORMATION

INTRODUCTION

PURPOSE AND SCOPE OF MANUAL

1-1. The purpose of this manual is to provide the design and developmental engineer of military electronic equipment with a reference handbook containing background material, circuit theory, and components data related to the application of piezoelectric crystals for the control of radio frequencies.

1-2. This manual is composed of the following sections:

I. GENERAL INFORMATION
II. CRYSTAL UNITS
III. CRYSTAL HOLDERS
IV. CRYSTAL OVENS
V. APPENDIXES

1-3. Section I contains a brief historical account of the discovery of the piezoelectric effect and of the application of crystal resonators as frequency-control devices, discussions covering the theory and physical properties of piezoelectric crystals, descriptions and performance characteristics of the more important quartz crystal elements, general discussions of the various crystal-unit fabrication processes and types of mounting, detailed discussions of the equivalent electrical parameters and performance characteristics of crystal units, and comprehensive qualitative and mathematical analyses of the various types of crystal oscillators, summarized with recommended design procedures.

1-4. Sections II, III, and IV provide the technical and logistical data, and information concerning the application of the crystal units, crystal holders, and crystal ovens currently recommended for use in equipments of new design.

1-5. The Appendixes contain the acknowledgments; a bibliography; a list of manufacturers associated with the piezoelectric crystal industry; a list of related U. S. Government specifications, standards, and publications; a table of definitions for the abbreviations and symbols used in the Handbook; conversion charts; and an alphabetical index.

CONTROL OF RADIO FREQUENCY

1-6. The greatly increased demand for military radio channels, with the consequent crowding of the radio-frequency spectrum, is, in the final analysis, a problem for the design engineer of frequency-control circuits. The problem is essentially one of providing a maximum frequency stability of the carrier at the transmitting station, and a maximum rejection of all but the desired channel at the receiving station. In each instance optimum results are obtained by the use of electromechanical resonators — maximum carrier stability is achieved by the use of crystal master oscillators, and maximum receiver selectivity is achieved by the use of crystal heterodyne oscillators and crystal band-pass filters.

1-7. The design of a constant-frequency generator has been an ideal of radio engineers almost from the beginning of radio science. Although many purely electrical oscillators have been devised which closely approach the ideal, none surpass the performance of the high-quality circuits employing mechanical oscillators. Temperature-controlled oscillators having a sonic-frequency tuning fork as the frequency controlling element and followed by a number of frequency multiplying stages were the first of the radio-frequency generators employing the high precision of mechanical control. The cumbersomeness and expense of the many multiplier stages, however, have made the tuning fork oscillators impracticable insofar as the control of any but sonic frequencies are concerned. Today, precision control of radio frequencies has been made possible through the development of piezoelectric resonators, where the frequency-controlling elements, usually quartz plates, have normal vibrations in the radio-frequency range.

THE PIEZOELECTRIC EFFECT

1-8. The word *piezoelectricity* (the first two syllables are pronounced pie-ee′) means "pressure-electricity," the prefix *piezo-* being derived from the Greek word *piezein,* meaning "to press."

1-9. "Piezoelectricity" was first suggested in 1881 by Hankel as a name for the phenomenon by which certain crystals exhibit electrical polarity when subjected to mechanical pressure.

1-10. That such a phenomenon probably existed seems to have been suggested first by Coulomb in the latter part of the 18th century. His suggestions prompted Haüy, and later A. C. Bequerel, into undertaking a series of experiments to see if electric effects could be produced purely by mechanical pressure. Although both Haüy and Bequerel reported positive results, there is some doubt as to whether these were not due to contact potentials rather than to piezoelectric properties of the substances investigated—particularly since electrical polarities were reported in crystals that are now known to be non-piezoelectric.

1-11. It is to the Curie brothers, Jacques and Pierre, that the honor goes for having been the first (in 1880) to verify the existence of the piezoelectric effect. (For the initial report of their discovery, see paragraph 1-56.)

1-12. The Curie brothers tested a number of crystals by cutting them into small plates that were then fitted with tin-foil electrodes for connection to an electrometer. When subjected to mechanical pressure, several of the crystals caused the leaves of the electrometer to be deflected. Among those crystals showing electrical polarities were quartz, tourmaline, Rochelle salt, and cane sugar. In the year following these experiments, a prediction by Lippmann that the effect would prove reversible prompted the Curies to further investigations. The results verified Lippmann's prediction by revealing that the application of electric potentials across a piezoelectric crystal would cause deformations in the crystal which would change in sign with a change in electric polarity. Furthermore, it was found that the piezoelectric constant of proportionality between the electrical and mechanical variables was the same for both the *direct* (pressure-to-electric) and the *converse* effects. In other words, the same polarization at the surface of the electrodes that results from a particular deformation of the crystal can, in turn, if applied from an external source, produce the deformation.

1-13. It should be mentioned that the piezoelectric effect, which occurs only in certain asymmetrical crystals, is not to be confused with *electrostricture,* a property common to all dielectrics. Although electrostricture is a deformation of a dielectric produced by electric stress, it is unlike the converse piezoelectric effect in that its magnitude varies, not linearly with the electric field, but with the square of the field, and is unaffected by a change in the applied polarity. Electrostricture is the type of deformation a capacitor undergoes on being charged. In piezoelectric crystals this effect is normally small compared with the piezoelectric properties.

DEVELOPMENT OF PIEZOELECTRIC DEVICES

1-14. From the time of its discovery until World War I, the piezoelectric effect found few practical uses. Those applications it did find appeared in the form of occasional laboratory devices for measuring pressure or electric charges. For the most part, however, little attention was attracted to piezoelectricity outside the crystallographer's study. Nevertheless, during this time considerable theoretical progress was made, due chiefly to the efforts of Lord Kelvin, Duhem, Pockels, and Woldmar Voigt. Voigt's comprehensive *Lehrbuch der Kristallphysik,* published in 1910, is still considered the reference authority on the mathematical relationships among crystal variables.

1-15. It was after the outbreak of World War I before serious attention was given to the practical application of piezoelectric crystals. During the war Professor Paul Langevin of France initiated experiments with the use of quartz crystal plates as underwater detectors and transmitters of acoustic waves. Although Langevin's immediate purpose was to develop a submarine detecting device, his research became of vital importance to many other developments. Not only did it attract the applied sciences to the possibilities of piezoelectric crystals, but also it initiated the modern science of ultrasonics.

1-16. The detecting apparatus that Langevin eventually devised employed quartz "sandwiches" which were coupled electrically to vacuum-tube circuits, and could be exposed under water where they would vibrate at the frequency of an applied voltage, or at the frequency of an incident acoustic wave. The first function was employed to emit ultrasonic waves, and the second function to receive and reconvert the echo into electrical energy for detection.

1-17. At the same time that Langevin was experimenting with quartz as a supersonic emitter and detector, Dr. A. M. Nicolson, at Bell Telephone Laboratories, was independently investigating the use of Rochelle salt to perform the same functions at sonic frequencies. Indeed, his first application for a patent on a number of piezoelectric acoustic devices, April 1918, preceded by five months Langevin's initial application for a French patent. Employing Rochelle salt instead of quartz, because of its greater piezoelectric sensitivity, Nicolson constructed a number of microphones, loudspeakers, phonograph pickups, and the like. Among the circuits included in his 1918 patent application, was one that later proved of particular interest — an oscillator employing a Rochelle salt crystal as shown in figure 1-1. With this exception, all the early applications of the piezoelectric crystal involved its use as a simple electromechanical *transducer*. That is, it was used either to transform mechanical energy in one system to electrical energy in another, or vice versa. Nicolson's oscillator was a distinct innovation in that it employed a piezoelectric crystal as a transformer of electrical energy to mechanical energy and back to electrical energy.

1-18. When Nicolson devised his oscillator, none of the possible functions of a piezoelectric vibrator had previously been investigated or discussed. His patent application offered no description of the crystal's function, although presumably the crystal performed in some way to transfer part of the plate circuit energy to the grid circuit. Evidence that the normal vibrations of the crystal actually controlled the frequency seems to have existed, but no mention was made of this fact. The circuit, however, embodies the combined principles of coupler, filter, and resonator. Obviously the crystal acts as a coupler between the plate and grid circuits; and, inasmuch as the crystal may block the feedback of all plate energy except that at the frequency of the crystal's normal mode of vibration, the crystal may be imagined to perform the function of a filter, even though the over-all operation is that of an oscillator. Finally, if the plate tap is connected at the bottom of the coil, so that the only feedback is through the plate-to-grid capacitance of the vacuum tube, the crystal may function as a conventional resonator, controlling the frequency as would a tuned grid tank circuit—the complete vacuum-tube circuit being the equivalent of a tuned-plate, tuned-grid oscillator. Thus, to Dr. Nicolson belongs the honor of being the first to employ the piezoelectric crystal purely as a circuit element, in all its principal circuit functions.

1-19. Although Nicolson was the father of the piezoelectric crystal circuit, Professor Walter G. Cady, of Wesleyan University, was its greatest prophet. In 1918 during a series of experiments being conducted to investigate the use of Rochelle salt plates for underwater signaling, Dr. Cady became interested in the electromechanical behavior of crystals vibrating in their normal modes. Out of the resonant properties that he discovered, he came to visualize the great possibilities that the piezoelectric crystal afforded as a resonator of high stability. After experimenting with several circuits, including the first quartz-controlled oscillator, Dr. Cady, in January 1920, not aware that Dr. Nicolson considered his oscillator controlled by the resonance of its crystal, submitted a patent application for the piezoelectric resonator, in which he reported its possibilities as a frequency standard, filter, and coupler, and described the principles

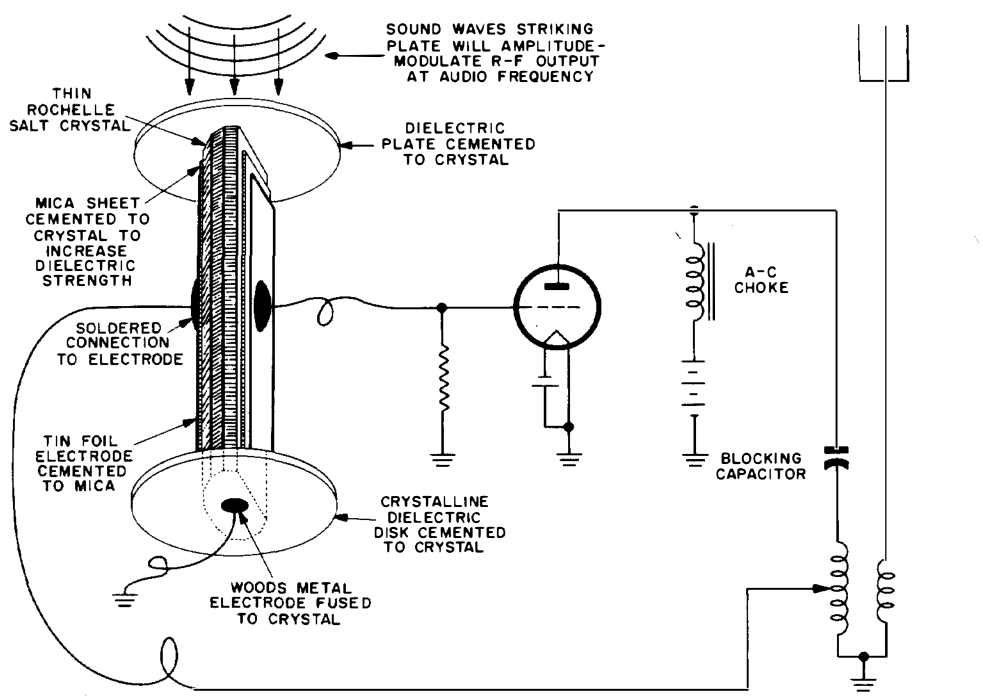


Figure 1-1. The first crystal- (Rochelle salt) controlled oscillator. Invented by A. M. Nicolson, 1918

of its operation. Although subsequent litigation concerning the discovery of the piezoelectric-resonator principle was decided in Dr. Nicolson's favor, it is distinctly to Dr. Cady's credit that he was the first to fully grasp the import of the piezoelectric resonator and to publish a public report of its principles and possibilities. His early pioneering in the field and his many later contributions have made Dr. Cady the American dean of piezoelectricity.

1-20. It soon became apparent that quartz crystals were the most stable and practical for use as resonators. Many investigators were attracted to the field, and progress was made both in the design and theory of crystal circuits. Professor G. W. Pierce of Harvard showed that quartz crystal oscillators could be constructed with a single amplifier stage, as Nicolson had already done using Rochelle salt. This marked a considerable improvement over Cady's oscillators, which had consisted of two or more vacuum-tube stages. Of particular note was the analysis by K. S. Van Dyke, in 1925, of the electrodynamic characteristics of a crystal resonator in terms of an equivalent electrical network; for the first time a way was opened to an understanding of the crystal resonator. In 1928, E. M. Terry showed that the frequency of a crystal oscillator was not entirely controlled by the crystal characteristics, but to a small degree was also dependent upon the other circuit constants. F. B. Llewellyn, in 1931, presented a classic analysis of oscillators showing the circuit impedance relationships that are necessary if the frequency is to be independent of variations in the voltage supply and vacuum-tube characteristics. Although the subject matter of this treatise deals with oscillators in general, the principles are applicable to the design of crystal oscillators, if the electrical parameters of the crystal are known.

1-21. The tuned-circuit oscillators of the early transmitters normally operated with heavy and variable loads. Many of the oscillators operated directly into an antenna, and in broadcast transmitters, modulation was performed in the oscillator stage. This resulted in considerable frequency instability, and broadcast reception was often unintelligible because of the frequency difference in radio waves arriving by different paths. It was in the determination of the cause and the correction of such interference that Messrs. R. Bown, D. K. Martin, and R. K. Potter of the Research and Development Department of the American Telephone and Telegraph Company recommended the use of lightly loaded crystal-controlled oscillators followed by amplifiers. Under their supervision, Station WEAF in New York, in 1926, became the first crystal-controlled broadcasting station.

1-22. The principal factor limiting the stability of the early quartz oscillator was the relatively large frequency-temperature coefficient of the crystal, which allowed small changes in the ambient temperature to cause excessive changes in the resonant frequency. The immediate method of obtaining stability, of course, was to mount the crystal in an oven where the temperature could be controlled thermostatically. However, to decrease the temperature coefficient of the crystal, itself, also became the goal of a number of researchers. Because some quartz plates exhibited positive temperature coefficients, whereas others exhibited negative coefficients, according to the orientation of the plate with respect to the axes of the mother crystal, the possibility arose that there should be some shape or median angle of cut which would have a zero coefficient. The first empiricists to enter the field were E. Giebe and A. Scheibe in Germany. In the United States, Mr. W. A. Marrison of Bell Telephone Laboratories turned his attention to the problem of achieving the maximum precision possible in frequency control, and by 1929 had perfected a 100-kc frequency standard using a doughnut-shaped crystal (originally pioneered by Giebe) with a nearly zero temperature coefficient. This success encouraged the Bell Laboratories research staff to launch a concerted investigation into all phases of quartz crystal physics. Out of this program have arisen most of the principal advances in the design and production of quartz crystal units in the United States; although the early pioneering of S. A. Bokovoy and C. F. Baldwin at RCA has also been of notable significance.

1-23. Originally only the Curie, or X-cut, quartz plate was used—a plate in which the thickness dimension is parallel to the crystal's X axis. Later the Y cut, where the thickness dimension is parallel to a Y axis, developed by E. D. Tillyer of the American Optical Co., began to compete with the Curie cut as the frequency-control element in commercial oscillators. By 1934, Messrs. F. R. Lack, G. W. Williard, and I. E. Fair of Bell Telephone Laboratories announced the discovery and development of two types of plates, called the *AT* and *BT cuts,* with such small temperature coefficients that they could operate stably under normal conditions without the use of temperature-controlled ovens. Concurrently, Bokovoy and Baldwin at RCA were experimenting with a series of crystals that they named the V cut, and their work, although of a less rigorous theoretical approach, substantially paralleled much of the research that was done at

Bell Laboratories. In 1937, Messrs. G. W. Williard and S. C. Hight announced the development of the CT, DT, ET, and FT cuts; and by 1940 Mr. W. P. Mason had discovered the GT cut, the most stable resonator ever devised. The time-keeping standards at both the Greenwich Observatory and the U. S. Bureau of Standards now use this crystal. Where other cuts exhibit a zero temperature coefficient only at certain temperatures, the GT cut has almost a zero temperature coefficient over a range of 100°C. Besides the cuts discussed above, a number of others have been investigated which have proved particularly applicable for special uses. Among these are the AC, BC, MT, NT, 5-degree X, and the —18-degree X cuts.

1-24. Paralleling the development of the new crystal cuts were the improvements made in the design of crystal holders. The early holders provided no means of "clamping" a crystal, for they were designed originally to accommodate X-cut plates whose favored modes of vibration required that the edges be free to move. Since the crystal in such a holder will slide about if used in equipment subject to mechanical vibrations, a method of clamping was needed before the crystal could be used in vehicular or airborne radio sets. Mr. G. M. Thurston of Bell Telephone Laboratories was led to the solution of this problem when he discovered that a crystal would not be restricted if clamped only at the mechanical nodes of its normal vibrations. The exact positions of these points, where the standing-wave amplitude is zero, depend, of course, on the particular mode of vibration. The low-frequency —18-degree X-cut crystal, for instance, can be held by knife-edged clamps running along its center, whereas AT- and BT-cut crystals can be clamped at their corners. Cantilever and wire supports which resonate at the crystal frequency have been devised for holding crystals at their centers. Although the mounting of crystals requires a far more exacting technique than formerly, the crystal holder today provides support and protection sufficient to insure high performance stability, even under the severe conditions of vibration that exist in military aircraft and tanks.

1-25. Unfortunately, the extremely critical nature of the design and production of crystal units has made it impracticable for manufacturers to mass-produce units with such exactitude that all the equivalent electrical parameters are standardized with an accuracy comparable to that now achieved in the case of vacuum tubes or other circuit components. However, definite progress has been made in this direction, and, if the need warrants the additional cost, reasonably exact characteristics may be obtained. For several years, each crystal unit had to be tested in a duplicate of the actual circuit in which it was to be used. This procedure was disadvantageous from the points of view of both the radio design engineer and the crystal manufacturer. On the one hand, the radio engineer, knowing little more than the nominal frequency of the crystal unit to be installed in his circuit, could not achieve that degree of perfection in oscillator design which was otherwise theoretically possible. On the other hand, the task of making a given oscillator perform correctly effectively became the responsibility of the crystal manufacturer, since it was necessary for him to fit each crystal unit by trial and error to the particular circuit for which it was intended. In recent years this cut-and-try procedure has been alleviated considerably by the development of standard test sets and by improvements in production techniques that permit more critical specifications. It is hoped that this handbook, by providing a more comprehensive description of the technical characteristics of the crystal units recommended for new design, will contribute in removing the limitations that too often in the past have forced the practical design engineer to approach his crystal circuits philosophically, rather than scientifically.

PHYSICAL CHARACTERISTICS OF PIEZOELECTRIC CRYSTALS

DESCRIPTIONS OF USEFUL PIEZOELECTRIC CRYSTALS

1-26. The piezoelectric effect is a property of a non-conducting solid having a crystal lattice that lacks a center of symmetry. Of the 32 classes of symmetry in crystals, 20 are theoretically piezoelectric, and the actual crystals which have been found in this category are numbered in the low hundreds.

1-27. Until the time of World War II only three crystals were commercially employed for their piezoelectric properties — quartz, Rochelle salt, and tourmaline. Today, the number is being increased by the development and application of synthetic crystals. Of these, the principal ones used in frequency selective circuits are ethylene diamine tartrate (EDT), dipotassium tartrate (DKT), and ammonium dihydrogen phosphate (ADP). See figure 1-2.

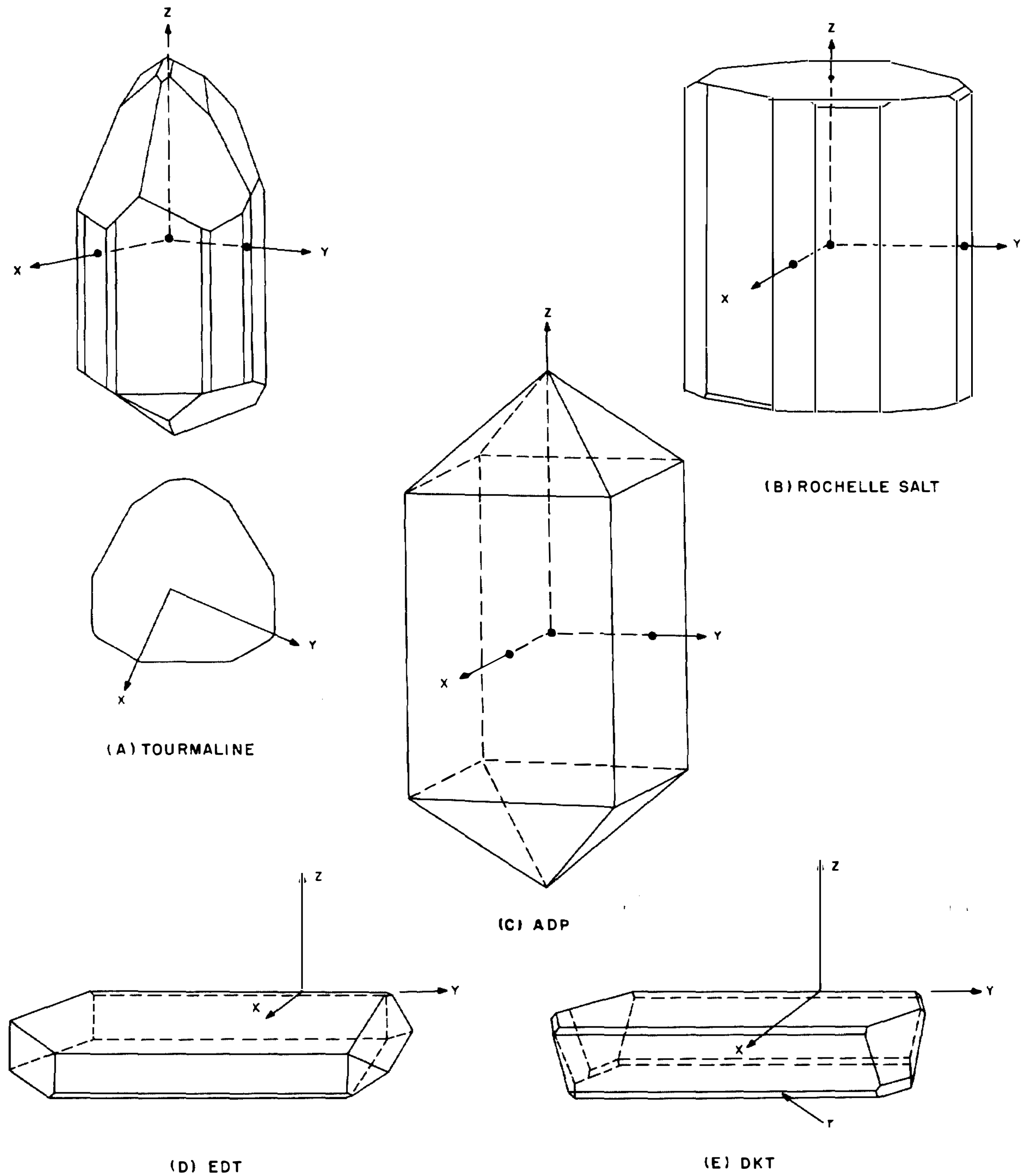


Figure 1-2. Commercially used piezoelectric crystals other than quartz

1-28. Piezoelectricity is still in its infancy, and until more data have been collected and coordinated into a comprehensive atomic theory of the phenomenon, the chemist will have few clues to direct his search for a crystal having the maximum possible piezoelectric effect.

Tourmaline

1-29. Tourmaline is a semiprecious stone which at one time was called the "Ceylon Magnet." This title seems to have been given it by early 18th century traders who introduced the stone to Europe, with the story of its strange magnetic property. If placed in hot ashes, tourmaline behaves as if it were electrified—first attracting ashes and then throwing them off. This is the phenomenon of pyroelectricity, closely associated with piezoelectricity, and was possibly the first electrical effect, other than lightning and St. Elmo's fire, ever to be noticed by man. According to the theory proposed by Lord Kelvin, the pyroelectric effect of tourmaline is due to a permanently polarized lattice in the crystal, so that when heated, an unneutralized increase in the dipole moment occurs, proportional to the change in temperature and the coefficient of expansion. It was this pyroelectric theory of permanently polarized crystals that eventually prompted the Curie brothers to test for the piezoelectric effect.

1-30. Tourmaline is unsuitable for wide commercial use because of its expense and the scarcity in the number of large-sized natural crystals. Also, the temperature coefficients are negative for all tested modes of vibrations, which fact rules out the possibility of zero-coefficient cuts.

1-31. Tourmaline does have the advantage of durability and a large thickness-frequency coefficient, so that for a given frequency it permits a more rugged crystal unit than quartz. For this reason it is sometimes used for the control of very high frequencies. However, the chief piezoelectric application of tourmaline is in the measuring of hydrostatic pressures.

Rochelle Salt

1-32. Rochelle salt ($NaKC_4H_4O_6$–$4H_2O$) is sodium potassium tartrate with four molecules of water of crystallization. The crystals are grown commercially by seeding saturated solutions of the salt and decreasing the temperature of the solutions a few tenths of a degree per day. They were first synthesized in 1672 by Pierre Seignette, an apothecary of La Rochelle, France, and until the time of Nicolson's inventions the salt was used primarily for its medicinal value. Its exceptionally great piezoelectric effect—a blow with a hammer can generate as much as five thousand volts—has made Rochelle salt the principal crystal for use as a transducer in acoustic devices, such as microphones, loudspeakers, pickups, hearing aids, and the like. As a stable resonator it is far inferior to quartz, not only because of a greater sensitivity to temperature variations, but also because of its tendency to disintegrate during extremes of humidity. If the ambient humidity drops below 35 per cent at room temperature, the water of crystallization will begin to evaporate, leaving a dehydrated powder on the crystal surface. Should the humidity rise above 85 per cent at room temperature, the salt will absorb moisture and begin to dissolve. For these reasons a Rochelle salt crystal should be mounted in a hermetically sealed container, or, if this is not possible, at least coated with wax. In the case of the former, if powders of both the crystalline and dehydrated forms of Rochelle salt are also enclosed within the sealed chamber, the humidity of the chamber will automatically increase or decrease with corresponding changes of temperature, and a stable balance between the crystal and chamber vapor pressures will be maintained. However, at a temperature of 55°C (130°F) the crystal, which is a double salt of tartaric acid, breaks down into sodium tartrate, potassium tartrate, and water. The solution formed will remain a viscous liquid for some time if super-cooled, and, as such, makes an effective glue for binding together plates of the crystal.

1-33. Although Rochelle salt, between the temperatures of −18°C and +24°C, has a greater piezoelectric effect than any other crystal, it seems that eventually it will be replaced by other synthetic crystals, in particular, ADP ($NH_4H_2PO_4$), which requires no water of crystallization. Nevertheless, as an electromechanical transducer, Rochelle salt is still the most widely used of the piezoelectric crystals.

ADP

1-34. ADP ($NH_4H_2PO_4$), ammonium dihydrogen phosphate, was discovered and used during World War II as a substitute for Rochelle salt in underwater sound transducers. Like Rochelle salt, ADP crystals can be grown commercially; but unlike Rochelle salt, it requires no water of crystallization, and hence has no dehydration limitations, being able to stand temperatures up to 100°C (212°F). Also, ADP is more durable mechanically than Rochelle salt.

1-35. Although the crystal's principal application has been in submarine-detecting apparatus, its greater stability suggests the probability that it will eventually replace Rochelle salt as the principal transducer in other sonic devices.

EDT

1-36. EDT ($C_6H_{14}N_2O_6$), ethylene diamine tartrate, was discovered and developed during World War II as a substitute for quartz in low-frequency filter units. Quartz crystals at this time were in such great demand for the frequency control of military communication equipment, that a shortage developed in the supply of large-sized natural crystals which were needed for cutting filter plates of 1½ to 2 inches in length. This shortage was acutely felt in the telephone industry, where there exists the chief demand for such plates for use in the band-pass filters of carrier systems. The discovery of EDT was the solution to this problem, for this crystal can be grown to any size desired, and it has the chemical stability (no water of crystallization), low mechanical loss, zero temperature coefficient, and small aging effects that make it a suitable substitute for quartz.

1-37. EDT is not as rugged mechanically, nor does it have quite as high a Q as quartz—although the EDT crystal units operating as filter elements in the 20- to 180-kc range do have Q's in the neighborhood of 30,000. Moreover, for use as the frequency-control element in high-frequency oscillators, EDT is inferior to quartz because of its greater sensitivity to temperature changes. Even though high-frequency modes of vibration have been found with zero temperature coefficients, the temperature shift to either side of the optimum value must be kept approximately one-fifth that for a comparable quartz plate (BT cut, for example) in order to maintain the same frequency tolerance. Where only a minimum of temperature control might be needed for quartz, EDT will require fairly accurate control. Because of these disadvantages, EDT does not threaten at this time to replace quartz in high-frequency oscillators, but it does have promising possibilities for use in oscillators of the frequency-modulated type. Here, EDT plates have the advantage of a relatively wide gap between their resonant and antiresonant frequencies, thus permitting a large percentage swing of the oscillator frequency. If temperature-controlled, the EDT crystal can thus give crystal stability to a frequency-modulated transmitter.

DKT

1-38. DKT ($K_2C_4H_4O_6$–$\frac{1}{2}H_2O$), dipotassium tartrate, is another synthetic crystal which was investigated at Bell Telephone Laboratories during World War II. The DKT molecule is similar chemically to that of Rochelle salt except that the sodium atom has been replaced by another potassium atom. The crystal, however, differs from Rochelle salt in that it contains only one molecule of water for each two DKT molecules, as compared with a water-to-salt molecular ratio of four-to-one in the Rochelle salt crystal, and it exhibits no tendency to dehydrate below 80°C (176°F). Also, the piezoelectric characteristics of DKT are less like those of Rochelle salt than of quartz. Indeed, in the lower-frequency filter circuits, DKT seems as promising as EDT as a substitute for quartz.

1-39. As compared with EDT, DKT has the advantage of better temperature-frequency characteristics. Zero temperature coefficients are possible where the frequency deviation on either side of the zero point is only one-third that for EDT. However, DKT crystals are more difficult to grow than the EDT crystals, and primarily for this reason the development of a small EDT industry has already been established, whereas the DKT crystals are still in the laboratory stage.

Raw Quartz

1-40. Quartz is silicon dioxide (SiO_2) crystallized in hard, glass-like, six-sided prisms. The normal crystal structure is called *alpha quartz*; if the temperature is raised above 573°C (1063°F) most of the piezoelectric property is lost with a crystal transformation to *beta quartz*. At 1750°C (3182°F) the crystal structure is permanently lost, and the melted quartz assumes the fused amorphous form of silica. The density of alpha quartz at 20°C (68°F) is 2.649 grams per cubic centimeter. The hardness of quartz is rated at 7 on Mohs' scale—a greater hardness than glass or soft steel, but less than hard steel.

1-41. Silicon dioxide is believed to constitute approximately one-tenth of the earth's crust. It occurs in many crystalline forms such as quartz, flint, chalcedony, agate, onyx, etc., and in the fused amorphous state of silica, called "quartz glass." Although quartz is an abundant mineral—sand and sandstone consist largely of quartz granules—large crystals of good quality are to be found in only a few areas. The chief source of supply has been Brazil, although large deposits of lower quality are also to be found in Madagascar and in the United States. Progress has been made in growing quartz crystals artificially. Such crystals are now commercially available, although this quartz source is still primarily in the developmental stage. The tremendous pressures required and the slow rate of growth have, until very recently, prevented quartz manufacture from being commercially feasible. Advances are now being made in growing imperfection-free quartz stones having major dimensions so oriented relative to the principal crystal axes that a desired type of quartz cut can be obtained with minimum waste. The future possibilities of quartz manufacture appear quite promising.

1-42. The large quartz crystals of geological origin are the products of long ages of growth under great pressure. The growing crystal assumes the shape of a hexagonal prism with each end pyramiding to a point. The prismatic faces are designated as m faces, see figure 1-4, and adjacent m faces always intersect at angles of 120 degrees. The opposite m faces of the prism are always parallel, but are rarely of the same dimensions. These faces are not perfectly planar, but are streaked with small horizontal growth lines, or striae. Parallel to the growth lines are the bases of the six end faces —three r and three z faces—which form a hexagonal pyramid, but with only the r faces meeting at the apex. The end faces are quite smooth, with the r, or major, faces usually appearing more polished than the z, or minor, faces. Figure 1-3 shows a mother crystal with one of the pyramidal ends missing. Complete crystals are rarely found except in very small sizes. More likely both pyramidal ends will be missing, and frequently crystals are found with all the natural faces broken or eroded away. The largest quartz crystal that has been recorded was found in Brazil. It is described as a crystal of smoky quartz, 7 ft 2 in. long, 11 ft 2 in. in circumference, and weighing more than 5 tons.

Figure 1-3. Raw quartz stone

1-43. Quartz is enantiomorphous—that is, it occurs in both right-handed and left-handed forms, which are mirror images of each other. The enantiomorphic faces of two ideal alpha-quartz crystals are represented in figure 1-4. The left-handed and right-handed forms are indicated by the direction in which the small upper x and s faces appear to be pointing. Note that this rule is valid regardless of which end of the crystal is turned up. However, the x and s faces are rarely found, so that the handedness of a crystal is usually determined by noting the optical effects when polarized light is passed through the crystal parallel to the optic (lengthwise) axis.

IMPERFECTIONS IN QUARTZ

1-44. Pure quartz of structural perfection is a transparent, colorless crystal—such that the early Greek physicists believed it to be a perfected form of ice. Through the centuries quartz has been cut and ground into many ornaments, and was mystically respected in the ancient art of crystal gazing.

1-45. The presence of impurities can convert quartz into a variety of gem-like colors. Amethyst, agate, and jasper are all quartz crystals colored by impurities. A different form of coloring is that which gives a smoky appearance to quartz. This effect differs in degree from crystal to crystal, and in extreme cases a crystal may be so dark that it cannot be inspected for defects nor for the alignment of axes. However, by heating a smoky crystal from 350°C (662°F) to 500°C (932°F) it becomes

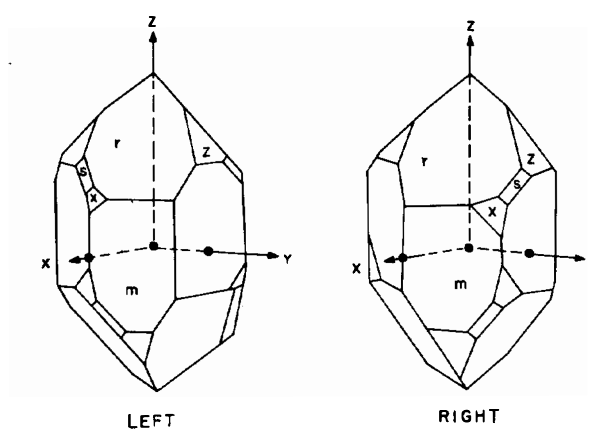


Figure 1-4. Left and right quartz crystals

quite as clear as the purest stone. Possibly the coloration is due to the dissociation of some of the silicon dioxide molecules, which recombine on heating; in any event, crystals which have been cleared of smokiness, remain clear, and have the same physical properties as the normal colorless crystals.

1-46. Other than those arising from chemical impurities, there are three types of structural defects to be avoided when cutting blanks from the raw quartz. These are *cracks, inclusions,* and *twinning.*

Cracks

1-47. All raw crystals contain cracks to some extent, particularly near their surfaces, where fractures are easily caused by impacts. Temperature variations and growth conditions are also causes of cracking. The larger cracks are readily visible, but not the separations with dimensions comparable to a wave-length of light. For this reason, any detected crevice should be assumed to extend somewhat beyond its visible length. Raw quartz should be handled with particular care, for the large crystals are more vulnerable to fracturing than are the small finished plates. No finished plate, however, should be permitted to contain a crack.

Inclusions

1-48. Inclusions are small pockets, often submicroscopic, holding foreign matter which was entrapped during the crystal's period of growth. The trapped material may be a gas, liquid, solid, or any combination thereof. The pockets are often too small to be seen individually, but are readily detected by the shapes and coloring of the clusters they form. Groups of the smallest-size inclusions have a bluish cast; groups of medium-size inclusions appear as a white frosting; and the larger inclusions are individually visible as small bubbles. Some of the clusters appear as small *clouds;* others appear as *needles,* which may be fine or feathery, and which may form parallel rows or spread comet-like from a bubble origin; still other groups are draped in sheets or folds like *veils;* and, finally, there are those inclusions that arrange themselves in surfaces parallel to the natural crystal faces, outlining former growths, and appearing as crystal *phantoms* within a crystal. See figure 1-5. Not a great deal is known concerning the effect of inclusions upon the performance of finished plates. However, the fine textured (blue) inclusions are the least objectionable, and the isolated bubbles are more to be tolerated than a veil or phantom. Blue needles are permissible in large, low-frequency plates that are not to be driven at high levels. Nevertheless, any inclusion weakens a crystal, and will not be present in a high-quality, finished plate.

Twinning

1-49. Twinning is the intergrowth of two crystal regions having oppositely oriented axes. This abnormality is rarely detectable by a casual visual inspection, and a crystal that appears homogeneous throughout may, indeed, have several twinned areas; in fact, almost all large crystals have twinning to some extent. There are two types of twinning common to quartz—*electrical twining* and *optical twinning.* In electrical twinning, only the electrical sense of the crystal axes is reversed, whereas in optical twinning, not only the electrical sense, but the handedness of the crystal

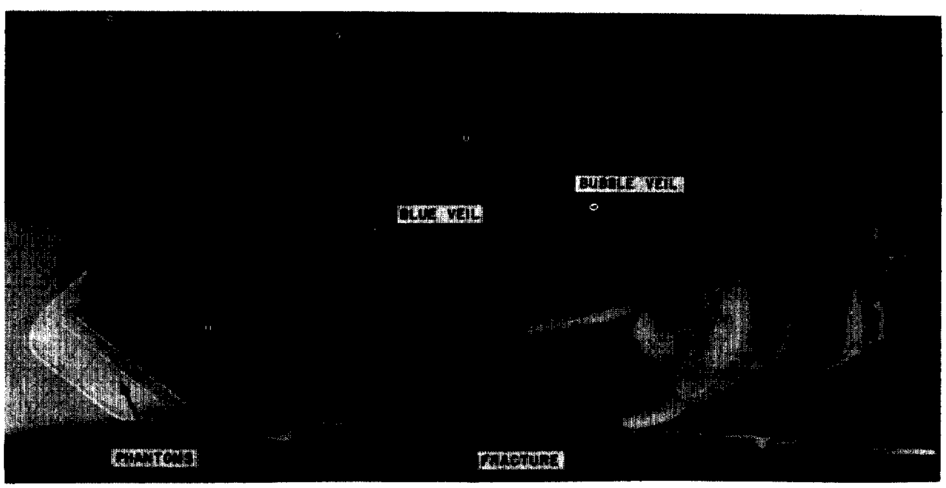


Figure 1-5. Quartz crystal containing inclusions and fractures *

structure is reversed—that is, one area will be right-handed and the other left-handed.

1-50. A finished plate, if it is to have predictable characteristics, must be cut entirely from a region having the same crystal structure; otherwise, the piezoelectric properties of one region will interfere with those of the other. Electrical twins are usually large, so both areas may be used separately for crystal blanks. Optical twinning, on the other hand, is usually confined to pockets, which are normally too small to provide crystal blanks, themselves, so that only the predominant crystal region can be utilized.

THE AXES OF QUARTZ

1-51. There are several crystallographic conventions by which the reference axes of crystals may be chosen, and much confusion has resulted in the past because of the various preferences of different crystallographers. Insofar as the over-all piezoelectric properties are concerned, the orientations of quartz have been universally measured according to rectangular sets of X, Y, and Z axes, with the XY, XZ, and YZ planes determined according to the crystal symmetries. However, even in this case, the choice of positive and negative axial and angular directions for right and left quartz remained more or less a matter of preference until the system proposed by the I.R.E. in 1949 became generally adopted. It is the I.R.E. system that will be followed here. It should be remarked first, however, that a crystal axis is not intended necessarily to coincide with a central point in the crystal, but may represent any straight line parallel to the axial direction. It might also be noted that the different types of crystal faces are designated in this manual by the small letters m, r, s, x, and z, and these should not be confused with the capital letters X, Y, Z which denote the axes, nor with the small letters, x, y, z, when used to denote dimensions of a crystal in the axial directions.

Z Axis

1-52. The Z axis is the lengthwise direction of the quartz prism and is perpendicular to the growth lines of all the m faces. It is an axis of three-fold symmetry, so that there are three sets of XY axes for each crystal (figure 1-6), with the direction of the Z axis common to all three. No piezoelectric effects are directly associated with the Z axis, and an electric field applied in this direction produces no piezoelectric deformation in the crystal, nor will a mechanical stress along the Z axis produce a difference of potential. Because the growth lines are generally missing, optical effects are usually employed to locate the Z azis in raw quartz. (See paragraphs 1-121 to 1-124.) Quartz properties are such that light waves passing through a crystal are effectively divided into two rectilinear components, with one component traveling faster than the other except when the light ray is directed parallel to the Z axis. The optical effects are found to be symmetric about the Z axis, and thus whereas optical instruments may be used to determine this axis, they cannot be used to distinguish an X from a Y axis. For this reason the Z axis is commonly designated as the *optic axis*. The optical effects associated with the propagation of polarized light parallel to the optic axis not only are used to locate the Z axis in unfaced quartz (crystals, such as river quartz, whose natural faces have been destroyed), but to identify left from right quartz, and to locate twinned regions. Plane polarized light traveling parallel to the optic axis will be rotated in one direction or the other according to whether the crystal is left or right. To an observer looking toward the light source the rotation will be clockwise for right-handed quartz and counterclockwise for left-handed quartz, with the amount of rotation depending upon the wavelength, being greater for blue light (short wavelength) and less for red light (long wavelength). Since the crystal lattice along the optic axis has no properties that distinguish one direction from the other, the choice of the +Z and the –Z reference directions are entirely arbitrary for either right or left crystals.

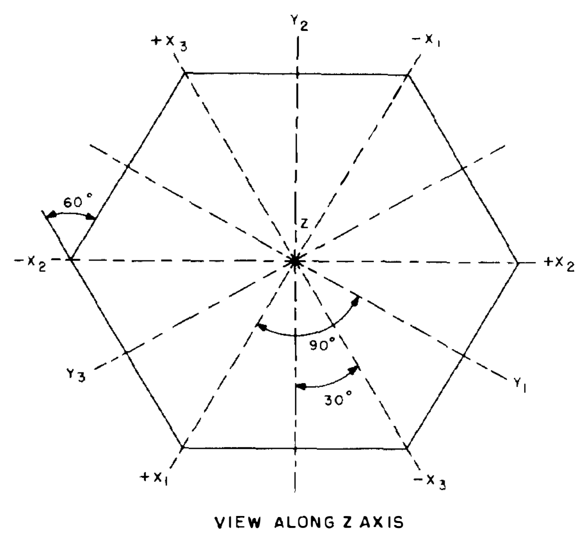


Figure 1-6. XY plane of quartz showing three sets of rectangular axes: X_1Y_1Z, X_2Y_2Z, X_3Y_3Z (Z axis is perpendicular to plane of paper)

Y Axis

1-53. The Y axes are chosen at right angles to the Z axis and to the growth lines of the m faces. See figure 1-4. For either left or right quartz, the positive end of a Y axis emerges from an m face that is adjoined by a z face at the end selected as the +Z direction. The Y axes are generally called the *mechanical axes* in contradistinction to the X axes, which are called the *electrical axes*. These names originated from the fact that simple compressional and tensional mechanical stresses along either an X axis or a Y axis would cause a polarization of the X axis, but not of the Y axis. The names are somewhat misleading, for polarization in the Y direction is also possible if a crystal undergoes shearing or flexural strains. In practice, the Y axis of a quartz stone is usually determined after the Z and X axes have been located.

X Axis

1-54. The X axes are parallel to the growth lines of the m faces, and to the lines bisecting the 120-degree prism angles. The positive end of an X axis is the direction that forms a right-handed coordinate system (see figure 1-4) with the Y and Z axes. This makes the directional sense of the X axis in right quartz the reverse, rather than the mirror image, of that in left quartz. Thus, in right quartz the negative ends of the X axes emerge from the prism corners that lie between the x faces, whereas, in left quartz the positive ends emerge from these corners.

1-55. In either right or left quartz when the X axis undergoes a tensional strain (stretching), a positive charge appears at the end emerging between the x faces; and when the X axis is compressed, this end becomes negatively charged. The X axis of raw quartz is usually determined by optical and x-ray methods. See paragraphs 1-126 and 1-127.

THEORY OF PIEZOELECTRICITY

Report Announcing Discovery of the Piezoelectric Effect

1-56. The theory of the cause of piezoelectricity stated in the most general terms is substantially the same today as it was at the time of its discovery. The following is the original report by Pierre Curie on the piezoelectric effect, which includes a statement of the theory that led to its discovery. The paper was read at the April 8, 1880, meeting of the *societe mineralogique de France*, and is recorded in the Bulletin, *soc. min. de France*, volume 3, 1880.

1-57. "Crystals which have one or more axes whose ends are unlike, that is to say, hemihedral crystals with inclined faces, have a special physical property, that they exhibit two electric poles of opposite names at the ends of those axes when they undergo a change of temperature: this is the phenomenon known as pyroelectricity.

1-58. "We have found a new way to develop electric polarization in crystals of this sort, which consists of subjecting them to different pressures along their hemihedral axes.

1-59. "The effects produced are analogous to those caused by heat: during a compression, the ends of the axis along which we are acting are charged with opposite electricities; when the crystal is brought back to the neutral state and the compression is relieved, the phenomenon occurs again, but with the signs reversed; the end which was positively charged by compression becomes negative when the compression is removed and reciprocally.

1-60. "To make an experiment we cut two faces parallel to each other, and perpendicular to a hemihedral axis, in the substance which we wish to study; we cover these faces with two sheets of tin which are insulated on their outer sides by two sheets of hard rubber; when the whole thing is placed between the jaws of a vise, for example, we can exert pressure on the two cut surfaces, that is to say, along the hemihedral axis itself. To perceive the electrification we used a Thomson electrometer. We may show the difference of potential between the ends by connecting each sheet of tin with two of the sectors of the instrument while the needle is charged with a known sort of electricity. We may also recognize each of the electricities separately; to do this we connect one of the tin sheets with the earth, the other with the needle, and we charge the two pairs of sectors from a battery.

1-61. "Although we have not yet undertaken the study of the laws of this phenomenon, we are able to say that the characteristics which it exhibits are identical with those of pyroelectricity, as they have been described by Gaugain in his beautiful work on tourmaline.

1-62. "We have made a comparative study of the two ways of developing electric polarization in a series of non-conducting substances, hemihedral with inclined faces, which includes almost all those which are known as pyroelectric.

1-63. "The action of heat has been studied by the process indicated by M. Friedel, a process which is very convenient.

1-64. "Our experiments have been made on blende, sodium chlorate, boracite, tourmaline, quartz, calamine, topaz, tartaric acid (right handed), sugar, and Seignette's salt.

1-65. "In all these crystals the effects produced by compression are in the same sense as those produced by cooling; those which result from relieving the pressure are in the same sense as those which come from heating.

1-66. "There is here an evident relation which allows us to refer the phenomena in both cases to the same cause and to bring them under the following statement:

1-67. "Whatever may be the determining cause, whenever a hemihedral crystal with inclined faces, which is also a non-conductor, contracts, electric poles are formed in a certain sense; whenever the crystal expands, the electricities are separated in the opposite sense.

1-68. "If this way of looking at the matter is correct, the effects arising from compression ought to be in the same sense as those resulting from heating in a substance which has a negative coefficient of dilation along the hemihedral axis."

Asymmetrical Displacement of Charge

1-69. The atomic lattice of piezoelectric crystals is assumed to consist of rows of alternating centers of positive and negative charges so arranged that the structure as a whole has no center of symmetry. When such a lattice undergoes a deformation, a displacement will result between the "centers of gravity" of the positive and negative charges. It is this displacement that results in a net unneutralized dipole moment, the polarity of which depends upon the previous equilibrium positions of the positive and negative centers of charge and the direction of the displacement, as indicated in figure 1-7.

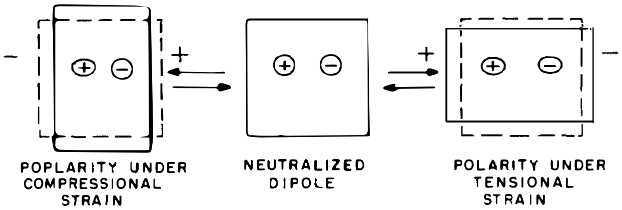


Figure 1-7. Effective polarities resulting from sudden displacements of the centers of charge of a neutralized dipole. (If after displacement, the crystal were maintained indefinitely in the strained position, the effective polarity would eventually be neutralized by an accumulation of ions at the poles. A sudden return from such a state would thus result in an effective polarization in the unstrained position)

1-70. In the case of a crystal with a center of symmetry, a uniform strain in the crystal will always result in as much displacement of like charges in one direction as in another, and hence there will be no net shift of the centers of opposite charge relative to each other. A distribution of charges having a center of symmetry is illustrated in figure 1-8. Note that the centers of charge, both positive and negative, are at the geometrical center. If a uniform stress—compressional, or shearing—is applied along any axis, it can be seen that the center of either type of charge will at all times remain undisturbed, and thus the net piezoelectric effect will be null.

1-71. Lord Kelvin was the first to propose a molecular model with a charge distribution designed to explain the physical and electrical characteristics of alpha quartz. See figure 1-9. This model was accepted generally by the crystallographers until the theory failed to conform to X-ray tests. When a beam of X-rays enters a crystal, the intersecting atomic planes can be likened to partially silvered mirrors, each passing part of the beam, but reflecting the rest. Since the distance between adjacent parallel planes is on the order of an X-ray wavelength, the waves reflected from adjacent planes will tend to alternately annul and reinforce each other as the angles of incidence vary. The interference pattern on a photographic film will show an array of spots indicating the angles at which the reflected waves from different planes arrive in phase. From such data, with the X-ray wavelength known, it is possible to determine the relative orientation of atomic planes, and hence to reconstruct the arrangement of the atoms in the crystal. The X-ray data on alpha quartz reveals a

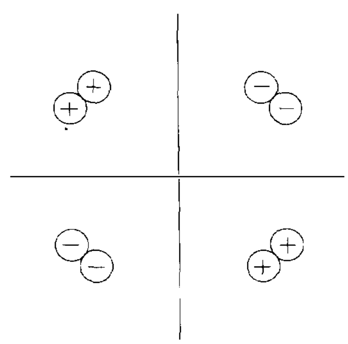


Figure 1-8. Example of distribution of charges having a center of symmetry *

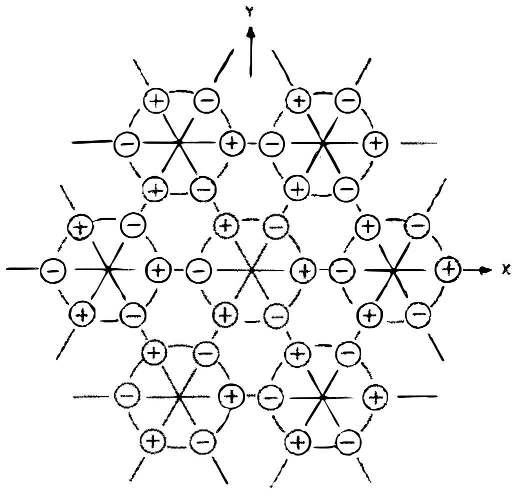
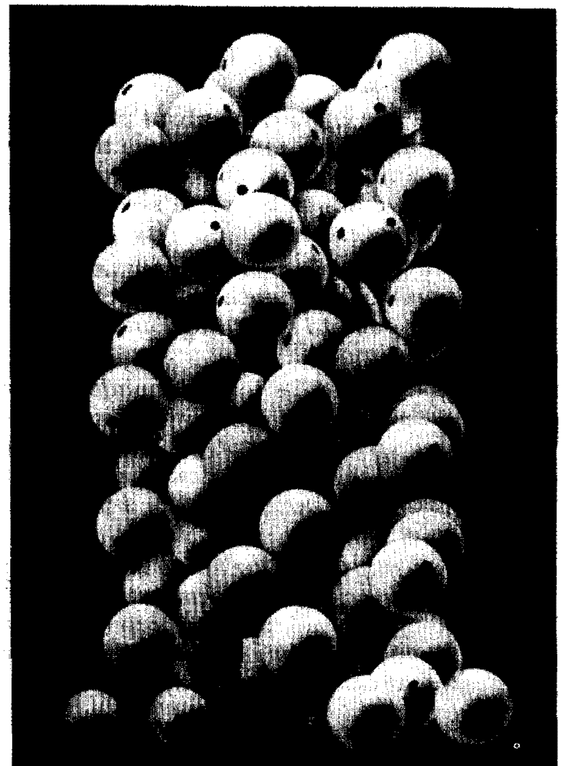


*Figure 1-9. Kelvin's molecular model of the charge distribution of alpha quartz. The positive direction shown for the X axis corresponds to that of right quartz, for a compression of the crystal along that axis will cause the piezoelectric polarities to coincide in sign with the X-axial directions. For left quartz, the sign of the Y, as well as the X, axis, must be reversed in order to maintain a right-handed coordinate system **



ARRANGEMENT OF ATOMS IN ALPHA QUARTZ, VIEWED ALONG AN X AXIS

Figure 1-10. Arrangement of atoms in alpha quartz. Plane of paper corresponds to YZ plane in crystal

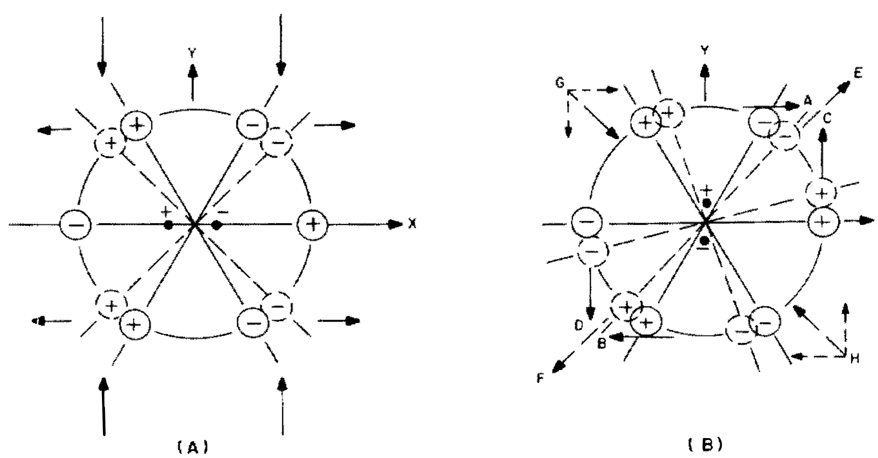


*Figure 1-11. Equivalent distribution of charges that account for observed piezoelectric effects of alpha quartz. (A) Piezoelectric polarity along X axis of right quartz due to compression along Y axis. (B) Piezoelectric polarity of Y axis of right quartz due to shearing stress, where the resultant strain is equivalent to a compression along the axis designated GH. (Note that in both A and B, the piezoelectric effect is due to a rocking of the axial dipoles, and not to their compression or extension. To achieve the same deformations by the converse effect, equal voltages, but opposite in sign to the polarizations indicated, are applied across the respective axes) **

more complex structure than was once suspected. See figure 1-10. Nevertheless, the early crystal model, as postulated by Lord Kelvin, still can be accepted as an approximation if we treat a single one of his molecules as representing simply the equivalent charge distribution within the lattice, as indicated in figure 1-11.

1-72. Figure 1-11A shows the displacement occurring when a compressional stress is applied along the Y axis, or a tensional stress is applied along the X axis of a right-handed crystal. Note that the center of positive charge shifts in the negative direction of the X axis, and that the center of negative charge shifts in the positive direction; however, there is no net displacement along the Y axis. If the direction of the stress is reversed, so also is the effective polarity.

1-73. The polarization of the Y axis due to a shearing strain is illustrated in figure 1-11B. Assume that vectors A and B represent a simple shearing stress applied at right angles to the Y axis. If A and B are equal and opposite forces, there will be no displacement of the center of mass; however, since these forces are not directed in the same straight line, they create a couple which would maintain a rotational acceleration about the center of mass unless opposed by an equal and opposite couple. This counter-couple is represented by vectors C and D. If now, the forces are combined vectorially, they may be represented as a longitudinal tension in the EF direction, or as a longitudinal compression in the GH direction. Consider the charge displacement from the point of view of a GH compression. Note that each of the positive charges is forced to shift slightly in the +Y direction, whereas each of the negative charges is displaced in the −Y direction. The net separation of the centers of charge thus causes the Y axis to become positively polarized at its geometrically positive end, and negatively polarized at its geometrically negative end.

1-74. Since the compression can be further analyzed into two components of equal magnitude—one horizontal, and the other vertical—it can be seen (figure 1-11A) that the polarities which these would induce along the X axis tend to cancel (for reinforcement to occur, one of the rectangular components would need to be tensional and the other compressional), and hence little or no polarization will appear in this direction.

MODES OF VIBRATION

1-75. If a piezoelectric crystal is suddenly released from a strained position, the inertia and elasticity of the crystal will tend to maintain a state of mechanical oscillation of constant frequency about one or more nodal points, lines, or planes of equilibrium, and alternating voltages will appear according to the particular mode of vibration. These are called the normal, or free, vibrations of a crystal, as distinct from the forced vibrations due to applied alternating mechanical or electrical forces that may differ in frequency from the crystal's natural resonance. The normal vibrations may, in turn, be of two general types: the free-free and the clamped-free vibrations. Free-free vibrations are those which would occur if a vibrating crystal were floating in empty space, where, regardless of the particular mode, the center of gravity is a nodal point. Clamped-free vibrations are those that would occur if a crystal were clamped at some point, or points, thereby preventing all normal modes except those at which nodes occur at the clamped points. For example, in a free-free vibration the ends of the crystal are free to move; however, if these ends are clamped, the resonant vibrations must be such that the ends become nodes. However, if a crystal is clamped only at those points which would be nodes in a free-free vibration, in the ideal case no interference results, and the resonance is still that of a free-free mode.

1-76. There are three general modes of vibration for which quartz crystal units are commercially designed: extensional, shear, and flexure. Fundamental vibrations of each of these modes are illustrated in figure 1-12. Higher harmonics up to and including the fifth are also widely used. Harmonic vibrations higher than the seventh have special high-frequency applications, but are rarely employed commercially.

1-77. A variation of the shear vibration is the torsional mode, which is readily excited in cylindrical crystals; however, except for laboratory

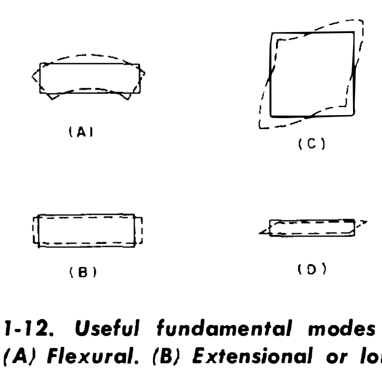


Figure 1-12. Useful fundamental modes of quartz plates. (A) Flexural. (B) Extensional or longitudinal. (C) Face (or length-width) shear. (D) Thickness shear.*

measurements of the properties of solids and liquids this mode is not in general use.

Frequency of Quartz Vibrations

1-78. The frequencies of the normal mechanical oscillations of a quartz plate may be considered as those at which standing waves will be established by reflection from the crystal boundaries. The positions of the nodes of the standing waves are predetermined by the geometry of the crystal, and by any difference that may exist in the velocities of propagation for the different wave components. The wavelength of a particular mode (but not the wave shape, if the velocity of one component differs from that of another) conforms only to the dimensions of the crystal faces. The frequency is related to the wavelength by the equation:

$$f = \frac{v}{\lambda} \qquad 1\text{—}78\ (1)$$

where: v = velocity of propagation
λ = wavelength

The fundamental equation of the velocity of propagation is:

$$v = \sqrt{\frac{c}{\rho}} \qquad 1\text{—}78\ (2)$$

where: c = stiffness factor in the direction of propagation
ρ = density

or: $$v = \sqrt{\frac{1}{\rho s}} \qquad 1\text{—}78\ (3)$$

where: $s = \frac{1}{c}$ = elastic compliance factor in the direction of propagation

Length- (or Width-) Extensional Mode

1-79. The motion of the atoms in an extensional mode is parallel to the direction of propagation. In the case of rectangular plates, stationary waves are established in the length direction by the interference of reflections from the opposite ends, where the wavelength is given by the formula:

$$\lambda = \frac{2l}{n} \qquad 1\text{—}79\ (1)$$

where l is the length and n is an integer (1, 2, 3, etc.) equal to the harmonic. Thus, the frequency of a length-extensional mode is:

$$f = \frac{nv}{2l} \qquad 1\text{—}79\ (2)$$

or, as expressed in terms of a frequency constant:

$$f = \frac{nk_1}{l} \qquad 1\text{—}79\ (3)$$

where: $k_1 = \frac{v}{2}$ = frequency constant for length-extensional mode

This formula, as well as the similar formulas for the shear and flexure modes, can be used to indicate the approximate dimensions required for a particular frequency when the appropriate frequency constant is known. Although the velocity of propagation decreases somewhat as the frequency increases, because of an increase in the frictional losses, this decrease is negligible for most purposes, and the same frequency constants that hold for the fundamental are also valid for the first few overtones. However, because of the coupling that exists between the length-extensional mode and other modes, the effective value of k will vary with changes in the w/l (width/length) ratio. Equation (3) also applies to width-extensional modes except that l is replaced by the width, w.

Thickness-Extensional Mode

1-80. This mode is little used today because of the close coupling that exists between it and the overtones of other modes. It is a mode that can be excited in a crystal whose thickness dimension is parallel to the electrical (X) axis (X-cut crystal) —the vibrations being such that the crystal alternately becomes thicker and thinner. Formerly, when X-cut crystals were widely used, the same crystal was often employed for the control of either a high- or a low-frequency circuit—using the thickness-extensional mode for the former and the length-extensional mode for the latter. Today, however, the more stable thickness-shear mode has almost entirely replaced the thickness-extensional mode in high-frequency circuits. The thickness-extensional frequency is given by the formula:

$$f = \frac{nv}{2t} = \frac{nk_2}{t} \qquad 1\text{—}80\ (1)$$

where v is the velocity of propagation in the thickness direction, n is the harmonic ($n = 1, 3, 5$,—for practical cases, although even harmonics of very small intensities have been observed), and k_2 is the generalized frequency constant. Actually, the effective thickness, t, decreases somewhat for the overtones, so that the correct value of k_2 is slightly greater for the harmonics than for the fundamental.

Thickness-Shear Mode

1-81. The motion of the atoms in a thickness-shear mode is parallel to the major (length-width) faces of the crystal, whereas the wave propagation is parallel to the thickness dimension. The equation for the fundamental frequency when the thickness is very small compared with the length and width is:

$$f = \frac{v}{2t} \qquad 1\text{—}81\ (1)$$

where: v = velocity of propagation along thickness dimension
t = thickness

or, as expressed in terms of a thickness-shear frequency constant:

$$f = \frac{k_3}{t} \qquad 1\text{—}81\ (2)$$

The thickness-shear is also called the "high-frequency shear" in contradistinction to the "face," "length-width," or "low-frequency" shear. The overtones of the thickness-shear mode may have components (reversals of phase) in the length and width directions as well as along the thickness. The more general formula for the frequency is:

$$f = k_3 \sqrt{\frac{m^2}{t^2} + \frac{n^2}{l^2} + \frac{p^2}{w^2}} \qquad 1\text{—}81\ (3)$$

where m, n, and p are integers representing the harmonic component in the t, l, and w directions, respectively. The above equation applies to an isotropic medium; however, since the elastic constants in quartz are not the same in all directions, the thickness-shear formula has been modified to:

$$f = k_3 \sqrt{\frac{m^2}{t^2} + a_1\frac{n^2}{l^2} + a_2\frac{(p-1)^2}{w^2}} \qquad 1\text{—}81\ (4)$$

where a_1 and a_2 are constants to be determined empirically. For most applications, t is much smaller than l and w, and n = p = 1, so that the formula, $f = \frac{k_3 m}{t}$, is sufficiently accurate.

Face-Shear Mode

1-82. The face-shear mode involves a more complex relation among the crystal dimensions. A complication arises from the fact that the wave is effectively divided into two components—one propagated along the length, and the other along the width. Each of these separate components has its own series of possible harmonics, so that the resultant frequencies of the face-shear modes are not necessarily integral multiples of the fundamental. The approximate-frequency equation is:

$$f = \frac{v}{2}\sqrt{\frac{m^2}{l^2} + a_1\frac{n^2}{w^2}} = k_4' \sqrt{\frac{m^2}{l^2} + a_1\frac{n^2}{w^2}} \qquad 1\text{—}82\ (1)$$

where m and n are integers representing the length and width harmonics, respectively. The symbol a_1 is a constant of proportionality, approximately equal to one, which is inserted when the velocity of propagation along w is not the same as that along l. If the face of the plate is square, the formula for the fundamental frequency is reduced to approximately:

$$f = \frac{k_4}{w} \qquad 1\text{—}82\ (2)$$

where: $k_4 = k_4'\sqrt{2}$

The fundamental vibration, where m = n = 1, is shown in figure 1-12C. Note that the shape of the deformation is not that of a parallelogram, as it would be if the plate were slowly compressed along a diagonal. Rather, the vibrational distortion is a dynamic one, and the resultant wave must be in the same phase at all points. Figure 1-13 represents the face-shear mode for m = 6, n = 3. Note that the number of nodes in each row is equal to m, and the number in each column is equal to n.

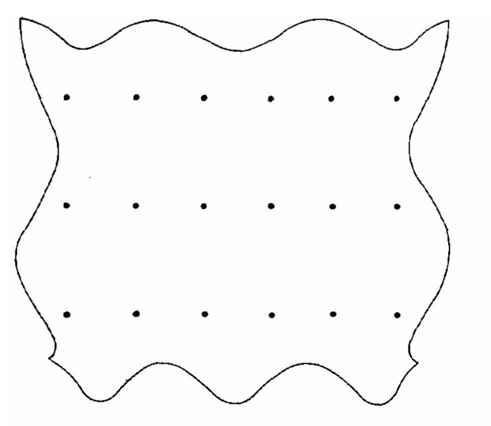


*Figure 1-13. Face-shear mode for m = 6, n = 3. Dots indicate nodes **

Length-Width-Flexural Mode

1-83. The length-width-flexural mode is a bending of the crystal in the length-width plane. Normally, the crystal is so mounted that the ends are free to vibrate in a free-free mode. The formula for the frequency involves the root of a transcendental equation, but expressed in terms of a frequency constant, the equation becomes:

$$f = \frac{k_5 w}{l^2} \qquad 1\text{—}83\ (1)$$

The convenience of a common frequency constant for all practicable harmonics is not realized in the case of length-width flexures, where the "constant" k_5 is a function not only of the particular harmonic, but also of l and w. However, for long, thin rods $\left(\frac{nw}{l}\right.$ less than 0.1, where n is the harmonic$\left.\right)$ k_5 is approximately independent of the dimensions, and fixed values of k_5 can be assumed for the particular harmonics of different types of cuts. Because of the elastic cross constants in quartz, which relate a field in one direction to a polarization in a perpendicular direction, a flexure may be accompanied by a torsion. To prevent this, the length of a crystal to be operated in a flexural mode should lie somewhere in a YZ plane.

Length-Thickness-Flexural Mode

1-84. Length-thickness flexures are used to control frequencies in the audio range. To obtain this mode, two long, thin plates of the same cut are cemented together with the electrical axes opposed, so that, when an alternating voltage is applied across the outer faces, one crystal strip expands as the other contracts, and vice versa—the over-all effect being a flexural vibration. The normal frequency of a free-free length-thickness flexure is given by an equation similar to that for the length-width flexure, except that the thickness, t, is substituted for the width, w. Thus:

$$f = nk_6 \left(\frac{t}{l^2}\right) \qquad 1\text{—}84\ (1)$$

Frequency Range of Normal Modes

1-85. Standard quartz crystal units are designed for frequencies from 400 cycles to 125 megacycles per second. Laboratory devices have employed thickness flexure crystals for the control of frequencies as low as 50 cycles per second, and, by exciting the higher thickness-shear modes, control of frequencies higher than 200 megacycles per second have been realized. At these high frequencies, however, so many interlocking modes are possible that it is difficult to prevent a crystal from jumping from one mode to another during slight variations of temperature, unless a very precise fabrication of the crystal unit has been achieved. The high-frequency limit of the lower harmonics is reached when the dimensions are so small that either the crystal cannot be driven without the risk of shattering, or that the impedances introduced by the mounting become proportionately too large for practicable operation.

1-86. The practical frequency ranges of the different modes are as follows:

Flexure Mode—
- Length-thickness: 0.4 to 10 kc
- Length-width: 10 to 100 kc

Extensional Mode—
- Length: 40 to 350 kc
- Thickness: 500 to 15,000 kc

Shear Mode—
- Face: 100 to 1800 kc
- Thickness (fundamental): 500 to 20,000 kc
- Thickness (overtones): 15,000 to 125,000 kc

ORIENTATION OF CRYSTAL CUTS

Right-Handed Coordinate System

1-87. With the positive sense of the quartz X, Y, and Z axes determined as in paragraphs 1-52, 1-53, and 1-54, the positive sense of rotation about the axes is fixed by the conventions of a right-handed

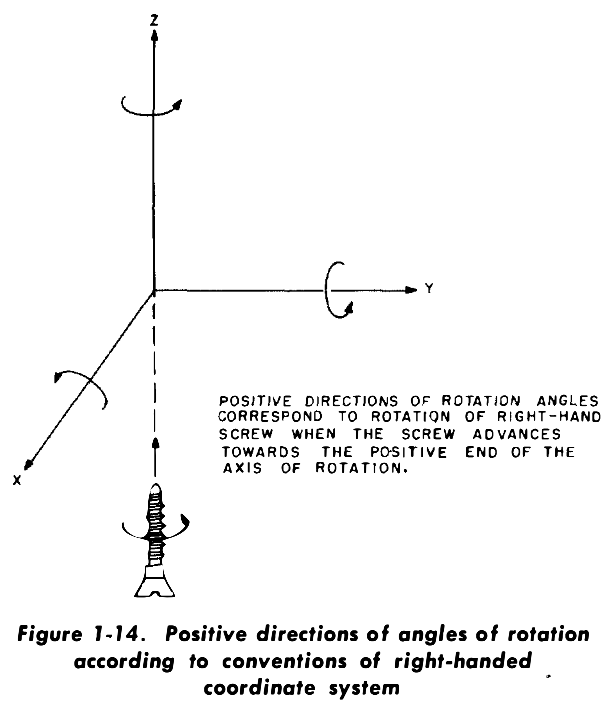


Figure 1-14. Positive directions of angles of rotation according to conventions of right-handed coordinate system

coordinate system for both right and left quartz. If one imagines a right-handed screw pointing towards the positive end of an axis of rotation, as represented in figure 1-14, the direction of an angle of rotation is considered positive if the rotation advances the screw in a positive direction—this corresponds to a clockwise rotation if observed when looking towards the positive end of the axis of rotation. The reverse, or counterclockwise, angles of rotation are taken as negative. The sense of the axes are such that the angles of rotation are positive when the directions of rotation are from +X to +Y, +Y to +Z, and +Z to +X. The axial and rotational conventions permit a particular cut of crystal to have the same rotation symbol for both right and left quartz.

Rotation Symbols

1-88. To specify the orientation of a piezoid cut, the following system, as recommended by the I. R. E. in 1949 is in general use. The crystal blank to be described is assumed to have a hypothetical initial position, with one corner at the origin of the coordinate system, and the thickness, length, and width lying in the directions of the rectangular axes. There are six possible initial positions, each of which is specified by two letters, the first letter indicating the thickness axis, and the second letter indicating the length axis. These positions are thus designated xy, xz, yx, yz, zx, and zy. The xy and yx positions are shown in figures 1-15 and 1-16, respectively. The starting position is so chosen that the final orientation may be reached with a minimum number of rotations. These rotations are taken successively about axes that parallel the dimensions of the crystal at the time of rotation. Only the first rotational axis will coincide with a rectangular axis; however, the positive direction of any axis of rotation is that defined by the XYZ system for the initial position. A single rotation is sufficient for describing the majority of standard cuts, and three rotations is the maximum in any case. The dimensions and axes of rotation are indicated by the symbols, t, l, and w, for thickness, length, and width, respectively. The Greek letters ϕ, θ, and ψ designate the first, second, and third angles of rotation, respectively. The following example, illustrated in figure 1-17, is a complete geometrical specification of a crystal plate:

$$\begin{aligned} &\text{yztwl } 30°/15°/25° \\ t &= 0.80 \pm 0.01 \text{ mm} \\ l &= 40.0 \pm 0.1 \text{ mm} \\ w &= 9.00 \pm 0.03 \text{ mm} \end{aligned}$$

The lettered combination at the beginning of the specification is called the "rotation symbol." The first two letters, yz, of the symbol indicate the initial position, and the next three letters, twl, state the axes of rotation and the order in which the rotations are taken. The three angles, all positive in this case, give the orientation and are listed in the same order as the respective rotations. The dimensions listed are those of the particular plate, and are not to be considered as necessary specifications for that type of cut. For circular plates, the initial position will indicate which directions are to be considered thickness and length, so that the same rotation symbol is used as for rectangular

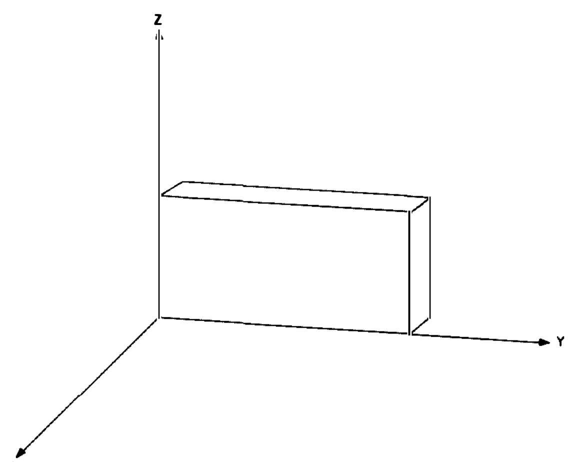


Figure 1-15 xy initial position for designating orientation of crystal cut

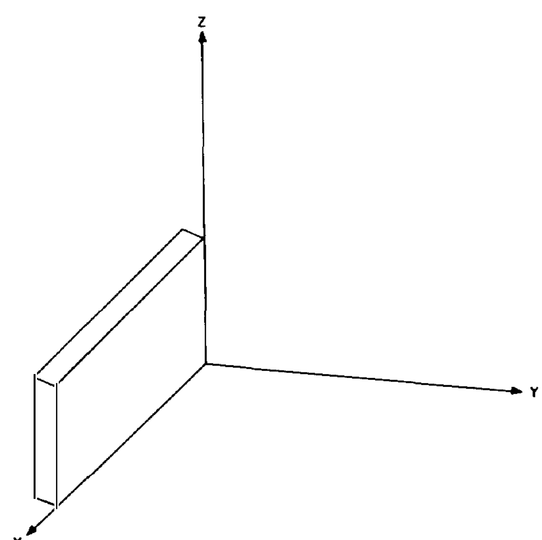


Figure 1-16 yx initial position for designating orientation of crystal cut

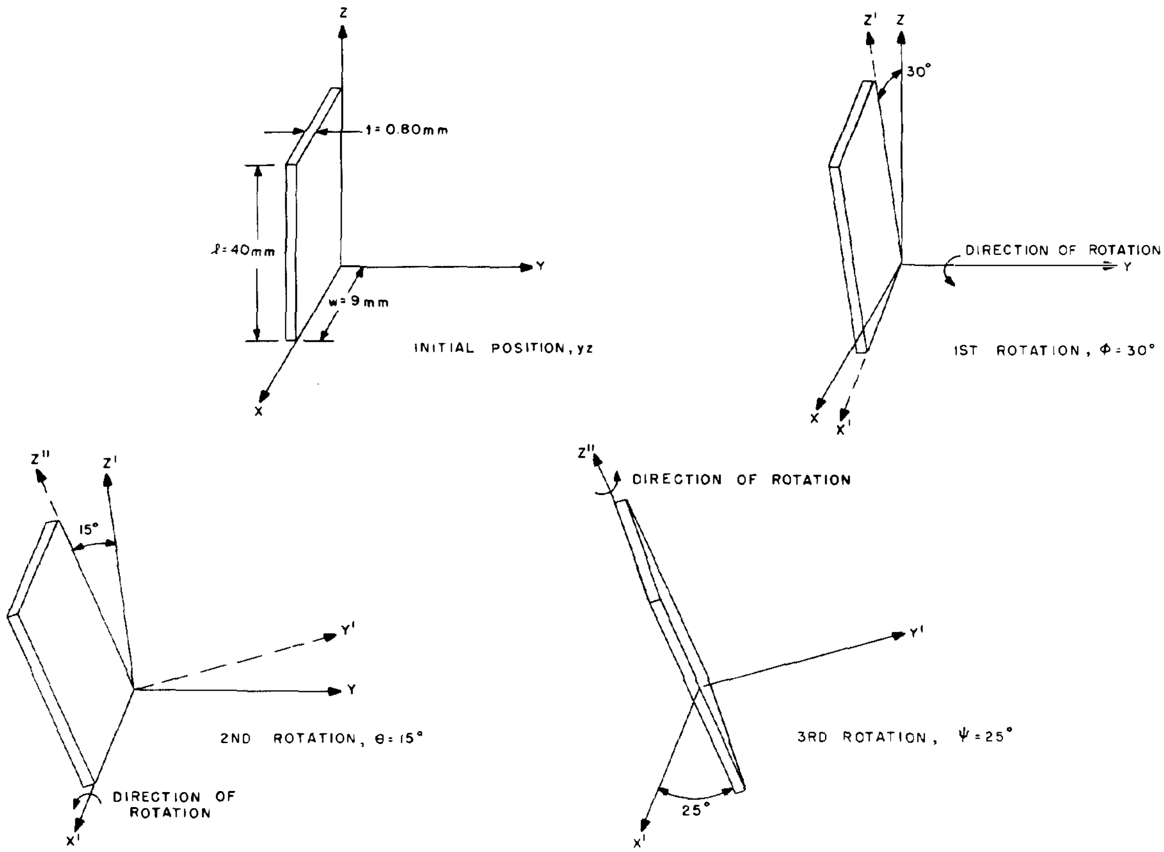


Figure 1-17 Orientation of crystal having the rotational specifications yztwl: 30°/15°/25°

plates; in specifying the dimensions, however, l and w are replaced by the diameter.

PIEZOELECTRIC ELEMENTS

1-89. The performance characteristics of a crystal plate are dependent on both the particular cut and the mode of vibration. For convenience, each "cut-mode" combination is considered a separate "piezoelectric element," and the more commonly used elements have been assigned a letter symbol. For example, the thickness-shear mode of the AT cut is designated as element A.

STANDARD QUARTZ ELEMENTS

1-90. The principal quartz elements are given below, with those which have been assigned element symbols listed first.

Element Symbol	*Name of Cut*	*Rotation Symbol and Orientation*	*Mode of Vibration*	*Frequency Range in KC*
A	AT	yxl 35°21′ or yzw 35°21′	thickness-shear	500 to 125,000
B	BT or YT*	yxl —49°8′ or yzw —49°8′	thickness-shear	1,000 to 75,000
C	CT	yxl 37°40′ or yzw 37°40′	face-shear	300 to 1,100
D	DT	yxl —52°30′ or yzw —52°30′	face-shear	60 to 500
E	+5°X	xyt 5°	length-extensional	50 to 500

* The YT cut, which is essentially the same as the BT cut, was developed independently by Yoda in Japan.

Element Symbol	*Name of Cut*	*Rotation Symbol and Orientation*	*Mode of Vibration*	*Frequency Range in KC*
F	−18.5°X	xyt −18.5°	length-extensional	60 to 300
G	GT	yxlt −51°7.5′/45°	width-extensional	100 to 550
H	5°X	yxt 5°	length-width flexure	10 to 50
J	Duplex 5°X	xyt 5° (right quartz) and xyt 5° (left quartz)	length-thickness flexure	0.4 to 10
M	MT	xytl 0° to 8.5°/±34° to ±50°	length-extensional	50 to 500
N	NT	xytl 0° to 8.5°/±38° to ±70°	length-width flexure	4 to 100
—	AC	yxl 31° or yzw 31°	thickness-shear	1,000 to 15,000
—	BC	yxl −60° or yzw −60°	thickness-shear	1,000 to 20,000
—	ET	yxl 66°30′ or yzw 66°30′	combination flexure and face-shear	600 to 1,800
—	FT	yxl −57° or yzw −57°	combination flexure and face-shear	150 to 1,500
—	V	xzlw or xywl 15° to 29°/−14° to −54° and 13° to 29°/27° to 42°	thickness-shear	1,000 to 20,000 (fundamental)
—	V	xzlw or xywl 0° to 30°/±45° to ±70°	face-shear	60 to 1,000
—	X	xy	length-extensional	40 to 350
—	X	xz	width-extensional	125 to 400
—	X	xy or xz	thickness-extensional	350 to 20,000
—	Y	yx or yz	thickness-shear	500 to 20,000

TYPES OF CUTS

1-91. The standard quartz elements can be divided into two groups: in the first group belong those crystals which are most conveniently described as being rotated X-cut crystals, and in the second group belong those crystals which are most conveniently described as being rotated Y-cut crystals. The first will hereafter be designated as the *X group*, and the second as the *Y group*.

1-92. The X and Y cuts have their thickness dimensions parallel to the X and Y axes, respectively, with the length and width dimensions parallel to the two remaining axes. See figure 1-18. Thus, in describing a crystal orientation, the X cut is the equivalent of the two initial positions xy and xz, and the Y cut is represented by the initial positions yx and yz. Belonging to the X and Y groups, then, are those crystals whose rotation symbols begin with the letters x and y, respectively. As a general rule, from the X group, the low-frequency crystal units are obtained, and from the Y group, the medium- and high-frequency units. A third group of crystals is theoretically possible, where the initial position is a Z cut (thickness parallel to the Z axis) ; however, because the piezoelectric effect

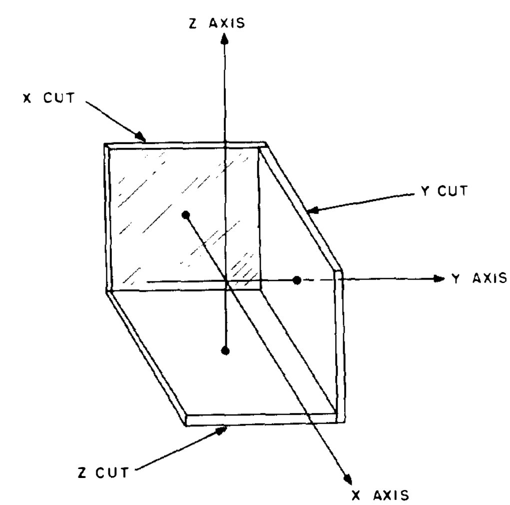


Figure 1-18 Orientation of X, Y, and Z cut plates

is restricted to the X and Y axes, the electrodes must be placed across one of these axes, which for the Z cut, would be at the edges—not a convenient location. Nor have other cuts, more or less simply oriented relative to a Z cut, been found to have optimum performance characteristics. However, there are experimental Z cuts, such as some of the ring-shaped crystals, which have proven of high quality, even though not practical for general use.

The X Group

1-93. The principal crystals of the X group are listed below with the frequency ranges for which they have found commercial application:

Name of Cuts	*Frequency Range in KC*
X	40 to 20,000
5°X	0.9 to 500
−18°X	60 to 350
MT	50 to 100
NT	4 to 50
V	60 to 20,000

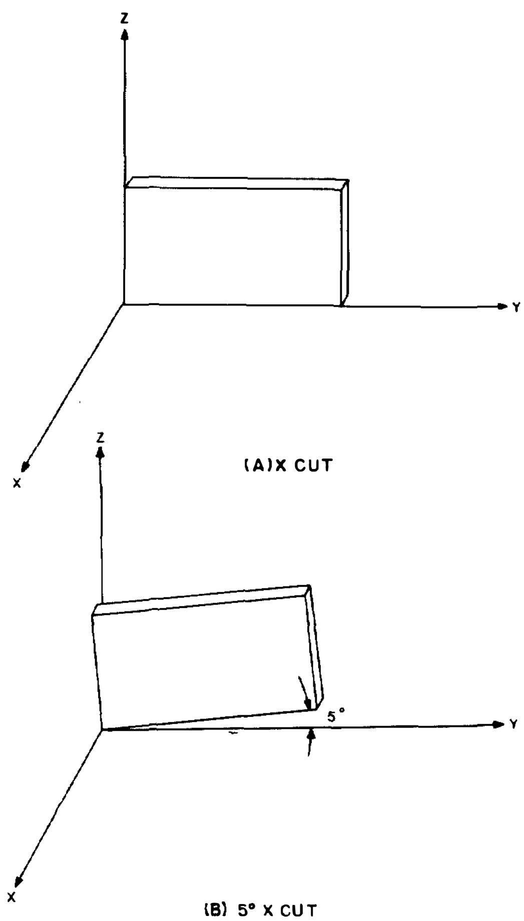


Figure 1-19 shows the orientations of an xy initial position (X cut with the length parallel to the Y axis) for the various cuts.

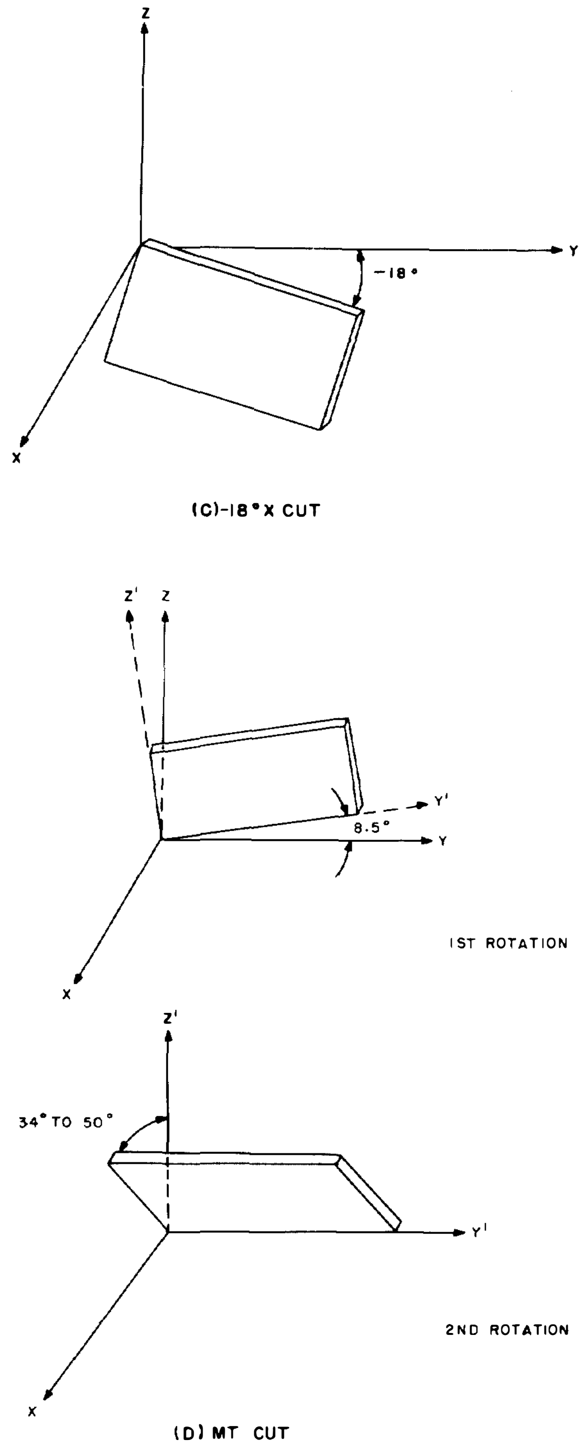
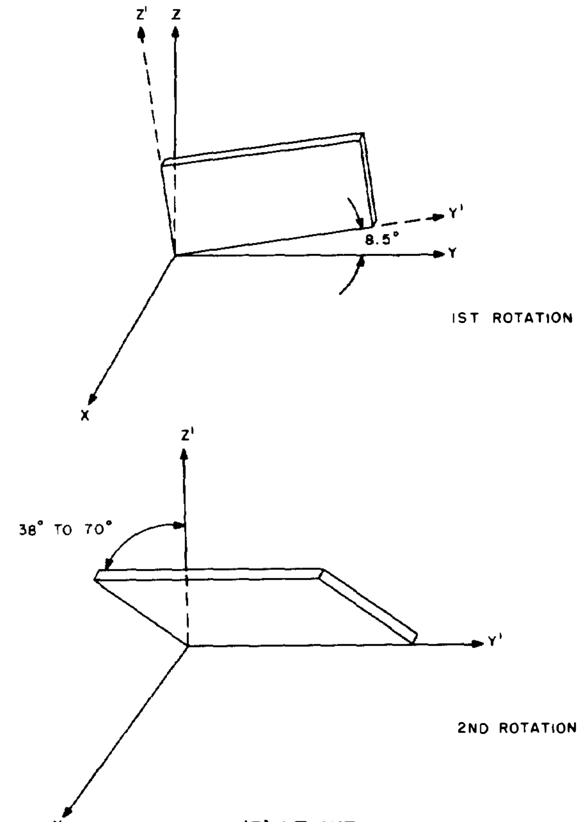
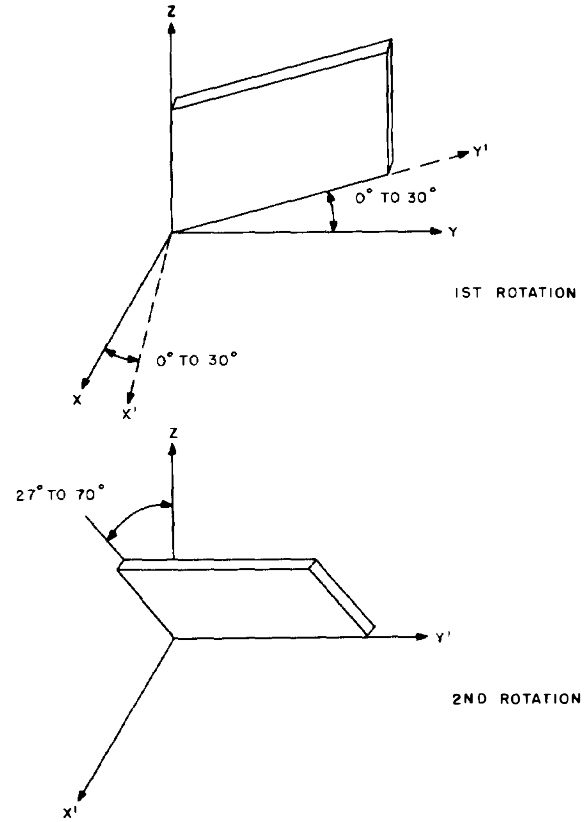


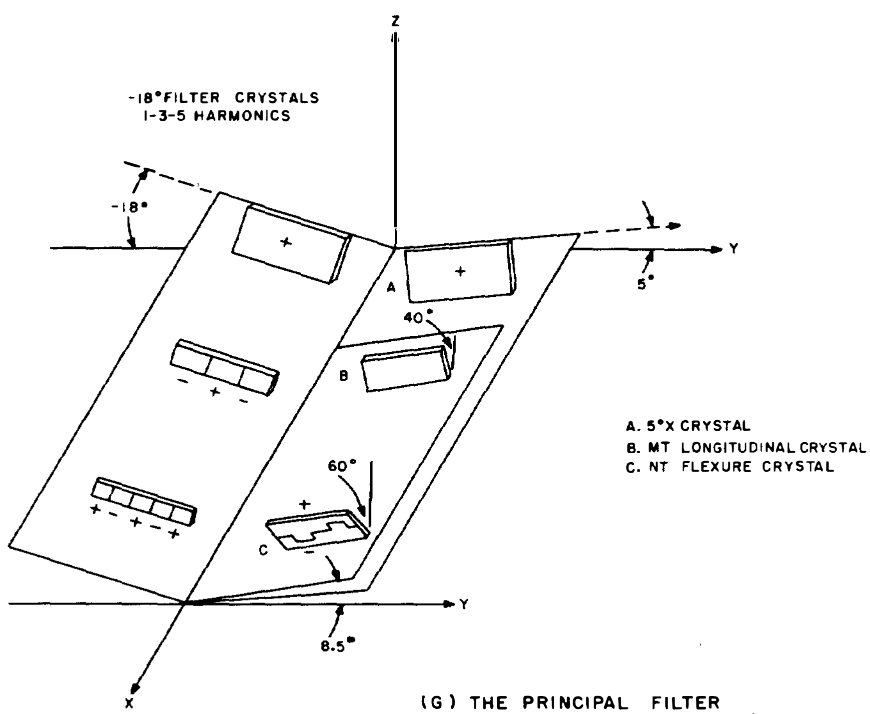
Figure 1-19. The X group. (The second rotations of the MT, NT, and V cuts are shown only for the positive angles) *



(E) NT CUT



(F) V CUT



(G) THE PRINCIPAL FILTER CRYSTALS OF THE X GROUP *

THE X CUT

1-94. The X cut was the original quartz plate investigated by Curie, and thus is sometimes called the "Curie cut." This cut was also the first to be used as a transducer of ultrasonic waves and as the control element of radio-frequency oscillators. However, because of its comparatively large coefficient of temperature, the X-cut plate is now rarely used in radio oscillators. As a transducer of electrical to mechanical vibrations, especially at high frequencies of narrow bandwidth, the X cut has a high electromechanical coupling efficiency, and is still widely used to produce ultrasonic waves in gases, liquids, and solids. These applications are largely for testing purposes, such as the measurement of physical constants and the detection of flaws in metal castings.

1-95. *CHARACTERISTICS OF X-CUT PLATES IN THICKNESS-EXTENSIONAL MODE*

Description of Element: X cut; xy or xz; thickness-extensional mode.

Frequency Range: 350—20,000 kc (fundamental vibration); lower frequencies when coupled as transducer for generating vibrations in liquids and solids.

Frequency Equation: $f = \frac{nk_2}{t}$ (n = 1, 3, 5, . . .)

Frequency Constant: $k_2 = 2870$ kc-mm.

Temperature Coefficient: 20 to 25 parts per million per degree centigrade; negative (i.e. for each degree increase or decrease in temperature, the frequency respectively decreases or increases 20 to 25 cycles for each megacycle of the initial frequency—a rise in temperature of 10°C would thus cause the frequency of a 5000-kc crystal to drop 1000 to 1250 cycles per second.)

**Methods of Mounting:* Sandwich and unclamped air-gap—for oscillator circuit; transducer mounting depends upon particular type of mechanical load.

Advantages: Mechanical stability, economy of cut, efficiency of conversion of electrical to mechanical energy, and large frequency constant make this piezoelectric element preferred for the radiation of high-frequency acoustic waves when the ratio of the highest to the lowest frequency need not exceed 1.1.

Disadvantages: Large temperature coefficient, tendency to jump from one mode to another, and the difficulty of clamping crystal in a fixed position without greatly damping the normal vibration prevent this element from being preferred for oscillator control. An electromechanical coupling factor of 0.095, which is only one-fourth that of the best synthetic crystals, makes this element inefficient as a radiator of a wide band of frequencies.

* See paragraphs 1-132 to 1-171.

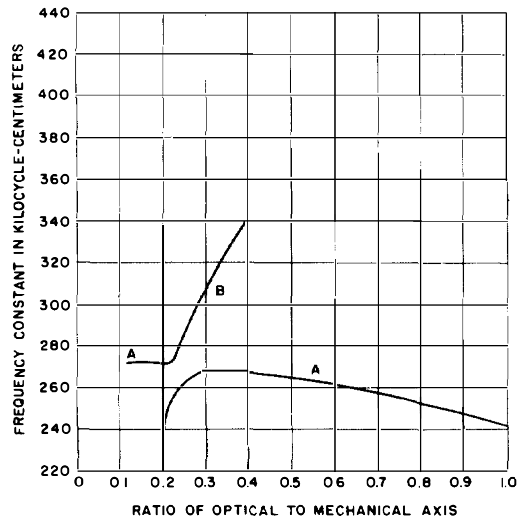


Figure 1-20. Frequency constant for length-extensional mode (curve A) of X-cut crystal where the width and length are parallel to the Z and Y axes, respectively. Curve B is the frequency constant of a face-shear mode coupled to a second flexural mode, whose interference makes the crystal useless for w/l ratios between 0.2 and 0.3, unless the thickness approaches the dimensions of the width *

1-96. *CHARACTERISTICS OF X-CUT PLATES IN LENGTH-EXTENSIONAL MODE*

Description of Element: X-cut; xy; length-extensional mode.

Frequency Range: 40—350 kc.

Frequency Equation: $f = \frac{nk_1}{l}$ (n = 1, 2, 3, . . .)

Frequency Constant: Varies with w/l ratio—see figure 1-20.

Temperature Coefficient: Negative**, varies with w/l ratio—see figure 1-21; zero coefficient if w/l = 0.272 and w = t.

Methods of Mounting: Sandwich, air gap, wire, knife-edge clamp, pressure pins, cantilever clamp; more than one pair of electrodes required for overtones; transducer mounting depends upon particular type of mechanical load.

** All quartz bars have negative temperature coefficients for *pure* length-extensional vibrations, although a zero coefficient is obtainable for certain cuts.

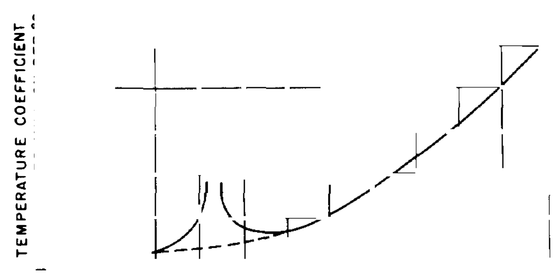


Fig
ten

= ***0.05***

Advantages: For w/l ratios from 0.35 to 1.0, the fundamental length-extensional vibration is not strongly coupled to other modes, and hence the resonance is easily excited and of good stability except for drift during temperature variations. Although not preferred over zero-temperature-coefficient cuts, this element, with temperature control, is reliable for use in low-frequency oscillators, and for long, thin bars, for use in filters. However, its most important application is to produce ultrasonic vibration in gases, liquids, and solids, when the ratio of highest to lowest frequency need not exceed 1.1.

Disadvantages: Inefficient as transducer of any but narrow frequency band, since electromechanical coupling is only one-fourth that of the better synthetic crystals. Strong coupling with a flexural mode makes the crystal useless at w/l ratios between 0.2 and 0.3 (see figure 1-20), and a weak coupling with a shear mode causes the frequency constant to decrease as the w/l ratio approaches 1.0. This coupling to other modes interferes with the frequency response of the element when used in filters, unless the w/l ratio is 0.1 or less. Although for long thin bars the temperature coefficient is only about 2 parts per million per degree, this is greater than the minimum obtainable with 5°X-cut bars.

1-97. *CHARACTERISTICS OF X-CUT PLATES IN WIDTH-EXTENSIONAL MODE*

Description of Element: X-cut; xz; width-extensional mode.

Frequency Range: 125 to 400 kc.

Frequency Equation: $f = \frac{nk_1}{l}$ (n = 1, 2, 3, . . .)

Frequency Constant: Varies with w/l ratio; see figure 1-22, which shows the face dimensions that will have a single frequency near the desired resonance. Plates with dimensions not included between the two outer curves will have interfering modes.

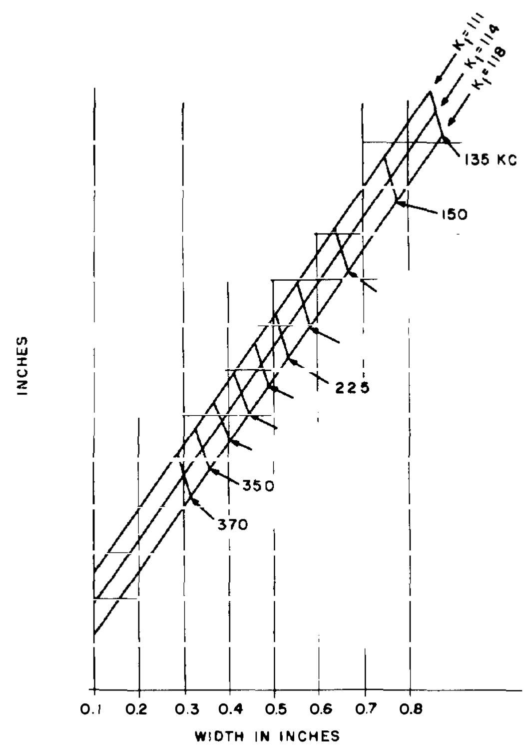


Figure 1-22. Frequency characteristics of X-cut crystal vibrating in width-extensional mode, where the width is parallel to the Y axis. w/l ratios not included between the two outer curves will have interfering modes. K_1 is in kc-inches

Temperature Coefficient: Negative; approximately 10 parts per million per degree centigrade, but varies with w/l ratio.

Methods of Mounting: Sandwich, air gap, wire, knife-edge clamp, pressure pins, cantilever clamp.

Advantages: If cut with dimensions within the single-frequency range shown in figure 1-22, this element can be used in temperature-controlled low-frequency oscillators and narrow-band-pass filters. With the thickness dimension ground for a particular high frequency, the same crystal unit may be used to generate either of two widely separate frequencies.

Disadvantages: Relatively large temperature coefficient prevents this element from being preferred over the low-coefficient cuts.

THE 5° X CUTS

1-98. The 5°X cut is the orientation that provides a zero temperature coefficient for the lengthwise vibrations of long, thin X bars, as shown in figure 1-23. Thus, this cut is preferred over the non-rotated X cut for use in low-frequency filters and control devices. Its length-extensional, length-width-flexural, and duplex length-thickness-flexural modes are defined as the elements E, H, and J, respectively; the last named element, J, providing the lowest frequencies. However, the 5°X elements are also coupled to the other modes, so that for w/l ratios much greater than 0.1 the frequency spectrum is little improved over that of the length-extensional mode of the X cut. Furthermore, as the w/l ratio increases, so also does the temperature coefficient. For these reasons the 5°X elements are especially advantageous only when the w/l ratio is 0.1 or less. These long, thin bars are used commercially for the control of low-frequency oscillators and as filters, and are particularly adaptable for use in telephone carrier systems.

1-99. *CHARACTERISTICS OF ELEMENT E*

Description of Element: 5°X cut; xyt: 5°; length-extensional mode.

Frequency Range: 50 to 500 kc.

Frequency Equation: $f = \frac{nk_1}{l}$ (n = 1, 2, 3, . . .)

Frequency Constant: Varies with w/l ratio (see figure 1-24).

Temperature Coefficient: Varies with w/l ratio (see figure 1-25, which holds for temperatures between 45 and 55 degrees centigrade). The

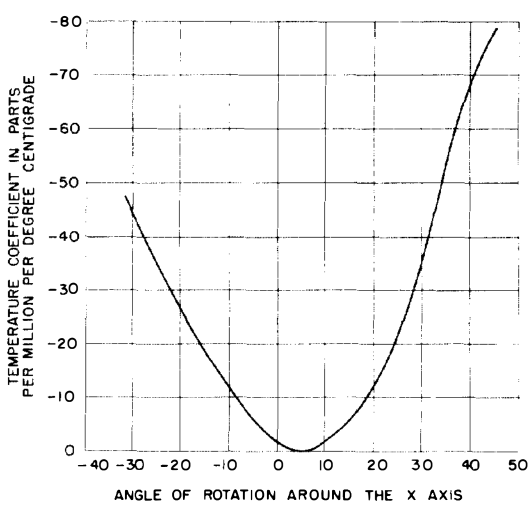


Figure 1-23. Temperature coefficient for length-extensional mode of long, thin X-group bars versus angle of rotation *

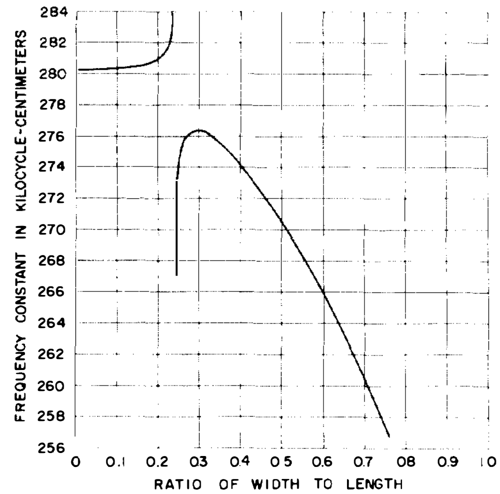


Figure 1-24. Frequency constant versus w/l ratio for element E *

frequency deviation of representative E elements of different w/l ratios is shown in figure 1-26, where the initial frequency is taken at 25°C.

Note that the temperature coefficient in parts per hundred per degree is the slope of a curve, and varies from positive to zero to negative as the temperature increases.

Methods of Mounting: Wire, knife-edge clamp, pressure pins, cantilever clamp; more than one pair of electrodes required for overtones.

Advantages: The low temperature coefficient and a large ratio of stored mechanical to electrical energy make this element preferred for filter networks. Long, thin bars have only a very

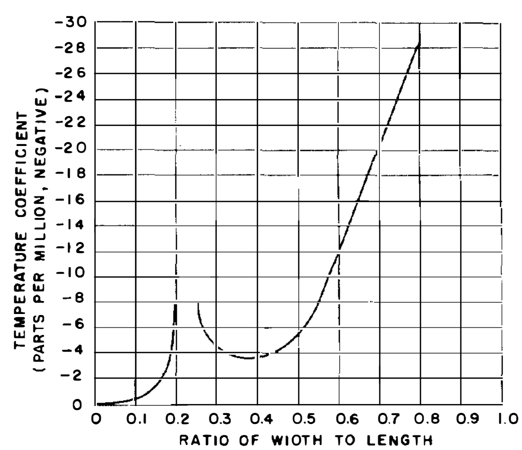


Figure 1-25. Temperature coefficient versus w/l ratio for element E at temperatures between 45° and 55°C *

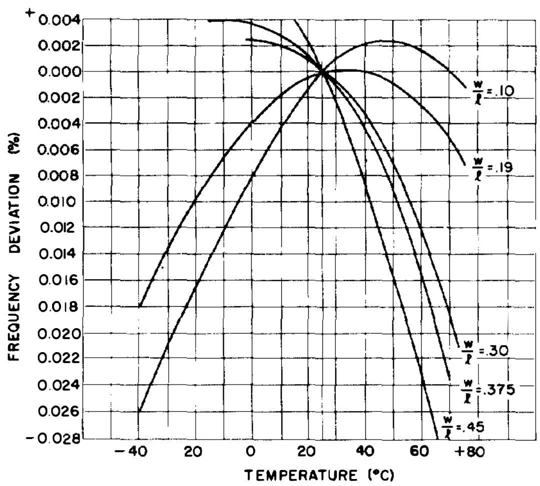


Figure 1-26. Percentage frequency deviation for E elements of various w/l ratios. Initial temperature = 25°C

weak coupling to other modes and are used for both filter networks and low-frequency oscillators. If a w/l ratio greater than 0.15 is desired, a ratio of approximately 0.39 is optimum insofar as a low temperature coefficient is concerned.

Disadvantages: At w/l ratios between 0.2 and 0.3 the length-extensional mode is so closely coupled to the length-width flexure that the crystal is useless; as the width is increased the coupling of the length-extensional to the face-shear mode becomes stronger, and the temperature coefficient becomes larger. However, because of the large electro-mechanical coupling of this element, w/l ratios of 0.35 to 0.5 can still be favorably used in filters if a temperature coefficient less than 4 parts per million is not required.

1-100. *CHARACTERISTICS OF ELEMENT H*

Description of Element: 5°X cut; xyt: 5°; length-width flexure mode.

Frequency Range: 10 to 100 kc.

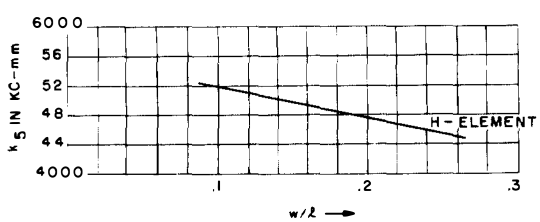


Figure 1-27. Frequency constant versus w/l ratio for element H

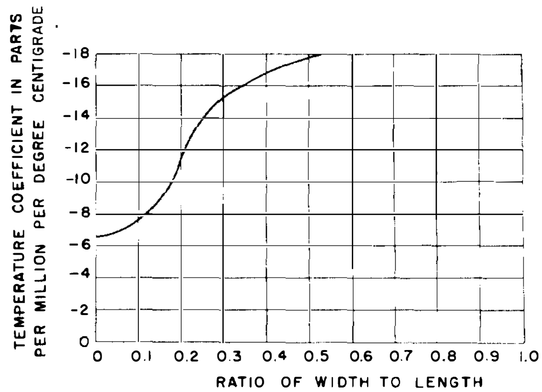


Figure 1-28. Temperature coefficient versus w/l ratio for element H *

Frequency Equation: $f = nk_5w/l^2$ $(n = 1, 2, 3, \ldots)$

Frequency Constant: Varies with w/l ratio (see figure 1-27).

Temperature Coefficient: Varies with w/l ratio (see figure 1-28).

Methods of Mounting: Wire, in vacuum; free-free flexures of long, thin bars have nodal points for the fundamental vibration at a distance of 0.224 x l from the ends; two electrically opposite pairs of electrodes are plated on each side of the YZ faces, with "ears" at the nodal points for soldering to the mounting wires. See figure 1-29. When the polarity of the lower electrodes causes a contraction of the bar, the polarity of the upper electrodes causes an extension, and vice versa—the over-all result being a flexural deformation.

Advantages: For long, thin bars the length-width flexural mode is resonant at much lower frequencies than is the length-extensional mode. This advantage, combined with the favorable electro-mechanical coupling, and reasonably low temperature coefficient, has made this element useful in very-low-frequency filters where only a single frequency is to be selected. When mounted in vacuum, a Q of 30,000 is obtainable.

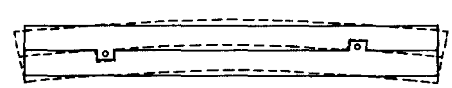


Figure 1-29. Element H, showing division of electrode plating for exciting fundamental mode. Similarly divided electrodes are on reverse side. The nodal "ears," where the mounting wires are attached, are at a distance of approximately 0.224 times the length from the ends *

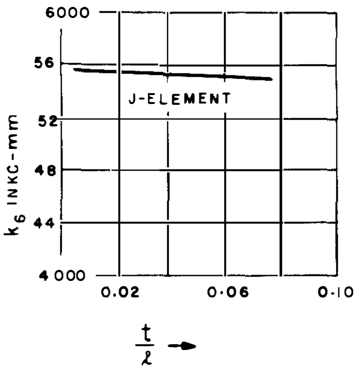


Figure 1-30. Frequency constant versus t/l ratio for element J

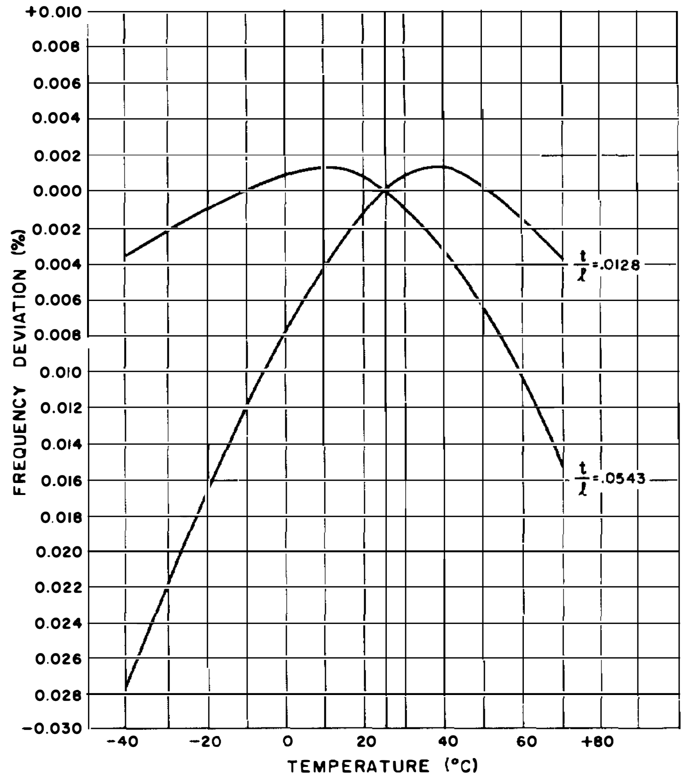


Figure 1-31. Percentage frequency deviation for J elements. The smaller t/l ratio is representative of a 1.2-kc element, and the larger t/l ratio is representative of a 10-kc element

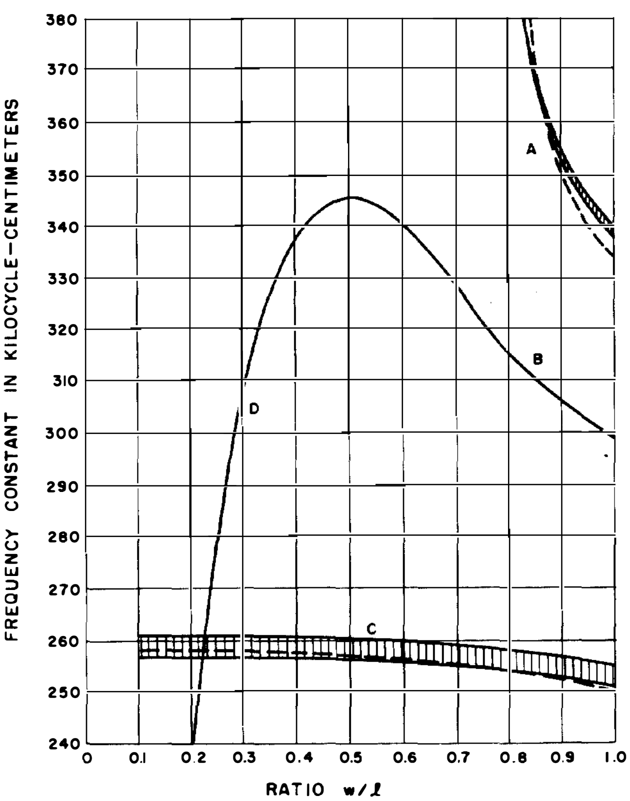


*Figure 1-32. Frequency constant versus w/l ratio for various resonances of —18° X-cut crystal. A is the width-extensional mode. B is the face-shear mode, which, at small w/l ratios, is strongly coupled to D, the second flexural mode. C represents the band between the antiresonant (upper curve) and the resonant (lower curve) frequencies of the length-extensional mode of element F. Note the weak coupling between C and D-B **

Disadvantages: The ratio of stored mechanical to electrical energy is not as large as that of the length-extensional mode, and because of this, the element does not give as broad a band-pass spectrum. Also, the effect of the shear stresses causes the temperature coefficient to become highly negative as the w/l ratio is increased. Finally, the damping effect of the air is greater for flexural than for other vibrations, so that flexure crystals should be mounted only in evacuated containers.

1-101. *CHARACTERISTICS OF ELEMENT J*

Description of Element: Duplex 5°X cut; xyt: 5° (right quartz), and xyt: 5° (left quartz); length-thickness flexure mode.

Frequency Range: 0.4 to 10 kc.

Frequency Equation: $f = nk_6t/l^2$ (n = 1, 2, 3, . . .)

Frequency Constant: Varies with t/l ratio (see figure 1-30).

Temperature Coefficient: Varies with both the t/l ratio and the temperature; figure 1-31 shows the total relative frequency deviation of two elements of different t/l ratios, the initial frequencies being those at 25°C. The temperature coefficients in parts per hundred at a given temperature are the slopes of the curves at that point. Note that the temperature at which a zero coefficient is obtained increases as the t/l ratio decreases. At temperatures below that of a zero-coefficient point, the coefficient is positive; at temperatures above, it is negative.

Methods of Mounting: Headed-wire, in vacuum; two thin plates are cemented together with polarities opposed so that only one pair of electrodes, plated on opposite YZ faces, are required; the crystal element is supported at the nodal points, which for the fundamental vibration are at a distance 0.224 x l from each end.

Advantages: Small temperature coefficient and low resonant frequencies (among the lowest obtainable with quartz) make this element useful in providing stable control for sonic-frequency oscillators, and as a component of single-frequency filters.

Disadvantages: Not economical for control of frequencies above 10 kc.

1-102. *CHARACTERISTICS OF ELEMENT F*

Description of Element: −18.5°X cut; xyt: −18.5°; length-extensional mode.

Frequency Range: 60 to 300 kc.

Frequency Equation: $f = \frac{nk_1}{l}$ (n = 1, 2, 3, . . .)

Frequency Constant: Varies slightly with w/l ratio (see figure 1-32).

Temperature Coefficient: 25 parts per million per degree centigrade—varies very little with changes in the w/l ratio.

Methods of Mounting: Wire, knife-edge clamp, pressure pins, cantilever clamp; more than one pair of electrodes required for overtones.

Advantages: The extremely weak coupling of this element to the face-shear and second flexure modes, represented by curves B and D, respectively, in figure 1-32, permits a better frequency spectrum than can be obtained with element E for w/l ratios greater than 0.1. For this reason, the F element used to be preferred over the E element as a filter plate, and was the principal quartz element in the channel filters of coaxial telephone systems. This is no longer true because channel filters now use +5°X plates which are smaller and conserve quartz.

Disadvantages: Relatively large temperature coefficient prevents this element from being preferred for oscillator control or as a channel filter if wide variations in temperature are to be expected. Also, the F plate is larger than the E plate of the same frequency and thus consumes more quartz.

1-103. *CHARACTERISTICS OF ELEMENT M*

Description of Element: MT cut; xytl: 0° to 8.5°/±34° to ±50°; length-extensional mode.

Frequency Range: 50 to 500 kc.

Frequency Equation: $f = \frac{nk_1}{l}$ (n = 1, 2, 3, . . .)

Frequency Constant: Varies with w/l ratio and angles of rotation (see figure 1-33).

Temperature Coefficient: Varies with w/l ratio and angles of rotation (see figure 1-34), and with the temperature. The total relative frequency deviation of an 8.5°/±34° M element, where the initial frequency is taken at 40°C, is shown in figure 1-35. Note that the temperature coefficient, which is the slope of the curve, changes from positive to negative as the temperature increases, with the zero coefficient occurring at 63°C.

Methods of Mounting: Wire, knife-edge clamp, pressure pins, cantilever clamp; more than one pair of electrodes required for overtones.

Advantages: The MT crystals were developed in an effort to overcome the large negative temperature coefficients of the X-cut and the 5°X-cut length-extensional modes for the larger w/l ratios. See figures 1-21 and 1-25. The unfavorable temperature characteristics are caused by the coupling of the extensional to the face-shear mode, the latter having a high negative temperature coefficient. However, if the crystal is rotated about its length, an orientation will be found where the face-shear mode has a zero temperature coefficient; that is, the coefficient will pass from negative to positive values. The low temperature coefficient of the length dimension will thus be preserved even though the coupling to the shear-mode has not, itself, been diminished. The low temperature coefficient makes the M element advantageous for oscillator control in the 50-to-100 kc range, and for use in narrow band filters, such as pilot-channel filters in carrier systems, where wide temperature ranges are to be encountered. The 8.5°/34° rotation with a w/l ratio of approximately 0.42 provides the greatest electromechanical coupling of the M elements, and hence the broadcast bandpass of the MT cut for use in filters.

Disadvantages: The electromechanical coupling rapidly decreases as the w/l ratio increases, so that at ratios greater than 0.7 the element is too selective for filter use, and of too small a piezoelectric activity to be advantageous for oscillator control. Maximum electromechanical coupling is obtained with w/l ratios of 0.39 to 0.42; but for a maximum bandwidth the E element is preferred. Although the interference of the face-shear temperature coefficient is reduced, the coupling to that mode remains relatively strong; so where the temperature varies very little, or where the secondary frequency effects are undesirable, the F element is preferred.

1-104. *CHARACTERISTICS OF ELEMENT N*

Description of Element: NT cut; xytl: 0° to 8.5°/±38° to ±70°; length-width flexure mode.

Frequency Range: 4 to 100 kc.

Frequency Equation: $f = \frac{nk_5w}{l^2}$ (n = 1, 2, 3, . . .)

Frequency Constant: Varies with w/l ratio (see figure 1-36).

Temperature Coefficient: For w/l ratios of 0.2 to 0.5, low coefficients are obtained by double rotations of 0° to +8.5°/±50°. Typical frequency deviation curves are shown in figure 1-37, where the initial temperature is taken at 25°C. Note that a zero temperature coefficient occurs at approximately 10°C. To produce a zero temperature coefficient at 25°C for w/l ratios of 0.05, the angles of rotation should be as shown in figure 1-38.

Methods of Mounting: Wire, in vacuum; special characteristics are the same as for the H element. See paragraph 1-100.

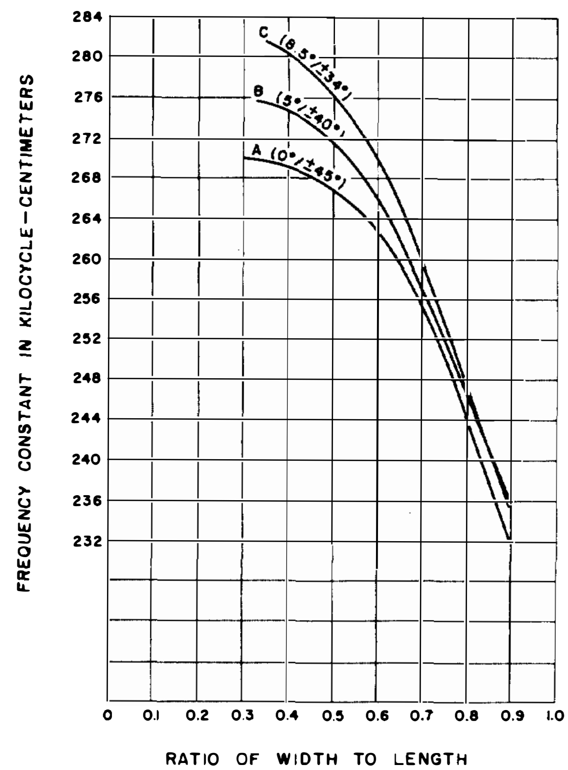


Figure 1-33. Frequency constant versus w/l ratio for M elements having low temperature coefficients. C is the curve of the most commonly used MT orientation*

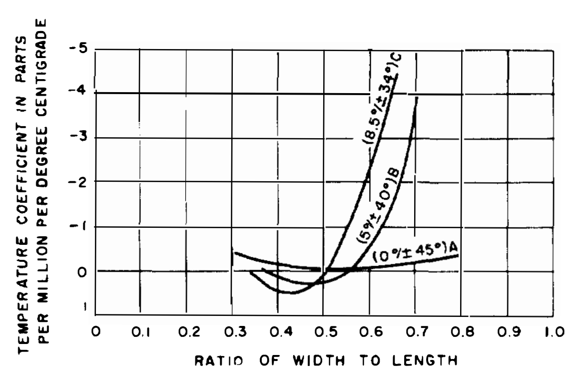


Figure 1-34. Temperature coefficient versus w/l ratio for M elements *

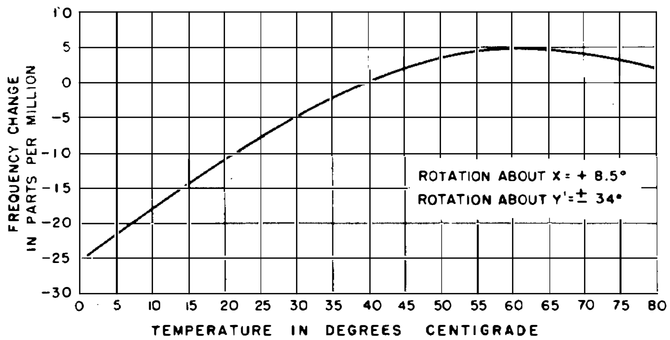


*Figure 1-35. Frequency-temperature characteristics of element M **

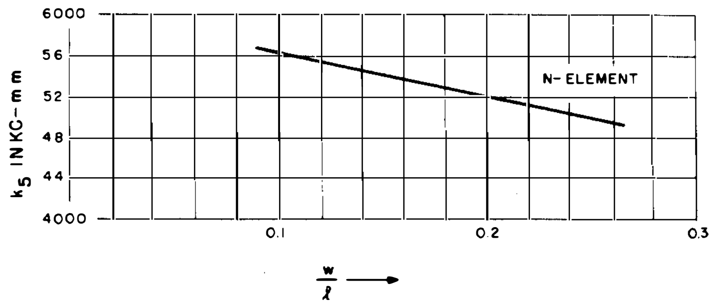


Figure 1-36. Frequency constant versus w/l ratio for element N

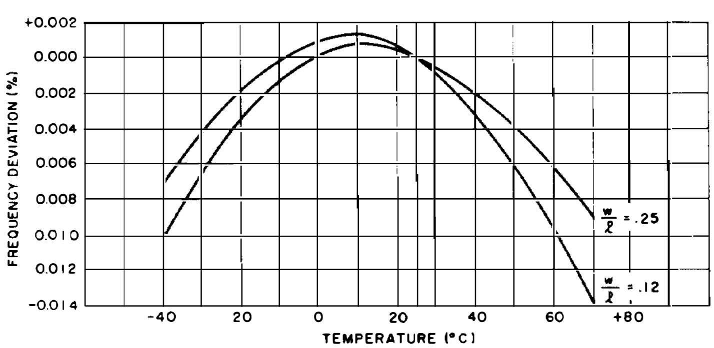


Figure 1-37. Frequency-temperature characteristics of element N. The larger w/l ratio is typical of 100-kc elements, and the smaller w/l ratio is typical of 16-kc elements

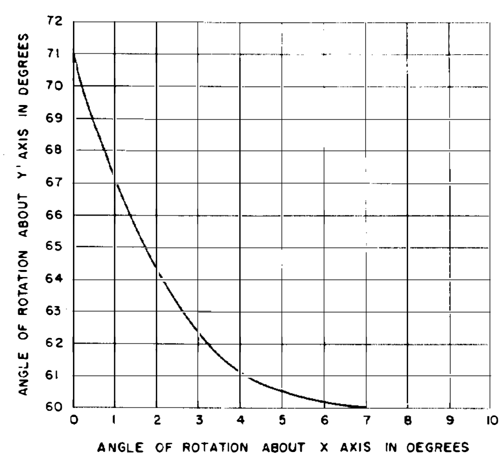


*Figure 1-38. Angles of rotation for N element with a w/l ratio of 0.05 which will provide zero temperature coefficient at 25°C **

Advantages: The principal advantage of the N element is that the second rotation reduces the temperature coefficient for the flexure vibration of long, thin crystals. This is accomplished by changing the width from near parallelism to the Z axis to near parallelism to the X axis. Theoretically the ideal rotation would be 90°, except that the piezoelectric effect would be reduced to zero. As a compromise, secondary rotations, about the length, of 39° to 70° are made. Besides reducing the flexure-mode temperature coefficient of the long, thin crystals, the rotation also reduces the negative coefficient for the shear modes at the higher w/l ratios, as in the case of the M element. Where wide temperature ranges must be met, this element is preferred for very-low frequency oscillators, and in single-frequency filter selectors. As the control element of an oscillator, it can maintain the frequency within ±0.0025% over a normal room-temperature range without temperature control.

Disadvantages: The electromechanical coupling is rather weak, more so for the larger than for the smaller w/l ratio. As a consequence, the bandwidth is too narrow for the element to be used as a band-pass filter of communication channels, and the piezoelectric activity is so low that special circuits are required for its use in oscillators.

THE V CUT

1-105. The V cut, developed by S. A. Bokovoy and C. F. Baldwin of RCA, is actually an entire series of cuts obtained by a sequence of double rotations of an initial X-cut plate. The first rotation angle, ϕ, is taken about the Z azis, and the second rotation angle, θ, is taken about the Y′ axis (the dimension of the crystal that is initially parallel to the Y axis). For each angle ϕ, there is an angle θ at which the crystal will have a given temperature coefficient for a particular mode of vibration. Normally, the combination of angles desired is one that will provide a zero temperature coefficient; however, it may be that a small positive or negative coefficient is required to counterbalance an opposite temperature coefficient inherent in the external circuit to which the crystal is to be connected. For this purpose curves of θ plotted against ϕ are shown in figures 1-40 to 1-41 for small positive and negative temperature coefficients, as well as for a zero temperature coefficient. Other ϕ and θ combinations may be extrapolated to give temperature coefficients differing from the actual values shown. It should be noted that when the rotation about the Z axis is equal to ±30°, the thickness dimension becomes parallel to a Y axis, and hence the crystal is in the position of the Y cut, with the Y′ axis coinciding with an X axis. Thus, if $\phi = \pm 30°$, the V cut is essentially the same as a rotated Y cut, and in this case would embrace practically the entire Y family. On the other hand, if $\phi = 0°$, the V cut becomes simply a singly rotated X cut—but with rotations about the Y axis, not the X axis as in the case of the 5°X and the −18°X cuts. However, when $\phi = 0°$, the V cut does overlap the MT and NT cuts.

1-106. *CHARACTERISTICS OF V-CUT PLATES IN THICKNESS-SHEAR MODE*

Description of Element: V cut; xzlw or xywl: 15° to 29°/−14° to −54° and 13° to 29°/27° to 42° (see temperature coefficient curves in figure 1-40 for exact ϕ and θ combinations); thickness-shear mode.

Frequency Range: 1000 to 20,000 kc (fundamental); higher frequencies on overtones.

Frequency Equation: $f = \frac{k_3}{t}$ (fundamental vibration when t << l and w). Figure 1-39 shows the frequency constant of the zero-temperature-coefficient series of V cuts as a function of the first rotation angle. The upper curve, designated k_3 $(+\theta)$, applies to positive angles of θ, the second rotation, whereas the lower curve, designated k_3 $(-\theta)$, applies to negative angles of θ.

Temperature Coefficient: Figure 1-40 shows the combinations of ϕ with positive values of θ that provide temperature coefficients of +15, 0, and −15 parts per million per degree centigrade, and those combinations of ϕ with negative values of θ that provide temperature coefficients of +5, 0, and −5 parts per million per degree centigrade.

Methods of Mounting: Sandwich, air gap, clamped air-gap, button.

Advantages: The principal advantage of the V cut is that a given temperature coefficient may be obtained from a large choice of orientations, and with a minimum in trial-and-error procedure. Not only can a series of zero-coefficient plates be obtained, but also plates with coefficients of desired sign and magnitude for annulling the known frequency-tem-

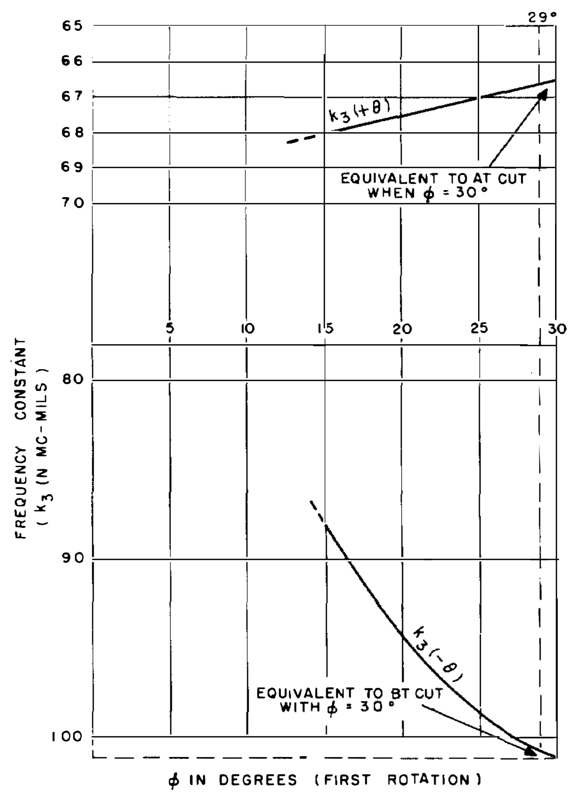


Figure 1-39. Frequency constant versus ϕ (angle of rotation about Z axis) for the thickness-shear mode of V-cut crystals when θ, the second angle of rotation, is so chosen that a zero temperature coefficient is obtained. The upper and lower curves are for positive and negative values of θ, respectively

perature effects of the circuits in which the plates are to be used. The V cut is the only member of the X group that provides a zero temperature coefficient for high-frequency vibrations; and because of the large choice of rotation angles, one or the other of the V orientations will frequently permit the maximum use of an unfaced or badly twinned mother crystal. Because their larger frequency constants permit a thicker and less fragile crystal, the orientations with a negative θ are preferred for the higher frequencies. Also, small deviations in negative values of θ produce less variation in the temperature coefficient than do the same deviations in positive values of θ. Hence, the negative orientations of θ are also generally more dependable for obtaining a desired temperature coefficient. On the other hand, positive values of θ permit a less bulky crystal for the lower frequencies, a less critical frequency constant, less interference from spurious frequencies, and for accurately determined orientations, a broader temperature deviation for a given deviation in frequency. At $\phi = 30°$, the values of $\theta = -49°$, $+31°$, and $+35°31'$ are substantially the same as the BT, AC, and AT cuts, respectively, of the Y group, as described in paragraphs 1-114, 1-111, and 1-112. The chief use of the thickness-mode V cut is for the control of high-frequency oscillators.

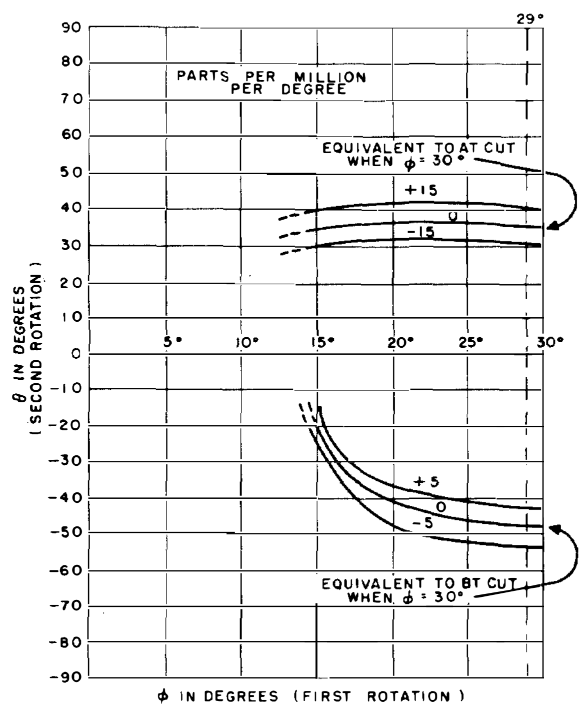


Figure 1-40. Relations of θ to ϕ, for thickness-shear mode of V cut, which provide the temperature coefficients indicated for each curve

Disadvantages: The possibility of spurious frequencies close to the desired fundamental is the most troublesome limitation of the V cut operating in a thickness-shear mode. As a general rule, the coupling between the desired and the stray modes diminishes as the initial rotation ϕ is increased. At values of ϕ less than 13°, the interference is too great for stable operation. Because of the relatively poor frequency spectrum, the V cut is not readily adaptable for use in selective networks. With a certain amount of cut-and-try experimentation, the more objectionable modes may be reduced by grinding down the width and length dimensions. For angles of ϕ close to 30° the length and width dimensions most important to avoid are approximately the same as those given in paragraphs 1-112 and 1-114 for the AT and BT cuts, respectively.

1-107. *CHARACTERISTICS OF V-CUT PLATES IN FACE-SHEAR MODE*

Description of Element: V cut; xzlw or xywl: 0° to 30°/±45° to ±70° (see temperature coefficient curves in figure 1-41 for exact ϕ and θ combinations); face-shear mode.

Frequency Range: 60 to 1000 kc.

Frequency Equation: $f = k_4/w$ (fundamental for square plates).

Frequency Constant: Insufficient data exist to plot the curve of k_4 for all the combinations of ϕ and θ corresponding to this element. However, in the case of the zero-coefficient plates, as the positive value of θ approaches 37.5°, k_4 approaches 3070 kc-mm, and as the negative value of θ approaches $-52.5°$, k_4 approaches 2070 kc-mm.

Temperature Coefficient: Figure 1-41 shows the combinations of ϕ and θ that provide temperature coefficients of +5, 0, and −5 parts per million per degree centigrade.

Methods of Mounting: Wire, cantilever clamp.

Advantages: The principal advantage is the low temperature coefficient, which makes the element useful for low-frequency oscillators and filters. The large choice of orientation angles is also advantageous for obtaining the maximum number of cuts from a given mother crystal, particularly if the presence of twinning or other defects limit the dimensions in

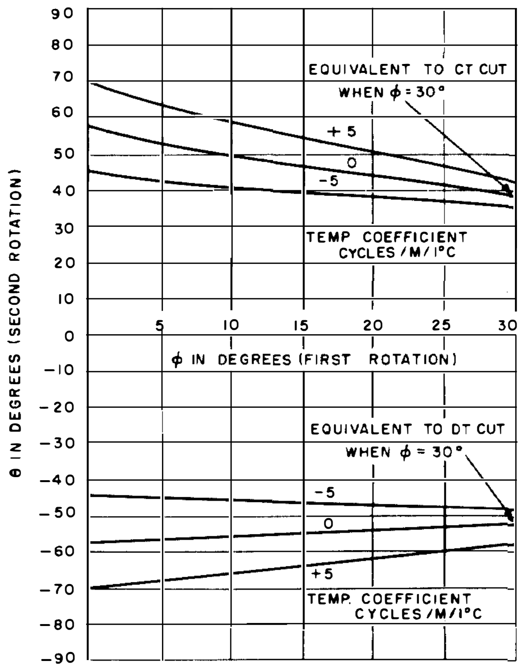


Figure 1-41. Relations of θ to ϕ, for face-shear mode of V cut, which provide temperature coefficients of 0, +5, and −5 parts, per million per degree centigrade

the directions at which rough bars would normally be cut. Also, the angles for small predetermined positive and negative coefficients permit a crystal to be cut which can exactly annul the known temperature effects of the external circuit. As indicated in figure 1-41, small deviations in the orientations angles will cause minimum deviations in the temperature coefficient when $\phi = 0°$ to $15°$, and θ is negative. On the other hand, maximum piezoelectric activity is obtained when ϕ is large, and θ is positive. As a general rule, the positive values of θ are used for the higher frequencies and the negative values of θ for the lower frequencies. The zero-temperature cuts for $\phi = 30°$ are substantially the same as the CT and DT cuts of the Y group. See paragraphs 1-115 and 1-116, respectively.

Disadvantages: Care must be taken that flexure modes are not strongly coupled to the face-shear mode. Such coupling may be reduced by making the plates square, or nearly so. For angles of ϕ approaching 30°, the thickness should be approximately within the limits given for the C and D elements in paragraphs 1-115 and 1-116.

The Y Group

1-108. The principal crystals of the Y group are listed below with the frequency range for which they have found commercial application:

Name of Cut	*Frequency Range in KC*
Y	1000 to 20,000
AC	1000 to 15,000
AT	500 to 100,000
BC	1000 to 20,000
BT	1000 to 75,000
CT	300 to 1100
DT	60 to 500
ET	600 to 1800
FT	150 to 1500
GT	100 to 550

Figure 1-42 shows the orientations of a yx initial position (Y cut with the length parallel to the X axis) for the various cuts. In special cases the width may be parallel to the X axis, but this is the exception rather than the rule, unless the plate is square or circular. With the exception of the GT cut, the crystals of the Y group are used in their shear modes—face shear for the low-frequency elements, and thickness shear for the high-frequency elements. The Y cut, itself, has a large positive temperature coefficient; and, because of coupling between the thickness-shear mode and the overtones of the face-shear mode, it also exhibits sharp irregularities in its frequency spectrum. However, by rotation about the X axis, zero temperature coefficients may be obtained, and the coupling between the shear modes can be greatly diminished. This coupling becomes zero at the angles of the AC and BC cuts, and the frequency constant of the thickness-shear mode has minimum and maximum values, respectively, for these two orientations. Figure 1-43 shows the thickness-shear frequency constant, and figure 1-44 the thickness-shear temperature coefficient, with each plotted as a function of the angle of rotation. For the face-shear mode, the frequency constant and the temperature coefficient are shown in figures 1-45 and 1-46, respectively, plotted as functions of the angles of rotation.

THE Y CUT

1-109. The Y cut was introduced commercially in the late 1920's, at which time its principal advantage was that it could be clamped at its edges, whereas the X cut would not oscillate if the edge movement were even slightly restricted. The use of a Y cut, vibrating in a shear-mode, was originally suggested by E. D. Tillyer of the American Optical Company, to whom a U. S. patent was

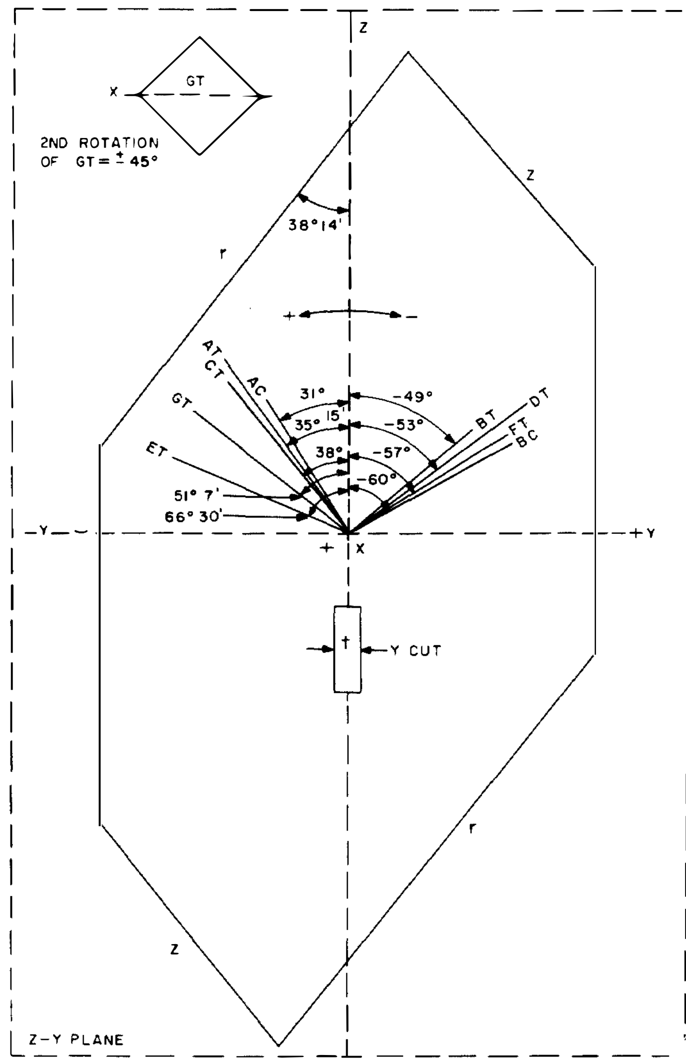


Figure 1-42. Rotation angles of Y cut about X axis which provide the principal members of the Y group. The GT cut is the only member having a second rotation (±45° about the Y′ axis). The +X sign indicates that the positive end of the X axis points toward the observer

issued in 1933. For this reason, the Y cut is sometimes called the *Tillyer cut.* For several years this crystal was used extensively in commercial and military transmitters mounted in mobile equipment, and also in commercial broadcast transmitters where the Y cut's readily excited oscillations permitted the use of crystal oscillators with low plate voltages. However, due to the strong coupling between the thickness-shear and the overtones of the face-shear and flexure modes, the Y cut's frequency spectrum is very poor. Also, small irregularities in the dimensions of the crystal readily produce abrupt changes in the frequency. A typical frequency-temperature curve of a Y-cut crystal is shown in figure 1-47. Today, the Y cut has been almost entirely replaced by the rotated cuts having small temperature coefficients, and the Y cut's only major application now is that of transducer for generating shear vibrations in solids.

1-110. *CHARACTERISTICS OF Y-CUT PLATES IN THICKNESS-SHEAR MODE*

Description of Element: Y cut; yx or yz; thickness-shear mode.

Frequency Range: 500 to 20,000 kc; much lower

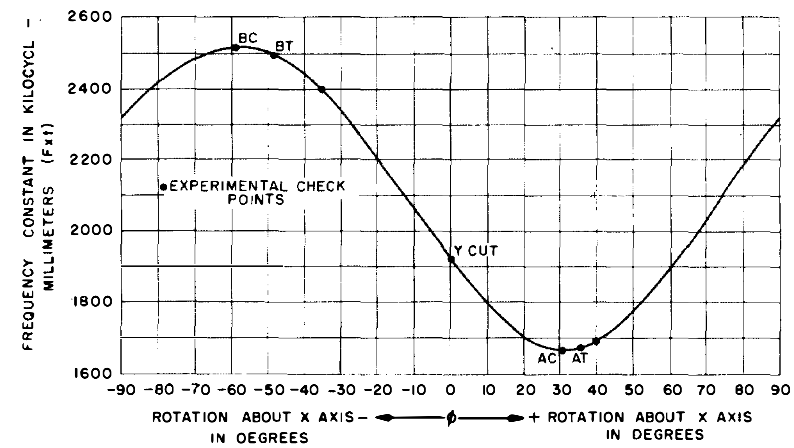


Figure 1-43. Frequency constant versus angle of rotation about X axis for thickness-shear elements of Y group. (Values shown for Y, BT, and BC cuts are smaller than the average) *

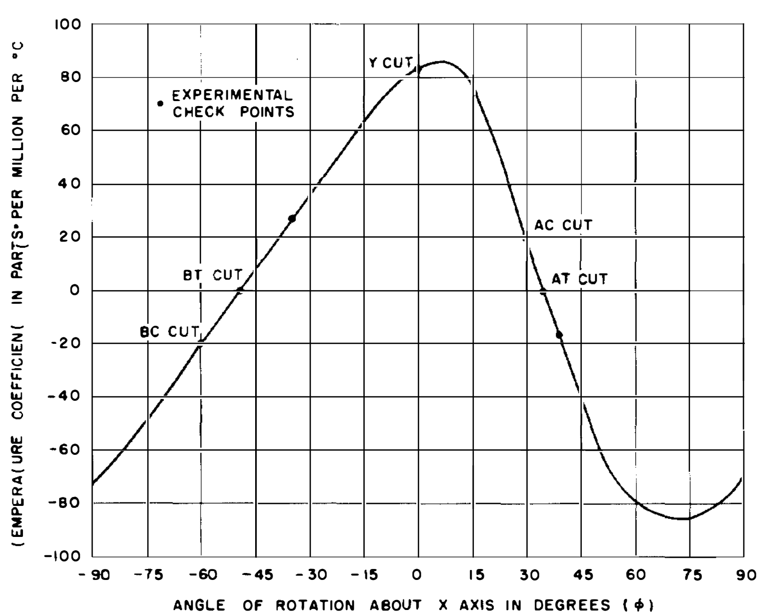


Figure 1-44. Temperature coefficient versus angle of rotation about X axis for thickness-shear elements of Y group *

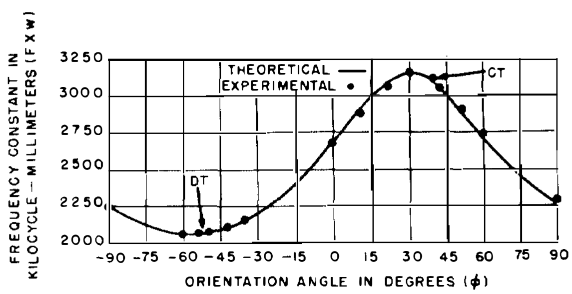


Figure 1-45. Frequency constant versus angle of rotation about X axis for face-shear elements of Y group

frequencies when bonded to solids for use as transducer.

Frequency Equation: $f = \frac{k_3}{t}$ (fundamental vibration).

Frequency Constant: $k_3 = 1981$ kc-mm (average value).

Temperature Coefficient: Varies with dimensions of crystal and with temperature but is usually between 75 and 125 parts per million per degree centigrade, and is positive, with an average value of 86 parts per million per degree centigrade.

Methods of Mounting: Sandwich; air gap, clamped air gap; bonded to solids when used as transducer.

Advantages: Ratio of stored mechanical to electrical energy is larger than that of any other quartz element; this large ratio, combined with the quartz crystal's superior strength, makes the Y cut desirable as a generator of shear vibrations in solids for the purpose of measuring or testing the solids' physical properties. This element is the easiest of all quartz cuts to excite into vibration, and thus requires the lowest voltages for operation. Large temperature coefficient makes element useful as a sensitive detector of variations in temperature.

Disadvantages: Large temperature coefficient, discontinuities of resonant frequencies, and poor frequency spectrum make this element a secondary choice for use in either oscillator or filter circuits. Special Y cuts, such as the block- and doughnut-shaped crystals in figure 1-48, vibrate in a combination mode composed of coupled shear and flexure modes, and have zero temperature coefficients at certain temperatures. However, because of the prevalence of spurious frequencies, the large volume of quartz used per cut, and the difficulties of mounting, these crystals have little practical use.

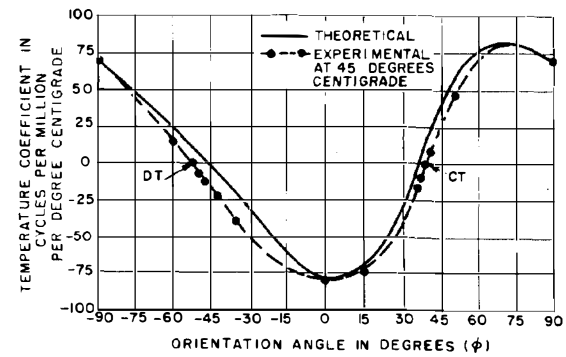


Figure 1-46. Temperature coefficient versus angle of rotation about X axis for face-shear elements of the Y group

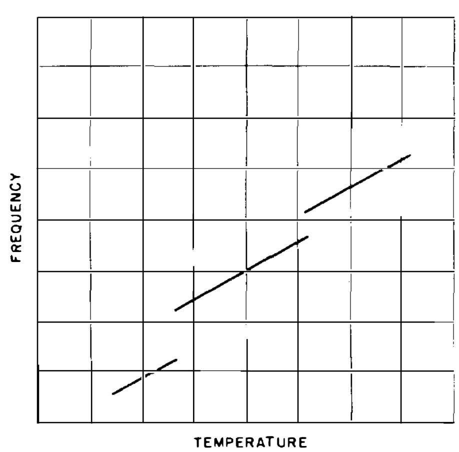


Figure 1-47. Temperature-frequency characteristics typical of the Y-cut, thickness-shear element. The frequency jumps are most apt to occur when small discrepancies are present in the thickness-dimension

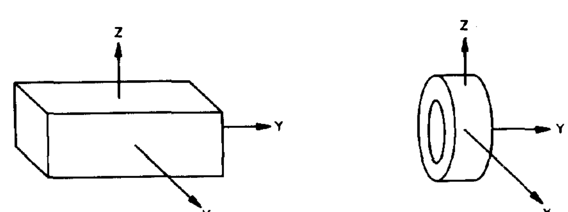


Figure 1-48. Y-cut block and doughnut-shaped crystals which can provide zero temperature coefficients for certain combination modes

1-111. *CHARACTERISTICS OF AC-CUT PLATES IN THICKNESS-SHEAR MODE*

Description of Element: AC cut; yxl: 31°; length-thickness-shear mode.

Frequency Range: 1000 to 15,000 kc (fundamental vibration).

Frequency Equation: $f = \frac{k_3}{t}$ (fundamental vibration when $t << l$ and w).

Frequency Constant: $k_3 = 1656$ kc-mm.

Temperature Coefficient: 20 parts per million per degree centigrade; positive.

Methods of Mounting: Sandwich, air gap, clamped air gap, button.

Advantages: This element vibrates in a very pure length-thickness mode with an excellent frequency spectrum. It has the lowest frequency constant of all the quartz thickness modes and thus permits a smaller thickness, and hence a more economical cut, for use at the low end of the high-frequency spectrum. For a given temperature, the electrical parameters of an AC crystal unit can be predetermined with an accuracy equal to, or greater than, that of the more commonly used AT units.

Disadvantages: The principal disadvantages of the AC cut is its relatively large temperature coefficient; because of this the element has found little commercial use, and the low-coefficient AT cut, with an orientation close enough to that of the AC for the coupling between the shear modes to be small, is generally preferred.

1-112. *CHARACTERISTICS OF ELEMENT A*

Description of Element: AT cut; yxl: 35°21′; length-thickness-shear modes; or, yzw: 35° 21′; width-thickness-shear mode.

Frequency Range: 500 to 1000 kc (special cuts); 1000 to 15,000 kc (fundamental vibration); 10,000 to 100,000 kc (overtone modes).

Frequency Equation:

$f = k_3/t$ (fundamental vibration when $t << l$ and w)

$$f = k_3 \sqrt{\frac{m^2}{t^2} + a_1\frac{n^2}{l^2} + a_2\frac{(p-1)^2}{w^2}}$$

where m, n, and p are integers.

Frequency Constant: $k_3 = 1660$ kc-mm.

Temperature Coefficient: 0.0 at 25°C; figure 1-49 shows the total relative frequency deviation for the normal maximum, minimum, and average angles of this element; the temperature coefficient at each point on a curve is the slope at that point in parts per hundred. At $\phi = 35°15'$, the temperature coefficient will vanish at 45°C, changing from negative to positive as the temperature increases. Optimum orientations for zero coefficients at other temperatures are given below:

Deg. C	ϕ
20	35°18′
20	35°27′ (overtones)
75	35°31′
75	35°33′ (overtones)
85	35°33′
85	35°36′ (overtones)
100	35°36′
190 (max.)	36°26′

Methods of Mounting: Sandwich, air gap, clamped air gap, button.

Advantages: The excellent temperature-frequency characteristics make this element preferred for high-frequency oscillator control wherever wide variations of temperature are to be encountered; it is particularly applicable

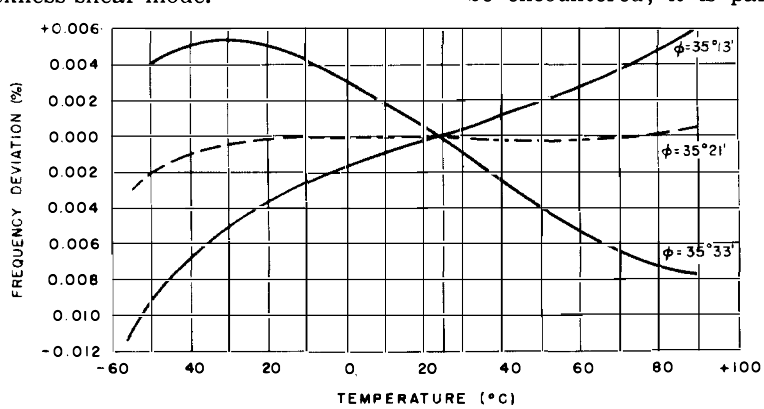


Figure 1-49. Temperature-frequency characteristics of element A

for aircraft radio equipment where sharp changes in temperature may be frequent, but where the added weight of constant-temperature ovens is undesirable. The angle of orientation is sufficiently close to that of the AC cut for the coupling between the shear modes to be weak, so that the resonant frequency can be isolated from that of other modes, except for certain dimensions of w and l. The A element also shares the low frequency constant of the AC element, and this is preferred for frequencies at the low end of the high-frequency spectrum. However, because of its superior temperature-frequency and piezoelectric characteristics as compared with the BT characteristics, the A element may well be preferred for the control of frequencies in the vhf range, even though the BT cut has the larger frequency constant.

Disadvantages: Because of the large thickness dimensions that would be required, the A element is generally not economical for the generation of frequencies below 1000 kc, although special circular cuts have been used at frequencies as low as 500 kc. In its normal high-frequency range, the most troublesome problem is to find the proper length and width dimensions which will permit the desired frequency to be widely separated from other modes. Although the orientation of the AT cut is close to that of the AC cut, there still exists a fair amount of coupling to the face-shear modes, and to the extensional and flexural modes along the X axis. Sufficient information is not available to avoid a certain amount of trial and error in grinding and finishing an AT blank to provide an optimum frequency spectrum at a desired frequency; however, there are certain X and Z′ (the dimension parallel to the Z axis before rotation) values that can be avoided by use of the frequency equations which hold approximately for the less complex of the unwanted modes. The following equations give the face dimensions of an AT cut which will produce unwanted resonances at the same frequency, f, as the thickness-shear mode.

For extensional modes along X:

$$X = \frac{2438\ n}{f} \quad (n = 1, 3, 5, \ldots)$$

For flexure modes along X:

$$X = \frac{1338.4\ n}{f} \quad (n = 2, 4, 6, \ldots)$$

For shear modes along X:

$$X = \frac{2542.0\ n}{f} \quad (n = 1, 3, 5, \ldots)$$

For shear modes along Z′:

$$Z' = \frac{2540.0\ n}{f} \quad (n = 1, 3, 5, \ldots)$$

With f expressed in kc, X and Z′ are given in mm. Either X or Z′ may be the length, with the other dimension being the width. It has been found that the unwanted modes are somewhat restricted by giving the plate a convex contour, and also by the use of circular plates. The convex contour is possible for all but the very thin plates that are used at frequencies above 15,000 kc. A 1000-kc crystal may have a contour of 3 to 5 microns. The equations above hold for flat plates, and become increasingly in error as the contour is increased.

1-113. *CHARACTERISTICS OF BC-CUT PLATES IN THICKNESS-SHEAR MODE*

Description of Element: BC cut; yxl: −60°; length-thickness-shear mode.

Frequency Range: 1000 to 20,000 kc (fundamental vibration).

Frequency Equation: $f = \frac{k_3}{t}$ (fundamental vibration when $t << l$ and w).

Frequency Constant: $k_3 = 2611$ kc-mm.

Temperature Coefficient: 20 parts per million per degree centigrade; negative.

Methods of Mounting: Sandwich, air gap, clamped air gap, button.

Advantages: The advantages and disadvantages of the BC cut are similar to those of the AC cut, except that the BC thickness-shear frequency constant is the highest obtainable for a rotated Y cut. A BC cut may thus have a greater thickness for the same frequency, and hence be less likely to be shattered from overdrive or mechanical shock—a distinct advantage at the higher fundamental frequencies where very thin crystals are used. Since the BC orientation is the negative angle of rotation which provides zero coupling between the shear modes, the element vibrates in a very pure length-thickness mode with an excellent frequency spectrum. For a given temperature, the electrical parameters of a BC crystal unit can be predetermined with an accuracy equal to or greater than that of the more commonly used BT units.

Disadvantages: As in the AC cut, the principal disadvantage of a BC cut is its relatively large temperature coefficient. Because of this, the element is not widely used, and the zero-coefficient BT cut, with an orientation sufficiently near to that of the BC cut to have a weak coupling between the shear modes, is used instead. An added disadvantage is that the magnitude of the rotation away from the Y axis is approximately double that for the AC cuts. For this reason the piezoelectric coefficient is smaller for the BC than for the AC or AT cuts, and, hence, somewhat higher voltages are required to maintain oscillations.

1-114. *CHARACTERISTICS OF ELEMENT B*

Description of Element: BT cut; yxl: —49°8′; length-thickness-shear mode; or, yzw: —49°8′; length-width-shear mode.

Frequency Range: 1000 to 20,000 kc (fundamental vibration).
15,000 to 75,000 kc (overtone modes).

Frequency Equation:

$f = k_3/t$ (fundamental vibration when $t << l$ and w)

$$f = k_3\sqrt{\frac{m^2}{t^2} + a_1\frac{n^2}{l^2} + a_2\frac{(p-1)^2}{w^2}}$$

where m, n, and p are integers.

Frequency Constant: $k_3 = 2560$ kc-mm.

Temperature Coefficient: 0.0 at 25°C; figure 1-50 shows the total relative frequency deviation for the normal maximum, minimum, and average angles of this element; the temperature coefficient in parts per hundred per degree centigrade at each point on a curve is the slope at that point. Zero coefficients are obtained at 20°C and 75°C when ϕ is —49°16′ and —47°22′, respectively.

Methods of Mounting: Sandwich, air gap, clamped air gap, button.

Advantages: The temperature-frequency characteristics make this element useful for high-frequency oscillator control where the temperature is not expected to vary too widely from the mean value. It is particularly applicable for use in radio equipment which is to operate at the high end of the high-frequency spectrum. Most of the high-frequency crystal oscillators employ either the BT or the AT cut, with the B element, because of its larger frequency constant, often preferred at frequencies from 10 to 20,000 kc. Since the orientation angle is near that of the BC cut, the shear modes are not too strongly coupled together; and, when ground to proper dimensions, the B element exhibits a reasonably satisfactory frequency spectrum.

Disadvantages: Like the A element, the B element is not suitable for use at the lower frequencies because of the large thickness dimensions

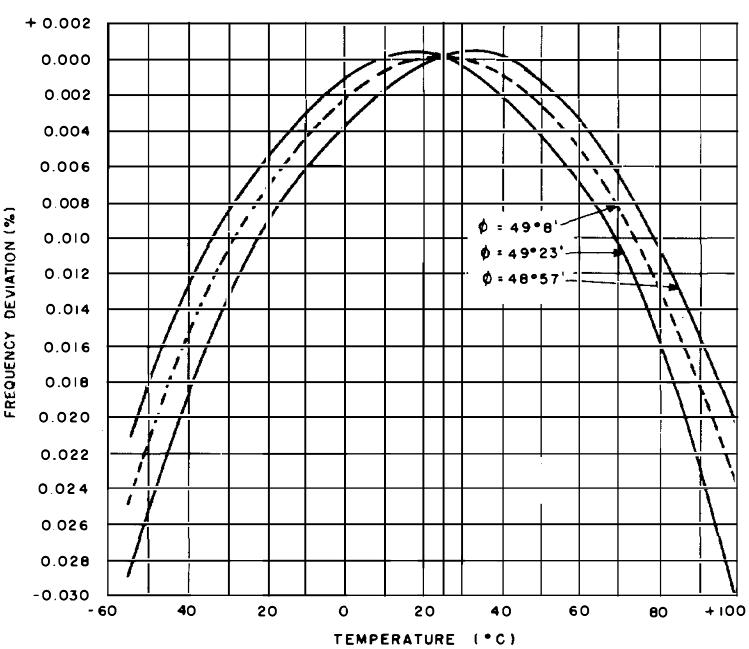


Figure 1-50. Temperature-frequency characteristics of element B

that would be required. Because of its greater angle of rotation from the Y axis, the BT has a smaller piezoelectric coefficient than the AT cut, and hence requires a higher voltage to maintain oscillations. Also, the temperature coefficient of the BT cut increases more rapidly than that of the AT cut when the temperature varies to either side of the zero point. Moreover, zero temperature coefficients cannot be obtained at as widely separated temperatures, as can be done with the AT cut by slightly varying the orientation angle. This limitation, however, becomes an advantage inasmuch as it reduces the percentage error when cutting a crystal to provide a given temperature-frequency characteristic. The greatest problem in preparing a B element is to avoid those length and width dimensions that cause the frequencies of unwanted modes to approach the frequency of the desired mode. As in the case of the AT cut, a BT blank with a good frequency spectrum will require a certain amount of trial and error in the finishing process. For the simpler modes of lower order, the following equations give the face dimensions of a BT cut which produce unwanted resonances of the same frequency, f, as that of the thickness-shear mode.

For flexure modes along X:

$$X = \frac{1810\text{ n}}{\text{f}} \quad (\text{n} = 2, 4, 6, \ldots)$$

For shear modes along X:

$$X = \frac{1635.14\text{ n}}{\text{f}} \quad (\text{n} = 1, 3, 5, \ldots)$$

For shear modes along Z′:

$$Z' = \frac{1664.5\text{ n}}{\text{f}} \quad (\text{n} = 1, 3, 5, \ldots)$$

With f expressed in kc, X and Z′ are given in millimeters. (Z′ is the dimension of the rotated Y cut that originally was parallel to the Z axis.) Either X or Z′ may be the length, with the other dimension being the width. As in the case of the A element, a convex contour of a plate will aid in restricting unwanted modes. At 1000 kc the contour may be as great as 5 microns; the thin, 20,000-kc plates, however, must be flat. The equations above hold only for flat plates, but are approximately correct if the contour is very small.

1-115. *CHARACTERISTICS OF ELEMENT C*

Description of Element: CT cut; yxl or yzw: 37° to 38°; face-shear mode.

Frequency Range: 300 to 1100 kc.

Frequency Equation: $f = k_4/w$ (fundamental of square plate).

$$f = k_4' \sqrt{\frac{m^2}{l^2} + a_1 \frac{n^2}{w^2}} \quad \begin{pmatrix} m = 1, 2, 3, \ldots \\ n = 1, 2, 3, \ldots \end{pmatrix}$$

Frequency Constant: k_4 = 3070 kc-mm. (Square plates are preferred since they have fewer secondary frequencies.)

Temperature Coefficient: 0.0 at 25°C for rotation angle of 37°40′. Figure 1-51 shows the total relative frequency deviation with temperature for maximum, minimum, and average angles of rotation for a nominal cut of 37°40′; the initial temperature is taken at 25°C. The slope of a curve at any point is the temperature coefficient in parts per hundred per degree centigrade at that point. Note that as the rotation angle is increased, the zero coefficient is shifted to a higher temperature;

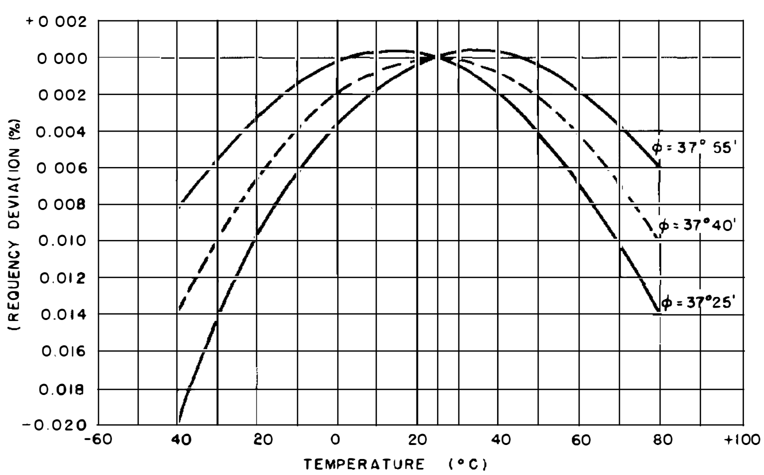


Figure 1-51. Temperature-frequency characteristics of element C

the same is true when the $\frac{Z'}{X}$ ratio is increased. For square plates, zero coefficients can be obtained at higher temperatures (50°C to 200°C) by rotation angles from 38°20′ to 41°50′, respectively.

Methods of Mounting: Wire, cantilever clamp.

Advantages: The CT cut is essentially a BT cut rotated approximately 90° so that the face shear of the C element corresponds to the thickness shear of the B element. This orientation provides a zero-temperature-coefficient shear mode for generating low frequencies, without requiring a crystal of large thickness dimension. The frequency characteristics of the C element, as compared with the D element, are roughly analogous to those of the B with the A element, except that the former pair vibrate at low frequencies, and the latter at high frequencies. The C element has the higher frequency constant, so is generally preferred over the D element at the high end of the low-frequency range. The C element is widely used both for low-frequency oscillator control and in filters, and does not require constant temperatures control under normal operating conditions. One of its principal applications has been as the control element in frequency-modulated oscillators.

Disadvantages: Because of its larger frequency constant, the C element must be cut with larger face dimensions than the D element to provide the same frequency of vibration. Thus, for the generation of very low frequencies the smaller DT cut is the more economical to use. Care must be taken that flexure modes are not strongly coupled to the face-shear mode. To prevent a coincidence of resonance between the two modes, the following thick-nesses have been used:

Frequency Range in KC	*Thickness in Mils*
370 to 428	18.5 to 19.9
428 to 475	16.0 to 17.5
475 to 540	18.5 to 19.9
730 to 875	12.0 to 14.0
875 to 1040	16.0 to 17.5

1-116. *CHARACTERISTICS OF ELEMENT D*

Description of Element: DT cut; yxl or yzw: −52° to −53°; face-shear mode.

Frequency Range: 60 to 500 kc.

Frequency Equation: $f = k_4/w$ (fundamental of square plate).

$$f = k_4' \sqrt{\frac{m^2}{l^2} + a_1 \frac{n^2}{w^2}} \quad \begin{matrix} (m = 1, 2, 3, \dots) \\ (n = 1, 2, 3, \dots) \end{matrix}$$

Frequency Constant: $k_4 = 2070$ kc-mm. (Square plates are preferred since they have fewer secondary frequencies.)

Temperature Coefficient: 0.0 at 25°C for rotation angle of −52°30′. Figure 1-52 shows the total relative frequency deviation with temperature for maximum, minimum, and average angles of rotation for a nominal cut of −52°30′, where the initial temperature is taken at 25°C. The slope of a curve at any point is the temperature coefficient in parts per hundred per degree centigrade at that point. Note that as the rotation angle is increased, the zero coefficient is shifted to a higher temperature. The upper limit for a zero coefficient is approximately 200°C, when $\phi = -54°$.

Methods of Mounting: Wire, cantilever clamp.

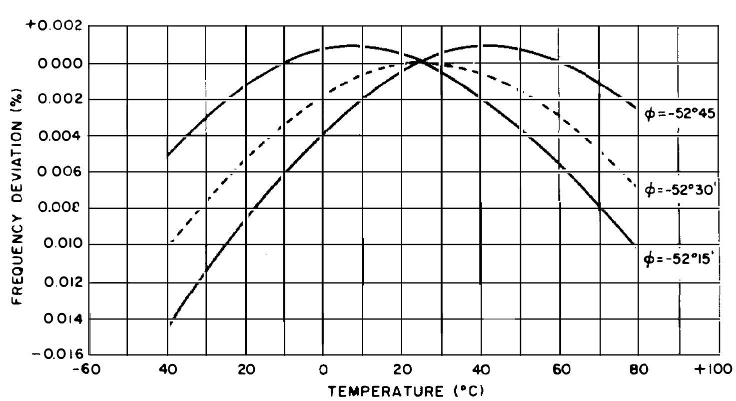


Figure 1-52. Temperature-frequency characteristics of element D

Advantages: The DT cut is essentially an AT cut rotated approximately 90° so that the face shear of the D element corresponds to the thickness shear of the A element. This orientation provides a zero-temperature-coefficient shear mode for generating low frequencies, without requiring a crystal of large thickness dimension. Because the frequency constant is less than that of element C, the face dimensions of element D are smaller for a given frequency, and hence the DT is the more economical cut for use at very low frequencies. Like the C element, the D element is widely used in both oscillators and filters, and does not require constant temperature control under normal operating conditions.

Disadvantages: At frequencies above 500 kc, the impedance effects introduced by the mounting become excessive, since the contact surfaces between the crystal and the supporting wires become rather large compared with the area of the crystal faces. Hence, the C is preferred over the D element in the 500—1000 kc range, since the higher frequency constant of the former permits a larger crystal face. To avoid strong coupling of the face-shear mode with flexure modes, certain thicknesses must be avoided. For most frequencies, however, a thickness of approximately 17 mils is satisfactory.

1-117. *CHARACTERISTICS OF ET-CUT PLATES IN COMBINATION MODE*

Description of Element: ET cut, with $\frac{w}{l}$ ratio approximately equal to 1.0; yxl or yzw: 66°30′; combination of coupled modes with second flexural vibration appearing to dominate a face-shear harmonic.

Frequency Range: 600 to 1800 kc.

Frequency Equation: $f = k/w$ (square plate).

$$f = \frac{2k}{l + w} \text{ (nearly square plate).}$$

Frequency Constant: $k = 5350$ kc-mm.

Temperature Coefficient: 0.0 at 75°C; see figure 1-53 for total relative frequency deviation.

Methods of Mounting: Wire; preferably mounted in vacuum.

Advantages: Besides its zero temperature coefficient, the principal advantage of the ET cut is its high frequency constant, which is almost 1.8 times that of the C element. This permits an effective extension of the frequency range for this type of plate and mounting. Optimum performance is obtained at temperatures near 75°C, so that the element is particularly advantageous where crystal ovens are used.

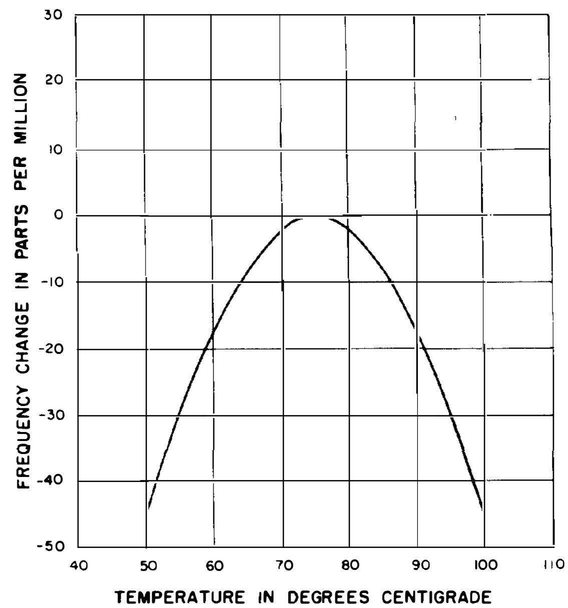


Figure 1-53. Temperature-frequency characteristics of ET-cut plate vibrating in combination mode *

Disadvantages: Stability and general performance are inferior to those that can usually be obtained by using, according to the particular frequency, either an A or a C element.

1-118. *CHARACTERISTICS OF FT-CUT PLATES IN COMBINATION MODE*

Description of Element: FT cut, with w/l ratio approximately equal to 1.0; yxl or yzw: —57°; combination of coupled modes with second flexural vibration appearing to dominate a face-shear harmonic.

Frequency Range: 150 to 1500 kc.

Frequency Equation: $f = \frac{k}{w}$ (square plate);

$$f = \frac{2k}{l + w} \text{ (nearly square plate).}$$

Frequency Constant: $k = 4710$ kc-mm.

Temperature Coefficient: 0.0 at 75°C; see figure 1-54 for total relative frequency deviation.

Methods of Mounting: Wire; preferably mounted in vacuum.

Advantages: The advantages of the FT cut are approximately the same as that of the ET, except that the FT has a lower frequency constant. The FT is related to the ET in approximately the same way that the DT is

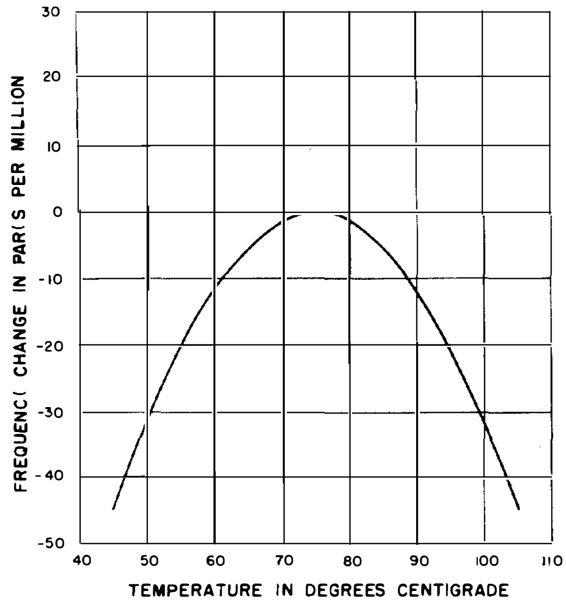


Figure 1-54. Temperature-frequency characteristics of FT-cut plate vibrating in combination mode *

related to the CT. However, the frequency constants of the ET and FT are approximately twice that of the low-frequency shear elements, so that these cuts can be made in practical sizes for twice the frequencies obtainable from the CT and DT crystals. Like the ET, the FT cut is particularly suited for use in ovens at temperatures between 70° and 80°C.

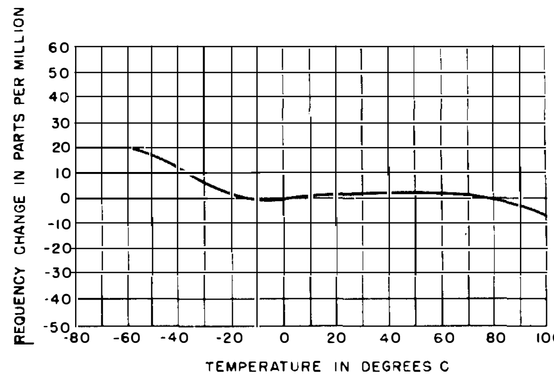


Figure 1-55. Temperature-frequency characteristics of element G *

Disadvantages: Stability and general performance are inferior to those that can usually be obtained by using, according to the particular frequency, either an A, C, or a D element.

1-119. *CHARACTERISTICS OF ELEMENT G*

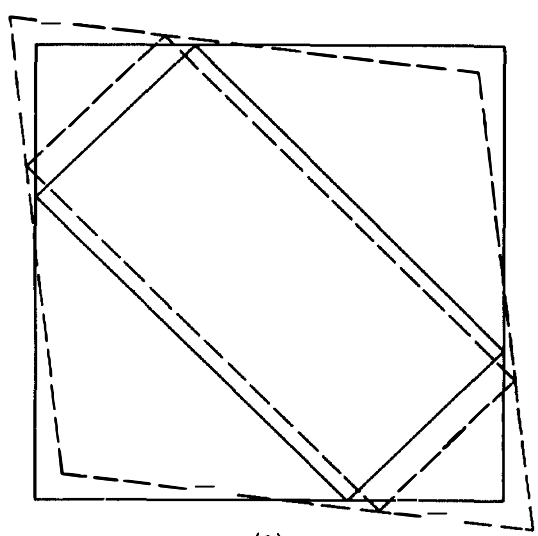
Description of Element: GT cut, with ratio $\frac{w}{l}$ = 0.859; yxlt: 51°7.5′/45°; width-extensional mode.

Frequency Range: 100 to 550 kc.

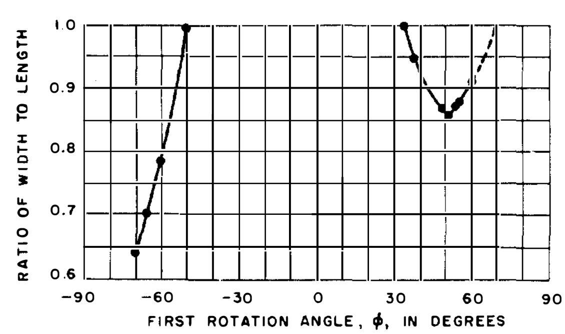
Frequency Equation: k_1/w (fundamental).

Frequency Constant: k_1 = 3370 kc-mm.

Temperature Coefficient: Very nearly zero over the range from −25° to +75°. Figure 1-55 shows the total relative frequency deviation at



(A)



(B)

Figure 1-56. (A) Diagram illustrating the equivalence between a face-shear mode and the length- and width-extensional modes of a rectangular plate which has been cut diagonally with respect to the face-shear element. (B) w/l ratio vs rotation angle, ϕ, of element G providing zero temperature coefficient *

for a span of 100°C the frequency does not vary more than one part in a million from the center frequency. The midpoint of the flat portion of the curve can be shifted from 25°C to 50°C by increasing the initial orientation angle from 51°7.5′ to 51°30′; the zero coefficient range will thus extend from 0°C to 100°C. A temperature variation of ±15°C on either side of the midpoint will not change the frequency more than 0.1 part in a million.

Methods of Mounting: Wire, knife-edge clamp, pressure pins, cantilever clamp.

Advantages: The GT cut provides the greatest frequency stability that has yet been obtained from a quartz plate. Where other quartz elements have zero temperature coefficients at only one temperature, the G element will not vary more than one part in a million over a range of 100°C. This element was originally suggested by the fact that a face-shear mode consists of two extensional modes coupled to-together. When a face-shear element is rotated 45° the vibrations lose their shear effect and appear as two extensional modes—one along the width, and the other along the length. See figure 1-56 (A). Since all *pure* extensional modes must have a negative or zero temperature coefficient, a positive coefficient of a face-shear mode must be due to the coupling between its two extensional components. If the cut of a face-shear crystal having a positive coefficient has been rotated 45°, the coupling between the extensional modes can be reduced by grinding down one edge so that one of the modes will increase in frequency. As the frequencies become more widely separated, the extensional modes will approach their true negative temperature coefficients. At some ratio of width to length a zero coefficient will be obtained. The GT cut is properly a ±45° rotation of any positive-coefficient face-shear crystal in the Y group. Although the most satisfactory cut is the one described above, a number of other GT cuts have been investigated where the initial angle of rotation, ϕ, has had negative as well as positive values. Figure 1-56 (B) shows the w/l ratios for both positive and negative angles that will provide a zero temperature coefficient. For negative values of ϕ, the dominant mode is the one of lower frequency, whereas for positive angles of ϕ, the higher-frequency mode is dominant. The G element is used for the control of oscillators where the most precise accuracy is required, such as in the frequency standards of loran navigational systems, the time standards at the U. S. Bureau of Standards and at Greenwich Observatory, and in both fixed and portable frequency standards of general use. Other than in frequency and time standards, the GT cut is employed in filters that are designed for use under very exacting phase conditions.

Disadvantages: The principal disadvantage of a GT cut is its expense. To obtain optimum temperature-frequency characteristics requires pains-taking labor in cutting and grinding to the exact orientation and dimensions. Furthermore, the excellence of a particular cut will be of little advantage unless the mounting and the external circuit are also of superior design. For these reasons a G element is not particularly practical except when the utmost frequency precision is mandatory.

FABRICATION OF CRYSTAL UNITS

1-120. The production stages during the fabrication of a crystal unit may differ somewhat from one manufacturer to the next because of variations in the instruments, techniques, and the type of mounting employed. However, the general procedure is fundamentally the same throughout the industry—first, the inspection and cutting of the raw quartz; next the lapping and etching of the diced blanks; and finally, the mounting and testing of the crystal unit in its holder.

INITIAL INSPECTION OF RAW QUARTZ

1-121. The manufacture of a crystal unit begins with the inspection of the raw quartz for impurities, cracks, and inclusions. For this purpose, the arc lamp of the inspectoscope shown in figure 1-57 is used.

1-122. The inspectoscope tank is filled with a clear, colorless oil mixture having an index of refraction approximately the same as the average in quartz (1.52 to 1.56). Such a medium for transmitting the light to and from the raw crystal, or "stone," is necessary in order to see the interior, for otherwise reflections and refractions at the rough surface will not only create an excessive glare, but will diminish the intensity of light penetrating beyond.

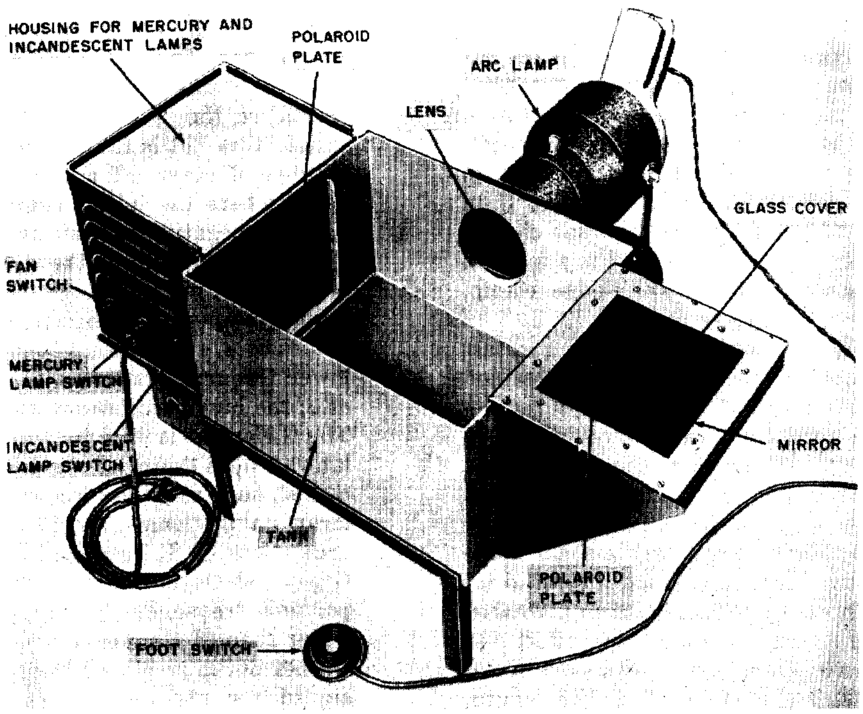


Figure 1-57. Polariscope-Inspectoscope. Used for examining raw quartz

The lamp incorporates a high-powered (500- to 1000-watt) projection system of white light, and inspection of the stone is performed by direct observation. The usable parts of the stone are marked; or if too many imperfections are present, the stone is discarded.

INSPECTION FOR OPTIC AXIS AND OPTICAL TWINNING

1-123. If a stone has retained some of its natural faces, the optic (Z) axis may be readily located by direct observation. In the usual case, however, it is necessary to use the plane-polarized light system that is provided by the inspectoscope. The light from a mercury or incandescent lamp is plane polarized by a polaroid plate placed between the lamp and the tank. On the opposite side of the tank is a second polaroid plate with its transmission axis perpendicular to that of the first, so that if a stone is not in the tank to rotate the polarity of the light, the rays will be stopped at the second plate. Light that does filter through, however, is reflected upward by a mirror, and the pattern may be observed through the glass cover shown in figure 1-57. When a stone is placed in the tank and oriented so that its optic axis is parallel to the rays, the polarity of the rays will be rotated and a bright image will be reflected from the mirror. If white light is used, a pattern of concentric rainbow colors will appear; and if monochromatic light is used, a pattern of concentric rings of light and darkness will appear. The optic axis will be exactly parallel with the light rays when the stone is in the position that yields the fewest and broadest bands. If optical twinning is present, it will be revealed by a fine-toothed pattern cutting across the rings, as indicated in figure 1-58. The twinning areas are more clearly indicated when white light is used, and when viewed slightly off the optic axis. On the other hand, monochromatic light produces ring patterns of maximum clarity for the determination of the optic axis itself. Flat surfaces are ground on opposite sides of the stone, parallel to the optic axis; and, with the stone resting on one of the flat surfaces at the bottom of the inspectoscope tank, a line is drawn on the upper surface to indicate the approximate Z-axis direction.

USE OF CONOSCOPE FOR EXACT DETERMINATION OF OPTIC AXIS

1-124. After the approximate optic (Z) axis is determined, the stone is cemented to a glass plate, and a small-end-section of the crystal is sliced off with a diamond saw, leaving a flat surface approximately perpendicular to the optic axis. The stone is then mounted on an adjustable orienting jig, which is placed against the reference edge in a conoscope tank. The conoscope (see figure 1-59) provides a polarized light system with which the optic axis may be accurately located by observing a concentric ring system. The principle of the conoscope is similar to that of the polarizing system of the inspectoscope, except that a converging lens system and a vernier system are provided that permit the optic axis to be determined with an accuracy of one degree. The handedness of the crystal is readily determined by rotating the second polaroid plate, or *analyzer*, of the conoscope. The quartz is right or left according to whether the concentric rings appear to expand or contract for a given direction of analyzer rotation. When the Z axis is accurately determined, each end is trimmed to form plane surfaces ("windows") exactly perpendicular to the Z direction.

SECTIONING THE STONE

1-125. There are three general methods of cutting the stone to obtain crystal blanks of desired orientation: the direct-wafering, X-block, and Z-section-Y-bar methods. In direct wafering, shown in figure 1-60, wafers are sliced directly from the stone at the desired orientation, and the blanks are diced from the wafer. The X-block method, as indicated in figure 1-61, is similar to that of direct

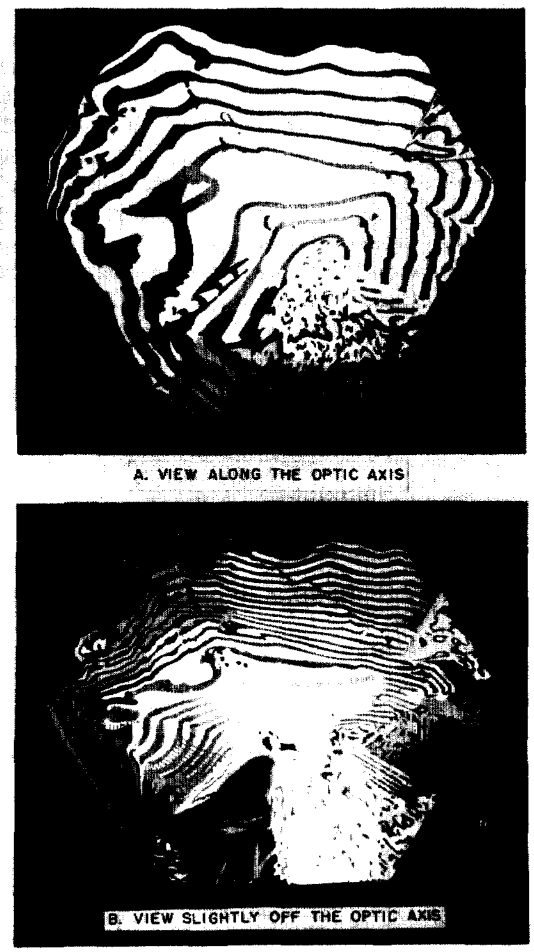


*Figure 1-58. Polarized-light view of pyramidal cap indicating optical twinning **

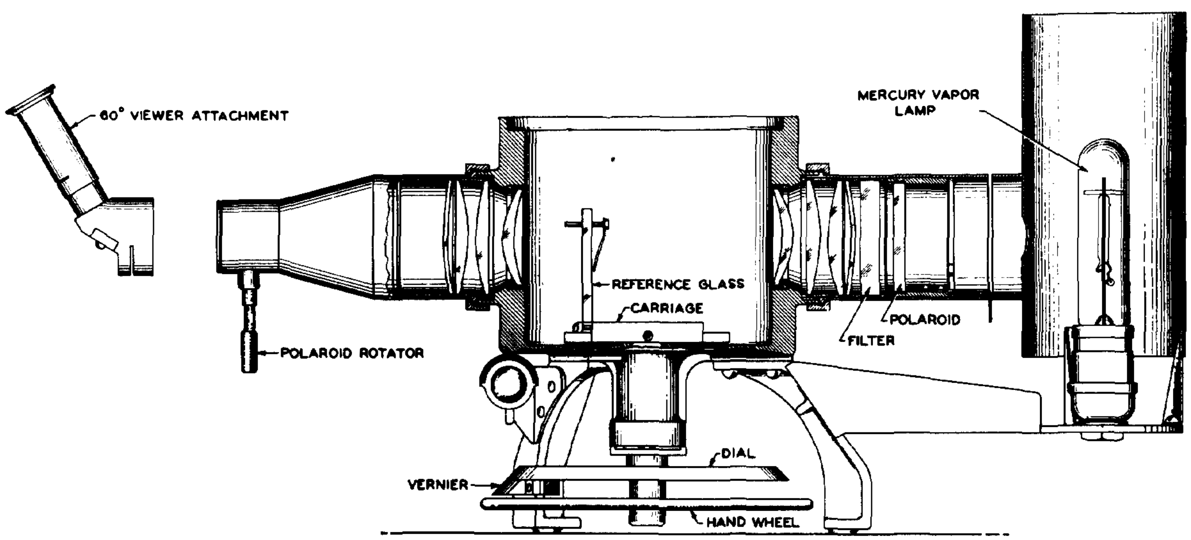
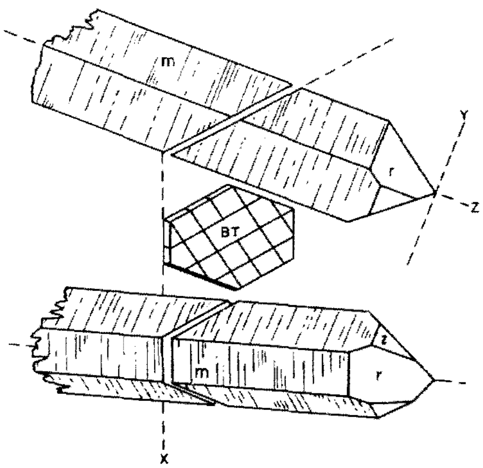


*Figure 1-59. Conoscope. Used for locating accurately the optic axis and for determining the handedness of quartz stones **



wafering, except that, before being sliced into oriented wafers, the stone is cut into one or more blocks with place surfaces at each end of the Z axis, and at the ends of one of the X axes; each surface is accurately cut at right angles to the axis it terminates. It is from these "X" blocks that the properly oriented wafers are cut and then diced into blanks. The third method of cutting proceeds as indicated in figure 1-62. The stone is sliced into Z sections (cross-sectional slabs with plane faces perpendicular to the Z axis) ; the Z sections are cut into Y bars (bars with the length parallel to the Y axis) ; and crystal blanks are sliced at the desired

Figure 1-60. Direct-wafering method of cutting crystal blanks

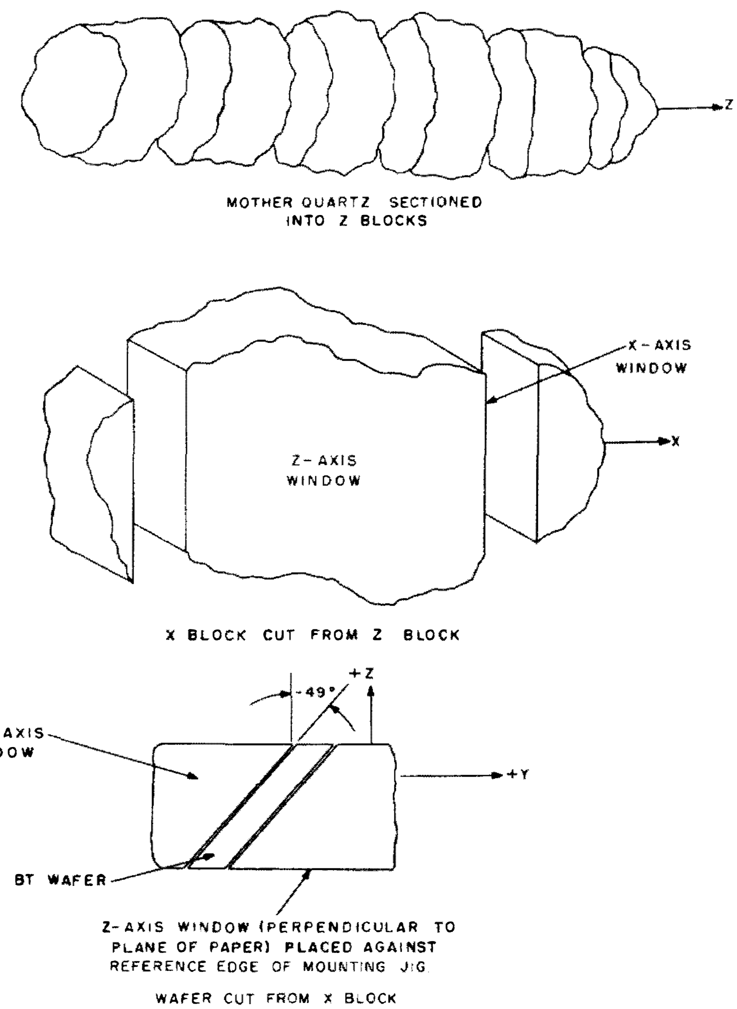


Figure 1-61. X-block method of cutting wafers from unfaced stone. Wafers, on being diced, provide crystal blanks of the proper orientation

orientation from the Y bar. Of the three methods of cutting, the X-block method is the one most commonly used, and will be the one assumed in the following paragraphs.

DETERMINATION OF X AXIS

1-126. A reasonably accurate method for a preliminary determination of the X axis of an unfaced quartz is by observing the cleavage lines of a thin Z section when it fractures after being heated and dropped into cold water. The intersections of the fractures with the XY plane of the Z section are normally parallel to an X axis. A more useful procedure, however, is to first etch a Z block (block or section with Z windows, before X windows have been cut) in a bath of 30% hydrofluoric acid, and determine the approximate X axis with a pin-hole oriascope, shown in figure 1-63. The oriascope provides a pin-point source of ordinary light which will cause a triangular image to appear on the upper XY window when the lower window of the Z block is placed over the pin hole. The sides of the triangle are approximately parallel to the X axes, and matching windows and a template are provided to aid in marking the crystal.

1-127. With the X axis approximately located by

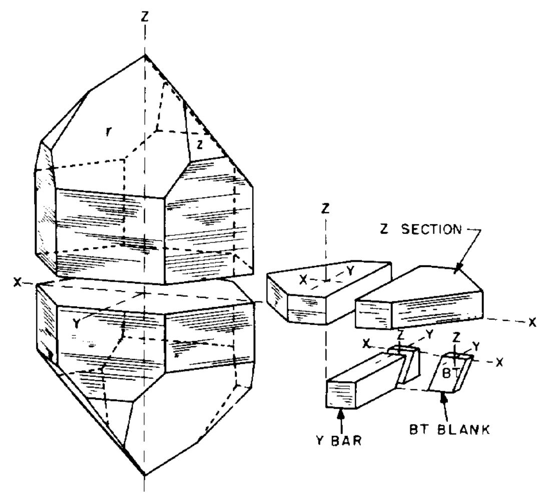


Figure 1-62. Z-section-Y-bar method of cutting properly oriented crystal blanks

the cleavage or oriascope methods, an exact orientation is determined by use of X-ray apparatus. The Z block is cemented to a glass plate, which in turn is placed on an adjustable orienting jig that can later be transferred to a saw. As indicated in

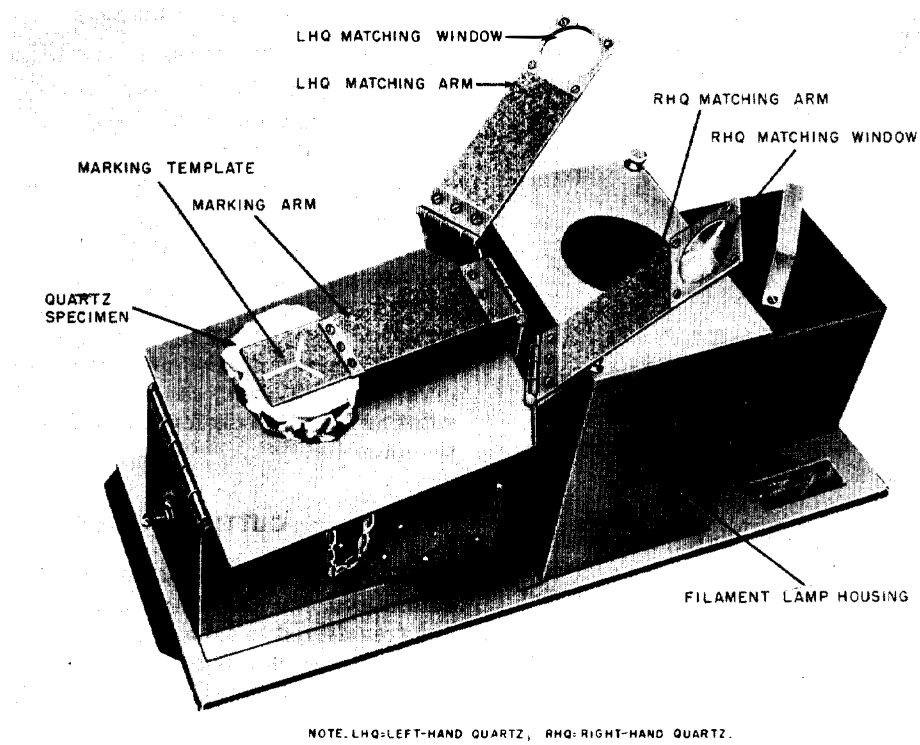


*Figure 1-63. Pinhole oriascope with matching and marking arms for use on Z sections **

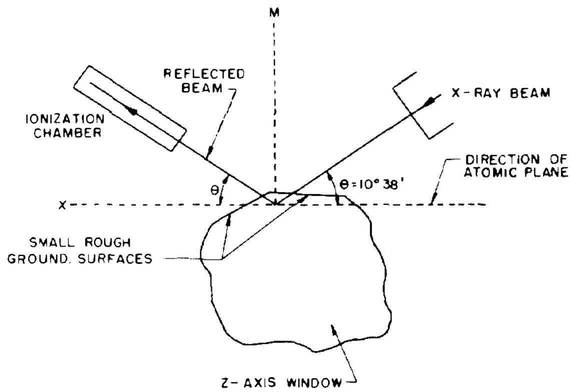


Figure 1-64. X-ray determination of X axis in Z block. M is horizontal bisector of angle that ray must make if reflected beam is to enter ionization chamber. θ, the Bragg angle of X-ray reflection for copper-anode K_α radiation, is predetermined according to the particular atomic plane to be identified. For plane that is parallel to an m face, and hence to an X axis, $\theta = 10°38'$. With positions of X-ray source and ionization chamber fixed, rotation of Z block about Z axis will cause maximum current to flow through ionization chamber when an X axis becomes perpendicular to M

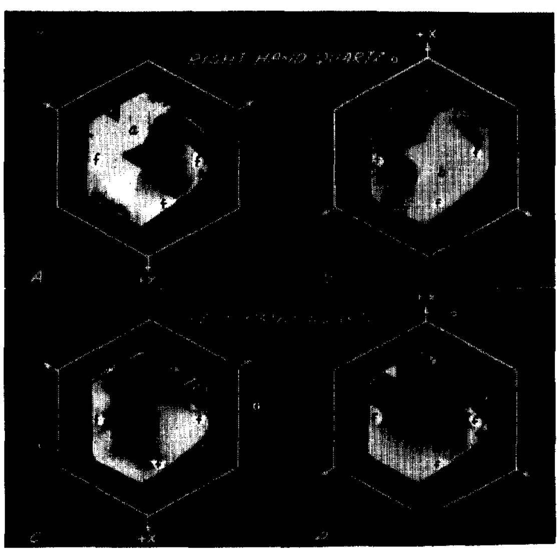


Figure 1-65. Reflection patterns of twinned Z section, showing both types of twinning. The section is predominantly right quartz, but is fairly evenly divided by the electrical twins a and b. The small regions of optical twinning of one electrical sense are shown in C, and those of the opposite sense are shown in D. The X-axis polarities indicated apply only to the respective bright regions. The regions marked f contain flaws

figure 1-64, an X-ray beam is directed toward the crystal's vertical surface, which deflects part of the beam into the window of an ionization chamber, causing a current to flow that has an amplitude proportional to the intensity of the rays entering the chamber. X-rays of constant wavelength are propagated in a narrow pencil from a properly filtered source, which consists of a special high-voltage cathode-ray tube having a copper anode. The X-rays are emitted by virtue of the high-energy electrons' striking the copper target, and a thin nickel plate is inserted in the X-ray path to eliminate unwanted wavelengths. The atomic planes of the crystal lattice effectively serve as reflecting surfaces, except that interference between the reflected rays from adjacent parallel planes eliminates all angles of reflection except those that permit the path lengths of coinciding rays to differ by an integral number of wavelengths. The above condition is satisfied when the distance between the atomic planes is related to the wavelength and the angle of incidence of the X rays in a manner that can be expressed by Bragg's law:

$$n\lambda = 2d \sin \theta$$

where: $n = 1, 2, 3, \ldots$

λ = wavelength of X rays

d = distance between parallel atomic planes

θ = angle of incidence of X rays with atomic plane

The ionization chamber is a gas-filled metallic cylinder having an electrode which is maintained at a voltage relative to the cylinder. X rays entering the chamber will ionize the gas, permitting a current to flow through the external circuit. With θ predetermined for a particular atomic plane, the exact direction of the plane, and hence of the crystal's orientation, can thus be determined by rotating the Z block for a maximum reading on the ammeter.

CUTTING X BLOCK

1-128. When the X direction has been precisely determined, the mounting jig is locked in position and transferred to a diamond saw, where windows are cut perpendicular to, and at each end of, the X axis—thereby forming an X block. After the alignment of the windows is rechecked with the X-ray apparatus, the X block is cleaned, and then etched in 48% hydroflouric acid or a saturated solution of ammonium difluoride.

DETERMINATION OF TWINNING

1-129. Electrical and optical twinning boundaries can be observed directly by shining a spot lamp upon the etched Z windows of the X block. The light should be directed at approximately a 30-degree angle with the surface being examined, with the line of sight of the observer perpendicular to the surface. As the block is rotated about the axis perpendicular to the surface, there will be four particular orientations, each corresponding to a reflection of maximum brightness from the etched area of one of the four possible twins—right-hand quartz of either electrical sense, and left-hand quartz of either electrical sense. See figure 1-65. Normally a crystal is predominantly right or left, so that optical twinning usually appears only in small scattered regions. Electrical twinning, however, normally divides a crystal into large regions of opposite electrical sense. The polarities of the various twinned areas can be readily determined by noting the angles of rotation at which maximum brightness is observed. The axial polarities of an X block may also be determined by examining the X windows with the aid of a pin-hole oriascope having matching and marking arms designed especially for X sections. The images observed will differ according to the electrical sense of the particular area—also, according to whether hydrofluoric acid or ammonium difluoride was used in etching. By a proper interpretation of the patterns, the axial directions of the twinned regions can be suitably marked. If there is an excessive amount of scattered twinning, the block must be discarded; otherwise, the observation permits a proper orientation for cutting slabs, so that optimum use of the quartz is possible.

PREPARATION OF WAFERS

1-130. The mounting jig, adjusted to the correct orientation, is transferred to a saw, and the X block is sliced into slabs of sufficient thickness for finishing. See figure 1-66. After being cleaned and etched, the slabs are inspected and marked for twinning, and the unusable portions are cut away by a diamond saw. Each slab is cemented to a holder and mounted in a jig for a final X-ray determination of the orientation. The adjusted slab holders are transferred to the jig of a lapping machine, and the slabs are lapped on one surface, using an abrasive of 400-grain carborundum, until the lapped faces have the desired orientation. The slabs are then cemented to a large plate with the corrected faces down, and the uncorrected faces are lapped until parallel with the bottom faces. The "wafers," as the slabs are now called, are next cemented to a glass-topped steel plate for dicing.

PREPARATION OF CRYSTAL BLANKS

1-131. The wafers are diced to the approximate crystal blank size with a dicing saw, as shown in

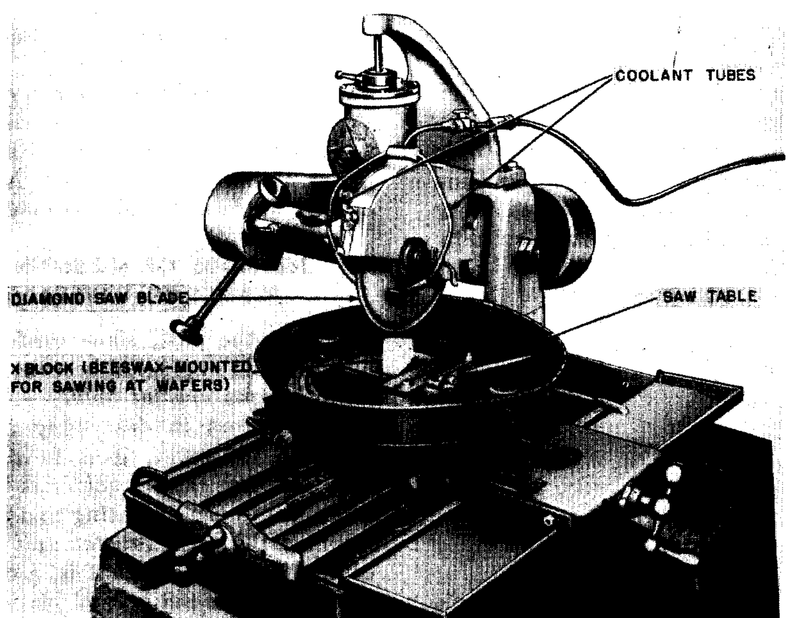


Figure 1-66. Diamond saw for cutting wafers from X block

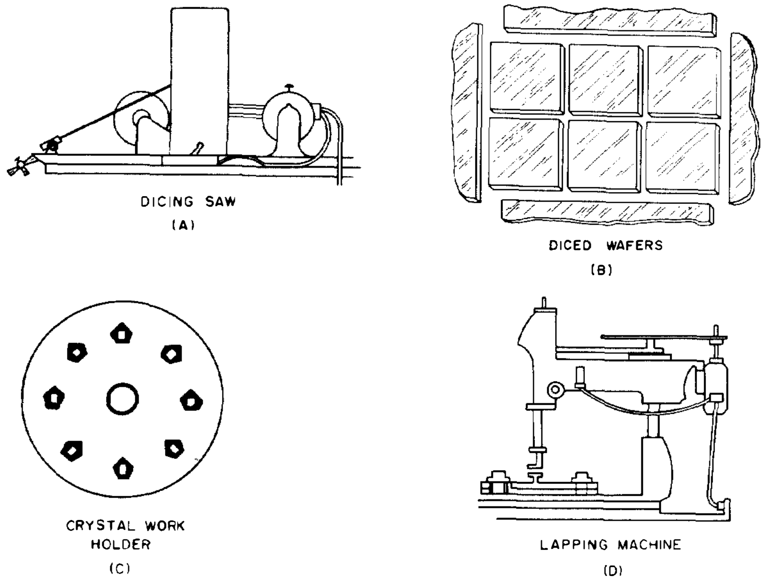


Figure 1-67. (A) Dicing saw. (B) Diced wafer. (C) Nest of lapping machine. (D) Lapping machine

figure 1-67. The dice are then transferred to the nest of a lapping machine, where they are lapped to a thickness equivalent to several kilocycles below the desired frequency. The lapping proceeds in three stages: rough, semifinishing, and finishing. However, the rough stage is accomplished prior to dicing, if the slabs are first lapped to wafers as described in paragraph 1-130. The semifinishing is done with 600-grain carborundum or equivalent, and the finishing requires 1000- to 1200-grain abrasive. Each of the last two stages should completely remove the surface left by the preceding stage. (Where extreme care is required, as when finishing thin harmonic mode plates, 3000 mesh aluminum oxide mixed with water, cosmetic talc and powdered white rouge provides high-precision results, with the ultimate operating dependability greatly increased if the final lapping is followed by a brief semi-polishing with a mixture of water, rouge, and small amounts of rust preventative and wetting agent.) In the case of high-frequency blanks, the final lapping should bring the blanks within 25 to 50 kilocycles of the desired frequency. Next, a stack of 25 to 100 dice are clamped into a loaf, with all the blanks oriented in the same direction. See figure 1-68. Two exposed edges are then ground parallel to locating surfaces by an edging machine. The loaf is then reversed and the two remaining edges are ground to square the blanks. Finally, the blanks are etched to the proper frequency. For high-frequency crystals, a frequency tolerance of ± 0.002 percent will require that the finished blanks be etched to within ± 0.00001 millimeter of the ideal thickness. After cleaning, the crystal blanks are ready for mounting.

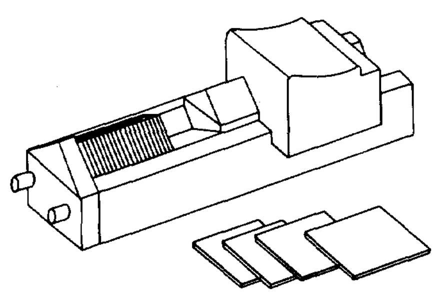


Figure 1-68. Loaf of crystal dice, all blanks oriented in the same direction in preparation for edge grinding by edging machine *

METHODS OF MOUNTING CRYSTAL BLANKS IN CRYSTAL HOLDERS

1-132. In the past, some confusion has resulted among radio engineers because of a mixed usage of the terms *crystal holder* and *crystal unit* by manufacturers in describing and naming their products. However, it is now generally agreed that the term *crystal holder* is to be used only in reference to the mounting and housing assembly, whereas the term *crystal unit* is to designate a complete assembly—that is, a crystal holder containing a mounted crystal plate.

1-133. Crystal holders have been variously classified by different specialists in the field, and in the absence of a standard nomenclature, a certain amount of overlapping has resulted among the different classifications. The procedure to be followed here is to treat each particular method of mounting as a separate category. The holders to be discussed are described according to the following types of mounting: gravity-sandwich, pressure-sandwich, gravity-air-gap, corner-clamped-air-gap, nodal-clamped-air-gap, dielectric-sandwich, plated-dielectric-sandwich, button-sandwich, knife-edge-clamp, pressure-pin, cantilever-clamp, solder-cone-wire, headed-wire, and edge-clamped. Only two general classifications of mounting, *wire* and *pressure*, are specified for Military Standard crystal units currently recommended for use in equipments of new design. The wire-mounted crystals are cemented directly to supporting wires. The pressure-mounted crystals are clamped in place by frictional contact with electrodes or other supporting elements. The wire mounts include the solder-cone-wire, the headed-wire, and the cemented-lead types, the latter being a particular kind of edge-clamped support cemented to the crystal. The pressure mounts include all other types listed above except the gravity-sandwich and the gravity-air-gap.

Gravity Sandwich

1-134. A "crystal sandwich" is simply a crystal plate sandwiched between two flat electrodes. In the simple gravity type of holder the crystal plate is placed on one electrode, with the second electrode resting on top and connecting to the external circuit through a flexible wire. See figure 1-69. A small clearance is provided around the sides to permit the crystal to vibrate freely. The clearance must not be too large, however, else the crystal will slide around in the holder, and may become chipped, or, at least, cause the electrical constants of the crystal unit to vary. The electrodes must be perfectly plane and made of noncorrosive metal, such as stainless steel, brass, or titanium. Brass is inferior to the other two metals, and titanium is largely a future possibility. The top electrode is considerably lighter than the bottom electrode, and is usually specifically dimensioned to fit a particular crystal size. If the top electrode is too heavy, the impedance it introduces will be excessive, preventing the crystal from vibrating near its normal mode; on the other hand, if the top electrode is too light, firm contact will not be made with the crystal, and the operation of the crystal unit will be unstable. The edges of the crystal are slightly rounded to insure that they are free of burrs. Both the crystal and the electrodes must be perfectly clean, and the entire unit must be mounted in a hermetically sealed chamber. Normally, the grid terminal of the unit connects to the flexible wire of the top electrode, and the ground or cathode terminal to the bottom electrode.

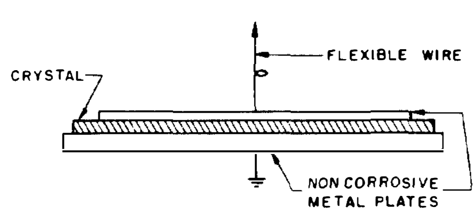


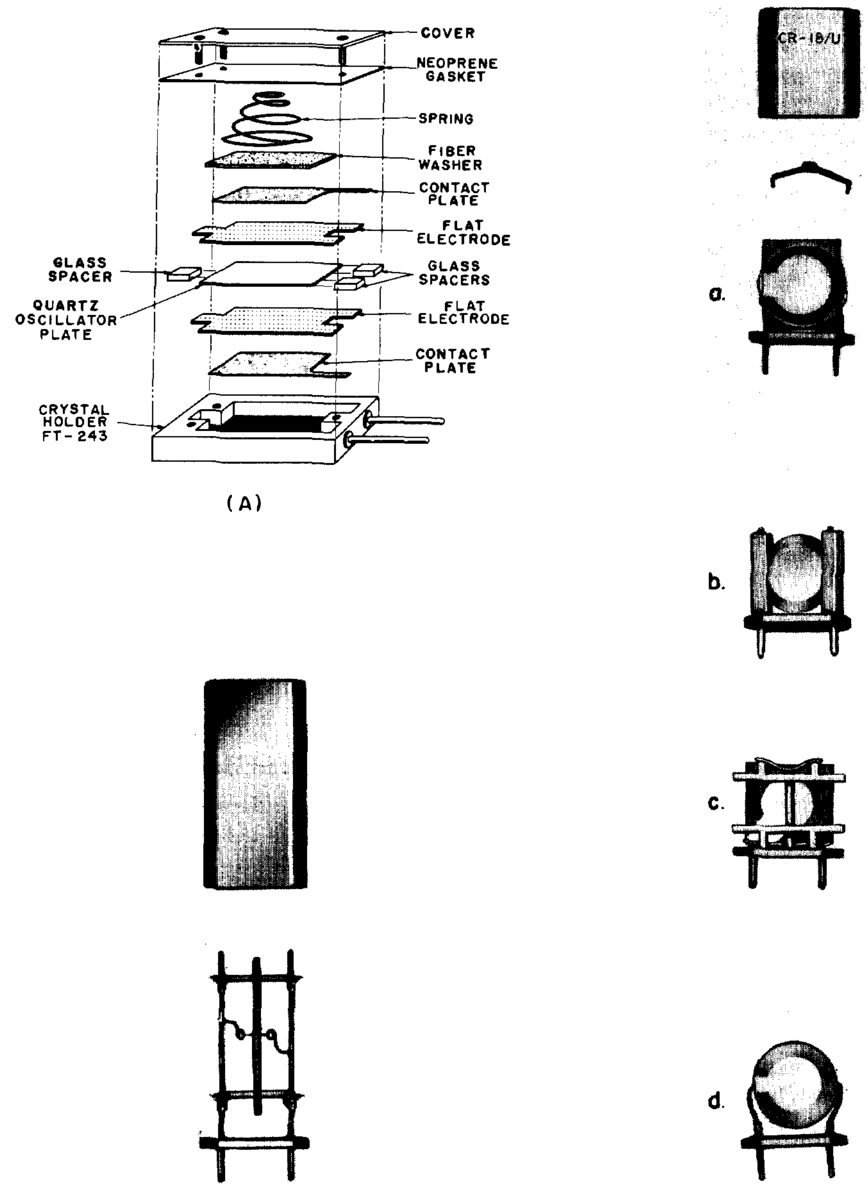
Figure 1-69. Gravity sandwich

1-135. The gravity holder was at one time widely used, but has now been largely replaced by holders that can maintain the crystal in a relatively fixed position if subjected to external vibration, such as might be encountered in vehicular or aircraft equipment. Occasionally, even when mounted in vibration-free equipment, a gravity crystal unit may fail to operate because one edge of the crystal has become closely pressed against one of the sides of the chamber. However, a light tap of the holder is usually sufficient to start oscillations. Compared with the holders in which flat inflexible electrodes are pressed against the crystal, the activity of the gravity holder is generally superior, and requires less voltage to maintain a state of oscillation.

Pressure Sandwich

1-136. In holders of the pressure-sandwich type, the crystal is held more or less firmly between two flat electrodes under the pressure of a spring. In a typical assembly, the electrodes, which normally are of identical size and shape, are in turn sandwiched between two very thin contact plates. The contact plates connect to two metal prongs that serve as electrical terminals and plugs for mounting the crystal holder in a socket. An insulating washer is placed over one of the contact plates, a coil spring is placed over the washer, and a neoprene gasket is placed between the spring and the cover to provide hermetic sealing. Except for the glass spacers, the pressure holder described above is very similar to the air-gap holder shown in figure 1-70.

1-137. Although the activity of a pressure-type crystal sandwich normally is not as great as that of the gravity type, it is much preferred because of its greater ruggedness and less critical requirements concerning the orientation and vibration of the equipment in which it is to be mounted. Another advantage of the pressure holder is its relative simplicity of design—fewer of its compo-



(B) (C)

Figure 1-70. Methods, old and new, for mounting crystal units. (A) Construction of early model crystal unit employing the gravity air-gap type of mounting, now largely outmoded. The crystal holder shown is the type FT-243. (B) Solder-cone wire mount for v-l-f length-flexure crystal. (Courtesy HEEMCO). (C) Recently developed techniques for mounting shock-proof, 1-mc, A elements in the miniature HC-6/U metal holder to meet the specifications for 1-mc CR-18/U crystal units: a. Reeves-Hoffman flexible nylon mount. b. Hupp loose-slotted edge-clamped mount. c. Bliley molded nylon bumper mount. d. RCA edge-clamping spring mount. (Courtesy McCoy Electronics.)

nents require separate and exact dimensions for each particular frequency than is true of the majority of holders. If the holder is constructed so that the spring pressure may be adjusted, very slight variations in the frequency are possible; the activity, however, will decrease proportionally as the pressure is increased. Although the pressure holder is widely used and is less expensive than most of the other types of holders, it has the disadvantage of low activity and comparatively large damping of the oscillations. Thus, crystal units of this construction are not as sharply selective, nor as electrically predictable, as crystal units of more critical design.

Gravity Air-Gap Mounting

1-138. A gravity air-gap crystal unit is essentially a gravity sandwich, but with an air space separating the crystal from the upper electrode. The air gap may be variable or fixed. In the variable type, the frequency can be adjusted slightly by raising or lowering the upper electrode, by means of a screw. The fixed air gap, however, is more commonly used. As shown in figure 1-70, the fixed gap is maintained by glass or other insulating spacers placed between the electrodes. It is important that the thickness of the air gap not be an even quarter-wavelength of the acoustic vibrations which the crystal will generate in the air. Otherwise, the air waves, on reflection from the upper electrode, will return to the crystal 180 degrees out of phase with the normal vibrations, thereby introducing a high impedance and greatly reducing the activity. Maximum activity is obtained when the air gap is an odd quarter-wavelength in thickness. The exact dimension, however, is not particularly critical in the case of shear modes, and a variation of $\pm 1/8$ wavelength will cause little change in the amplitude of the vibrations. The quarter-wavelength formula is $\frac{\lambda}{4} = \frac{v}{4f}$, where v is the sound velocity in air, and f is in cycles per second. At room temperature and pressure, v = 330,000 mm/sec = 12,992,000 mils/sec. The gap thickness should not exceed 3 mils, else the piezoelectric coupling will be too weak to maintain oscillations at reasonable voltages. Where it is necessary to have as large a piezoelectric coupling as possible, the air gap must be reduced to the minimum of $1/8$ wavelength.

1-139. The advantage of the air-gap mounting is that it shares the simplicity of design of the sandwich holders, but eliminates the damping effect caused by the frictional contact of the upper electrode with the crystal. The presence of the air gap also effectively inserts in the crystal circuit a series capacitance equal to that between the upper electrode and the crystal face, thereby increasing the ratio of the stored electrical to the stored mechanical energy, and thus decreasing the electromechanical coupling. The reduction of the frictional losses (i.e., the effective electrical resistance) and the electromechanical coupling serves to give the crystal unit a higher Q, and thus to make it more selective and stable, and less affected by small variations on the external circuit. However, with the decrease in electromechanical coupling, the tendency of the crystal to vibrate is reduced; and also reduced is the bandwidth for use in filters. On the credit side, the gravity-air-gap mounting is particularly convenient for the preliminary testing of crystals at the time they are being ground, and it has also been widely used in fixed-plant equipment with thickness-mode crystals operating at frequencies between 200 and 1500 kc. The principal disadvantages of the gravity-air-gap holder are the loose mounting of the crystal, the reduced piezoelectric coupling, and the possibility that a momentary overdrive will cause arcing across the air gap, or a corona discharge, thereby damaging the crystal and electrode surfaces, or even puncturing the crystal completely.

Corner-Clamped, Air-Gap Mounting

1-140. The principal features of the corner-clamped, air-gap mounting are indicated in figure 1-71. Note that air gaps exist at both the major faces of the crystal, except at the corners where the electrode risers, or lands, clamp the crystal in position. If the length of a thickness-shear plate, such as an AT or BT cut, is not less than twenty times the thickness, a firm pressure may be applied at the corners without greatly reducing the activity. The same precaution for avoiding an air column with dimensions approaching a multiple of a half-wavelength is necessary for the clamped as for the unclamped holder. The lands at the corners normally provide a gap of 0.5 to 2 mils.

1-141. The corner-clamped, air-gap method is widely used for mounting high-frequency thickness-shear crystals. Its operating characteristics

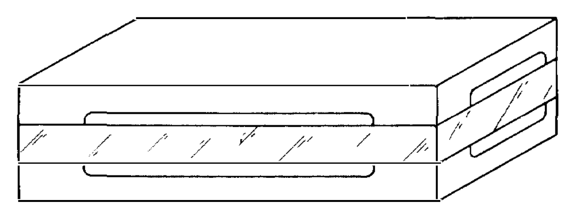


*Figure 1-71. Typical corner-clamped, air-gap method of mounting crystals **

are similar to those of the unclamped holder; and, in addition, it has the important advantage of clamping the crystal in a fixed position, thus permitting its use in aircraft and vehicular equipment. However, the clamping at the corners introduces an excessive amount of impedance when used for the lower-frequency, thickness-shear crystals where the l/t ratio is less than 20; hence, this type of mounting is generally confined to crystals with frequencies above 1500 kc.

Nodal-Clamped, Air-Gap Mounting

1-142. The principal features of the nodal-clamped, air-gap mounting are indicated in figure 1-72. This method may be used for mounting low-frequency piezoelectric elements vibrating in an extensional mode and having a nodal area at the center of the crystal. Each electrode has two risers for clamping the crystal at each end of its nodal axis, and thus provides a secure mounting with a minimum of damping from direct contact between the crystal and the electrodes. A general advantage of any "zone-type" clamping, such as the nodal or corner methods where particular areas of a crystal are subjected to pressure, is that spurious frequencies requiring free vibrations in the clamped zones will be suppressed.

Dielectric Sandwich

1-143. This type of holder is essentially a crystal sandwich with a "lettuce" of high dielectric material inserted between the crystal and the electrodes. The sandwich and air-gap holders previously described do not permit a crystal to operate

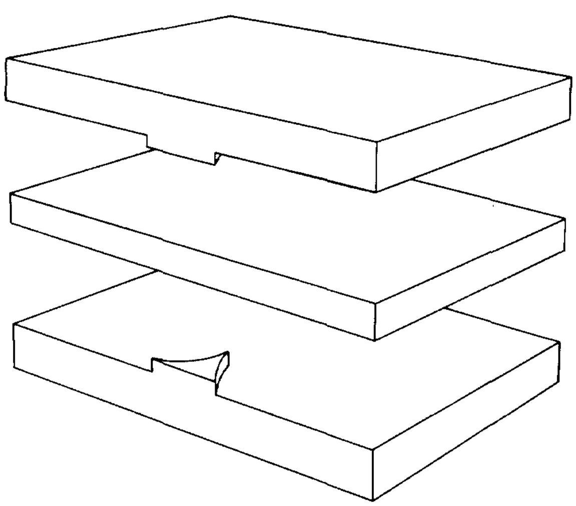


Figure 1-72. Typical nodal-clamped, air-gap method of mounting crystals

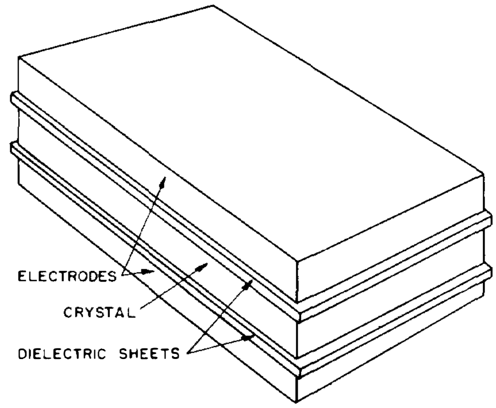


Figure 1-73. Pressure type of dielectric sandwich for mounting crystals

near its elastic limit, for otherwise arcing would occur between the electrodes and the crystal. Low drive levels are particularly necessary at frequencies above 4000 kc, for the likelihood of corona discharge or arcing increases with the frequency. Even if the arcing is insufficient to puncture the crystal, its presence will cause either wet or dry oxides to form on the crystal and the electrodes, thereby reducing the activity and greatly shortening the crystal's useful lifetime, if not preventing its operation entirely. Many factors contribute to the possibility of a breakdown: type of holder, presence of sharp edges, smoothness and parallelism of crystal and electrode faces, type of cut, air-gap dimensions, d-c and r-f voltages, frequency, and the like. However, since the arcing in all cases is the direct result of ionization of the air between the electrodes and the crystal, this danger may be removed if the more vulnerable air spaces are filled with an elastic cushion that has little tendency to ionize. It is necessary that the material have a dielectric constant much higher than that of air, and preferably higher than that of the crystal, and that it have low dielectric losses at the operating frequency; otherwise the special advantages of the particular types of mounting with which dielectric material is used would be destroyed by an increase in damping. The dielectric filler may consist of insulating sheets cut to fit a particular mounting, or it may be coated over the electrode faces. In either case, a material of high dielectric constant will permit a crystal to be driven near its elastic limit without the danger of corona effects, and with much less restraint of the normal vibrations than occurs when the crystal is in direct contact with the electrode faces. Suitable dielectric materials are mica, thin sheets of glass or fused quartz or other ceramics, "Cellophane," nonsul-

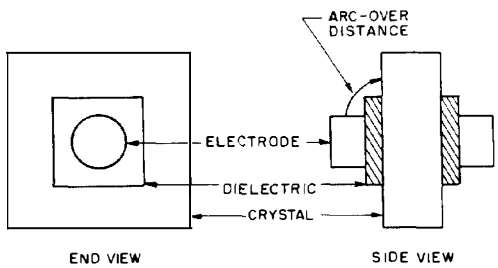


Figure 1-74. Center-pressure type of dielectric sandwich for mounting crystals

furous rubber sheeting, cellulose esters and ethers, varnishes, lacquers, vitreous enamels, metallic oxides, rubber coatings applied by electro-deposition, rubber containing resin and other fillings, or fused coatings of natural or synthetic resins. The sheets or coatings should be from 1 to 5 mils in thickness, but care should be taken that the thickness of the insulating material does not approach a multiple of a half-wavelength of the acoustic waves that will be generated. In any event, the addition of the dielectric material will tend to raise the impedance and frequency slightly, so that in extreme cases it may be necessary to grind the crystal to a frequency lower than that at which it is to operate.

1-144. Figures 1-73 to 1-77 indicate different methods in which the dielectric fillers may be used in mounting a crystal. Figure 1-73 illustrates a pressure type of mounting with two sheets of dielectric material—mica, for instance—inserted between each electrode and crystal. Note that the mica extends beyond the edges of both electrodes. This feature is important, for although in a well-designed pressure sandwich no air spaces exist between the crystal and electrode faces, so that ionization and arcing do not occur at the major surfaces, corona discharges can and do occur at the edges, particularly if the sides of the chamber are close in, as is usual, causing the alternating field around the edges to be more intense. However, with the insulating sheet of high dielectric constant overlapping the electrode edges, the intensity of the electric field will be greatly diminished. That part of the dielectric directly between the crystal and the relatively inelastic electrode, acts as an elastic cushion, and thus serves much the same function as an air gap, but without increasing the possibility of corona or arcing effects.

1-145. Figure 1-74 shows a top and a side view of a center-pressure type of mounting, where two circular electrodes of small cross section are separated from the crystal faces by insulating sheets of high dielectric constant. Again, it is important that the insulation extend well beyond the edges of the electrodes. This arrangement increases the length of the shortest possible arcing path, and, in so doing, diminishes the chance of the occurrence of a discharge.

1-146. Figure 1-75 illustrates two methods by which a corner-clamped air-gap holder can be converted into a dielectric sandwich while still retaining the principal advantages of the air-gap mounting. Figure 1-75A shows an insulating sheet cut to the dimensions of the air gap, and figure 1-75B shows a corner view of the assembled sandwich. Figure 1-75C is a side view of a similar sandwich,

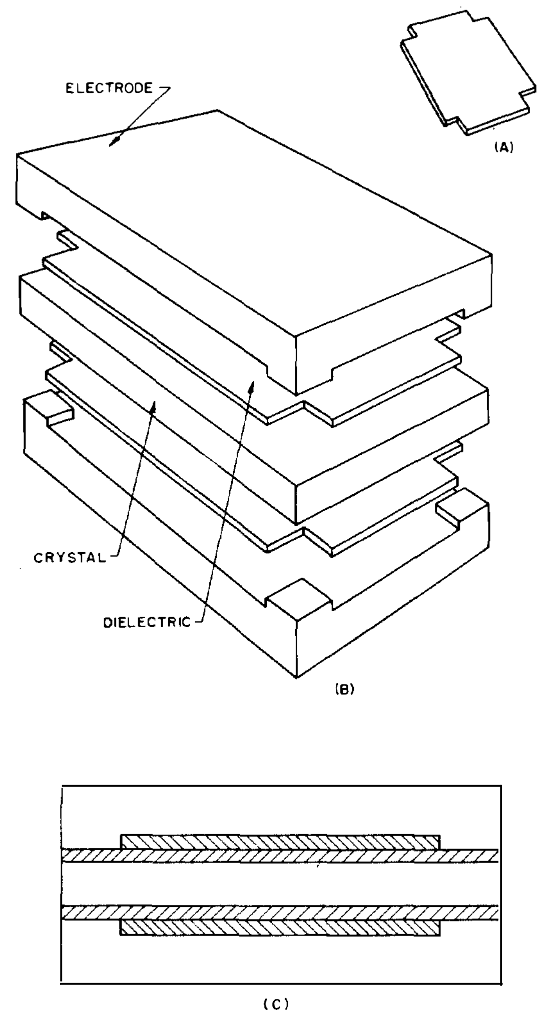


Figure 1-75. Two methods (B and C) by which a corner-clamped, air-gap holder can be converted into a dielectric sandwich. Figure A shows a dielectric sheet cut to the dimensions of the air gap

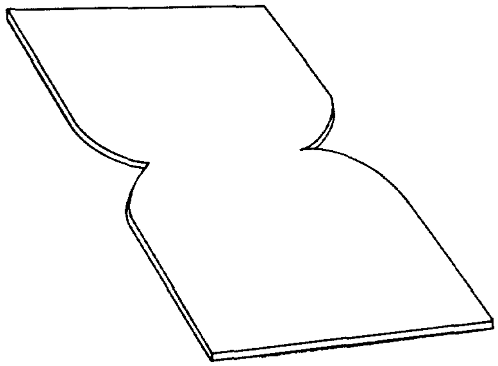


Figure 1-76. Dielectric sheet cut to fill air gap of nodal-clamped mounting

but with two additional insulating sheets inserted to cushion the crystal entirely from direct contact with the electrode risers.

1-147. Figure 1-76 illustrates the cut of an insulating sheet for converting a nodal-clamped, air-gap mounting into a dielectric sandwich. Two niches in the edges of the sheet are cut to fit the two risers of an electrode. When assembled, the sandwich is similar to the corner-clamped model of figure 1-75B; or, if additional rectangular sheets are inserted next to the crystal, the assembly will resemble the arrangement in 1-75C. With either method, maximum rigidity is obtained for the nodal mounting with a minimum in damping.

1-148. Figure 1-77 is a cross-sectional view of a gravity type air-gap mounting with the electrodes coated with an insulating material of high dielectric constant. It is characteristic of air-gap holders that the smaller the thickness of the air gap, the higher the r-f voltage that can be applied before arcing occurs between the crystal and electrode faces. When the electrodes make perfect contact with the crystal, not only are the opposing surfaces theoretically at the same potential, but no ionizable substance lies between them in which an arc can form. However, the introduction of an air gap not only inherently reduces the electromechanical coupling of a crystal unit, but also effectively lowers the voltage that can be practicably applied. To remove the latter restriction without diminishing the advantages the air gap provides, the arrangement shown in figure 1-77 can be used. Note that the coating covers the edges of the electrodes—an important consideration since it is at the points of sharpest curvature that ionization is most likely to arise.

1-149. The use of insulating sheets and coatings of high dielectric constant permits a crystal to be operated near its elastic limit without the danger

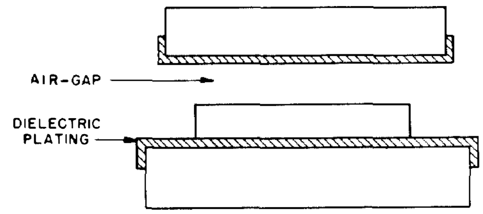


Figure 1-77. Cross-sectional view of gravity-air-gap mounting with electrode surfaces protected by coating of high-dielectric material

of arcing, and hence this type of crystal unit can be operated at higher drive levels than would otherwise be possible. The dielectric sandwich would be advantageous in filter circuits where high amplitude signals are to be encountered; or in small portable transmitters where several amplifier stages are not possible, and the excitation level must be as high as possible; or in any radio transmitter designed to be keyed in the oscillator stage where it is important that the oscillations built to peak amplitude in a minimum number of cycles. The insertion of the dielectric sheets also improves the stability and selectivity of the sandwich-type holders, inasmuch as they eliminate direct contact between the crystal and the relatively inelastic electrodes. The principal disadvantages of the dielectric sandwich are the reduced piezoelectric coupling caused by the separation of the electrodes from the crystal, and the damping effect of the frictional and small dielectric losses which are slightly greater than those of the air—provided the crystal is operated well below its elastic limit.

Plated-Dielectric Sandwich

1-150. This type of mounting is essentially the same as the previously described dielectric sandwich except that a thin layer of conducting material is interposed between the dielectric sheets and the electrodes, or between the dielectric sheets and the crystal, or both. The conductive surfaces may be strips of metal foil not more than 1 mil in thickness, or they may be plated, painted, or sprayed directly on the insulating material. Suitable conducting substances are copper, nickel, silver, gold, platinum, and their alloys. The conducting layer may be coated on one or both major surfaces of the insulating sheet, or it may completely cover the edges as well as the major surfaces, thus effectively converting the sheet into a highly compliant metal plate.

1-151. Figure 1-78A illustrates the corner-clamped-air-gap mounting using dielectric plates having conducting films on both major surfaces. The two

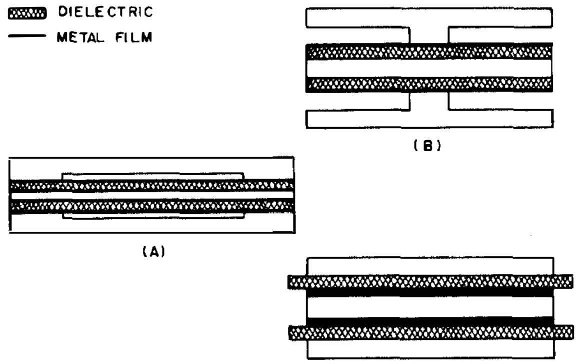


Figure 1-78. (A) Air-gap mounting using dielectric sheets having conducting films on each major surface. (B) Air-gap mounting using dielectric sheets having a conducting film on major surface in contact with electrode. (C) Dielectric sandwich in which leaves of metal foil are inserted between dielectric and crystal

films that are in direct contact with the electrode risers prevent the establishment of differences of potential across the air gaps, and hence obviate the possibility of arcing or corona discharges in these spaces. Figure 1-78B illustrates the nodal-clamped, air-gap mounting using dielectric plates having a conducting film on only one surface. In the case of air-gap mountings, if only one conducting surface is to be interposed between a crystal and each electrode, it is preferable that this surface make contact with the electrode rather than the crystal, so that possible electric fields will be "shorted" around the air gap. Figure 1-78C illustrates a dielectric sandwich mounting in which leaves of metal foil are inserted between the dielectric plates and the crystal. The metal foil, being very thin and flexible, snugly fits the crystal surface and interferes but little with the crystal's vibrations. On the other hand, its presence insures a uniform potential at all points on the crystal's surface, thus protecting the surface from the effects of excessive electric stresses.

1-152. The plated-dielectric sandwich combines the advantages of the plain dielectric sandwich with improvements in the frequency stability, crystal life, frequency spectrum, and piezoelectric coupling. The improvement in frequency stability is greatest in the case of the air-gap crystal units, for the danger of arcing or corona discharge in an air gap is removed without the insertion of a dielectric sheet to fill the gap. Since the damping effect of the air is less than that of the insulating material, the use of conducting film permits a closer approach to the high selectivity of the pure air-gap mounting for crystals which are to be driven near their elastic limit. The insertion of a metallic film next to the crystal surface serves to reduce possible electrical stresses at the surface which might indirectly aid the production of small fractures, or cause ionization and chemical effects that would lead to a weathering of the crystal's face. The insurance of a uniform potential at all points on the surface of the crystal also improves the frequency spectrum, particularly at very high frequencies, where many possible overtone modes can vibrate at frequencies close to that of the desired mode. However, the majority of the unwanted modes will have changes of phase and differences in amplitude along the major plane of the crystal, so that the resulting eddy currents that they induce in the conducting surfaces will aid in damping them out. Where the interfering modes might otherwise lead to a frequency drift or jump, the damping effect will be reflected principally as an increase in the effective electrical resistance and as a decrease in activity. Finally, closer piezoelectric coupling is achieved if the entire insulating material is given a metallic coating. The dielectric sheet thus effectively becomes an extension of the electrodes, and the close coupling of the simple sandwich mounting is approached, but without the heavy damping caused by friction between the crystal and solid metal.

Button Mounts

1-153. The ceramic button crystal mount represents the ultimate in crystal-holder design yet to be reached via the sandwich and air-gap evolutionary chain. Originally, the button electrode was developed as an all-metal modification of the corner-clamped, air-gap type. As illustrated in figure 1-79A, the all-metal electrode is provided with conventional lands at the corners, but the effective center area has been reduced by surrounding the center with a relatively deep circular groove. The effect is to reduce the shunt capacitance across the crystal while retaining a central area of sufficient size for adequate excitation; the reduction in shunt capacitance is particularly desirable if the crystal is to be operated in the v-h-f range. Also, since the principal excitation is confined to a central circular area, the likelihood of spurious modes is somewhat reduced, because the vibrating part of the crystal tends to exhibit the properties of a circular plate. The superior frequency spectrum of the circular plate is probably even more closely approached by using electrodes having solid ring-shaped lands that completely surround the circular

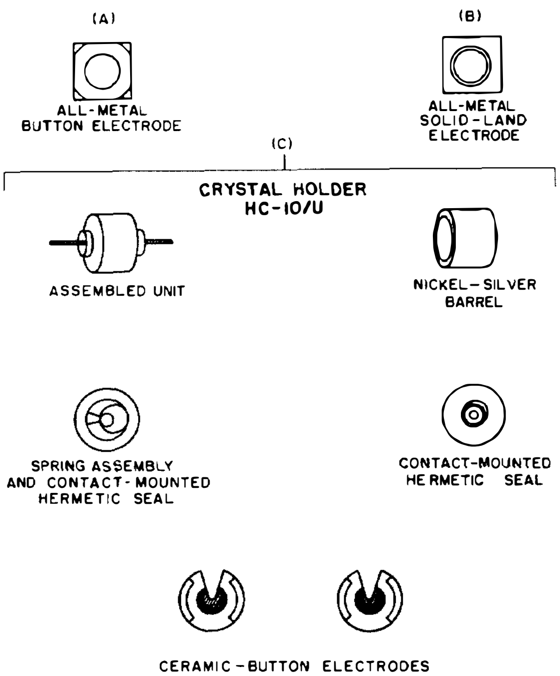


Figure 1-79. Button electrodes and methods of mounting

air gap at the center. See figure 1-79B. However, it is by combining the advantages of the button mounting with those of the plated-dielectric sandwich and circular quartz plates that optimum performance is obtained for thickness-shear modes at very high frequencies. Figure 1-79C shows the principal parts and the complete assembly of Crystal Holder HC-10/U. The shunt capacitance is held to a minimum, first, by the use of ceramic supporting plates in place of all-metal electrodes, and second, by the use of a coaxial electrode system in place of the conventional method where the crystal leads parallel each other through the base assembly. The ceramic-button electrodes are usually very thin metallic platings that cover a small circular area at the center of each ceramic plate. Although the lands may be provided by forming thickened sections at the rim of the ceramic disks, usually they are obtained by plating metal risers on the surface of the ceramic; these plated risers are not connected electrically to the center metallic section. The air-gap thickness is normally between three and five microns. A notch in each ceramic button permits an extension of the electrode plating to the opposite side, so that contact with the crystal leads can be made with a minimum of increase in electrode capacitance. This type of crystal holder is unequalled in performance when used with harmonic-mode crystals in the very-high-frequency range. It should be noted, however, that one of the original advantages of the plated-dielectric sandwich mounting is not fully realized in the case of ceramic-button electrodes — namely, the protection against arcing or corona discharges. For this reason, the ceramic-button crystal units will not withstand as high a drive level as might otherwise be possible. On the other hand, the thin air gap that can be obtained, the presence of a high-dielectric material almost flush with the edges of the plated electrodes, and the firm mechanical support by which the crystal is held and cushioned against shock make this unit more durable under high drive levels than conventional air-gap holders. One of the more important advantages of the ceramic-button is the reduction of spurious modes through the use of circular quartz plates and small electrodes. The small electrode dimensions serve to concentrate the activity at the center of the crystal, where the crystal is most likely to be of uniform thickness; thus, sudden frequency jumps are prevented, for these seem to be due primarily to abrupt shifting of the center of activity between areas having slightly different average thicknesses.

Plated Crystals

1-154. Since 1940 the designers of crystal units have increasingly favored the use of electrodes in the form of extremely thin metal films deposited directly on the crystal. Coatings of silver and gold have been successfully applied by spraying and baking, but in general the most advantageous method is by evaporating the metal in a vacuum and allowing it to condense on the exposed surfaces of the crystal. Sputtering processes are being used increasingly, particularly for base plating, where the crystal is plated in vacuum by ionic bombardment from high-voltage negative electrodes composed of the desired plating metal. Electroplating of crystals also finds application. The noble metals, gold and silver, are the elements most commonly used in plating crystals because of their resistance to oxidation, their relative ease of plating, and the strength of their soldered junctions. Other metals that are used in plating are nickel and copper. Aluminum plating is preferred if a crystal is to be held in position by pressure pins or knife-edge clamps. This is because aluminum is more durable to frictional wear, and because its lesser density permits an electrode of lighter weight. However, aluminum is the more difficult to apply, has a tendency to oxidize, and its soldered connections are not as strong as those of silver or gold. For these reasons, silver is more widely used if the crystal is to be soldered between wire supports, and gold is used if the wire-supported unit requires maximum stability and resistance to aging. Aluminum coatings are commonly applied at 1 milligram per square inch, which is equivalent to a thickness of approximately 0.0225 mil; silver is applied at 4 milligrams per square inch, a thickness of approximately 0.0232 mil; and gold is applied at 3 milligrams per square inch, a thickness of approximately 0.0114 mil. The actual plating procedure may be divided into two or more steps involving more than one plating process. As an example, the Signal Corps Engineering Report E-1108 by J. M. Roman recommends as many as three different plating stages during the fabrication of low-resistance, 50-mc harmonic-mode crystal units of the CR-23/U type. The base plating is accomplished by a sputtering machine in which a group of crystal blanks are mounted in a rectangular metallic mask midway between two gold electrodes, which are 5½ inches square and 6¼ inches apart. A bell jar is placed over the electrode assembly and is evacuated to 0.05 to 0.02 millimeters of mercury. 2200 volts dc are applied between the crystal mask and the electrodes; first for 30 minutes at 100 ma with the mask negative in order to clean the crystals by ionic bombardment, and next, for 37 minutes at 100 ma with the electrodes negative for the actual gold plating operation. A second sputtering machine is used to clean the crystal mask of the gold deposited upon it during the plating procedure. This latter operation requires an hour at 100 ma. After being mechanically mounted on HC-6/U bases between 9-mil, edge-clamping spring wires, the crystals are given a preliminary performance test. If a crystal is more than 0.1 per cent off its nominal frequency it is subjected to an additional plating process. This time it is plated electrolytically with nickel. The electroplating of 50-mc crystals proceeds at a rate of 0.9 ma, which is equivalent to a harmonic frequency change of 100 kc per minute. The electrolytic solution consists of chemically pure nickel ammonium sulphate, boric acid, ammonium chloride, and water in a weight ratio, respectively, of 75/15/15/1000. After mounting, testing, and electrically bonding the plated crystal electrodes to the supporting wires with silver cement, the crystals are given a final spot plating with gold in an evaporation type plater to bring them to their specified frequency. This final plating process is accomplished in vacuum while the crystal is connected in a test oscillator circuit.

1-155. The advantages of using metal-film electrodes are several fold: maximum piezoelectric coupling is achieved; the possibility of arcing between the electrodes and the crystal is reduced to a minimum; variations of frequency due to a shift of the position of the crystal relative to the electrodes are eliminated; frictional losses and wear due to inelastic contact between the crystal and the electrodes are removed; the metallic film aids in protecting the crystal from erosion; the film is readily adaptable for various types of nodal mounting, and is easily divided into several electrodes for use in exciting particular harmonic modes. All in all, the plated crystal is the most practical for obtaining optimum crystal performance at low and fundamental-mode high frequencies. The metal-film electrodes, however, have certain disadvantages: the metal has a tendency to absorb moisture, thereby causing the frequency to change; when clamp supports are used, friction at the clamped points will eventually wear away the metal coating; and generally, the mounting techniques are more critical for plated crystals.

Pressure-Pin Mounting

1-156. Pressure-pin holders (see figure 1-80) are used to support low-frequency (up to 200 kc), electrode-plated crystals, particularly those crystals used in telephone filters. Each crystal is clamped at the center of a nodal zone by one or more pairs of opposing pins. For crystal plates one-half inch square and smaller, the diameter of the pins is approximately 10 mils, and the clamping force varies from one to two pounds; for larger plates, pins of larger dimensions exerting somewhat greater clamping forces are used. It is important that the plated electrode be of aluminum, for the greater hardness of aluminum is required to resist the wear at the points of contact with the pins. Normally, these holders are designed for mounting a complete set of filter crystals within

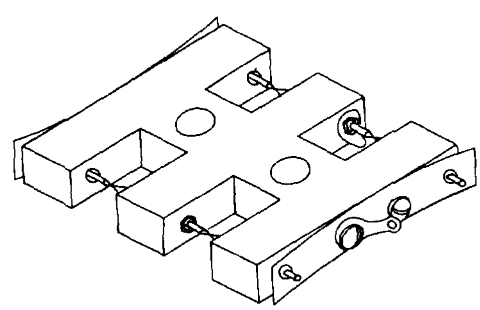


*Figure 1-80. Pressure-pin mounting, with provisions for supporting four plated filter crystals **

a single hermetically sealed container. The holder shown in figure 1-80 mounts four crystal elements. The pressure is applied by the springs mounted at the ends, and the pins serve as electrical connections as well as mechanical supports. For greater mechanical stability, slight niches may be made in the quartz at the clamped points.

1-157. The pressure-pin holder has the advantages and disadvantages of the plated electrodes, and is used primarily for low- and medium-frequency filter crystals. It is particularly applicable for use with face-shear elements, since these have but one practicable nodal spot for clamping. The chief limitation of the pressure-pin mounting is the mechanical impedance it introduces. If the diameter of the pin is made too small, the crystal will tend to rotate about its axis of support; however, the larger the diameter is made, the more the clamping area will extend beyond the nodal point. To obtain optimum performance with this type of mechanical support, a resonant-cantilever design for pressure pins was invented by J. M. Wolfskill of the Bliley Electric Company (U.S. Patent 2,240,453, 1941). This step was quite significant, not only in its own right, but because it provided a forerunner of the resonant-wire type of mounting. The following discussion is based on an analysis of the cantilever clamp by R. A. Sykes (Bibliography No. 741).

The Cantilever Clamp

1-158. The cantilever clamp is a pressure-pin support designed to resonate at or near the crystal frequency. Figure 1-81 illustrates a pin mounted as a cantilever, and figure 1-82 indicates the motion of the cantilever as a quarter-wavelength bar with a node at the fixed base and a loop at the point of contact with the crystal. It is important that even quarter-wavelengths be avoided, else the mechanical energy returning to the crystal will

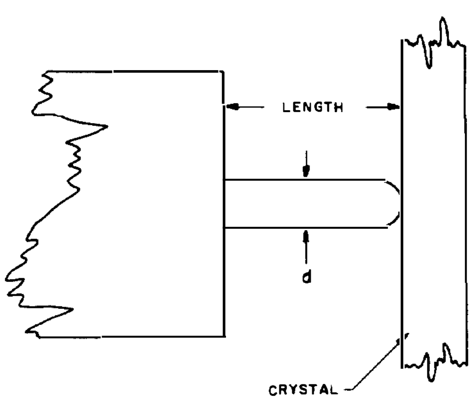


Figure 1-81. Cantilever clamp for providing a resonant-pin support for the crystal *

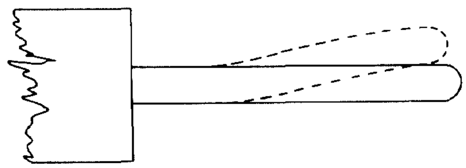


Figure 1-82. Resonant motion of cantilever pin when its length is equal to one-quarter wave-length of clamp-free flexural vibration. Note that the effective free end of the pin is that end supporting the crystal (not shown) *

oppose its motion, thereby greatly increasing the impedance and lowering the activity. The length of a cantilever pin that will present a loop to the crystal can be determined approximately from the frequency formula of a clamp-free rod in flexural vibrations:

$$f = \frac{m^2 d v}{8\pi l^2}$$

where: $m = 1.875$ for the 1st node of vibration of the rod (pin)

$m = \left(n - \frac{1}{2}\right)$ for $n = 2, 3, \ldots$

d = diameter of pin

v = velocity of propagation along pin

l = length of pin

For phosphor-bronze pins, $v = 3.6 \times 10^5$ cm/sec; therefore, to support a 100-kc crystal, pins 1 mm in diameter should be 2.25 mm long to resonate in the mode indicated in figure 1-82. To resonate as a three-quarter-wavelength rod, $n = 2$, and $l = 5.67$ mm for a pin of 1-mm diameter. The pin should be rounded at the end, as shown in figure 1-81, so that firm contact is made without the risk of having all the clamping force concentrated momentarily at a sharp point.

1-159. A properly designed cantilever clamp should extend the useful range of the pressure-pin type of mounting to somewhat higher frequencies, and this has proved to be true in actual practice; however, at the present time no data is available concerning its frequency application above 350 kc. Theoretically, a pair of pins could be used at any of their clamp-free harmonic modes, and thus pins of the same design need not be restricted to use at a single frequency. The principal promise of the cantilever clamp, however, is that it can provide a firm mechanical support while presenting a minimum of interference to the normal vibration of the crystal.

Knife-Edged Clamp

1-160. The knife-edged clamp is similar to the pressure-pin method of mounting, except that the

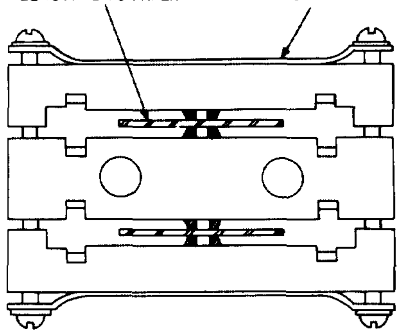


Figure 1-83. Knife-edge clamp support for two crystal plates. Each crystal has two pairs of plated electrodes, and is so mounted that each pressure blade makes electrical contact with a different electrode. This arrangement effectively provides four crystal elements for use in a balanced filter circuit*

clamping prongs are blade-shaped, as indicated in figure 1-83. The dimensions of the clamping points are, on the average, about 35 mils in length, and 10 to 15 mils in width. These blades are used with crystal elements that have a nodal axis parallel to the plane of the major faces, and care must be taken to make certain that the blades are centered along the nodal line. Pressure is applied by phosphor-bronze springs, with the blades serving as electrical connections as well as mechanical supports for the crystal. The holder shown in figure 1-83 mounts two crystals, but, because the plated metal films are divided to provide two electrode pairs for each crystal, the equivalent of four crystal plates is effectively available for use in a balanced filter circuit.

1-161. The advantages of the knife-edge clamp are essentially the same as those of the pressure-pin mounting, except that the greater surface of contact between the crystal and the clamp permits a firmer mechanical support. However, the knife-edge clamp is limited to use with those crystal elements that have well-defined nodal lines. Its most important application has been as a mounting for the −18° X-cut filter crystal, a crystal that can vibrate in a very pure length-extensional mode, and which has a nodal axis at the center parallel to the width dimension. The knife-edge clamp is generally useful only at frequencies below 120 kc.

Wire Mounting

1-162. Wire-mounted crystal units are of two kinds: those that employ wire supports designed to resonate at the crystal frequency in a manner similar to that described in paragraph 1-158 for cantilever clamps, and those that clamp the crystal at the edges by non-resonant spring wire. This latter type of wire support is the cemented-lead mount, which is classified here as an edge-clamped mount. The wire mounting provides a firm but flexible support that serves to cushion the crystal from external vibration and shock. In addition, it can combine the advantages of the metal-film electrode with the low impedance of resonant supports, and can be used to mount any of the crystal elements, both high- and low-frequency plates, vibrating in extensional, shear, or flexural modes. Because of these advantages, the wire-type mounting is generally favored for crystal units used in military equipment.

1-163. There are two principal types of resonant wire mounts, the solder-cone and the headed-wire. In general, the solder-cone support is restricted to relatively small crystal plates—for example, to frequencies above 300 kc for C elements. The headed-wire type is more suited to larger plates.

Solder-Cone Wire Support

1-164. A diagram of the solder-cone type of wire mounting is shown in figure 1-84, and a mounted crystal is shown in figure 1-85. The crystal to be mounted is first spotted with small silver footings, 40 to 90 mils in diameter, at the nodal points where the wires are to be attached. Next, the electrodes are plated on the crystal by an evaporation or other process. Silver is generally used, although gold may be preferred where resistance to corrosion is paramount. Aluminum has not been widely used in wire-mounted units, because of the weak junction it makes with the solder. However, recent experiments indicate that an aluminum junction with a solder of indium (a rare, fusible metal, chemically similar to aluminum) is quite strong, so that eventually greater application may be found for aluminum-plated crystals. The mounting wires are normally of phosphor bronze, because of its high tensile strength and resistance to fatigue. A eutectic tin-lead solder is used that would normally be an alloy of approximately 63 percent tin and 37 percent lead by weight; however, to prevent an excessive diffusion of silver molecules from the silver spot into the solder during the soldering operation, the solder should contain 0.1 percent silver if the soldering is performed by hot-air blast, or a 59.5—34.5—6 percent tin-lead-silver combination if performed by hot iron. A solder cone in the shape of a bell (see figure 1-84) has been found to provide the best performance characteristics, and is the type of cone that

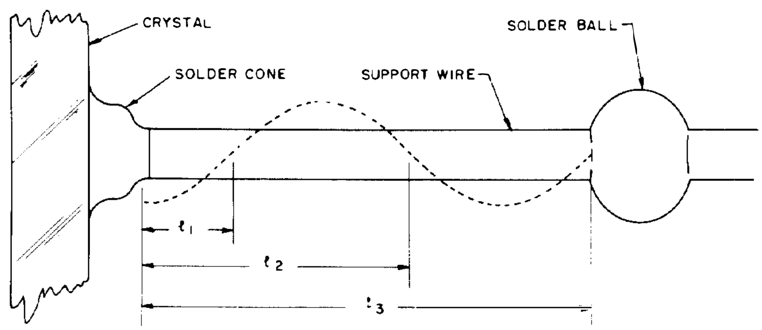


*Figure 1-84. Solder-cone resonant-wire support. The solder ball "tunes" the wire to the crystal frequency if it is placed at a distance equal to an odd multiple of a quarter wavelength (l_1, l_2, etc.) from the peak of the solder cone **

is least likely to rupture at the peak to form a "crater." For small crystals, the part of the wire enclosed by the cone may be straight, but for larger crystals sufficient anchorage requires that the end of the wire form a small hook. The wire is tuned to resonance by fixing the position of a solder ball at an odd quarter-wavelength from the peak of the cone; the solder ball serves as a "clamped" point for reflecting the wave energy back to the crystal. The "free" end of the wire is effectively at the point where it enters the solder cone. The distances l_1, l_2, and l_3 indicated in figure 1-84 mark "free lengths" of wire that will be resonant at the given wavelength. Note that each of the lengths defines a distance from the "free" end of the wire to a node where the solder ball should be placed.

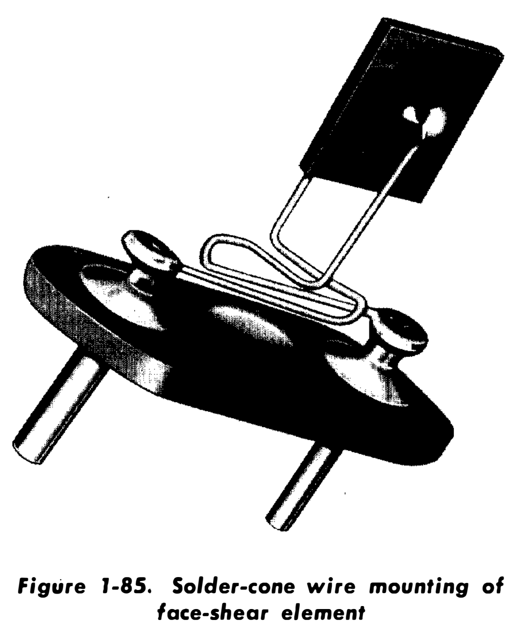


Figure 1-85. Solder-cone wire mounting of face-shear element

1-165. Theoretically, the resonant lengths l_1, l_2, l_3, . . . obey the same clamp-free frequency equation that is given for the cantilever clamp in paragraph 1-158. Experiment, however, has demonstrated that somewhat longer lengths are required for optimum performance. Normally, the free length of the wire is made a quarter-wave section, l_1, in the frequency range of 20 to 250 kc, and a three-quarter-wave section, l_2, in the range of 250 to 1000 kc. For phosphor-bronze wire, the empirical formulas for these distances are:

$$l_1 = 5.42 \sqrt{\frac{d}{f}} \text{ inches}$$

$$l_2 = 12.4 \sqrt{\frac{d}{f}} \text{ inches}$$

where: d = diameter of wire in inches (usually 0.0035, 0.005, 0.0063, or 0.008 in.).
f = frequency in kc.

1-166. After soldering to the crystal, the supporting wires are bent to make them serve as springs. One, two, or three bends are carefully spaced and directed so that the displacement per unit force will be the same for all directions. The ends of the wires are then soldered without tension to metal rods, or "straights," which in turn are welded to eyelets staked in a mica or bakelite base. In mounting small crystal plates, the straights are little more than short, metal stubs, but larger crystals are mounted in "cages" having a mica roof as well as a mica base. Figure 1-86 shows the cage assembly of a 40-kc length-width flexure crystal. The cage is formed by two mica plates at each end, and four straights. Besides providing for the proper mounting of the straights, the mica plates also serve as "bumpers" for the crystal. The inner and outer plates limit the horizontal and vertical displacements respectively, thereby protecting the unit from wire or crystal damage in the event of severe shock or vibration. The spacing between bumper and crystal is normally between 25 and 30 mils. Where the operating frequency is below 3 kc, the wavelength is usually sufficiently long for the entire wire to be cut to a resonant length, so that the soldered junction at the straight can serve as the nodal terminal. However, the optimum free length of wire becomes increasingly critical as the frequency is raised. Solder balls are used to establish resonance, but at low frequencies small metal disks are threaded on the wire to provide greater mass while permitting a precise adjustment to the correct position; after adjustment, the disks are loaded at the back with the correct amount of solder. Better control is obtained at the higher frequencies without the disk. The solder weights range from approximately 80 milligrams for 8-mil wire, for large crystals, to 6 milligrams for 3.5-mil wire, for small crystals.

1-167. The principal disadvantages of the solder-cone wire support arise from the effects of the solder cone upon the electrical characteristics of the crystal. To provide a junction of given mechanical strength, a certain quantity of solder is required. The solder, however, considerably increases the effective resistance of the crystal circuit as the temperature becomes high; if a high crystal Q at high operating temperatures is required, the solder cone must be small, and, consequently, the crystal unit cannot be as rugged mechanically as would otherwise be possible. Conversely, if the crystal unit is to withstand severe mechanical vibrations and high operating temperatures, the solder cone must be of maximum size, so that the Q and frequency stability will necessarily be at a minimum. Furthermore, as the volume of solder is increased appreciably, the temperature-frequency characteristics of the crystal may be considerably changed. Normally, the tendency will be for the zero temperature coefficient to shift to a lower temperature; in extreme cases, the zero point may be lost altogether. The temperature-frequency effects of hooked wire are generally more pronounced than those of straight wire, when equal volumes of solder are used. Another consideration is the difficulty experienced in making two cones of the same dimensions.

Headed-Wire Support

1-168. The headed-wire support (see figure 1-87) was developed to obviate the disadvantages of the solder cone, while preserving all the advantages of the wire type of mounting. The head of the wire, which resembles that of a common pin, has a diameter of approximately 22 mils for 6-mil wire. It is pretinned, and a small globule of solder is left at the end for sweating to the crystal; the volume of solder varies from 1000 to 7000 cubic mils, according to the size of the crystal. Phosphor-bronze wire is used, and all other mounting details are substantially the same as those for the solder-cone type of support.

1-169. The headed-wire is superior to the solder-cone mounting, because it provides a greater and more uniformly distributed mechanical support with a smaller quantity of solder; in the case of low-frequency crystals, the Q is improved by as much as twenty-five percent. Furthermore, the

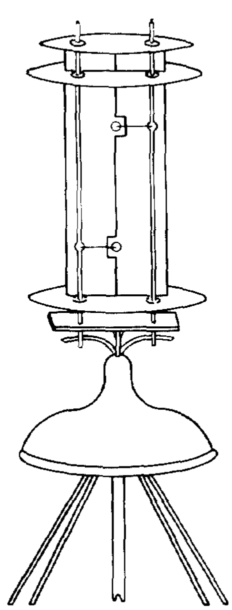


Figure 1-B6. Cage assembly for solder-cone wire mounting of low-frequency length-width flexure crystal *

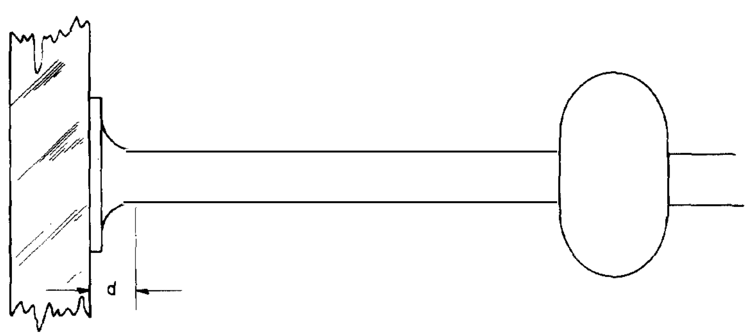


Figure 1-B7. Headed-wire crystal support *

distance d (figure 1-87) is a constant for all crystal units of the same design, so that the resonant free length of the wire can be predetermined accurately, thus permitting smaller tolerance in the rated characteristics. An additional advantage is that the headed wire diminishes the mechanical coupling between the vibrating systems represented by the crystal and the wires. Standing waves are caused, not only by reflections between solder ball and crystal, but also to a certain extent by reflections from one solder ball, through the crystal, to the solder ball on the opposite side. By reducing the coupling between crystal and wires, the impedance effects due to the interfering through-waves are reduced, and a purer frequency spectrum is possible. Headed wire may be used to replace any other type of low- and medium-frequency crystal mounting, and a well-designed headed-wire crystal unit will generally surpass the other types in all-round performance. However, at the higher frequencies a clamped air-gap holder is still to be preferred for greater activity and frequency stability, and at low frequencies, ultimately the cantilever clamp may prove superior for general use.

Edge-Clamped Mounts

1-170. Two variations of the edge-clamped type of mounting are illustrated in figure 1-88. The mount

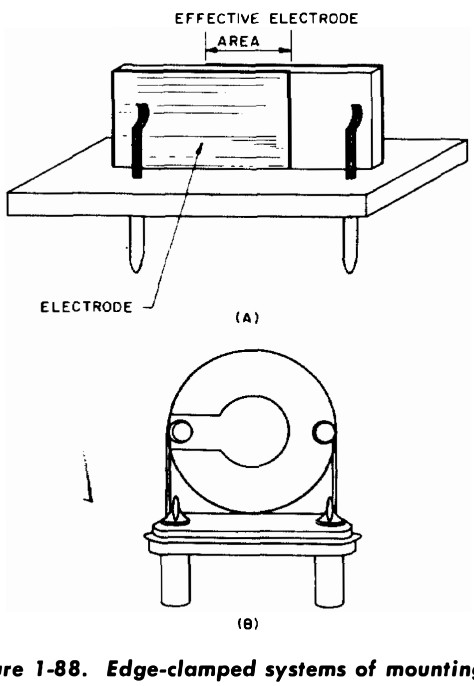


Figure 1-88. Edge-clamped systems of mounting. (A) Mounting for low-frequency crystals. (B) Cemented-lead mounting for high-frequency crystals

shown in (A) has been used with low-frequency crystals vibrating in extensional or flexure modes; the mount in (B), known as the cemented-lead mount, is widely used as an alternate to air-gap holders in mounting high frequency, thickness-shear elements. Although edge-clamped mounts have been successfully used in the production of high-activity crystal units for both high- and low-frequency applications, this type of mounting when used with low-frequency crystals, is probably somewhat inferior to well-constructed headed-wire or resonant-pin supports. However, a special feature of the edge-clamped mounting system is the method of dimensioning the electrodes (a method also adaptable for use with resonant pins), by which optimum performance characteristics can be obtained with high-frequency crystals. Plated electrodes (or metal foil cemented to the crystal) are used, but, as shown in figure 1-88(B), the crystal faces are only partially plated, and the plating on opposite faces is extended to opposite edges only, so that the effective electrode area is concentrated within a small circular region at the center of the crystal. By this means the capacitance is kept small, and the principal activity is confined to the central region, where the crystal is most likely to be of uniform thickness. Both of these factors are advantageous in improving the frequency stability. Also by reducing the activity in the vicinity of the edges, much of the damping due to the impedances of the supporting structure is obviated. Mechanical support and electrical connection is supplied by tinned, high-quality spring piano wire, which is clamped and cemented to the crystal at the edge where electrical contact can be made with the lead-outs from the electrodes. The cementing is used principally for the purpose of insuring good electrical connection, and not for supplying mechanical support, which should be provided by spring-wire clamps. The base ends of the spring wire are coiled around and soldered or welded to the base stubs. Although the supporting wires are not designed to be resonant elements of the crystal unit, they do provide the protection against shock and external vibration afforded by the other types of wire mounting. As compared with the performance of fundamental-mode, thickness-shear crystals, such as elements A and B, that are mounted in corner-clamped air-gap holders, the performance of the same elements, when wire-mounted, will generally be superior. In addition, the wire mounting permits the use of smaller crystal holders. The elimination of the air gap reduces the likelihood of arcing, but this does not mean that the wire-mounted units can be operated at higher voltages than the conventional air-gap crystal units. This is because the wire-mounted crystal is more isolated thermally and tends to become hotter. The advantages of the cemented-lead over the ceramic-button mounting system are less pronounced than the advantages over the other air-gap systems. For operation at frequencies from 1 to 10 mc, the wire-supported crystal usually has the better operating characteristics. As the frequency increases, however, the metal plating of the wire-mounted element becomes an increasingly greater factor in damping the oscillations; and in the upper very-high-frequency range, above 100 mc, non-plated crystals that are pressure-mounted between ceramic buttons are definitely to be preferred. Even in the fundamental frequency range, the ceramic-button mounts, which provide the better mechanical protection, may be used with good effect, and optimum performance characteristics for given operating conditions might better be achieved by combining the merits of wire-mounted edge clamps with those of plated dielectric buttons.

HOUSING OF CRYSTAL UNITS

1-171. The principal function of the housing is to provide a hermetically sealed, moisture-resistant container. Plastic housings of sandwich, air-gap, and clamp-type holders are normally sealed with neoprene gaskets. Natural rubber is not recommended, as the sulphur used in processing the rubber will ultimately contaminate other parts of the holder. Wire-mounted units are readily adaptable for housing in metal or glass tubes, employing standard radio parts; however, small, two-pin holders are generally preferred. Before sealing, a crystal unit is exposed to high temperature in an evacuated oven, in order to drive off adsorbed gasses. The sealing itself is usually performed in dry air, although certain crystals, particularly the flexure-elements, are sealed in vacuum. Optimum performance is obtained when a crystal is mounted in an evacuated container, since the damping effect of the air is eliminated.

1-172. If metal, rather than glass, housing is employed, it is difficult and expensive to seal a crystal unit so perfectly that not even minute leaks will develop due to stresses on the pins and the glass-sealing of the eyelets. For this reason most crystal units are sealed in dry air, so that if very small leaks are present, the crystal characteristics will not be appreciably affected for a long period of time. Leakage is minimized if the base is rigidly protected against deformation, and if the glass sealing fills the entire eyelet cavity uniformly. However, if a crystal is to be mounted in vacuum, a glass housing is to be preferred.

AGING OF CRYSTAL UNITS

1-173. "Aging" is a general term applying to any cumulative process which contributes to the deterioration of a crystal unit and which results in a gradual change in its operating characteristics. There are, of course, many interrelated factors involved in aging—minute leakage through the container, adsorption of moisture, corrosion of the electrodes, ionization of the air within the container, wire fatigue, frictional wear, spurious electrolytic processes, small irreversible alterations in the crystal lattice, outgassing of the materials composing the unit, over-drive, presence of foreign matter, various thermal effects, pin strain due to socket stresses, and erosion of the surface of the crystal. However, if a crystal unit is well designed and carefully constructed, the rated operating characteristics may well outlast the equipment in which the crystal is used.

1-174. Usually the first effects of aging can be traced to changes at the surface of the crystal. These changes may be due directly or indirectly to almost any combination of the factors mentioned in paragraph 1-173, and their occurrence can be avoided or greatly diminished only if proper precautions and techniques are employed during manufacture, and if low driving voltages are employed during operation. To produce a crystal unit of long life, the final stages of production require particular precautions. These concern the finishing processes of lapping, etching, cleaning, mounting, heat cycling, and protecting against moisture.

Lapping to Reduce Aging

1-175. Whether a crystal is being ground with abrasives which are cemented or imbedded in a grinding disk, or lapped with loose abrasives under a lapping disk, the cutting proceeds by virtue of the small fractures and chips which result when the hard, sharp edges of the abrasive particles are rubbed against the surface of the crystal. Commercial crystals are usually produced by lapping with loose abrasives, instead of grinding by "grindstones," except in the initial cutting stages and the final edging process, where diamond saws are commonly used. Each succeeding lapping stage employs a finer grade of abrasive, and must completely remove the surface left by the preceding stage. The final lapping requires very fine abrasive particles, such as 1000- to 1200-grain carborundum, and should preferably be performed in a mixture of abrasive, castile soap, and water. To reduce aging, soap and water are preferred as the coolant in the finishing stage, rather than kerosene or other oils, although kerosene permits a faster cutting rate for the same abrasive and lapping speed. Apparently, the residue of fractures remaining after a soap-water-abrasive lapping does not penetrate as deeply as that remaining after a kerosene-abrasive lapping. Regardless of how fine the abrasive, small fractures and cracks will be left in the surface of the crystal after the final lapping, and in time these cracks will spread, absorb moisture, and ultimately result in a weathering of the surface. Additional care must be taken to ensure that the crystal is not finished with slight concavities in the surfaces, or with one end lapped down more than the other, making the crystal wedge-shaped. Although optimum performance is to be obtained with perfectly planar surfaces, greater insurance against unwanted non-parallelisms is gained if the lapping is controlled to give the plates a symmetrical convex contour of approximately 5 microns for lower-frequency crystals, and approximately $10^7/f$ (cycles) microns for crystals above 3 mc.

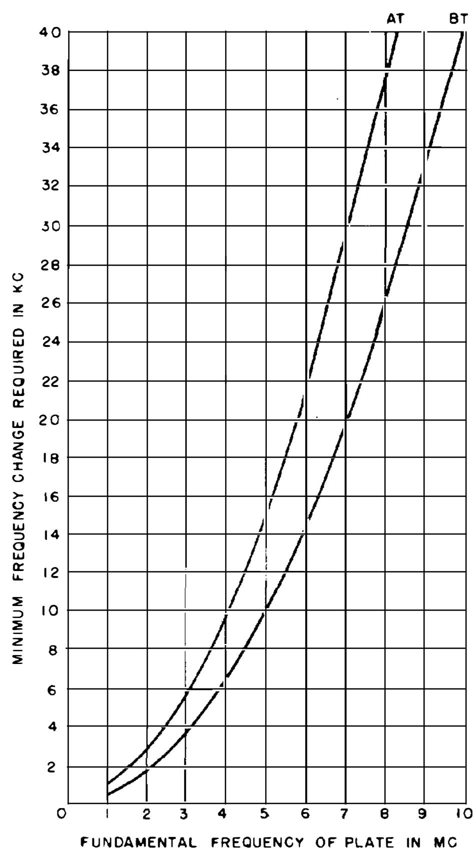


Figure 1-89. Minimum change in frequency that AT and BT plates must undergo due to etching, if the etching is to be sufficient to remove all surface cracks and fissions remaining from the final lapping stage. Note that, as the crystal becomes thinner, a given change in the thickness dimension means a greater change in the frequency. The frequency change for a BT cut is less than that for an AT cut of the same initial frequency, since the larger frequency constant of the BT cut permits a thicker plate

Etching to Reduce Aging

1-176. After the final lapping stage, the crystal is normally given an etching bath to remove all foreign particles. An eight-minute bath in forty-seven percent hydrofluoric acid is sufficient for the average crystal, and will permit a firm contact between the crystal and its electrode coating. An etching time of at least thirty minutes is necessary, however, if a minimum aging and a maximum Q, stability, and drive level are desired. The longer etching period is required to ensure that the deeper fissions in the surface caused by the final lapping are thoroughly removed. However, the deep etch is difficult to control, and particular care must be exercised if the desired dimensions of the crystal are to be achieved. It is customary to divide the deep-etching process into two, or more, steps: (1) to etch the crystal to within 1 kc of the desired frequency; and (2) in the succeeding steps, to bring the crystal within its tolerance limits. Figure 1-89 indicates the degree of etching required to prevent aging in AT and BT cuts.

Cleanliness to Reduce Aging

1-177. The protection of a crystal from foreign matter and moisture is of paramount importance if the crystal is to operate with stability and long life. Only minute traces of dirt, dust, or fingerprints on a crystal will cause the performance to be erratic. Cleanliness is necessary throughout the final production period, but particular emphasis is required during the stages immediately prior to sealing. Before and after etching, each crystal blank should be scrubbed thoroughly in soap, or trisodium phosphate, and water with a soft brush; rinsed in 0.5 percent ammonium hydroxide solution, and again washed thoroughly in running water; dried in an oven heated to 100 degrees centigrade, or in a warm, clean, air stream; washed again in distilled carbon tetrachloride or other solvent; rinsed in hot distilled water; and finally, carefully dried in an oven. The electrodes and holder must be similarly cleaned, and neoprene tweezers should be used in handling the parts during the final assembly. If the crystal is to be metal-plated, the complete mounting must be cleaned again before sealing. A hot spray of distilled trichloroethylene for one-half minute is sufficient. The plated crystal will normally require a small amount of edge-grinding with fine emery paper to bring the mounted unit to the proper frequency. This step unfortunately weakens the aging resistance of the treated surfaces at a stage when further etching is no longer feasible for commercial crystals. However, before testing and sealing, a retouched crystal unit should be thoroughly washed and scrubbed, with every precaution taken to ensure that no foreign matter remains on the crystal or mounting. Where the facilities are available, cleaning can be performed by exciting the bath with supersonic acoustic waves, which can clean the crystal by shaking all loose fragments off its surface. In fact, a supersonic bath can be quite as effective as an etching bath in reducing aging.

Mounting to Reduce Aging

1-178. As a general rule, any deviation in the mounting which causes an increase in the frictional losses will shorten the useful life of a crystal unit. Thus, in the nodal types of mounting, small deviations from the nodal point in the position at which a crystal is held will shorten the life of the crystal. Wire-mounted crystals require additional precautions during fabrication to avoid local changes or stresses at the surface of the crystal. Particular care must be taken to avoid electrical "twinning," which will occur if the temperature is raised above the inversion point, 573°C, and then lowered again; or twinning may be induced at a much lower temperature if a sharp temperature gradient is present in the crystal. These precautions are necessary during the baking of the silver spots, the division of the electrode coating by electric stylus, and the soldering operation. In baking the silver spots, the temperature should be kept forty to fifty degrees centrigrade below the inversion point, and care must be taken to make certain that the crystals are heated uniformly. Some twinning is inevitable when using an electric stylus to divide an electrode coating; however, if straight-line division is required, the twinning may be avoided by using an abrasive tool or sand blasting in place of the stylus. To avoid thermal stresses during the soldering operation, a heated support should be provided for heating the crystal uniformly to a temperature of approximately 100°C. Twinning, regardless of its cause, primarily affects the steady-state electrical characteristics of a crystal element, and only indirectly contributes to gradual changes in the performance of the crystal. At least, no statistical data has been collected to show a correlation between twinning and aging; nevertheless, a series of small twinned spots at the surface is likely to make the area more susceptible to erosion. Readjustments of the crystal lattice at the twinning boundaries after long periods of electrical, mechanical, and thermal stresses might be expected; but if these are due to occur, they can probably be made to take place by a process of artificial aging before the crystal is placed into operation. Twinning, however, raises the inductance and effective resistance of an element, and hence, decreases its activity for a given operating voltage. Since the ultimate requirement of a higher operating voltage can lead to a shortening of the life of the crystal unit, an undue amount of twinning indirectly becomes a factor in the aging. Twinning will also raise or lower the frequency, according to the particular type of element. If the twinning is introduced during the final stages, this may require a substantial amount of edge-grinding during the final frequency-adjustment stage, and more of the etched surface may need to be removed than otherwise. Thus, although "heat" twinning is considered primarily in connection with its immediate effect upon the characteristics of the crystal, it should also be avoided as an indirect factor in aging. A more direct factor in shortening the life of a wire-mounted crystal unit is a nonuniformity in the soldered junction, which is more likely to occur in a solder-cone than in a headed-wire support. When the stresses are unevenly distributed, the soldered junction itself will tend to age; and even if mechanical breakage does not occur, the changes in the electrical characteristics will lead to poor performance and instability. Special care must be taken to make certain that the silver spots are of uniform density. The containers of liquid silver should be agitated for several hours immediately prior to application. Also, the critical nature of the soldering operation requires the aid of a machine and accessories of special design.

Heat Cycling to Reduce Aging

1-179. A newly mounted crystal will normally appear to age more rapidly than one that has been in operation for a long period of time. This effect is not due to an actual deterioration of the crystal unit, but merely to an initial adjustment of the crystal, particularly at its surface, to its operating environment and changes in temperature. The stabilization period can be reduced to one of very short duration by subjecting the crystal unit to a series of slow heating and cooling cycles varying between 24°C and 116°C. Metal-plated elements are frequently heat-cycled during the final frequency-adjustment period, and again after sealing. In a series of tests at the Hunt Corporation, it was found that negative aging (frequency decreases with time) is generally due to insufficient cleaning of the crystal unit. When this was remedied, it was found that the crystal units would then age positively. The cause of the positive aging was traced to the outgassing of the metal plating of the crystal, and its elimination has been achieved by pre-aging the plated crystal for 3 minutes in a 300°C oven. After a sufficient period of artificial aging, a properly fabricated and operated crystal unit will maintain its final temperature-frequency characteristics indefinitely.

Low Relative Humidity to Reduce Aging

1-180. A low relative humidity is of paramount importance if excessive aging is to be prevented. Even if a crystal is perfectly mounted and clean, an ambient relative humidity higher than 40 percent will sharply increase the insulation resistance, and will greatly accelerate the weathering of the surface of the crystal and the corrosion of the electrodes. For optimum performance and long life, every precaution must be taken to ensure that the interior of the sealed crystal unit is as free as possible from moisture. Prior to sealing, all components of the crystal unit should be heated in vacuum to drive off absorbed water vapor and other gases; and if the sealing is performed in air, the atmosphere should not have a relative humidity higher than 5 percent.

Low Drive Level to Reduce Aging

1-181. As a general rule, the lower the drive level, the longer will be the useful life of a crystal unit. This is true because the cumulative effects of almost all of the previously discussed aging factors are considerably more pronounced when the crystal is operated at high drive levels. Also, the higher operating voltages greatly increase the tendency toward corona discharge and other ionization effects, and the vibrations of greater amplitude are more likely to result in crystal or wire fatigue. To ensure maximum lifetime, a piezoelectric resonator should be driven at the lowest practicable level consistent with the circuit requirements.

ELECTRICAL PARAMETERS OF CRYSTAL UNITS

EQUIVALENT CIRCUIT OF CRYSTAL UNIT

1-182. A crystal unit may be represented by the equivalent electrical circuit shown in figure 1-90. R_I represents the terminal-to-terminal r-f insulation resistance of the crystal unit. C_L, L_L, and R_L represent, respectively, the distributed capacitance, inductance, and resistance of the electrical leads and terminals of the mounted crystal. C_L, in addition, includes the capacitance across any electrode parts that extend beyond the quartz. C_{H1} and C_{H2} represent the distributed capacitance of the crystal circuit to the holder H. C_A represents the capacitance between the electrodes and the crystal faces when they are separated by an air gap or other dielectric. If a dielectric exists on both sides of the crystal, C_A would equal the total capacitance of the two capacitances in series. Thus, if the air-gap capacitances on the opposite sides of the crystal were equal, as would normally be the case, C_A

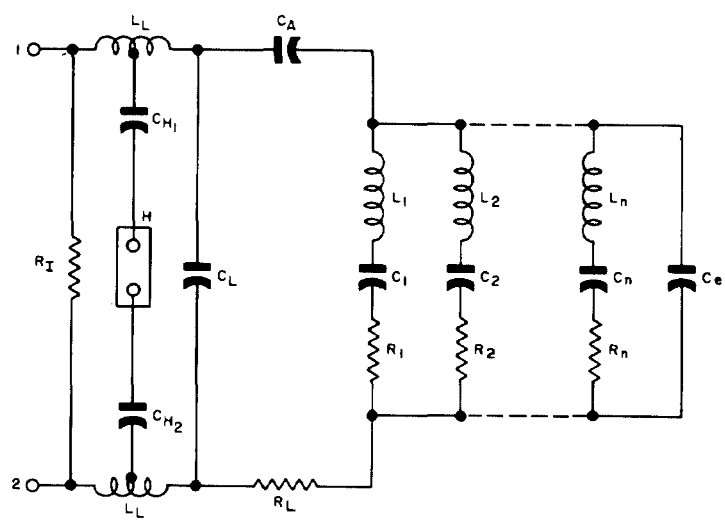


Figure 1-90. Equivalent circuit of crystal unit

would be equal to one-half the value of either one. C_e is the electrostatic capacitance across the quartz plate, where the quartz serves as the dielectric. The series LCR branches represent the piezoelectric properties of the crystal as they appear to the external circuit when the crystal is undergoing mechanical vibrations. For this reason, these values are called the "motional-arm" (also, "series-arm") parameters, in contradistinction to the parameters such as C_e that are not of piezoelectric origin.

1-183. The motional-arm values of L are closely associated with the mass of the crystal, those of C are closely associated with the elasticity of the crystal, and the motional-arm values of R indicate the tendency of the crystal to dissipate heat during vibration. Each of the motional-arm branches is associated with a different mode or harmonic of vibration, and the normal frequency of each of the modes coincides with the series-resonant frequency of the respective LCR branch. It will be assumed that the branch indicated by L_1, C_1, and R_1 represents the equivalent circuit of the desired mode, and that all of the higher subscript branches L_k, C_k, R_k, represent unwanted modes.

1-184. Since a crystal unit is normally intended for use only within a very narrow frequency range centered at a specified nominal frequency, the equivalent circuit may be greatly simplified to that shown in figure 1-91. If the crystal is mounted so that the electrodes are in direct contact with the crystal faces, C_A will not be effective, and the values of L, C, and R in figure 1-91 will normally be approximately the same as those of L_1, C_1, and R_1 in figure 1-90, and C_o will approximately equal $C_e + C_L + \dfrac{C_{H1}\,C_{H2}}{C_{H1} + C_{H2}}$. For these assumptions to hold, R_1 must be much greater than the impedance of the crystal when parallel resonance is established between the motional arm and C_o. Also, the operating frequency must not be so high that the reactance of L_L becomes significant; and the normal frequencies of all the unwanted modes must be sufficiently removed from the nominal frequency, if each of the unwanted branches is to present a high impedance at the desired operating frequency.

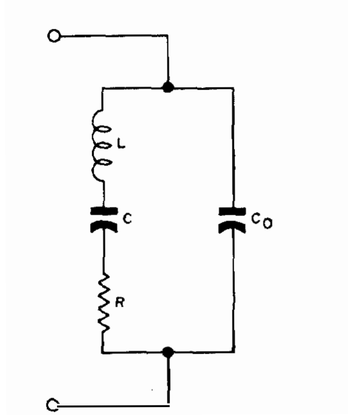


Figure 1-91. Simplified equivalent circuit of crystal unit

SIMPLIFIED EQUIVALENT CIRCUIT OF AIR-GAP CRYSTAL UNIT

1-185. C_e is normally much greater than the distributed capacitance across the leads, so an air-gap or dielectric-sandwich type of crystal unit may be represented by the equivalent circuit shown in figure 1-92. This circuit, in turn, may be reduced to the equivalent circuit of figure 1-91 by assigning the following values to L, C, R, and C_o:

$$L \quad \left(\frac{C_A + C_e}{C_A}\right)^2 L_1$$

$$C = \frac{C_A^2\, C_1}{(C_A + C_e)(C_1 + C_e + C_A)}$$

$$R = \left(\frac{C_A + C_e}{C_A}\right)^2 R_1$$

$$C_o = \frac{C_A\, C_e}{C_A + C_e}$$

THE EFFECT OF R-F LEAKAGE RESISTANCE

1-186. The principal effect of R_I, the terminal-to-terminal r-f leakage resistance shunting the crys-

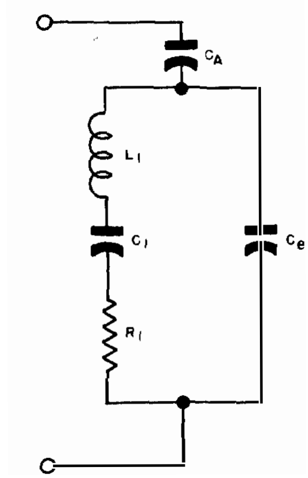


Figure 1-92. Simplified equivalent circuit of air-gap crystal unit

tal, is to reduce the effective Q of the crystal unit. For all practical purposes this effect is negligible when the crystal is being operated at, or very near, the resonant frequency of the series arm. Under these conditions the electrical impedance of the crystal is so small by comparison that R_I can be ignored. On the other hand, as the frequency rises above the resonant point, the impedance increases sharply, and the greater the impedance becomes, the greater is the effect of a given R_I. Insofar as the equivalent circuit of figure 1-91 is concerned, the effect will be to increase the value of R. The extent of this increase will depend upon how large the effective reactance of the crystal becomes, relative to R_I. For the sake of simplification, most of the discussion given later concerning the equivalent circuit assumes that the increase in R due to R_I is negligible, or at least, is constant, regardless of the frequency, an assumption that can produce reasonably accurate results in the case of well-fabricated crystal units. The leakage resistance of military crystal units has a specified minimum d-c value of 500 megohms. As long as this minimum d-c value is maintained, R_I at low frequencies will be comparable to this value. However, if an accumulation of moisture, dirt, or the like seriously reduces the d-c insulation resistance below the allowed minimum, the off-resonance characteristics will undergo a noticeable change. For instance, low-frequency filter crystals may have impedances at antiresonance in the neighborhood of 50 to 100 megohms. If R_I decreases below 500 megohms, the equivalent R of the motional arm will increase markedly. In the case of high-frequency crystal units, the effective dielectric losses may become relatively large, particularly when plastic holders are used, so that, at off-resonant frequencies, R_I can become a significant parameter of the over-all effective resistance. However, for high-frequency crystal units employing modern methods of mounting and construction, R_I can generally be ignored. In the very-high-frequency range, crystal units are almost always operated at series resonance, so that, even if R_I were on the order of 100,000 ohms, as might easily be the case, the effect would still be relatively minor. However, where the shunt resistance cannot be ignored, a more concrete analysis of its effect is to let R_I, in figure 1-90, represent only the d-c leakage resistance, and to account for the r-f dielectric losses by inserting other equivalent resistances in series with the various shunt capacitances. In the simplified equivalent circuit in figure 1-91, the d-c leakage resistance could still be ignored, but non-negligible r-f shunt losses could be interpreted as being due to a single resistance in series with the shunt capacitance, C_o. The Q of the equivalent shunt arm will effectively equal the Q of the crystal unit when the unit is operated at frequencies well removed from resonance. The crystal units that are mounted in metal or glass holders of the type described in Section II, and are recommended for use in equipments of new design, can be expected to have shunt-arm Q's greater than 1000 at all frequencies within their specified range. This assurance, however, cannot be given for the crystal units mounted in plastic holders, particularly the old-style phenolic holder, or for those employing all-metal sandwich or air-gap electrodes. However, the lower Q's of the older types of crystal holders are not entirely due to greater dielectric losses and larger values of shunt capacitance. An equally important factor is the effective inductance of the circuit effectively in series with the shunt capacitances. For example, a corner-clamped air-gap mounting, such as that provided in a DC-31 crystal unit, has an effective shunt-arm Q of approximately 80 or 180 at 30 mc, depending upon whether the clamping pressure is applied by a coiled or a flat spring, respectively. Apparently, the reactance and resistance of a coil spring can be quite detrimental to the quality of a crystal holder at very high frequencies, since it can cause not only an effective increase in the shunt capacitance, but also an increase in the effective dielectric losses. These losses would become prohibitive if the inductance of the spring and its stray capacitance should approach the properties of a series-resonant arm shunting the crystal. However, except in such abnormal cases, and in cases where the insulation is weakened by extremes in humidity and temperature, the shunt resistance will have a negligible effect upon the performance of a crystal circuit.

EFFECT OF DISTRIBUTED INDUCTANCE

1-187. The effective self-inductance of the crystal leads, L_L, is normally not sufficient to seriously affect the crystal parameters, except in the case of very high operating frequencies where it is necessary to operate the crystal at series resonance. At resonance, the reactance of the crystal unit will be zero, so that the crystal, in combination with its shunt capacitance, must have a net equivalent series X_C equal in magnitude to the X_{LL} of the distributed inductance. This means that the resonant frequency will be slightly lower than would be the case if there were no distributed inductance. The net effect on the equivalent circuit of figure 1-91 is that the LC product is increased very slightly (lower resonant frequency), and that C_o

is increased to a greater extent. If the distributed inductance is completely negligible, the resonant frequency of the crystal will be slightly higher than the normal resonant frequency of the series arm, because of the reactive component of current through C_0. However, the distributed X_L of the lower-frequency crystal units may be sufficient to approximately cancel the reactance due to the true C_0. Under these conditions, the resonant frequency of the crystal unit as a whole would coincide with the natural vibration frequency of the crystal—an ideal operating state. In the case of the higher-frequency crystal units, the distributed inductive reactance may be sufficient to lower the frequency below the natural resonance point by several cycles. If the crystal unit were being operated at series resonance in a capacitance-bridge circuit, for example, such an effect would lead to frequency jumps with slight changes in the tuning adjustments. Under such conditions it would be desirable to add a capacitance in series with the crystal, with a reactance just sufficient to cancel the unwanted X_{LL}. The distributed inductance, L_L, of the lower-frequency crystals, and of practically any crystal unit which is to be operated above series resonance, has only a minor effect. The maximum effect will always be at very high frequencies near series resonance. In analyzing the behavior of a crystal unit where the distributed X_{LL} cannot be neglected, the simplest approach is to consider X_L as a separate fixed reactance in series with the crystal unit. From this point of view, as long as X_{LL} is very small, as compared with X_{C_0}, it can be seen that L_L will not seriously affect the rate at which the net crystal reactance will change with frequency, and, therefore, will not influence the stabilizing effect of the crystal on the frequency. Crystal oscillators can operate successfully up to frequencies as high as 200 mc. However, crystal-control of the frequency can be stable only when the impedance at series resonance is much smaller than the reactance of the effective shunt capacitance C_0. The larger the value of X_{LL}, the smaller this ratio will be. Thus, the higher the frequency, the greater the importance of keeping the crystal leads as short as possible, not only to reduce L_L, but also to reduce the distributed capacitance and the r-f resistance of the wires. The small coaxial-electrode type of mounting, such as the HC-10/U, is the most satisfactory for achieving a minimum effective C_0, and hence, a maximum frequency stability in the very-high-frequency range. It should be remembered, however, that since the distributed X_{LL} will increase with the frequency, the effective C_0 will also increase with the frequency, so that a measurement of C_0 at a frequency far lower than that of resonance will not alone give a reliable indication of the effective parameter near the operating frequency.

EFFECT OF DISTRIBUTED CAPACITANCE

1-188. The effect of the distributed capacitance on the parameters of the simplified equivalent circuit is merely to increase the value of C_0. However, it should be noted that the amount of the increase will depend somewhat upon how the crystal unit is connected in the external circuit. For example, assume that the holder and terminal 1 in figure 1-90 are grounded. C_{H1}, which would otherwise be in series with C_{H2}, is now effectively short-circuited, so that the total shunt capacitance C_0 is increased. If C_{H1} were assumed to be equal to C_{H2}, the amount of the increase due to grounding terminal 1 and the holder would equal $C_{H1}/2$. On the other hand, grounding the holder can result in an effective decrease in C_0. Assume, for instance, that a crystal unit is connected in a circuit equivalent to that shown in figure 1-93. With the metal holder ungrounded, C_{H1} and C_{H2} are effectively connected in series, so that, if $C_{H1} = C_{H2}$, the total capacitance of the series combination is $C_{H1}/2$. If switch S is closed, thereby grounding the holder, the effective total C_0 becomes larger or smaller, depending upon the point of view of the observer. Since C_{H1} is no longer in series with C_{H2}, but, instead, is shunted across the entire circuit, whereas C_{H2} is shunted across the load Z, the total capacitance facing the generator is increased (assuming that Z is the reactance of a capacitor). When S is closed, the current through M_1 increases; however, the current through M_2 decreases. An observer at M_1 would say that grounding the holder increased C_0, whereas an observer at M_2 would say that C_0 has decreased. At frequencies well removed from the nearest resonant frequency of the motional arms, the branch impedances are so

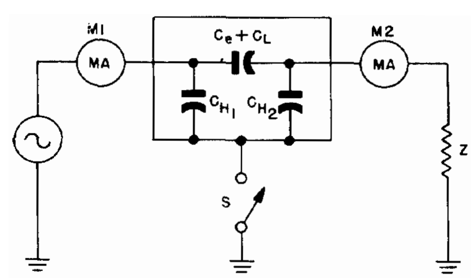


Figure 1-93. Circuit diagram indicating the effect that grounding a metal holder may have on shunt capacitance

high that the crystal unit behaves essentially as a capacitor of value C_o. It can be seen that if a measurement were being made of the change in C_o due to the grounding of the holder, it would be important to know exactly how the measurements were made. For example, low-frequency crystal units mounted in the HC-13/U have been reported as having 0.8 $\mu\mu$f less shunt capacitance, and medium-frequency crystal units mounted in the HC-6/U holder have been described as having 0.5 $\mu\mu$f less shunt capacitance with the holder grounded. However, it should be noted that crystal units so specified are intended primarily for use in circuits where the crystal operates in a series-resonant rather than a parallel-resonant circuit. Even so, the grounded holder alters the entire circuit, not simply C_o. Thus, in the circuit of figure 1-93, suppose that it is necessary for the current through Z to be in phase with the generator voltage. If the circuit is properly adjusted with an ungrounded holder, grounding the holder will detune the circuit by effectively decreasing C_o, on the one hand, and on the other, by shunting the load with C_{H2}. The over-all effect cannot be predicted simply by specifying an effective change in C_o, since the end result will depend upon the impedance characteristics of the entire circuit. If the frequency of a crystal oscillator is being measured by beating its output with the output of a frequency standard, it is common practice to touch the crystal holder with the hand in order to determine whether the crystal unit which is touched is operating at a frequency above or below that of the standard oscillator. The oscillator frequency will be higher or lower than that of the standard according to whether the hand capacitance causes the beat frequency to fall or to rise, respectively, provided that the effective C_o is actually increased by the touch of the hand, as is invariably assumed. Before this assumption is made with complete assurance, however, the response of the circuit to a grounded holder should be known.

EFFECT OF DISTRIBUTED RESISTANCE

1-189. R_L is assumed to include only the ohmic resistance of the electrical leads and the reflected resistance due to eddy currents in the holder and ground connections. At normal frequencies R_L is quite small, as compared with R_1; even the small-sized supporting wires of wire-mounted crystals have r-f resistances that are measurable in tenths of an ohm. As in the case of the other distributed parameters, the effect of R_L upon the equivalent circuit of figure 1-91 becomes more pronounced at the higher frequencies. To a first approximation R is simply $R_1 + R_L$. However, at frequencies above 10 mc, the r-f resistance of the leads increases directly as the square root of the frequency, so that, in the v-h-f range R_L may be greater than one ohm. R_L will also increase somewhat if the holder is grounded, as the increased eddy-current losses in the shielding will be reflected as additional resistance losses in the crystal circuit. At normal frequencies, however, the effect of R_L is of minor importance; and even at frequencies above 100 mc, its consideration is secondary to the effects of the distributed capacitance.

RULE-OF-THUMB EQUATIONS FOR ESTIMATING PARAMETERS

1-190. The crystal parameters for a given frequency vary rather widely from one crystal unit to the next. Even crystal units of similar dimensions and fabrication made by the same manufacturer may show significant differences between corresponding parameters. These differences arise from the sensitivity of the quartz plate to slight changes in its angle of cut, surface state, electrode area, soldered connections, and the like. The parameter with the greatest percentage variation is R, and it is not uncommon for the larger values of R to be from 300 to 900 percent greater than the minimum values. The most predictable parameter is C_e, since it is primarily a linear function of the electrode area, the thickness of the quartz dielectric, and the dielectric constant, all of which are reasonably constant for a given fabrication technique, although variations may be expected in crystal units of the same nominal frequency and type of mounting, when made by different manufacturers. For the same manufacturer, nominal frequency, and type of crystal unit, however, C_e rarely varies by more than $\pm 5\%$ of its nominal value. With a reasonably constant C_e as a starting point, approximate values for the major parameters L, C, R, and C_o may be predicted for the principal types of crystal elements and holders. First, C_e is computed from the known values of plate area, dielectric thickness, and dielectric constant. Next, C can be found, since it is theoretically equal to C_e times a constant of proportionality. L can next be computed, since the LC product must conform to the nominal frequency. Next, an approximate range of the values of R may be estimated from the empirical values of the crystal quality factor, Q. Since Q is the ratio X_L/R (or $-X_C/R$), R is thus equal to X_L/Q. Finally, C_o can be estimated by simply adding to C_e the approximate total distributed capacitance common to the particular type of holder and mounting.

Estimating C_e, Static Capacitance of Crystal

1-191. Although the dielectric constant of quartz varies somewhat according to the angle of cut, the following formula will be approximately correct for plated electrodes:

$$C_e = 0.402\ A/t\ \mu\mu f \qquad 1—191\ (1)$$

where A is the effective electrode area in square centimeters, and t is the thickness in centimeters.

1-192. In the case of partially plated A elements, where t is a function of the nominal frequency and the harmonic, equation 1-191 (1) may be expressed as:

$$C_e = 2.42\ Af/n\ \mu\mu f \qquad 1—192\ (1)$$

where f is the nominal frequency in mc/sec, and n, an odd integer, is the harmonic of the thickness-shear vibration. Although the quartz plates range from 1 to more than 2 sq cm in plate area, the electrode area normally covers only a fraction of the total quartz surface. The RTMA Standards Committee on Quartz Crystals has recommended the following approximate electrode areas for the fundamental frequencies of this type of crystal unit.

Frequency in mc/sec (n = 1)	*Electrode Area ±10%* (sq cm)
1 — 2	0.504
2 — 5	0.385
5 — 9	0.283
9 — 15	0.159
15 — 20	0.126

For the overtone modes, where n is greater than 1, the electrode area will be the same as that of the fundamental mode of frequency equal to f/n. The harmonics for various ranges of f are as follows:

$$f = 10—\ 45\ \text{mc};\ n = 3$$
$$f = 45—\ 75\ \text{mc};\ n = 5$$
$$f = 75—105\ \text{mc};\ n = 7$$

1-193. In the case of crystals vibrating in a face-shear mode, it is the electrode area A that is a function of the frequency. For fully plated C elements, equation 1-191(1) may be expressed as:

$$C_e = 0.038/tf^2\ \mu\mu f$$

where t has an average value of 0.05 cm, and f (mc/sec) lies between 0.3 and 1 mc/sec.

1-194. For fully plated D elements, equation 1—191(1) may be expressed as:

$$C_e = 0.0172/tf^2\ \mu\mu f$$

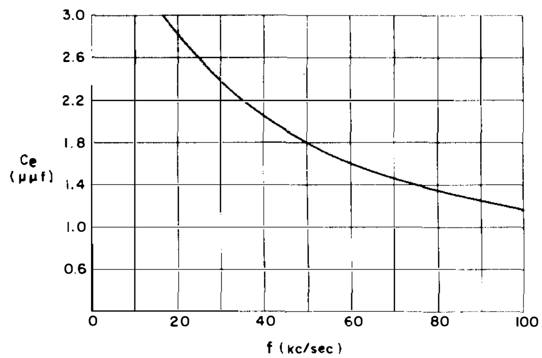


Figure 1-94. C_e versus frequency for typical wire-mounted N elements

where t has an average value of 0.05 cm, and f (mc/sec) lies between 0.2 and 0.5 mc/sec.

1-195. For a typical wire-mounted J element, equation 1—191 (1) may be expressed as:

$$C_e = k/f\ \mu\mu f$$

where: k = 38 for f = 1.2 to 2.5 kc/sec
= 45 = 2.5 to 4.0 kc/sec
= 58 = 4.0 to 6.6 kc/sec
= 77 = 6.6 to 10.0 kc/sec

Note that f in this case is to be expressed in kc/sec.

1-196. Typical values of C_e for an N element are shown in figure 1-94.

Estimating C, Equivalent Motional-Arm Capacitance

1-197. After C_e is known, an approximate value for C at the fundamental frequency can be readily obtained from the following equation:

$$C = C_e/r_e \qquad 1—197\ (1)$$

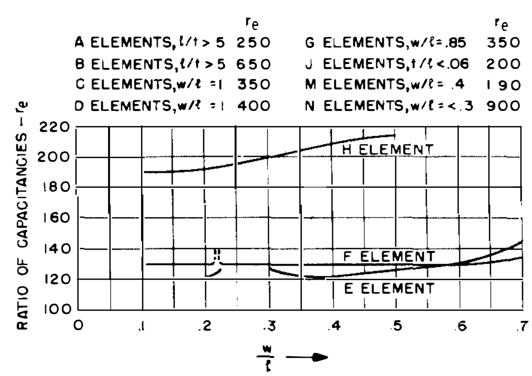


Figure 1-95. Approximate values of the ratio of capacitances, $r_e = \frac{C_e}{C}$, for various plated crystal elements

where r_e is simply the ratio of the electrostatic capacitance C_e to the motional capacitance C, with C_e and C expressed in the same units. The values of r_e for the more important elements are given in figure 1-95. For the odd harmonics (n) of the thickness-shear modes:

$$C = C_e/r_e n^2 \qquad 1\text{—}197\ (2)$$

Estimating L, Equivalent Motional-Arm Inductance

1-198. Since X_L is equal to X_C at the series-resonant frequency of the motional arm, L is found quite simply, once f and c are known. Thus:

$$L = \frac{1}{4\pi^2 f^2 C} \text{ henries} \qquad 1\text{—}198\ (1)$$

Remember, however, that f is expressed in cycles/sec, and C in farads.

Estimating R, Equivalent Motional-Arm Resistance

1-199. A theoretical equation for R would not be practical, since this parameter is much too sensitive to slight variations during the fabrication process and to changes in the crystal drive. An approximate estimate is gained from observations of the value of Q for the various frequency ranges. Thus:

$$R = \frac{2\pi f L}{Q} \text{ ohms} \qquad 1\text{—}199\ (1)$$

where f is in cycles/sec, and L is in henries. The values of Q will range from 10,000 to 200,000, and in exceptional cases will have much higher values. Generally, the higher Q's are to be found at the higher frequencies. For face-shear elements, the average Q is approximately 30,000, with most values falling between 10,000 and 40,000. Thickness-shear elements will have average Q's of approximately 75,000, and most of the values will lie between 35,000 and 100,000.

1-200. The Q is not a dependable parameter, and will vary from frequency to frequency, and from manufacturer to manufacturer, for the same type of crystal unit. For example, when expressed as $Q = -X_C/R = \frac{1}{2\pi f C R}$, it can be seen that Q is inversely proportional to C, and thus might be considerably increased by simply reducing the area of the electrodes. On the other hand, the resistance, R, is at least limited in practice by military specifications. For this reason, the typical

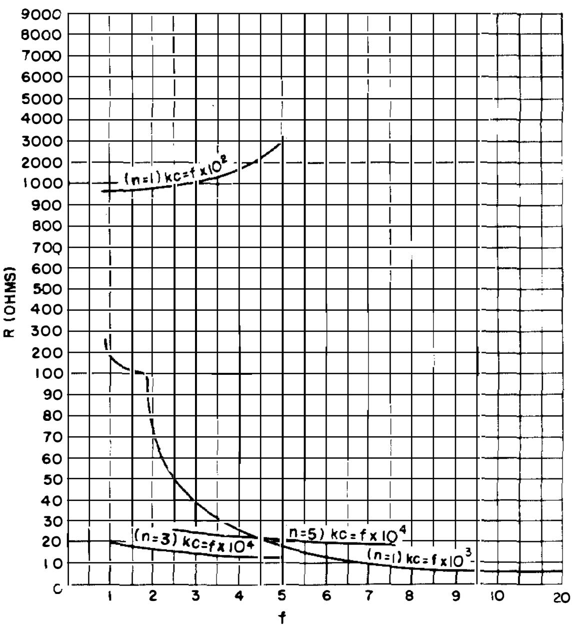


Figure 1-96. Typical curves of the series-arm resistance of plated crystals versus frequency. Actual series-arm resistances will vary between R/3 and 3R, where R is the value shown, except when R is less than 10 ohms, in which case the minimum resistance will be approximately one-half the value indicated. Values indicated are average for fundamental modes and approximately ½ the average for overtone modes

values of R versus f, shown in figure 1-96, are more likely to be found in randomly selected crystal units than is a given value of Q. The values of R indicated in figure 1-96 are merely typical, however, and a small percentage of actual Military Standard crystal units will have series-arm resistances as small as one-third, or as large as three times the amounts shown.

Estimating C_o, Total Static Shunt Capacitance

1-201. The equation for C_o is

$$C_o = C_e + C_d$$

where C_d is the total distributed capacitance of the crystal leads and terminals. Approximate values of C_d for plated crystals in ungrounded holders are given below:

Crystal Holder	C_d *(μμf)*
HC-6/U	0.7
HC-10/U	0.3
HC-13/U	1.0
HC-15/U	1.5

IMPEDANCE CHARACTERISTICS VERSUS FREQUENCY

1-202. The superiority of the quartz crystal as a frequency stabilizer lies in the fact that a small change in the frequency will cause a much larger change in the impedance of the equivalent circuit than can be obtained with conventional inductor-capacitor networks. Where an ordinary r-f tank coil would have an inductance measured in microhenries, and an effective Q of 10 to 250, the equivalent circuit in figure 1-91 will have an inductance measured in henries and a Q of 10,000 to 250,000 or more. C, of course, is extremely small, since its reactance must equal X_L at resonance, and is commonly expressed in milli-$\mu\mu$f (thousandths of a micromicrofarad). R is expressed in ohms, and although at low frequencies it may have values higher than 3000 ohms, depending upon the particular crystal element and method of mounting, the more common values lie between 10 and 100 ohms. C_o normally lies between 3.5 and 14 $\mu\mu$f, although much larger values are encountered where electrodes of large surface area are employed. Among the smaller holders, such as types HC-6/U and HC-10/U, values of 5 to 6 $\mu\mu$f are quite common.

1-203. Since $X_L = 2\pi fL$

$$\text{and } X_C = \frac{-1}{2\pi fC}$$

then, the rates at which X_L and X_C change with frequency will be, respectively:

$$\frac{dX_L}{df} = 2\pi L$$

$$\frac{dX_C}{df} = \frac{1}{2\pi f^2 C}$$

Note that both of these derivatives indicate a positive change in reactance with an increase in frequency. However, it should be remembered that X_C is negative, so that a positive change in X_C means that its magnitude becomes smaller as the frequency increases. On the other hand, the reactance of the inductance increases by an amount $2\pi L$ for each additional cycle per second. At the series-resonant frequency of the series arm, the total reactance

$$X_L + X_C = 0$$

or

$$2\pi f_s L = \frac{1}{2\pi f_s C}$$

or

$$2\pi L = \frac{1}{2\pi f_s^2 C}$$

However, note that this last equation not only implies that the two reactances have equal magnitudes at the series-resonant frequency, f_s, but also, that f_s is the one frequency at which both reactances will change with frequency at the same rate. Therefore, for small changes in frequency near series resonance:

$$\Delta X_L = \Delta X_C$$

And since the total change in the reactance of the series arm is

$$\Delta X_s = \Delta X_L + \Delta X_C$$

then

$$\Delta X_s = 2\Delta X_L = 4\pi L \Delta f$$

If f_s is taken as the reference frequency, so that $\Delta f = f - f_s$, then, since $X_s = 0$ at resonance, the total reactance of the series arm, X_s, will be equal to ΔX_s. That is:

$$X_s = 4\pi L \Delta f \qquad 1\text{—}203\ (1)$$

Thus, for all frequencies near f_s, the equivalent circuit of a crystal unit may be represented as shown in figure 1-97, where X_{C_o} and R may be assumed

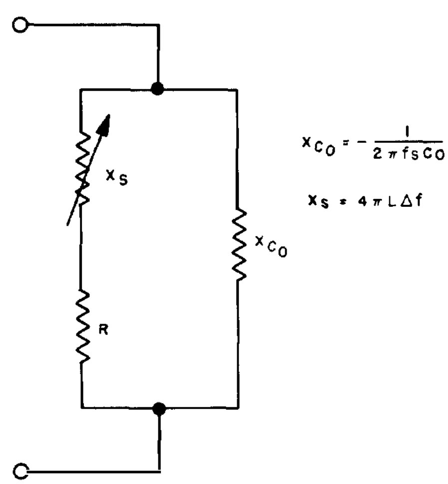
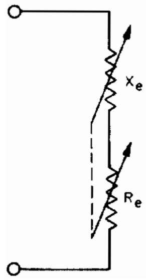


Figure 1-97. Impedance diagram of equivalent circuit of crystal unit



① $$X_e = \frac{X_{C_0}\left[R^2 + X_s^2 + X_{C_0} X_s\right]}{R^2 + \left[X_s + X_{C_0}\right]^2} \approx \frac{4\pi L \Delta f}{1 + \frac{4\pi L \Delta f}{X_{C_0}}}$$

② $$R_e = \frac{R X_{C_0}^2}{R^2 + \left[X_s + X_{C_0}\right]^2} \approx \frac{R}{\left[1 + \frac{4\pi L \Delta f}{X_{C_0}}\right]^2}$$

③ $$Z_e = \sqrt{R^2 + X_e^2}$$

④ $$\Delta f = f - f_s$$

Figure 1-98. Equivalent circuit of crystal unit when represented as an effective reactance in series with an effective resistance. The ganged arrows indicate that X_e and R_e will vary together with changes in the frequency, as indicated by the approximate formulas given as functions of Δf. (X_{C_0} is negative.)

to be constant, but with X_s a variable that changes linearly with Δf, and has the same sign as Δf.

1-204. The series-parallel circuit of figure 1-97 may be reduced to an equivalent circuit of X_e and R_e in series, as shown in figure 1-98. It should be remembered, however, that X_{C_0} is negative, whereas X_s is either negative or positive, according to the sign of Δf. The values of R_e and X_e, expressed as functions of Δf, are not exact, but are close approximations, well within the accuracy of the normal test procedure, except when the numerators reduce to zero. Note, however, that with $\Delta f = 0$, the approximate expressions equate X_e to 0, and R_e to R. This is equivalent to assuming that X_{C_0} is infinite by comparison with R, so that at series resonance of the motional arm the crystal unit as a whole behaves as a pure resistance equal to R. Although this is a close approximation, it is not exact. For X_e actually to be zero, the term $(R^2 + X_s^2 + X_{C_0} X_s)$ must be zero. There are two frequencies at which this will occur. One is called the resonant frequency of the crystal unit, f_r, and the other is called the parallel-resonant, or antiresonant frequency, f_a.

RESONANT FREQUENCY OF CRYSTAL UNIT

1-205. First, it should be remembered that f_r, the resonant frequency of the crystal unit, is almost, but not exactly, identical with f_s, the series-resonant frequency of the motional arm. If there were no shunt capacitance, C_0, then f_r would indeed be the same as f_s; but, as it is, C_0 introduces a reactive component to the current which must be cancelled by a reactive component of opposite phase through the motional arm, if the crystal unit is to appear as a pure resistance. These conditions are illustrated (not to scale) in the vector diagram of the currents through the two arms of the crystal unit, shown in figure 1-99. *Note that the frequency at which the crystal unit has the lowest impedance (maximum current) is f_s. Since $X_s = 0$ at this frequency, the equivalent circuit of figure 1-98, according to

* This sentence applies only to the relative impedances suggested by the current vectors in figure 1-99. It can be shown that the true minimum impedance of the crystal unit occurs at a frequency, f_m, that is as far below f_s as f_r is above f_s.

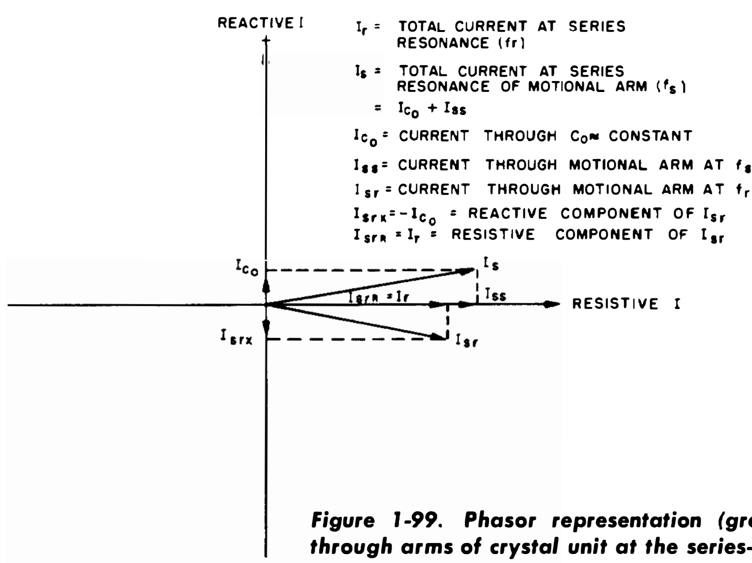


Figure 1-99. Phasor representation (greatly exaggerated) of current through arms of crystal unit at the series-resonant frequencies, f_r and f_s. Distributed inductance of the crystal leads is assumed to be negligible

equations 1 and 2, becomes

$$X_e = X_{Co}\left(\frac{R^2}{R^2 + X_{Co}^2}\right)$$

and
$$R_e = R\left(\frac{X_{Co}^2}{R^2 + X_{Co}^2}\right)$$

Except at the very high frequencies, X_{Co} is much larger than R, so that $R_e \approx R$, and X_e is so small that it may well be more than annulled by the distributed inductance of the external wiring. Even at frequencies in the neighborhood of 100 mc, X_{Co} will have a magnitude in the vicinity of 400 ohms, or approximately 10 times or more than that of R, so that R_e will equal R within ± 1 percent. The true frequency at which a "series-resonant" crystal circuit is intended to operate, however, is f_r, where all the reactive components of crystal current cancel. Actually, the term "series-resonance" is somewhat misleading, for the conditions of crystal resonance are those of a parallel, and not a series circuit. It should be understood that when we speak of series-mode circuits and oscillators, the operating frequency is generally assumed to be f_r.

1-206. By equation 1, figure 1-98, in order for X_e to be zero, the frequency must be such that:

$$R^2 = -(X_s^2 + X_{Co} X_s)$$

Since X_{Co} is negative, this equality can only exist when X_s is positive, i.e., X_s is inductive, and $f > f_s$. At frequencies very close to f_s, $X_{Co} >> X_s$, so that X_s^2 may be considered negligible. Thus, f_r will be the frequency at which

$$R^2 = -X_{Co}X_s = -4\pi L X_{Co} \Delta f_r$$

where $\Delta f_r = f_r - f_s$

Since X_{Co} is negative,

$$\Delta f_r = \frac{-R^2}{4\pi L X_{Co}} \qquad 1\text{—}206\ (1)$$

1-207. As a concrete example, assume that a partially plated A element, mounted in an HC-6/U holder according to RTMA recommendations, operates at resonance in its fundamental mode at a nominal frequency of 10 mc. Approximately, what value of Δf_r could be expected? Referring to paragraph 1-192, we find that A = 0.159 sq cm. On substitution in equation 1—192(1):

$$C_e = 2.42 \times 0.159 \times 10 = 3.85\ \mu\mu f$$

According to figure 1-95, $r_e = 250$. Thus, by equation 1—197 (1):

$$C = \frac{3.85}{250} = 1.54 \times 10^{-2}\ \mu\mu f$$

By equation 1—198 (1):

$$L = \frac{10^{14}}{4 \times 3.14^2 \times 10^{14} \times 1.54}$$
$$= 1.65 \times 10^{-2} \text{ henries}$$

According to paragraph 1-201:

$$C_o = 3.85 + 0.7 = 4.55\ \mu\mu f$$

So that

$$X_{Co} = \frac{-1}{2\pi f_r C_o} = \frac{-10^{12}}{6.28 \times 10^7 \times 4.55}$$
$$= -3.5 \times 10^3\ \Omega$$

From figure 1-96, a typical value of R is found to be 8 Ω. On substitution of the foregoing values of R, L, and X_{Co} in equation 1—206(1), we find that:

$$\Delta f_r = \frac{8^2 \times 10^{-1}}{4 \times 3.14 \times 1.65 \times 3.5}$$
$$= 0.088 \text{ cycle/sec.}$$

With such an extremely small difference between the two resonant frequencies of the crystal unit (less than 1 part of 10^8), for all practical purposes it can be assumed that $f_s = f_r$. Indeed, it would be academic to seek to distinguish between them. Remember, however, that the discussion has only concerned the equivalent circuit, in which the effects of the distributed inductance have been assumed to be reflected in a lower series-arm frequency, and a larger C_o. If the parameters in the example above are assumed to be the "true" values, so that the inductance of the leads must be represented separately, then a slightly more realistic interpretation will be possible. Assume, for instance, the $L_L = 10^{-8}$ henries. Then

$$X_{LL} = 2\pi L_L f_s = 0.628$$

In order for X_e to cancel this reactance, then, by equation 1 in figure 1—98:

$$-4\pi L \Delta f_r = 0.628$$

or

$$\Delta f_r = \frac{-0.628}{4\pi \times 1.65 \times 10^{-2}} = 3 \text{ cycles/sec}$$

The inductance of the external connections could easily increase this value of Δf_r ten-fold, so that for optimum frequency stability, an external series capacitance would be necessary. It is important to note the negligible effect that a small change in R or C_o will have on the frequency of a crystal unit at series resonance. In equation 1—206(1), as applied to the 10-mc crystal unit, even if R should triple in value, the frequency would not change by more than 1 part in 10^7. Although the power transferred through the crystal would be diminished, as would the Q, and hence, the effectiveness of the crystal as a frequency stabilizer, a reasonable increase in R will not, in itself, cause the frequency of a series-resonant crystal oscillator to drift.

ANTIRESONANT FREQUENCY OF CRYSTAL UNIT

1-208. Returning again to equation 1 of figure 1-98, it can be seen that the term $(R^2 + X_s^2 + X_{Co} X_s)$ can also be zero at some higher frequency than f_r, namely, when $X_s = X_{Co}$ (R^2 being negligible). This would represent the high-impedance, parallel-resonant state of the equivalent circuit in figure 1-97. Letting $\Delta f_a = f_a - f_s$ then, at f_a, the antiresonant frequency

$$X_{sa} = 4\pi L \Delta f_a \approx |X_{Co}|$$

so that

$$\Delta f_a = \frac{|X_{Co}|}{4\pi L} \qquad 1\text{—}208\ (1)$$

On substitution of the typical values of X_{Co} and L that were found for the 10-mc crystal unit:

$$\Delta f_a = \frac{3.5 \times 10^3}{4 \times 3.14 \times 1.65 \times 10^{-2}} = 16.9 \text{ kc/sec}$$

For a 10-mc crystal, this value of f_a represents a 0.169 percent frequency range in which the crystal may be used as a frequency-control device. At all frequencies within its range, except at f_r and f_a, the unit will appear to the external circuit as an inductive reactance, X_e, in series with a resistance, R_e. There is a very simple relation between the fractional frequency range, $\frac{\Delta f_a}{f}$, and the ratio of the capacitances, $r = \frac{C_o}{C}$, that can be derived from equation 1. Thus:

$$\Delta f_a = \frac{1}{4\pi L 2\pi f C_o}$$

so

$$\Delta f_a / f = \frac{1}{2\omega^2 LC_o}$$

where $\omega = 2\pi f$. Now,

$$\omega^2 = 1/LC$$

so, on substitution:

$$\Delta f_a / f = C/2C_o = 1/2\,r \qquad 1\text{—}208\ (2)$$

In the case of plated crystals, r is usually somewhat less than that predicted by theory. Where it should be slightly greater than the values of r_e in figure 1-95, since $C_o > C_e$, it is usually somewhat less. However, as a practical rule-of-thumb, it can be assumed that $r = r_e$, but only in those cases where $C_o \approx C_e$. The ratio of capacitances, r, is quite an important parameter of the crystal unit in its own right, not only as an indication of the maximum percentage width of the frequency band in which a particular crystal element can operate, but, as will be discussed later, as a measure of the electromechanical coupling, and also, because of its relation to the frequency stability.

IMPEDANCE CURVES OF CRYSTAL UNIT

1-209. Figure 1-100 shows the typical characteristics of the equivalent impedance circuit of figure 1-98, but with the frequency scale greatly expanded near the resonance point of the crystal. At frequencies sufficiently removed from resonance, both above and below f_r, where the motional impedance is large compared with X_{Co}, the X_e curve is essentially the same as the reactance curve of a capacitance equal to C_o. X_e is inductive only between its two zero points, f_r and f_a. Note that R_e

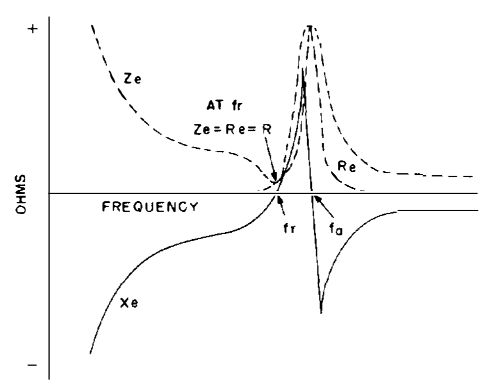


Figure 1-100. Impedance characteristics versus frequency of crystal unit. Neither the frequency nor the impedances are drawn to scale

rises sharply to a maximum at f_a, where it is equal to the parallel-resonant impedance of the equivalent circuit of figure 1-97. Since R is much smaller than X_{Co}, at antiresonance

$$R_e = Z_e \approx (X_{Co})^2/R$$

In the case of the particular 10-mc crystal unit where $X_{Co} = -3.5 \times 10^3 \Omega$, and $R = 8\Omega$, R_e at antiresonance will be approximately 1.5 megohms. $Z_e (= \sqrt{R_e^2 + X_e^2})$ at most frequencies is simply equal to X_e. Only in the immediate regions of f_r and f_a, where X_e becomes negligible, is the magnitude of Z_e affected greatly by R_e. The impedances, of course, are not drawn to scale. For example, if Z_e at antiresonance were drawn to the scale used for Z_e at resonance, the curve could extend more than a mile above the horizontal axis.

PARALLEL-RESONANT FREQUENCY, f_p, OF CRYSTAL CIRCUIT

1-210. Although an oscillator may depend upon a crystal operating at its series-resonant frequency, it is not practicable for a crystal unit to control an oscillator at the antiresonant frequency, f_a. The crystal will either be operated to pass a maximum current (series-resonant circuit), or to develop a maximum voltage (parallel-resonant circuit) at some proper phase and frequency. It would seem that these latter conditions could best be met by operating the crystal unit at its antiresonant frequency, for it is in this region that the effective impedance is most sensitive to small changes in the frequency. However, another circuit, such as the input of a vacuum tube, will necessarily be

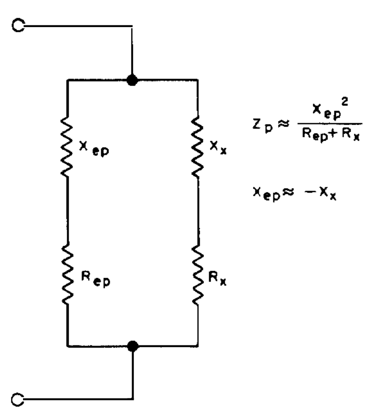


Figure 1-101. Equivalent parallel-resonant circuit of crystal unit (X_{ep}, R_{ep}) shunted by load (X_x, R_x). Normally $f_r < f_p < f_a$, so that X_{ep} is inductive and X_x is capacitive

shunted across the crystal. The shunt, or load circuit into which the crystal operates will have a much lower impedance than that of the crystal at antiresonance, so that the total impedance will be relatively insensitive to small frequency variations in the region of f_a. In determining the actual frequency stability, the entire circuit must be considered as a whole. The operating frequency may be considered as the resonant frequency, f_p, of an equivalent parallel circuit, as shown in figure 1-101. X_{ep} and R_{ep} are simply the reactance and resistance of the equivalent circuit of the crystal unit at f_p, and X_x and R_x are the equivalent shunt reactance and resistance, respectively. Since X_e is more frequency-sensitive above series resonance than below, there is normally no advantage in using a crystal in circuits that require X_e to be capacitive. Thus, in practice, f_p, will be some intermediate frequency between f_r and f_a, so that X_{ep} is always inductive and X_x is always capacitive. The distinction made between "parallel resonance" and "antiresonance" in this discussion is somewhat arbitrary, and it is not uncommon to use the term "antiresonant" to describe any parallel-resonant crystal unit.

Effects of Changes in Shunt Capacitance on f_p

1-211. In discussing Δf_r, the difference between the motional and the effective resonant frequency, it was found that

$$\Delta f_r = \left| \frac{R^2}{4\pi L X_{Co}} \right|$$

That this quantity is normally insignificant is fortunate, for it varies directly with the square of R, a parameter quite likely to change during operation. On the other hand, it was later found that

$$\Delta f_a = \left| \frac{X_{Co}}{4\pi L} \right|$$

could amount to more than 0.1 percent difference in frequency. In this case, since, Δf_a is relatively large, it is also quite fortunate that, *to a first approximation,* the antiresonant frequency of a given crystal unit is independent of operational changes in R. However, it is not the antiresonant frequency of the crystal unit itself, but rather, the actual parallel-resonant frequency at which the crystal unit will operate that is of primary interest. Let $f_p - f_s = \Delta f_p$. Now, it can be imagined that Δf_p is simply the Δf_a of a crystal unit whose shunt capacitance C_o has been increased by an

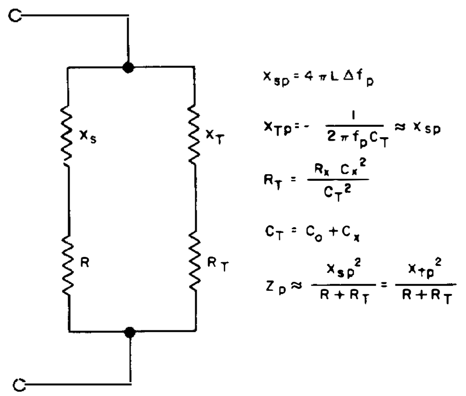


Figure 1-102. Equivalent parallel-resonant tank circuit, in which the motional impedance of the crystal unit is the inductive arm of the tank, and the total shunt impedance is the capacitive arm

amount C_x and which has an effective resistance added to the shunt arm equal to $R_x \left(\frac{C_x}{C_o + C_x}\right)^2$. This last assumption can be made without introducing an appreciable error as long as R_x is small compared with X_x. The multiplying factor is needed, since only a fraction, $\left(\frac{C_x}{C_o + C_x}\right)$, of the total equivalent tank current will flow through R_x. The equivalent tank circuit is shown in figure 1-102, where C_T and R_T are the values of the shunt parameters. Now, since Δf_p is equivalent to the Δf_a of a crystal unit that has $C_o = C_T$, Δf_p will be expressed by the same general formula that holds for Δf_a. Thus, $\Delta f_p = \left|\frac{X_{Tp}}{4\pi L}\right|$. Also, since $X_T = -1/2\pi f C_T$, it can be seen that Δf_p will be inversely proportional to the total shunt capacitance. Although C_o, itself, is the most stable of all the crystal parameters, the stability of the effective external capacitance C_x will depend upon the over-all design of the oscillator circuit. The crystal unit may be considered a device that determines the limits within which the frequency may be varied; that is, f_p must lie somewhere between f_r and f_a. However, it is primarily the parameters of the external circuit in conjunction with the equivalent L, C, and C_o of the crystal that fix the exact frequency; and although the stability of the crystal parameters is fundamentally a problem for the crystal manufacturer, the stability of the effective C_T is largely the concern of the radio designer.

1-212. Figure 1-103 (A) shows the reactance curve of X_s versus frequency, and figure 1-103 (B) shows the reactance curve of X_T versus C_T. The values of X_s are those of the 10-mc crystal unit which has previously been taken as an example, and where L is assumed to be 1.65×10^{-2} henry. Note that the variations in X_T with frequency have been neglected, and f is simply assumed to equal the nominal frequency of 10 mc, insofar as the capacitive arm is concerned. Since X_s and X_T are drawn to the same scale, a horizontal line drawn through

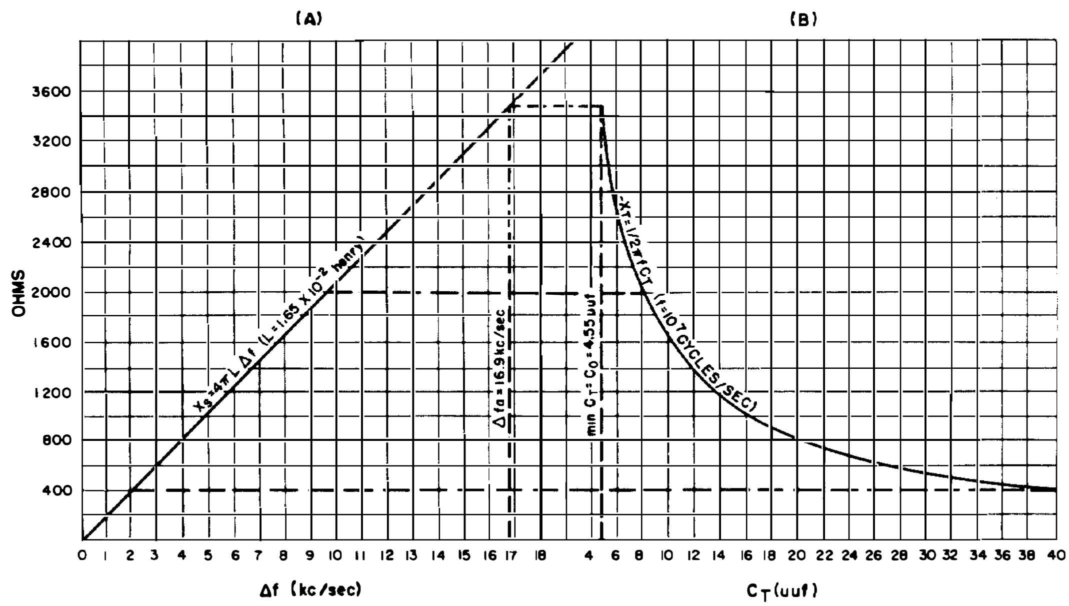


Figure 1-103. Reactance curves of: (A) X_s versus Δf, (B) $-X_t$ versus C_t

both curves will intersect points of equal but opposite reactances. These points of intersection will, in turn, indicate the value of Δf required for a given C_T, if the two arms are to be resonant. For example, at $C_T = 7.69\ \mu\mu f$, $X_T = -2000\ \Omega$; so that in order for X_s to be 2000 Ω, Δf_p must equal 9.65 kc. Likewise, a C_T of 40 $\mu\mu f$ will mean approximately a Δf_p of 2 kc. Now it so happens that the part of C_T represented by C_x will have a component that tends to vary with changes in the plate voltage applied to the vacuum tube, changes in the temperature or the tuning, changes in the coupling and neutralizing adjustments, and any changes in the vacuum-tube characteristics or other circuit parameters due to other causes. Such a change in X_x will cause not only a change in the resonant frequency, but also a change in the amplitude of the oscillations. If a given change in C_x is to have a minimum effect upon the frequency and power expenditure of the oscillator, then C_T must be as large as possible without seriously reducing the stabilizing effect of the crystal. In other words, C_T should have a value where the slope of the X_T-vs-C_T curve is not steep. For the 10-mc crystal of figure 1-103, maximum stability would be obtained with C_T between 36 and 40 $\mu\mu f$. With C_o equal to 4.55 $\mu\mu f$, this would mean a load capacitance, C_x, between 32 and 36 $\mu\mu f$. As much of C_x as is possible should be supplied by a fixed or adjustable capacitor connected directly across the crystal unit or in some other part of the circuit, so that its effective capacitance with respect to the crystal terminals will remain constant, and not be affected by changes in the tube characteristics. This would reduce the variable part of C_x to a minimum. C_T, however, should not be made so large that X_T will approach the magnitude of R, otherwise the crystal will not only lose some of its stabilizing effectiveness, but will require an excessive drive level to maintain oscillations.

Stabilizing Effect of Crystal on f_p

1-213. Although C_T plays an important role in the final determination of the frequency, it is the crystal itself that must be primarily responsible for the stability of the frequency—that is, if the use of a crystal is to be justified. For this reason, care should be taken to make certain that the apparent Q_s of the crystal series arm $\left(\frac{X_s}{R}\right)$ is as large as 10, if possible, and preferably mucn larger during operation. Otherwise, the series-arm impedance will not respond with maximum sensitivity to changes in C_T. However, since X_T, and hence X_s, must be kept small to reduce the effects of a change in C_T, it might appear at first thought that a conventional coil could serve quite as well as a crystal. The reason why this is not true is that the frequency stability is dependent upon the magnitude of the change in reactance for a given change in frequency, and not primarily upon the total magnitude of the reactance. It will be recalled that, in the conventional L-C circuit, the instantaneous rate of change of X_L with frequency is

$$\frac{dX_L}{df} = 2\pi L$$

and that, at resonance

$$\frac{dX_L}{df} = \frac{dX_C}{df}$$

In the parallel-resonant crystal circuit, however, these equalities do not hold, for $X_s = 4\pi\Delta fL$, and not $2\pi fL$. Thus, $\frac{dX_s}{df} = 4\pi L$, where L of the crystal is greater than L_c of a coil of the same reactance by a factor of $f/2\Delta f$. Since at resonance, the rate of change of X_T would equal that of X_{Lc}, it follows that $4\pi L$, the change in the motional reactance with frequency, will be $f_p/\Delta f_p$ times as great as the change in X_T with frequency. Consequently, the stabilizing effect of the crystal is much greater than that of the shunt reactance, so that, for all practical purposes, the crystal can "automatically" annul the effect of small changes in C_T, but not vice versa. It can be seen that, even with X_s relatively small, the stabilizing effect of the crystal for a fixed change in X_T is not diminished, provided, of course, that X_s is sufficiently large, as compared with R, so that the total impedance of the series arm is essentially equal to, and varies linearly with, X_s. (See paragraphs 1-238 to 1-245.)

Effect on Parallel Crystal Circuit Due to Variations in Resistance

1-214. As long as the apparent Q of the parallel-resonant circuit $\left(\frac{X_s}{R + R_T}\right)$ is at least as great as 10, a change in either R or R_T will not, in itself, have a large effect upon f_p. However, depending upon the design of the particular circuit, a change in the resistance may indirectly affect the frequency by causing a change in C_T, since, to a certain extent, the effective C_T will be a function of the other circuit parameters. The most critical effect due to changes in the resistance parameters is the effect on the power required for excitation of the oscillator in order to obtain a given output. The impedance, Z_P, of the parallel circuit at resonance will be approximately $\frac{X_{sp}^2}{R + R_T}$. An increase in the total resistance of 100 percent would thus decrease Z_P by one half. If, for example, the output of the oscillator depended directly upon the r-f voltage across Z_P (i.e., across the crystal), a decrease in Z_P by one half would require twice as much power in the crystal circuit to maintain the output at the same level as before. A part of R_T will be the result of reflected resistance losses in the output circuit. An increase in the load will thus be reflected as an increase in R_T. This is unfortunate, for if the load should increase it would be desirable to have an increase in Z_P, to raise the excitation voltage automatically, or at least to keep it constant. As it is, the effect is to decrease the excitation, unless special circuits, such as the Tri-Tet, are employed to increase the feedback directly. If a principal component of the losses in R_T are due to the losses in the grid circuit, and if the oscillator design is such that the grid current is not linear with the excitation voltage, but rises at a much greater rate, then R_T can rapidly increase or decrease with the excitation voltage, and Z_P will vary inversely. Under these conditions, Z_P will always change in a direction that will tend to annul any change in the excitation voltage. The greater that part of R_T reflecting the grid losses, as compared with that part reflecting the output losses, the greater will be the amplitude stabilizing effect for counteracting changes in the plate voltage or the effective load resistances. Another characteristic of a crystal circuit in which R_T varies automatically is that the effect resulting from a variation in R is minimized. Assume, for example, that the desired output at a constant load will require a certain effective value of Z_P. If, for some reason, R should change, thereby changing the excitation voltage, R_T would tend to change by an equivalent amount in the opposite direction, thus maintaining Z_P, and hence the output, essentially constant. However, a change in R or R_T will almost certainly be accompanied by a change in the crystal power losses, thereby causing a frequency drift if the particular crystal unit is frequency-sensitive to the drive level. At this point, however, the important items to note are: (1) X_{Tp} and hence X_{sp} preferably should not be smaller than $10(R+R_T)$, or the maximum stabilizing effect of the crystal will not be realized; (2) the direct effect of a change in $(R + R_T)$ is to change Z_P; (3) the effects of a change in Z_P primarily will involve changes in the excitation voltage, in the power expended in the crystal circuit, as well as that delivered to the load, and in the equivalent value of C_T, thereby also changing the frequency; (4) if changes in either R or R_T are such that the power expended in the crystal unit itself is caused to vary, then a significant change in the frequency characteristics of the crystal may result; and (5) for maximum frequency stability, the oscillator should be lightly loaded, and the drive level of the crystal should be as small as is practicable.

Effect on Parallel Crystal Circuits Due to Variations in Motional-Arm C or L

1-215. Crystal circuits operated at the resonant frequency of the crystal units may be only slightly affected by variations in C or L from one crystal unit to the next, or even during the operation of a particular unit, provided that the effective LC product remains constant, so that the frequency does not change. In the parallel-resonant circuit, however, even if f_s is the same, a different C and L means a change in Δf_p. For a given C_T and nominal frequency, X_{Tp}, and hence, X_{sp}, must remain approximately constant, so that Δf_p, equal to $X_{sp}/4\pi L$, will tend to vary inversely with L. The exact value of L for a given crystal unit will depend upon the effective electrode area, the orientation of the cut, the thickness of the crystal, whether twinning is present in the quartz, and the degree to which spurious modes are coupled to the desired mode. Insofar, as the variations in L from one crystal unit to the next are concerned, no problem arises unless it is necessary to adjust f_p to an exact value; in which case the problem of the design engineer is to ensure that C_T will be sufficiently adjustable so that the desired f_p may be obtained with any reasonable value of L. Since such adjustments must be provided for anyway, in order to allow for different values of f_s, no new problems are introduced, except that a greater deviation in Δf_p must be met than otherwise. Unless spurious modes are closely coupled to the desired mode, the variations in L that might occur during the operation of a particular crystal unit will be too small to affect the magnitude of Δf_p, as long as X_{Tp} remains constant. However, the operational variations of L and/or C may be such that f_s will change, in which case f_p will also change. Such a deviation in frequency, i.e., in the equivalent LC product, would occur during changes in temperature or drive level, or because of fatigue or other aging effects. Minimum variations in L and C are obtained by the use of low temperature-coefficient crystals and constant-temperature ovens, and by ensuring that the drive level will remain both low and constant. In any event, a reasonable operational variation in f_p can be compensated for by an adjustment in C_T.

Minimum Value of Δf_p

1-216. Returning to equation 1 in figure 1-98, let it be imagined that X_e represents the effective reactance of the motional arm of a crystal unit in parallel with a total capacitance C_T, instead of simply the C_o of the crystal unit, itself. Furthermore, assume that R_T is negligible. As before, the condition of resonance is that X_e be zero, which will occur only when

$$R^2 + X_s^2 + X_s X_T = 0$$

(Note that X_T now replaces X_{Co}.) Now $X_s = 4\pi L \Delta f$, and on substitution in the preceding equation and rearranging, it is found that

$$(\Delta f)^2 + \frac{X_T}{4\pi L}\Delta f + \frac{R^2}{16\pi^2 L^2} = 0$$

Note that this is simply a quadratic equation of the type $AX^2 + BX + C = 0$, so by the quadratic formula

$$\Delta f = \frac{-\frac{X_T}{4\pi L} \pm \sqrt{\frac{X_T^2 - 4R^2}{16\pi^2 L^2}}}{2}$$

The $\pm$ term indicates that there are two possible solutions for Δf at which resonance will occur. One of these is equivalent to Δf_r (but with C_o replaced by C_T), and the other is equivalent to Δf_p. For these solutions of Δf to be real, X_T^2 must be greater than $4R^2$; otherwise, the expression under the radical sign becomes negative, and Δf will be imaginary. However, in the special case where $X_T^2 - 4R^2 = 0$, there is only one solution for Δf. In other words,

$$\Delta f_r = \Delta f_p = \frac{-X_T}{8\pi L}$$

This represents the minimum value obtainable for Δf_p; or, from the point of view of series resonance, it may be considered the maximum value obtainable for Δf_r. The important point to note is that neither parallel nor series resonance is possible unless X_T^2 is equal to, or greater than, $4R^2$. At the minimum Δf_p,

$$X_T^2 = 4R^2$$

or

$$|X_T/R| = X_s/R = 2$$

It should be remembered that all the resistance has been assumed to be in the motional arm, and so has the effect of limiting the maximum component of lagging current for a given voltage; but parallel resonance could be achieved at any frequency between f_s and f_a if R_T were equal to R. However, since C_o limits the minimum amplitude of leading current, R_T cannot be made equal to R for all values of f_p, and there will still be a minimum f_p greater than f_s. (A minimum which can be shown to be identical with the natural f_r of the crystal unit.) With R_T assumed to be negligible, the ratio of reactance to resistance equal to 2 represents the minimum apparent Q_s of the parallel crystal circuit, if resonance is to be obtained. As stated previously, if the full stabilizing properties of the crystal are to be in use, Q_s should be at least 10. However, if the power delivered to the crystal circuit is sufficient, oscillations can be maintained as long as the apparent Q_s does not fall below 2. This occurs at the frequency at which the amplitude of the lagging component of current through the series arm is the maximum obtainable.

TYPICAL OPERATING CHARACTERISTICS OF CRYSTAL UNIT

1-217. Figure 1-104 shows the effective impedance characteristics of the 10-mc crystal unit which has been assumed to have the following parameters:

$L = 1.65 \times 10^{-2}$ henry
$C = 1.54 \times 10^{-2}$ $\mu\mu$f
$R = 8$ ohms
$C_o = 4.55$ $\mu\mu$f

X_e, R_e, and Z_e are given by equations 1, 2, and 3, respectively, in figure 1-98; X_{Co} is assumed to be equal to -3.5×10^3 ohms for all values of Δf. Note that the normal operating range covers only about one fourth of the total range between f_r and f_a. Of course, if C_o were greater than the value assumed, Δf_a would be smaller and the normal operating range would be a larger percentage of the total. As explained previously, C_T ($= C_o + C_x$) must be relatively large, so that small variations in C_x will not greatly affect the frequency, and it is this consideration that limits the practical operating range to low values of Δf. At parallel resonance, X_e must approximately equal $-X_x$, for the same reason that X_s must equal $-X_T$. Standard military high-frequency crystal units are normally tested with a value of $C_x = 32$ $\mu\mu$f. At 10 mc, a capacitance of 32 $\mu\mu$f will have a reactance of approximately -500Ω, as indicated in figure 1-103(B). With $C_o = 4.55$ $\mu\mu$f, C_T will be 36.55 $\mu\mu$f, which corresponds to a value of $X_T = -440$ Ω, and a $\Delta f = 2.1$ kc/sec. Δf can also be found from the reactance curve of figure 1-104 at the point where

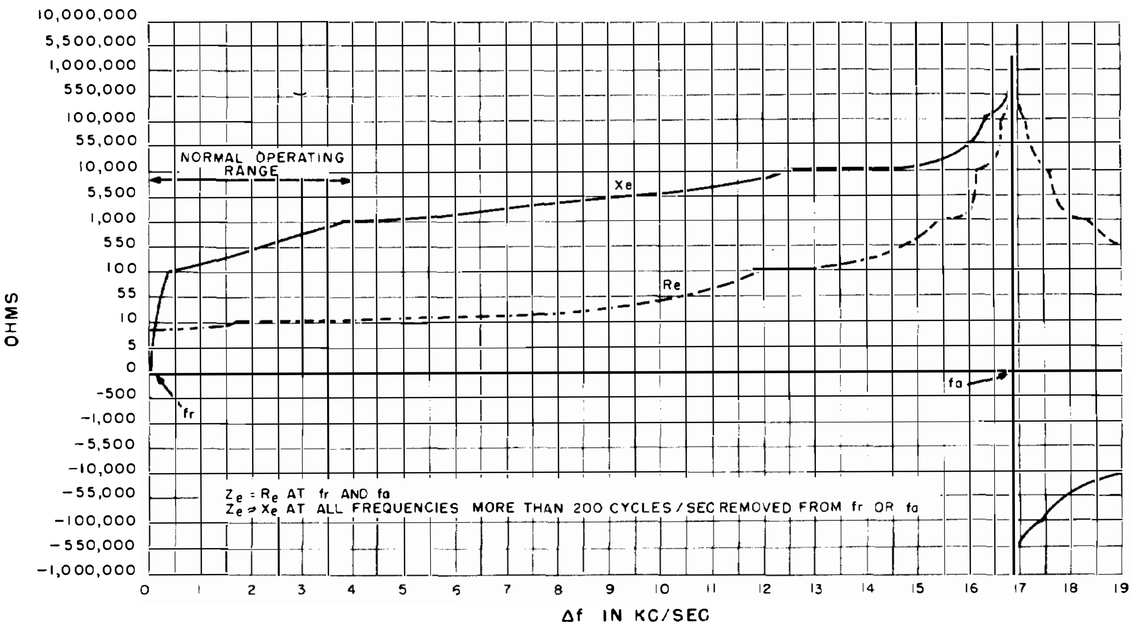


Figure 1-104. Typical characteristic curves for X_e and R_e of 10-mc crystal unit. (Shunt resistance, R_I, across crystal is assumed to be negligible)

$X_e = -X_x = 500\ \Omega$. Crystal units are sometimes operated in series with an external capacitor, C_x, as indicated in figure 1-105. Slight variations in the frequency can be compensated for by adjustments of C_x, and resonance will occur at the frequency at which X_x is exactly annulled by X_e. If the ratio of X_e/R_e is sufficiently large, then for all practical purposes the series-resonant frequency, f_{rx}, is the same as the f_p of the crystal unit in parallel with the same C_x. At resonance, the crystal unit and C_x in series have an effective impedance equal to R_e. Although there is an effective maximum $Q_{em} = \frac{X_e}{R_e}$ when $\Delta f = \frac{\Delta f_a}{2}$, it has no special significance directly concerning the frequency stability of the circuit, but does tend to increase the activity. In general, if a series capacitor is used, its reactance will be small, as compared with X_{C_0}. Indeed, it may be used for no other purpose than to annul the self inductance of the crystal leads. It can be seen in figure 1-104 that R_e does not increase nearly as rapidly as does X_e, except in the region of f_a. With $C_x = 32\ \mu\mu f$, Δf_{rx} ($= f_{rx} - f_s$) will be 2.1 kc for the crystal unit of figure 1-104, and R_e will be between 11 and 12 ohms.

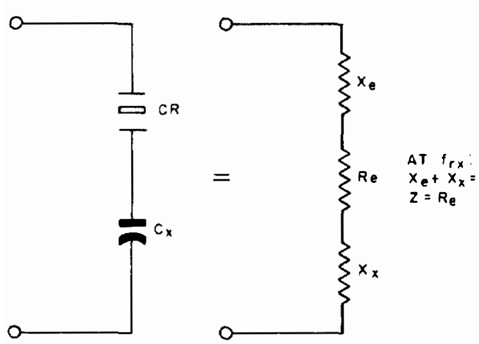


Figure 1-105. Equivalent circuit of crystal unit connected in series with capacitor

MEASUREMENT OF CRYSTAL PARAMETERS

1-218. The parameters L, C, R, and C_0 of any crystal unit chosen at random are effectively four independent variables, so that a minimum of four measurements are required to determine the values of these variables. Probably the four easiest measurements to make are those for f_r, R, C_0, and f_{rx}. The measurement for the last quantity is made when a known load capacitance C_x is connected in series with the crystal unit. Since f_{rx}, the resonant frequency of the crystal and C_x in series, for all practical purposes will be equal to the f_p of the crystal in parallel with C_x, we shall normally not make a distinction between the two frequencies in the following discussion, but shall use the symbol "f_p" in referring to either.

Measurement of the Shunt Capacitance, C_0

1-219. At all frequencies sufficiently removed from resonance the crystal unit will have the characteristics of a capacitance equal to C_0. Thus, at these off-resonance frequencies, C_0 can be measured by a conventional Q meter or an r-f bridge. The frequency at which C_0 is to be measured should be lower than, but reasonably close to, the operating frequency, particularly if the crystal unit is to be operated at a high harmonic mode in the v-h-f range. Otherwise, the effect of the distributed inductance of the leads will not be properly taken into account.

Measurement of the Series-Arm Resistance, R

1-220. R is normally measured with the aid of a CI meter (crystal impedance meter). (See also paragraphs 2-60 through 2-65.) There are four standard CI meters with which the crystal units described in Section II of this handbook have been tested:

Crystal Impedance Meter	Frequency Range (kc/sec)
TS-710/TSM	10 to 1100
*TS-537/TSM	75 to 1100
TS-330/TSM	1000 to 15,000
TS-683/TSM	10,000 to 75,000

* (Crystal Impedance Meter TS-537/TSM may soon be replaced entirely by the recently developed Crystal Impedance Meter TS-710( )/TSM.) A CI meter is essentially an r-f oscillator provided with a feedback circuit in which a crystal unit or a resistor, or a crystal unit in series with a calibrated capacitor, can be connected. A simplified schematic diagram of a typical CI meter is shown in figure 1-106. The circuit shown is a modified Colpitts oscillator in which the tank inductor has been effectively divided into two equal sections, L_1 and L_2, between which a resistor, R_o, equal to R_e, the effective resistance of the crystal unit, can be connected. The ganged tuning capacitors C_1 and C_2 are at all times equal. C_B is simply a blocking capacitor to isolate the plate voltage from the crystal terminals, and adds only a negligible reactance to the tuned circuit. The potentiometer, P, is used to control the screen grid voltage, and hence the r-f output of the tube and the drive level of the crystal. With S_1 connected as shown, and R_o adjusted to a value typical of the motional-arm R for the type of crystal unit being measured, the circuit will oscillate at approximately the resonant frequency of the tank. If C_1 and C_2 are adjusted so that the natural frequency of the oscillator is near the nominal frequency of the crystal, then, on connecting the crystal into the circuit, oscillations will continue, but with the frequency determined

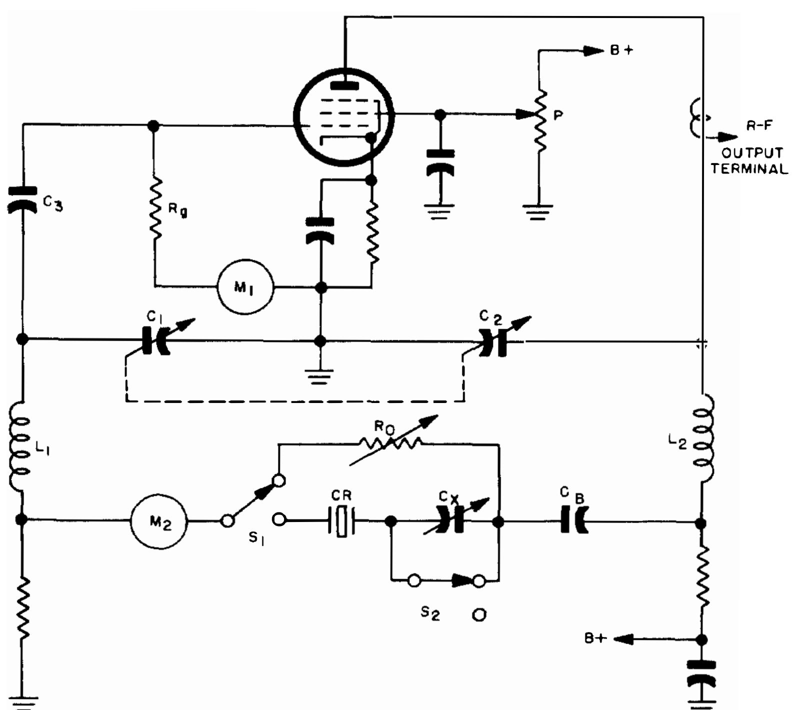


Figure 1-106. Simplified schematic diagram of CI meter

by the total reactance of the tank, including X_e of the crystal unit. There is no standard practice as to grounding the crystal holder; but whether grounded or ungrounded, the method of connecting the crystal unit should be noted. The drive is adjusted so that a very small grid current is indicated on M_1. Under these conditions the control grid is positive with respect to the cathode only at the peaks of the positive swings of the excitation voltage developed across C_1. Since electrons flow from cathode to grid only at these instants, a small percentage change in the excitation voltage, as illustrated in figure 1-107 can cause a very large percentage change in the cathode-to-grid electron flow. In an actual circuit the idealized constant bias that is indicated in figure 1-107 does not occur because of the gridleak action. However, if the gridleak contribution to the bias is very small compared with that part developed across the cathode resistance, the increase in bias due to an increase in excitation occurs almost entirely across R_g, not across the cathode resistance. Hence, a five or ten per cent increase in the total bias can result from hundred per cent increase in the gridleak IR drop. In this way, the grid current meter is a very sensitive indicator of slight changes in the r-f voltage across C_1, and hence of any change in the tank current. With S_1 in the crystal position and S_2 closed, as C_1 and C_2 are varied, a peak in the grid-current reading indicates a maximum current through C_1. This in turn means that the effective resistance of the crystal unit has reached the minimum value equal to the series-arm R. In other words, the oscillator frequency is coinciding with the resonant frequency, f_r, of the crystal unit. R_o can now replace the crystal in the circuit and be adjusted to give the same meter readings at the same frequency. At this point R_o will equal R, and, since R_o is known, R will have been measured. The crystal current meter, M_2, is not sufficiently sensitive to permit an accurate observation of the small changes in tank current that occur as the circuit is tuned through f_r. The purpose of the meter is to decrease the possibility of overloading the crystal and to provide a ready means for determining the exact drive level at which the crystal is being tested. Since the crystal parameters may change with the drive, it is necessary to specify the drive level at which the measurements are made. Expressed in milliwatts, the drive level equals $I^2R \times 10^{-3}$, where I (in milliamperes) is the current through M_2 at f_r. If a crystal-current meter is not supplied, two vacuum-tube voltmeters can be used to measure the voltage from each crystal terminal to ground without seriously affecting the circuit. Where E is the difference in potential across the crystal, equal to the difference between the two terminal voltages, the drive level is equal to E^2/R. The temperature at which the measurements are made should also

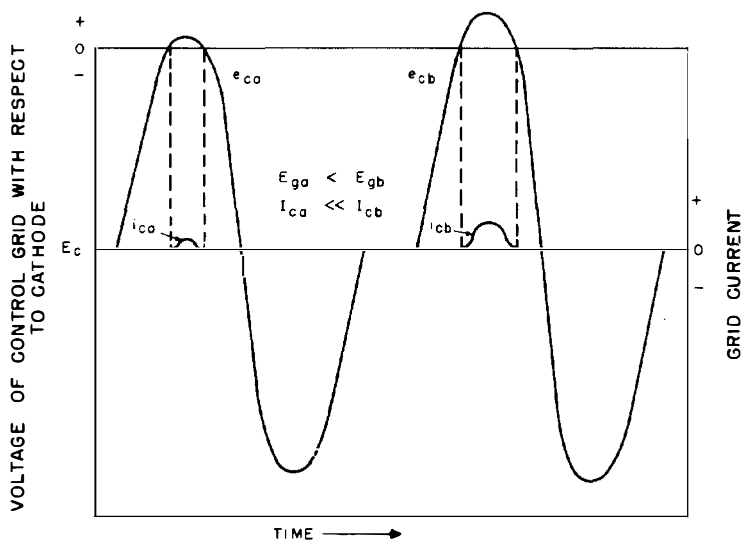


Figure 1-107. How a small percentage change in excitation voltage can cause a large percentage change in grid-leak current of CI meter. Bias does not actually remain constant as indicated, but follows the percentage changes of excitation voltage. Nevertheless, the relative percentage variations of grid current and excitation voltage can be approximately as shown when the greater part of the bias is developed across the cathode resistance

be specified, although, in general, the variation of R with ambient temperature is much less than its variation with the amplitude of the crystal vibrations.

Measurement of the Resonance Frequency, f_r

1-221. To measure f_r, a c-w radio receiver, a radio-frequency standard, a calibrated audio-frequency source (interpolation oscillator), and either a loudspeaker, a pair of ear phones, or an oscilloscope are used in conjunction with the CI meter. With the crystal connected in the CI-meter circuit and the oscillator tuned to series resonance, the r-f output can be loosely coupled through a coaxial cable to the antenna post of the c-w radio receiver. After the receiver is tuned to the frequency of the crystal unit, the CI meter is turned off, and the receiver is connected and tuned to receive the particular harmonic of the frequency standard that is nearest to f_r. The bfo of the receiver is then cut off, and the CI meter is turned on. With both the standard and the CI-meter signals being fed to the receiver input, the output of the receiver will be an audio beat note equal to the difference between the known standard frequency and the unknown crystal frequency. By momentarily switching a fairly large value of C_x in series with the crystal, so that the CI-meter frequency increases slightly, the audio beat frequency will rise or fall according to whether f_r is respectively greater than or less than the standard signal. The audio beat frequency is next mixed with the audio output of the interpolation oscillator, which in turn is adjusted to bring the beat frequency of the two audio signals to zero—the zero beat being observed by phones, loudspeaker, or oscilloscope. At zero beat, the crystal frequency will have been measured to be equal to the selected r-f standard frequency $\pm$ the interpolation oscillator frequency. The accuracy of the measurement depends primarily upon the accuracy of the frequency standard, and secondarily on that of the interpolation oscillator. As in the case of the resistance measurement, both the temperature and the drive level should be specified, and these should be the same as when the measurement of R was made.

Measurement of the Parallel-Resonance Frequency, f_p

1-222. To measure f_p, it is first necessary to adjust the CI meter to oscillate at f_r, by the same steps employed previously. With the oscillator so adjusted, a known value of C_x is switched in series with the crystal. The new frequency will be approximately equal to f_p. The more common values of C_x are 32 $\mu\mu$f for high-frequency crystals, and 20 $\mu\mu$f for low-frequency crystals. In testing to determine whether the crystal frequency is above or below the frequency of the test standard, it may be more convenient to add capacitance across the crystal unit than to change the setting of C_x. In this event the CI-meter frequency is decreased rather than increased, so that the effect on the beat note will be the opposite of that previously described. Simply touching the crystal holder with the hand is normally the quickest method of increasing the shunt capacitance; however, care should be taken that the method employed does not effectively decrease the capacitance by grounding the holder.

1-223. Theoretically, the foregoing method of measuring f_p is not exact, for if the LC circuit is correctly tuned when the crystal appears as a pure resistance, the same feed-back phase relations cannot hold at the higher frequency, f_p, unless CR and C_x in series introduce a negative reactance to compensate for the increase in X_{L1} and X_{L2} and the decrease in X_{C1} and X_{C2}. In other words, X_e of the crystal unit is approximately, but not exactly, equal in magnitude to X_x of the load capacitance. Actually, X_e is less than $|X_x|$ by an amount approximately equal to the change in reactance around the LC loop, exclusive of CR and C_x. This change is approximately equal to $4\pi(L_1 + L_2)\,\Delta f_p$, where Δf_p is the difference between f_p and f_r. At the true f_{rx}, $X_e + X_x = 0$; at the observed f_p,

$$X_e + X_x + 4\pi\,(L_1 + L_2)\,\Delta f_p = 0 \qquad 1\text{—}223\ (1)$$

Now a small change in $X_e + X_x$, equal to $\Delta X_e + \Delta X_x$, as a result of a small change in frequency, is practically equal to ΔX_e alone. If the frequency is sufficiently close to the resonant point, f_r, we may set $X_e + X_x$ (at observed f_p) $= \Delta(X_e + X_x) \approx \Delta X_e \approx \Delta X_s = 4\pi L\Delta f$, where $\Delta f =$ observed f_p $-$ true f_{rx}. By substitution in equation (1)

$$\text{True } f_{rx} \approx \text{Observed } f_p + \left(\frac{L_1 + L_2}{L}\right)\Delta f_p \qquad 1\text{—}223\ (2)$$

Measurement of the Effective Resistance, R_e, at Parallel Resonance

1-224. In the measurement of f_p, the drive level and the temperature should be the same as in the measurement of R and f_r. To determine the drive level, either the voltage across the crystal unit, or the effective resistance R_e should be known—or both, if a crystal-current meter is not provided. A measurement of R_e is also important for its own sake and as a check to see whether the motional-arm parameters are the same at f_p as at f_r. In the case of a crystal unit which is intended to be operated only at parallel resonance, R_e is generally treated as a primary parameter of more immediate importance than the motional-arm R. R_e is measured in a manner similar to the measurement of R, except that on substituting R_o for the crystal the circuit must be retuned so that oscillations are being maintained at f_p. For a very precise drive-level measurement, additional precautions must be taken if the power dissipation is to be the same in both the series- and parallel-resonant measurements. The best assurance that the f_r and f_p drive levels will not be greatly different is to be had when the crystal current is kept near the minimum necessary to maintain oscillations. Thus, even though the relative differences in drive level may be large, the absolute differences will be small. This is not a completely reliable method, for some crystal units exhibit very sharp increases in resistance when the drive level approaches a minimum.

Computing the Series-Arm C and L from the Measured Parameters

1-225. From the formulas for f_r, X_e, and X_x, it is quite easy to derive the following approximate equations for the series-arm parameters, C and L:

$$C = \frac{2(C_o + C_x)\Delta f_p}{f_r} \qquad 1\text{—}225\ (1)$$

$$L = \frac{1}{(2\pi f_r)^2 C} \qquad 1\text{—}225\ (2)$$

where, C, C_o, and C_x are in farads, L is in henries, and Δf_p and f_r are in cps.

METHODS FOR EXPRESSING THE RELATIVE PERFORMANCE CHARACTERISTICS OF A CRYSTAL UNIT

1-226. If the four equivalent electrical parameters (L, C, R, C_o) are accurately known for a given state of operation, no other independent data concerning a crystal unit can increase the radio engineer's knowledge of how the crystal will perform under the given conditions. However, the radio engineer has been slow in requesting specific information concerning the electrical characteristics of the crystal units available, and as a result the problem of making a given circuit perform correctly has often in the past effectively become the responsibility of the crystal manufacturer, who, by cut-and-try methods, has been more or less required to design the crystal unit around the particular circuit. Fortunately, progress toward greater standardization of crystal units has been considerably accelerated during recent years because of the increased demands of the military services; but there is still a tendency on the part of the design engineer to regard a crystal unit, as one production engineer has expressed it, as a "mystery box," rather than the equivalent circuit that it is. Contributing to this tendency has been a hesitancy upon the part of the manufacturer to describe his crystal units in terms of the most probable equivalent electrical parameters. At the present state of the art, wide variations from the most probable values can occur, and the manufacturer quite naturally wishes to avoid the chance that typical values of the parameters will be misinterpreted as specified values. For similar reasons, a description of a crystal unit in terms of its most probable parameters is not at present desirable from the point of view of the military services, lest a crystal circuit be designed upon the assumption that the typical crystal parameters will always be available, rather than upon the assumption that the crystal unit cannot be depended upon to meet other than its minimum performance specifications. If the former, rather than the latter assumption were made, a carefully designed circuit might fail to operate properly if used with a borderline crystal unit. Thus, the purpose of the standardization of types—to ensure a complete interchangeability among the crystal units of the same type number and nominal frequency — would be defeated. Nevertheless, the lack of emphasis upon the basic parameters has served to cloak the crystal in an air of mystery, and to instill in the radio engineer an impression that a crystal circuit is possessed of properties that cannot be expressed in the normal idiom of LCR networks. Contributing somewhat to this point of view is the special terminology that has been developed for the purpose of comparing the performance characteristics of one crystal unit with those of another particularly where the definitions of the terms contain certain ambiguities or conditional interpretations, or are presented as mathematical relationships without concrete qualitative meanings. What may be implied as a property of the crystal unit alone, may well be a function of the particular circuit in which the crystal unit is mounted. Much of the difficulty can be avoided if it is kept in mind that a crystal unit has no important circuit performance qualities that cannot be expressed in the everyday terminology of radio engineering as it might apply if the crystal unit were replaced by an equivalent network of L, C, R, and C_o.

1-227. There are five general categories in which crystal units can be placed for comparison insofar as their relative merits are reflected by their performance in a standard test-oscillator circuit: (1) activity, (2) frequency stabilization, (3) bandwidth, (4) quality factor, and (5) parameter stability. *Activity,* as applied to a crystal, is a general term, rather loosely defined, that refers to the relative ease with which a crystal may be caused to maintain oscillations. The basic parameter most closely associated with the crystal activity is the motional-arm resistance, R. Besides R, or R_e, there are certain performance parameters that can be used as indices of relative activity quality. These are the effective Q (Q_e), the maximum effective Q (Q_{em}), the figure of merit (M), and the performance index (PI). The term, *frequency stabilization,* as used in this context, refers only to the ability of a crystal to minimize any change in the frequency due to variations in the parameters of the external circuit. In this sense, those performance parameters that can be used as indices of the frequency-stabilization quality are the series-arm L/C ratio, the coefficient of frequency stability (F_X), and the capacitance ratios C_T/C and C_T/C_X. The *bandwidth* of a crystal unit refers to the frequency range over which the crystal unit is considered operable. The performance parameters indicating this quality are the capacitance ratio, $r = C_0/C$, and the electromechanical coupling factor, k. The *quality factor* is simply the crystal Q, which is, itself, a major performance parameter, but one that is not exclusively identified with any one of the other four performance categories. The term *parameter stability* is used here to refer to the relative stability of the crystal parameters during changes in the temperature, drive level, tuning adjustments, and the like. The *frequency stability* of the crystal unit, which is included in this category, should not be confused with the function of *frequency stabilization* which is the characteristic we have arbitrarily assigned to the second performance category. The frequency stabilization is dependent upon the *magnitudes* of the equivalent-circuit parameters; whereas, the frequency stability is dependent upon the *stability* of the equivalent-circuit parameters. The stability of a crystal oscillator circuit is dependent upon *both* the crystal stabilization and the parameter stability. Performance indices or terms indicating the relative parameter stabilities are represented by the temperature coefficients of frequency and resistance, drive-level coefficients of frequency and resistance, frequency tolerance, frequency deviation, resistance deviation, relative freedom from unwanted modes, and general expressions indicating durability and aging characteristics. Since most of the characteristics identified with the five performance categories can be expressed as functions of the same basic equivalent-circuit parameters, a performance parameter in one category quite often serves as an indication of the crystal quality in another. It cannot be said that those properties most closely identified with the activity, for instance, are not also related to the frequency stabilizing effect. Nevertheless, classifying the various methods for rating the performance of a crystal unit is helpful in interpreting the different performance parameters in terms of the basic equivalent-circuit parameters.

Activity Quality of Crystal Unit

1-228. The "activity" of a crystal oscillator is a qualitative expression referring to the amplitude of the oscillations. It is a term that came into use during the early days of crystal resonators, but one that seems never to have been vigorously defined. For example, it is not always certain whether the "activity of an oscillator" is intended to refer to the amplitude of current in the feedback, or in the output circuit, or to the voltage across some particular circuit component, or to the output power, or to the excitation power, or to the ratio of these powers, or simply to the amplitude of the crystal's mechanical vibrations. Were the expression not already so strongly entrenched in the crystal terminology, its use would probably be discouraged. As it is, crystal units are commonly described as having high or low activities, or more specifically, as having high or low potential activities or activity qualities. It will be found that the crystal parameter most directly indicative of the activity quality is the motional-arm conductance, 1/R. In crystal oscillators employing gridleak bias, when one crystal is replaced by another of the same nominal frequency, one of the crystals is usually found to produce stronger excitations and hence a larger grid current under similar operating conditions. Frequently the relative grid currents are defined to be equal to the relative activities of the crystals. With this method of measurement it can be seen that, if a crystal is connected directly across the grid-to-cathode input the excitation, and hence the activity, will depend upon the amplitude of the r-f voltage across the crystal. On the other hand, if the crystal is connected in series with the oscillator input, the activity will depend upon the amplitude of the current through the crystal unit. Since, in any event, the grid current depends upon the values of every parameter in the oscillator circuit, such a measurement is ambiguous unless a standard test circuit can be referred to for each frequency. Only in this way can the crystal unit, itself, be considered the only significant variable. Even under the assumption of ideal test conditions, however, the exact mathematical relationship among the crystal parameters, which provides the most direct measure of a crystal unit's inherent activity quality, has been a subject of some controversy. A number of suggestions have been made, but the usefulness of each of these depends considerably upon the method by which the crystal is to be used to control oscillations. As the crystal terminology becomes more rigorously defined we can imagine that the word "activity" will fall into disuse eventually, with "effective resonance resistance" or "conductance" taking its place.

ACTIVITY QUALITY FOR SERIES RESONANCE

1-229. As an example of series-mode operation, we refer to the test circuit in figure 1-106. It can be seen that the grid excitation will be approximately equal to IX_{C1}, the r-f voltage developed across C_1. X_{C1} depends upon the frequency and the value of C_1, whereas, I, the current through C_1 depends upon the B^+ voltage, the setting of P, the tube characteristics, etc., as well as the tank impedance and hence the resistance of the crystal. With all the circuit parameters constant, the only variable that the crystal introduces at resonance is its resistance R. Rather than specify all the parameters of the test circuit for each nominal frequency, it is clear that the measurements of R provide sufficient indication of the relative activities of different crystal units under any similar conditions of resonance. Since the current and voltage amplitudes vary inversely with R, the series-resonance activity of any crystal unit can be assumed to be directly proportional to 1/R, the motional-arm conductance.

ACTIVITY QUALITY FOR PARALLEL RESONANCE

1-230. The interpretation of the activity quality of a crystal unit becomes more complicated when the crystal is to be operated at parallel resonance. But even as in the case of series resonance, the inherent property of the crystal unit that most readily indicates the relative activity is the motional-arm conductance, 1/R. In a parallel-resonant oscillator circuit, the excitation is normally directly proportional to the voltage developed across the crystal in parallel with its effective load capacitance C_x. This voltage in turn is proportional to $Z_p \approx X_T^2/R$, where Z_p is the parallel-resonance impedance, X_T is the reactance of C_T, the total shunt capacitance, and R is the series-arm resistance. As long as X_T remains constant, the only significant crystal variable that affects the activity is R, or more directly 1/R. A complication arises from the fact that $C_T = C_o + C_x$, so that if C_x, the effective capacity of the external circuit is to be held constant, then X_T, and hence the activity, changes with C_o. Another complication arises when a measure of crystal quality is desired that will hold between crystals of different nominal frequency. If C_T or C_x is to be held constant, then Z_p tends to vary inversely with the square of the frequency. Finally, the complications are multiplied several fold when one begins to take into account the many ways in which a crystal can be connected to stabilize or to control a parallel-resonant type of circuit. Unless the term "activity" is to refer to some desired and well-defined end result that can be measured quantitatively, the word may have little practical meaning. With this in mind, we note that since the usefulness of the oscillator depends entirely upon its output, the useful oscillator activity can be said to concern only the amplitude of oscillations in the load circuit. *Thus, the relative ease with which a crystal unit enables a given output to be achieved under specified conditions can be regarded as the relative activity quality of the crystal unit.* From this point of view, the relative activity of a crystal unit can be considered inversely proportional to the driving power that the crystal requires in order to maintain a given output level in a fixed load. Now, there is a special case of parallel-mode operation—where the series arm operates into a constant C_T—for which the activity, as interpreted above, requires only a measurement of R. Assume that a small, variable shunt capacitance, C_v, can be connected directly across each crystal unit whose activity quality is being tested, so that the effective shunt capacitance (eff $C_o = C_o + C_v$) can be adjusted to give the same value for all crystals. With this arrangement, the crystal power dissipation for a given power output will vary positively with the motional-arm resistance, R. In other words, two crystal units of the same series-resonance frequency and the same motional-arm resistance can produce the same oscillator activity at the same parallel-resonance frequency regardless of the particular values of C_o, or of the motional-arm C and L, but only if a variable shunt capacitor is provided by which the effective C_o can be held constant. Thus, for gauging the potential activity of a particular crystal unit to be used in any parallel-mode oscillator circuit having a variable capacitor connected directly across the terminals of the crystal unit, the motional-arm conductance, 1/R, can usually be considered a sufficient and proper activity parameter.

1-231. Where the load capacitance, C_x, that the entire crystal unit faces (not necessarily $C_T = C_o + C_x$ that the series arm faces) is to remain constant, the parameter 1/R is not a sufficient index of the activity quality. If the proper measure of crystal activity is defined to be the ratio of output power to crystal power, such a definition is general enough to be applicable for any type of crystal oscillator. Certainly, the crystal unit that requires the least expenditure of energy to perform its task should be considered the one of greatest activity. Unfortunately, activity quality, from the point of view of a power ratio, becomes a function of each particular oscillator and load circuit, so that the generalization gained in the definition is completely lost on application, unless a standard test oscillator is available for each type of circuit. Only when the crystal unit is operated at its series-resonant frequency or is connected directly in parallel with a variable capacitor, can the relative activity qualities of two or more crystal units be considered constant and independent of the particular design of the external circuit. In all other cases, C_o, as well as R, becomes a significant parameter of the activity, and the exact relation of C_o to the activity will depend upon the circuit design. For a parallel-mode activity parameter to apply in the general case, the oscillator circuit, itself, must be considered from a generalized point of view. By this approach, the effective Q $\left(Q_e = \frac{X_e}{R_e}\right)$ is often considered a more reliable activity quality factor, than 1/R alone. The reason for this belief is most readily indicated when a generalized crystal oscillator is represented by the negative-resistance method — in particular, by diagrams (A), (B), and (C) in figure 1-108.

Q_e AS AN INDEX OF ACTIVITY QUALITY

1-232. If the oscillations of a crystal are to be sustained, energy must be supplied at a rate equal to the power losses in the crystal. This state is indicated in figure 1-108 (A), where the power source is represented as a generator with an emf equal, but opposite in sign, to the voltage across the effective resistance of the crystal unit. The power input thus is $|EI_c| = I_c^2 R_e$. Furthermore, since the total voltage drop around the circuit must be zero, the external circuit must appear as having a reactance equal, but opposite in sign, to the effective reactance of the crystal. Now, since the current through each component of the equivalent series circuit is the same, the voltage may be represented as being the result of a current flowing through a circuit of zero total impedance, as shown in figure 1-108 (B). Note that the generator is replaced by a negative resistance, ρ_s, numerically equal to R_e. Figure 1-108 (C) shows the same operating conditions, but with X_1 and ρ_s of the external circuit replaced by an equivalent capacitive reactance, X_x, in parallel with a negative resistance, ρ, equal to $-Z_p$, where $Z_p \approx \frac{X_e^2}{R_e} \approx \frac{X_x^2}{R_\rho}$. At all instants the impedance across the terminals at 1 and 2, whether that of the crystal unit or of the external circuit, is the same for both the (B) and the (C) equivalent circuits. Imagine now that after equilibrium has been reached, R_e suddenly decreases by one half. At this instant the power being supplied, $I_c^2\rho_s = I^2\rho$, is greater than that being dissipated in R_e. In (B), the amplitude of oscillations will increase until a new equilibrium is reached, at which time ρ_s will also have decreased by one half and will be once again numerically equal to R_e. In (C), the halving of R_e means that Z_p is approximately doubled. The same increase in current through R_e must be shown to occur in (C) as in (B). Thus, the amplitude of the oscillations must increase until ρ has doubled its value and is again numerically equal to Z_p. (The changes in ρ are caused by the limiting elements in the oscillator. For example, R_p of the vacuum tube will increase with an increase in gridleak bias, and this will be reflected as an increase in ρ.) From the generalized circuit approach, we can reach the general conclusion intuitively that oscillations build up as long as $|\rho_s| > R_e$, or $|\rho| < Z_p$, and that oscillations diminish in amplitude under the reverse conditions. It may not be at once apparent why oscillations should not build up if ρ were numerically larger than Z_p, in the same way that they do when ρ_s is greater numerically than R_e. A rigorous proof can be obtained by a differential equation of the current through the inductance, applying Kirchoff's laws and keeping in mind that resistance, negative or positive, is mathematically an instantaneous rate of change of voltage with current. Qualitatively it can be seen that the amplitude increases or decreases, depending upon

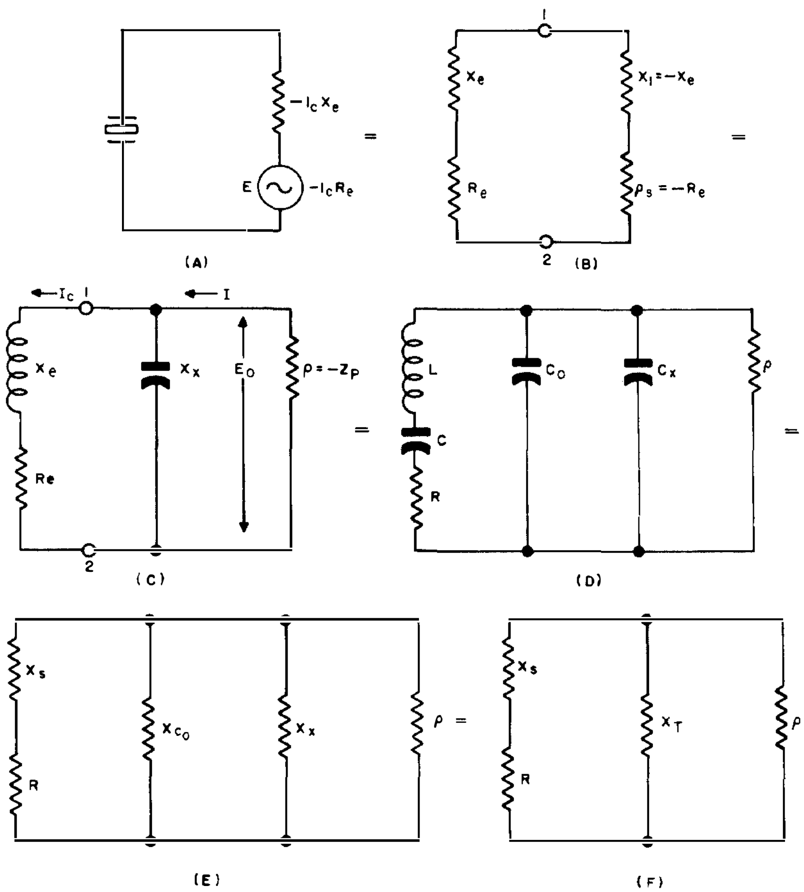


Figure 1-108. Generalized crystal oscillator circuit

whether the ratio of the power input to the power dissipation is, respectively, greater or less than 1. For circuit (B) the current, I_c, is the same for both the crystal unit and ρ_s, so that the power ratio is:

$$\left|\frac{I_c^2\ \rho_s}{I_c^2 R_e}\right| = \left|\frac{\rho_s}{R_e}\right| \qquad 1\text{—}232\ (1)$$

Oscillations thus build up as long as ρ_s is greater than R_e. For circuit (C) the voltage, E_o, is the same across the crystal unit as across ρ, so that the power ratio is:

$$\frac{E_o^2/\rho}{R_e\, E_o^2/Z_e^2} \approx \left|\frac{Z_p}{\rho}\right| \qquad \text{where } Z_e = \sqrt{R_e^2 + X_e^2} \text{ and } X_e >> R_e$$

$$1\text{—}232\ (2)$$

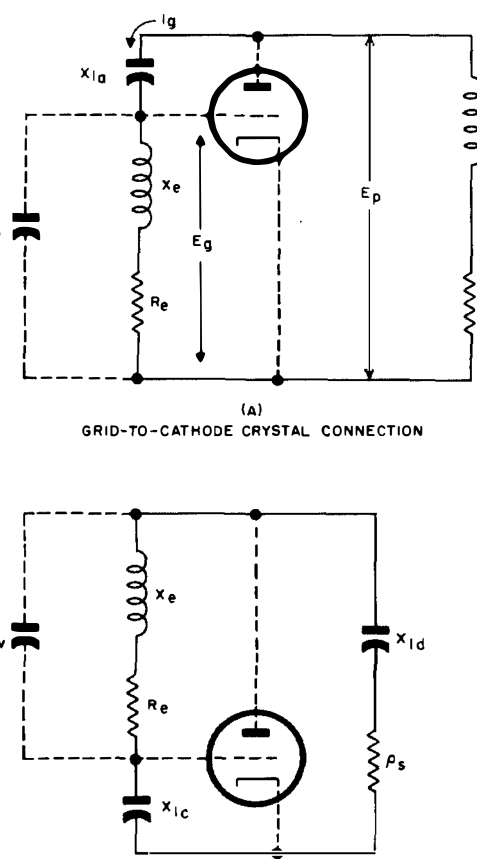
Oscillations thus build up as long as ρ is less than Z_p, the equivalent parallel-resonance impedance at equilibrium. The initial values of ρ_s and ρ for a given frequency can be assumed to be fixed parameters characteristic of the particular oscillator circuit, although the exact magnitudes may be extremely complicated functions involving all the circuit variables. Nevertheless, it is reasonable to assume that the more the negative resistance must change, the greater will be the activity of the oscillator by the time equilibrium is reached. From the point of view of circuit (B), it would seem that with a given initial ρ_s maximum activity is to be obtained with a minimum R_e; but in circuit (C), on starting with a given value of ρ, maximum activity is to be obtained with a maximum Z_p. Note that these two conditions are not entirely equivalent. For a given crystal unit, Z_p, for instance, can be increased by increasing X_x, which in turn requires that R_e, as well as X_e, become greater. (Since X_e must increase to match the increase in X_x, so must the frequency, and hence also R_e.) Remembering that the activity that is assumed to be proportional to the change in negative resistance is that in the oscillator output and is not necessarily the current amplitude in the crystal circuit, it can be seen that R_e alone, in spite of the implications to be drawn from figure 1-108 (B), may not be a sufficient parameter to indicate the relative activity quality of a crystal unit in the general case, i.e., X_e must also be considered. For these reasons, the effective Q of the crystal unit, $Q_e = X_e/R_e$, is usually considered the more reliable index of the crystal activity quality for parallel-resonant oscillators. There are exceptions, however, where R_e, or rather $\frac{1}{R_e}$, is the proper activity parameter. These occur when the crystal is actually operated at series resonance with an external capacitance. An example is the CI-meter circuit in figure 1-106, when C_x is connected in series with the crystal. (See paragraph 1-585 for a more detailed analysis of negative-resistance limiting.)

1-233. The crystal Q_e is a more direct index of the potential activity in some oscillator circuits than in others. The first consideration is the effect that a change in Q_e has upon the excitation voltage. Normally, an increase in Q_e means an increase in excitation, but this is not true in every case, even in the conventional parallel-resonant circuits. In these oscillators, the feed-back network may consist of a crystal unit shunted by one capacitance and in series with another. Referring to figure 1-109, assume at first that the capacitance, C_v, shunting the crystal unit in both (A) and (B) is negligible. The generalized circuits are thus equivalent to that of figure 1-108 (B). X_1 of figure 1-108 (B) is represented by $(X_{1a} + X_{1b})$ and by $(X_{1c} + X_{1d})$ in circuits (A) and (B), respectively, of figure 1-109. Referring now to figure 1-109 only, the crystal unit in circuit (A) is connected between the control grid and cathode, so that the principal activity consideration is to obtain the desired excitation voltage across the crystal unit with a minimum power dissipation in the crystal unit. Similarly, in circuit (B), the higher the crystal quality, the less the crystal power that would be required to obtain a desired excitation voltage across X_{1c}. As a first approximation, assume that X_e is much greater than R_e, so that the voltage across the crystal unit can be assumed to equal I_gX_e, where I_g is the feed-back current and where I_gX_e, is 180° out of phase with the voltage across the series reactance, X_{1a}, or X_{1c}, as the case may be. In circuit (A), the excitation voltage is thus equal to I_gX_e, and the crystal power dissipation is $I_g^2R_e$. If the ratio of the r-f output voltage of the tube to the excitation voltage, E_p/E_g, is assumed to be k, then the ratio of the output voltage to the crystal power, P_c, is:

$$\frac{E_p}{P_c} = \frac{kI_gX_e}{I_g^2R_e} = \frac{kQ_e}{I_g} \quad \text{Circuit (A) (figure 1-109)}$$

The magnitude of the total impedance of the feed-back circuit must be kZ_g, where Z_g is the grid-to-cathode impedance. As long as it can be assumed that the impedance of the crystal unit is 180° out of phase with the reactance of the series capacitance, then the magnitude of the plate-to-grid impedance must be approximately $(k + 1)Z_g$. Thus, in circuit (B), $|X_{1c}| \approx \frac{X_e}{k+1}$. Substituting this value of X_{1c} in place of X_e in the equation above:

$$\frac{E_p}{P_c} = \frac{kQ_e}{(k + 1)\, I_g} \quad \text{Circuit (B) (figure 1-109)}$$



(B)
PLATE-TO-GRID CRYSTAL CONNECTION

Figure 1-109. Generalized crystal oscillator circuits, showing two conventional methods for connecting crystal unit (X_e, R_e) in feed-back circuit

The equations for circuits (A) and (B) show that for a given k and a given crystal current, a maximum ratio of E_p/P_c is to be obtained when Q_e is a maximum. This assumes that the circuit capacitance, C_v, directly shunting the crystal unit is negligible. When this assumption cannot be made, the effective Q of the parallel combination must be substituted for Q_e. The Q of the combination is approximately equal to $Q_e\left(\frac{X_v + X_e}{X_v}\right)$ as long as $(X_v + X_e)$ is numerically large compared with R_e, X_v being the negative reactance of C_v. The larger the magnitude of the ratio X_v/X_e, the more directly does Q_e become the principal activity index. It should be remembered that the direct proportionality between Q_e and the activity in the example above holds only upon the assumption that I_g is to be held constant, regardless of the value of Q_e. Another instance in which the activity of an oscillator is a direct function of Q_e would be the unconventional case of an oscillator so designed that the crystal unit is operated in series resonance with an external capacitance and with the excitation voltage equal to, or directly proportional to, the voltage across either the crystal unit or the series reactance. In such a circuit, use would be made of the resonant rise in voltage that is developed when a component impedance is greater in magnitude than the total impedance. Since the current through the component is the same as that through the total impedance, the step-up voltage ratio is the same as the impedance ratio. At series resonance the total impedance of the crystal circuit would equal R_e (assuming no other resistance in the circuit), so that if Z_e of the crystal unit were approximately equal to X_e, the ratio of the voltage across the crystal unit to the feed-back emf would be $I_cZ_e/I_cR_e = Q_e$. The standard crystal units which are intended for use at parallel resonance are tested for operation with definite values of load capacitance, C_x. Thus, the recommended operating value of X_e may be assumed to be equal to $\frac{1}{\omega C_x} = |X_x|$. The maximum value of R_e that is permissible with this value of X_x is also specified. Hence, in the design of an oscillator that must operate satisfactorily with any randomly selected crystal unit of a given type, allowance must be made in the circuit design to ensure that satisfactory activity is obtained for the minimum $Q_e\left(= \frac{1}{\omega C_x R_e\ (\max)}\right)$.

MAXIMUM EFFECTIVE Q (Q_{em})

1-234. Where it is desirable to have an activity parameter that is not a function of the particular external load capacitance, the maximum effective Q (Q_{em}) offers a convenient index of the maximum potential activity of a crystal unit which is to be operated in a type of circuit whose activity depends primarily upon Q_e. The maximum effective Q can be expressed solely in terms of the basic crystal parameters, since the maximum occurs midway between f_r and f_a, so that $\Delta f = \frac{\Delta f_a}{2}$. Thus, from equations (1) and (2) of figure 1-98, it can be seen that

$$Q_{em} = \frac{X_e}{R_e} = \frac{2\pi L \Delta f_a}{2R}$$

Since

$$\Delta f_a = \frac{fC}{2C_o} \quad \text{[by equation 1—208 (1)]}$$

Then

$$Q_{em} = \frac{2\pi fLC}{4RC_o} = \frac{\sqrt{LC}}{4RC_o} = \frac{1}{4R\omega C_o} = \left|\frac{X_{Co}}{4R}\right| \qquad 1\text{—}234\ (1)$$

Q_{em} provides a convenient activity factor combining all the crystal parameters. It is equal to the maximum step-up voltage ratio that can be obtained by operating the crystal in series with a negative reactance. Where another capacitor, C_v, is shunted directly across the crystal unit, the maximum effective Q of the combination becomes

$$\sqrt{LC}/4R(C_o + C_v).$$

FIGURE OF MERIT, M

1-235. In paragraph 1-216, it was shown that the minimum f_p obtainable with a crystal unit occurs when the apparent Q_s of the motional arm, $\frac{X_s}{R}$, is equal to 2; that is, unless R_T, the effective resistance in the shunt arm, is significant. Within the frequency range at which the crystal unit appears as a positive reactance, the maximum value of Q_s occurs at the antiresonant upper limit. This maximum theoretical value of Q_s has been selected as a convenient figure of merit to indicate the relative activity quality of a crystal unit, and has been assigned the symbol M. In general, the larger the value of M, the less will be the feed-back energy required to sustain a given activity. If M is less than 2, the crystal unit cannot exhibit a positive reactance, and hence cannot be used in conventional oscillator circuits. To sustain oscillations at a desired level, an oscillator will require that the crystal exhibit some minimum value of Q_s, equal to 2 or greater, depending upon the oscillator, so that a knowledge of the M of a crystal unit is of value in determining whether or not the crystal can be used. Formulas for M are:

$$M = \left|\frac{X_{Co}}{R}\right| = \frac{X_{sa}}{R} = \frac{4\pi L\Delta f_a}{R} = \frac{2\pi fLC}{RC_o}$$

$$= \frac{Q}{r} = \frac{\sqrt{LC}}{RC_o} = 4\,Q_{em} \qquad 1\text{—}235\ (1)$$

where Q is the series-arm Q and $r = C_o/C$. Note that M is equal to four times the maximum Q_e of the crystal unit, so that the measurement of either will indicate approximately the same performance characteristics. Actually, as M approaches 2 the value of Q_{em} as given by equation 1-234 (1) becomes unreliable, because of the approximations made in its derivation. If $M = 2$, Q_{em} is zero, although its approximate formula would indicate a value of 0.5. In practice, however, crystal units with such low values of M are normally far below specified standards, except possibly in the case of v-h-f crystal units operating on harmonics higher than the fifth, so that Q_{em}, which can be meassured directly as the maximum step-up voltage ratio obtainable with the crystal unit in series with a capacitor, provides a reasonably accurate indication of $\frac{M}{4}$. M was originally chosen as a figure of merit because it can be shown to be a constant of proportionality in the equation for Q_e, and because it is expressible in terms of the crystal parameters alone. As performance parameters of a crystal unit, M and Q_{em} are practically equivalent, but M is the parameter more commonly encountered in treatises discussing crystal activity.

PERFORMANCE INDEX

1-236. The fact that the Q_e of a crystal unit is the most direct factor influencing the activity first became apparent through consideration of the requirements necessary for oscillations to build up in the generalized oscillator circuit in figure 1-108 (C). When X_e and R_e are assumed to be the reactance and resistance of an actual coil, it can be rigorously shown that oscillations build up as long as $|C_x\rho| < \frac{L_e}{R_e}$. Here, C_x and ρ are both functions of the external circuit, and L_e and R_e can be assumed to be constants of the coil. As the amplitude of oscillations increases, the plate resistance of the tube increases, which in turn causes ρ to increase. (C_x may also vary, but usually to a much smaller degree.) Multiplying both sides by ω: $|\omega C_x\rho| < \frac{\omega L_e}{R_e}$ or $\frac{\rho}{X_x} < Q_e$. From this point of view it would appear that the change in ρ/X_x which must be undergone before equilibrium is reached, or, equivalently, the rise in amplitude necessary to bring the plate resistance to the equilibrium point, will always increase or decrease with Q_e, and that with a given R_e, the amplitude increases or decreases with X_e. These implications can be misleading. First, with a given X_x, X_e is no longer a significant variable if Q_e is equal to 10 or more, but must remain equal in magnitude to X_x. In this event the only variable of the activity is R_e. Secondly, the rise in amplitude is more accurately a function of the difference between Q_e and the starting value of $\frac{\rho}{X_x}$ rather than of the magnitude of Q_e alone. Furthermore, Q_e is, itself, a function of X_x, for as C_x is varied from a relatively large value of capacitance and made to approach zero, X_e must increase with X_x. However, Q_e does not increase indefinitely with X_e, but reaches its maximum value, Q_{em}, when $C_x = C_o$ and then steadily decreases as C_x becomes smaller. Yet the difference between Q_e and the starting value of $\frac{\rho}{X_x}$ does continue to increase even though Q_e has passed its maximum, for the change in $\frac{X_e}{R_e}$ is less than the change in the value of $|\rho\omega C_x|$. Now, C_x and even more so, C_t ($= C_o + C_x$), can be considered reasonably constant parameters as compared with ρ during the build-up of oscillations. Referring to figure 1-108 (F), it can be seen that oscillations build up as long as $\frac{X_s}{R} > \frac{\rho}{X_T}$ (for the same reasons that hold in the case of $\frac{X_e}{R_e}$ and ρ/X_x). Since X_T can be assumed to be relatively constant, both sides of the function can be multiplied by X_T, so that it can be said that as long as $\left|\frac{X_sX_T}{R}\right| \approx Z_p > |\rho|$, the amplitude of the oscillations continues to increase. Thus, for a given starting value of ρ and with X_T relatively independent of the amplitude, the most direct index of the activity is the equivalent parallel-resonance impedance, Z_p, of the generalized oscillator circuit. For this reason, Z_p is called the Performance Index and has been given the symbol PI. PI, unlike M and Q_{em}, is not a parameter of the crystal unit alone, but of the crystal unit effectively in resonance with some specified load capacitance, C_x. Military Standard crystal units intended to be operated at parallel resonance have a recommended load capacitance specified. PI meters have been developed for measuring the performance index directly, but there are very few such meters available. Where the PI of a crystal unit is desired, it can readily be computed from measurements made with standard CI meters. Various expressions of PI are given below:

$$PI = \frac{X_T^2}{R} = \frac{1}{\omega^2 (C_o + C_x)^2 R} = \frac{LC}{RC_T^2}$$

$$= \frac{|X_{Co}|^2}{R\left(1 + \frac{C_x}{C_o}\right)^2} = \frac{|MX_{Co}|}{\left(1 + \frac{C_x}{C_o}\right)^2} \qquad 1\text{—}236\ (1)$$

Note that PI is not a function of the crystal alone, but of C_x as well, and care should be taken that the capacitance ratio of C_x/C_o is not mistaken for $r = C_o/C$. It can be seen that the maximum PI occurs at antiresonance, where $C_x = 0$, so that (max) $PI = |MX_{Co}|$.

1-237. The PI of a crystal unit, or more properly, of a crystal circuit is usually found to be an important parameter entering the equations of an oscillator circuit, particularly when the equations express the conditions required for a given output. As a simple example, consider again the two generalized oscillator circuits in figure 1-109. In paragraph 1-233, it is shown that the ratio of E_p to crystal power, P_c, is equal to kQ_e/I_g for circuit (A), and equal to $\frac{kQ_e}{(k+1)I_g}$ for circuit (B). As long as I_g is considered predetermined, the voltage-to-power ratio is primarily a function of k and Q_e. Normally, however, it is not I_g that is to be predetermined, but rather the crystal power that must not exceed the maximum value. When $Z_e \approx X_e$, the magnitude of the impedance of the feed-back arm in circuit (A) can be assumed to be equal to kX_e, and to $\frac{kX_e}{k+1}$ in circuit (B). I_g is thus equal to E_p/kX_e in (A) and to $\frac{E_p(k+1)}{kX_e}$ in (B). On substitution in the E_p/P_c equations:

$$E_p/P_c = \frac{k^2 Q_e X_e}{E_p}$$

or $$E_p^2/P_c = \frac{k^2 X_e^2}{R_e} = k^2 PI \qquad \text{Circuit (A)}$$

Similarly, $$\frac{E_p^2}{P_c} = \frac{k^2 PI}{(k+1)^2} \qquad \text{Circuit (B)}$$

Inasmuch as the power output of the oscillator can be assumed to be directly proportional to E_p^2, then, for a given drive level of the crystal unit, the power output will vary directly with the PI. It should be noted that the E_p^2/P_c equations above are not affected by the assumption of shunt capacitances, C_v, across the crystal unit. Just as $PI = \frac{X_e^2}{R_e} = \frac{X_s^2}{R}$, the same value holds even if the shunt capacitance is assumed to be increased by C_v.

Frequency Stabilization Quality of Crystal Unit

1-238. The over-all frequency stability of a crystal oscillator is dependent upon the stability of all the parameters influencing the crystal circuit; these in turn are dependent upon the stability of the power source and the load, as well as the ambient conditions under which the oscillator is required to operate. The over-all stabilizing ability of the crystal is dependent upon both the stability of the crystal parameters when the crystal is exposed to changes in temperature or drive level, and the ability of the crystal to minimize the change in frequency that is necessary when the parameters of the external circuit deviate. It is this latter quality of the crystal that makes the crystal oscillator superior to oscillators that use only coils and condensers to control the frequency, and is the type of frequency stability that concerns us now.

1-239. The frequency stabilizing property of a crystal is normally expressed as the rate at which its reactance changes with frequency. In figure 1-110 are shown the curves of $\frac{dX}{df}$ for the series-arm parameters L and C. Resonance happens to occur at the frequency at which X_C is changing at the same rate as X_L. Since the rate of change of X_L is a constant, at frequencies near resonance it can be said that the total change in reactance with frequency is primarily a function of L, for the absolute rate of this change is the same whether C is large or small. Normally, however, it is not the absolute change in reactance that is important—it is, rather, the change in reactance per percentage change in frequency, or, more usually, the percentage change in reactance per percentage change in frequency. When the frequency stability is expressed in percentage, it is no longer primarily a function of L, but becomes dependent upon the other crystal parameters as well. Only where the major concern is to produce a definite shift in reactance or frequency for a given change in the external circuit does the major attention center upon the parameter L. Just as the relative activity potential of a crystal depends somewhat upon the type of circuit in which it is used, so also does the frequency stabilizing characteristic of a crystal depend upon the external-circuit de-

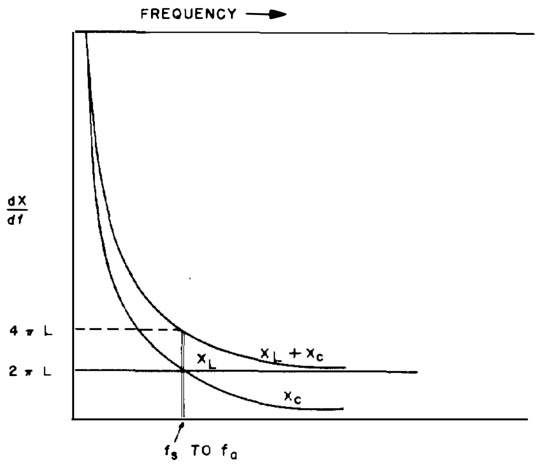


Figure 1-110. Rates of change of reactance of equivalent series-arm parameters, L and C, with frequency

sign. A relative stability index will be discussed briefly for each of three general types of circuits: where the crystal is operated at its normal series-resonant frequency, where it is operated in parallel with a negative reactance, and where it is operated in series with a negative reactance.

FREQUENCY STABILITY AT SERIES RESONANCE

1-240. Since the total series-arm reactance at f_s is equal to zero, it is not convenient to express the relative frequency stability in terms of the percentage rate of change in reactance. Approximately the same considerations apply for the resonance frequency, f_r. Also, the effective stability in a given circuit may well depend more upon the rate of total impedance change or the rate of phase shift with frequency than upon the actual rate at which the reactance changes. Suppose, for example, that the feed-back energy must pass through the crystal unit and return to the oscillator input in a certain phase. If, because of a change in the circuit parameters, the feed-back energy is returned slightly out of phase, the frequency will have to shift away from the normal resonant point exactly enough for the crystal to correct the change in phase. If the change in phase has originally been caused by a change in the reactance of, say, a capacitor connected directly in series with the crystal, it is only necessary for the frequency to shift the amount necessary for the crystal reactance to exactly counteract the change in the series reactance. In this case, the resistance of the crystal circuit is not effective in degrading the stability. It is true that the greater the resistance that the crystal faces, the greater must be the frequency change to produce a given phase shift. But on the other hand, since the series capacitance faces the same resistance, the initial phase shift due to a change in the capacitance is correspondingly reduced. In this case, the frequency stability of one crystal as compared with that of another depends almost entirely upon its relative rate of change of reactance. At series resonance the frequency stability factor can be defined as

$$F_s = \frac{dX_e}{df}f = 2\omega L = 2\sqrt{L/C} \qquad 1\text{—}240\ (1)$$

where F_s is the rate of change of reactance per fractional change in frequency. Thus, in comparing one crystal unit with another, the one with the larger L/C ratio can be assumed to provide the greater frequency stability at series resonance. However, if the change in reactance occurs at a point in the circuit only loosely coupled to the crystal, the resistance of the feed-back circuit is relatively ineffective in reducing the phase shift of the feed-back voltage, but instead, tends to increase the change in frequency necessary for the crystal to correct the phase. In the case of a feed-back network where the crystal must compensate for a change in phase that is relatively independent of the resistance in the crystal circuit, the frequency stability is more directly measured by the rate of change of phase in the crystal circuit as a whole than by the rate of change of reactance alone.

1-241. Figure 1-111 shows that a small phase displacement, $\Delta\theta$, at series resonance is approximately equal to $\Delta X_e/R_e$, where ΔX_e is a small

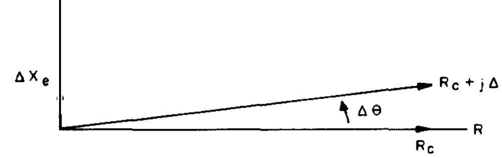


Figure 1-111. Phasor diagram, showing change in reactance, ΔX_e, of series-mode crystal required to produce a small change in phase, $\Delta\theta$, where the resistance of the crystal circuit is equal to R_e

change in the effective reactance of the crystal, and R_c is the total effective resistance in the crystal circuit (equal to R_e of the crystal plus R_x of the external circuit). For convenience, the frequency is usually expressed in terms of angular frequency, $\omega = 2\pi f$ radians per second, instead of cycles per second. In equations (1) and (2) of figure 1-98, it can be seen that for small values of Δf, the denominators of the approximate equations of X_e and R_e are approximately equal to unity, in which case $X_e \approx 2L\Delta\omega$ and $R_e \approx R$. Thus,

$$\Delta\theta = \frac{\Delta X_e}{R_c} = \frac{2L\Delta\omega}{R + R_x}$$

The frequency-stability index can be defined to be

$$\frac{d\theta}{d\omega} = \frac{2L}{R + R_x} \qquad 1\text{—}241\ (1)$$

Expressed as the change in phase angle per percentage change in frequency, equation (1) becomes:

$$\frac{\omega d\theta}{100\, d\omega} = \frac{2\omega L}{100\,(R + R_x)}$$

or more simply:

$$\frac{\omega d\theta}{d\omega} = \frac{2\omega L}{R_c} = \frac{2X_L}{R_c} = \frac{2}{R_c}\sqrt{L/C} \qquad 1\text{—}241\ (2)$$

The last term on the right shows that where the fractional rate of change is concerned, the frequency stability is directly proportional, not simply to L, but to the square root of the L/C ratio. Equation (2) also shows that the frequency stability is inversely proportional to the crystal circuit resistance. But it must be remembered that this is true only to the extent that the original phase shift of the input to the crystal circuit can be considered independent of R_c.

1-242. As an exaggerated example, we can see that minimum stability is to be expected if the input to a high-resistance, series-resonant, feedback circuit is supplied through a weak coupling from a plate tank circuit sharply tuned to the resonant frequency but having an impedance small compared with the R_p of the tube. Since a slight change in the parameters of the tank circuit could shift the phase of the feed-back input almost 90 degrees, such an oscillator would obviously be completely unstable, even if it were assumed that oscillations could be maintained.

FREQUENCY STABILITY AT PARALLEL RESONANCE

1-243. When it can be assumed that a crystal unit is effectively operating in parallel with an external capacitance, C_x, the value of which is relatively independent of small changes in the frequency, the frequency stability can be assumed to be directly proportional to the rate of change in the reactance of the motional arm of the crystal. In this case, it would seem that a frequency-stability factor for parallel resonance

$$F_p = \frac{dX_s}{d\omega}\cdot\omega = 2\sqrt{L/C} = F_s$$

would be appropriate—just as in the case of series resonance—and F_p would be identical with F_s. However, it will be recalled (see figure 1-103) that the higher the reactance, i.e., the smaller the value of C_T, the less stable will the oscillator become. Taking this into account, a more accurate indication of the stabilizing quality of a parallel-resonant crystal is given by what is called the *frequency-stability coefficient,* which is the percentage rate of change in reactance for a percentage change in frequency. Thus,

$$F_{Xs} = \frac{dX_s}{d\omega}\cdot\frac{\omega}{X_s}\cdot\frac{100}{100} = \frac{2\omega L}{X_s} = \frac{f}{\Delta f}$$

This result is quite interesting, for it indicates that for crystal units of the same frequency equal stabilities can be achieved simply by operating the crystal units at the same value of Δf above series resonance. Since the same equation holds for any value of L and C, it is not immediately apparent as to why a crystal is so much more stable than a conventional inductor and capacitor. In paragraph 1-208 it was found that the ratio $\frac{\Delta f_s}{f}$ is equal to $\frac{C}{2C_o}$. If Δf_p is substituted for Δf_s, and C_T is substituted for C_o, then

$$F_{Xs} = \frac{f}{\Delta f_p} = \frac{2C_T}{C}$$

It can be seen that in order for a conventional series-parallel inductor-capacitor network to have the same theoretical stability as a crystal, the shunt capacitance must be thousands of times greater than the series-arm capacitance. The series arm of such a network would require an extremely small L/C ratio. The parallel impedance would be small, and the net series-arm reactance much smaller still. The crystal, on the other hand, has such a small value of C that the reactances are reasonably large even for small values of Δf. Although the equation for the frequency-stability coefficient indicates an unlimited stability if Δf_p is simply made small enough, this would be theoretically true only for a circuit resistance equal to zero. In pratice, as X_s approaches the motional-arm R in magnitude, a given change in X_T will cause a greater change in the phase of the over-all circuit impedance than will the same change in X_s. Of significance is the fact that the frequency-stability coefficient, F_{X_s}, represents the stabilizing effect of a crystal for the percentage change in the reactance of the total effective shunt capacitance, C_T ($= C_o + C_x$). With C_o considered constant, a given percentage change in the total capacitance becomes less than the actual percentage change in the variable component, which we can assume to be the equivalent external capacitance, C_x. For a given C_T, the larger the ratio C_o/C_x the smaller will be the percentage change in C_T for a given percentage change in C_x, and the greater will be the oscillator stability in the face of changes in the external circuit. In other words, the effective frequency-stability coefficient of the crystal unit as a whole is greater than that of the motional-arm alone if it can be assumed that the percentage changes in C_o will be negligible compared to those in C_x. When R is small compared with X_s, $X_e \approx X_{Co}X_s/(X_{Co} + X_s)$ and the effective frequency-stability coefficient becomes:

$$F_{Xe} = \frac{dX_e}{d\omega} \cdot \frac{\omega}{X_e} - \frac{2LX_{Co}^2}{(X_{Co} + X_s)^2} \cdot \frac{\omega(X_{Co} + X_s)}{X_{Co}\, X_s}$$

$$= \frac{2\omega L}{X_s} \cdot \frac{X_{Co}}{X_{Co} + X_s} = \frac{fC_T}{\Delta f C_x}$$

Since

$$F_{Xs} = \frac{f}{\Delta f} = \frac{2C_T}{C}$$

then

$$F_{Xe} = F_{Xs} \cdot \frac{C_T}{C_x} = \frac{2C_T^2}{CC_x} \qquad 1\text{—}243\ (1)$$

It should be remembered that equation (1) is based upon the assumption that C_o is constant and that any change in X_e will be due to a change in C_x. If C_o is effectively increased by a *fixed* capacitance, C_v, directly shunting the crystal unit, the effective variable C_x becomes smaller. The effective C_x, insofar as the frequency stability is concerned, will equal $C_T - (C_o + C_v)$. Substituting this effective value of C_x in equation (1) will provide a more accurate frequency-stability coefficient for a crystal unit directly shunted by a fixed C_v.

FREQUENCY STABILITY AT SERIES RESONANCE WITH EXTERNAL CAPACITANCE

1-244. The effective frequency-stability coefficient, $F_{Xe} = \frac{2C_T^2}{CC_x}$, provides an appropriate index of the frequency-stability quality of a crystal unit operated in series resonance with an external capacitance C_x, for the same reasons that make the coefficient applicable in the case of parallel-resonant circuits. C_T, here, represents the sum of two actual capacitances, $C_o + C_x$, and has a more concrete meaning than simply a generalized parameter. F_{Xe} gives the percentage change in X_e per percentage change in frequency. The reciprocal, $1/F_{Xe}$, can be interpreted as equaling the percentage change in frequency that will occur per percentage change in the negative reactance X_x. In unconventional circuit designs, where a significant phase shift can occur as a result of changes in the impedances in the oscillator output circuit (which is only weakly coupled to the feed-back input), the resistance of the feed-back circuit may need to be taken into account in a manner similar to that discussed in the case of crystal units operating in series resonance.

FREQUENCY STABILITY OF OVER-ALL CIRCUIT

1-245. Although the absolute values of the frequency-stability indices discussed in the foregoing paragraphs depend upon generalized parameters of the external circuit, the values are primarily useful for indicating the relative stabilization quality of one crystal unit as compared with another when operated under similar circuit conditions. The actual frequency drift due to changes in the supply voltages, tube characteristics, circuit impedances, etc., depends upon the particular oscillator design as well as upon the performance characteristics of the circuit components. The percentage variation that can be expected in the generalized parameter, C_x, is of equal importance in gauging the frequency stability of the oscillator as a whole. The crystal-unit frequency-stability coefficients appear as single parameters among others in the frequency-stability equations for each particular type of oscillator circuit. In general, the series-resonant type of oscillator has the greater frequency stability, permitting tolerances from four to twenty times as narrow as those normal for parallel-resonant oscillators. Indeed, in a well-designed series-resonant oscillator where the reactive components are negligible in their effect on the phase of the feed-back voltage, the frequency stabilization of the crystal unit can be very nearly perfect, so that the most significant factor to consider is the stability of the equivalent-circuit parameters of the crystal unit, itself. One source of frequency instability common to both series-resonant and parallel-resonant oscillator circuits is the presence of harmonics in the output. For certain applications, such as in crystal calibrators, these harmonics are desirable, but in most cases it is preferable that they be kept to a minimum. Harmonics are unwanted not only for the sake of a sine-wave output as such, but also because they introduce reactive components in the crystal circuit, thereby increasing the chances of frequency instability. The harmonics can be reduced by designing the oscillator plate circuit to provide a low-impedance bypass path for them, and by using low plate and grid voltages. Unwanted reactive effects in the oscillator circuit also occur as a result of feedback from the amplifier stages following the oscillator. These can be minimized by the use of proper shielding, buffer amplifiers, neutralizing circuits, and by careful attention to the physical layout in designing the equipment, to ensure that all leads are as short as practicable and that the oscillator is electrically isolated from circuits carrying high amplitudes of r-f voltage or current. The effective load capacitance, C_x, with which the crystal unit resonates is usually a function of the vacuum-tube parameters, the load resistance, the effective grid resistance, as well as the reactive impedances in the feed-back and output circuits. All these variables are, in turn, functions of the oscillator output and the grid and plate d-c voltages. Thus, the frequency stability is dependent upon the degree of voltage regulation, the constancy with which the oscillator load is maintained, and in the care taken in the original design to ensure that the circuit components are so proportioned that the effects of variations in the tube parameters are minimized. Silvered mica capacitors normally are to be preferred for fixed capacitances in the tuned circuits. Those capacitors having dielectrics composed of titanium compounds can be used for r-f bypass purposes, but are too variable under changes in temperature and voltage for use as tuning components. Air-dielectric capacitors are almost always adjustable. With the exception of the vacuum-dielectric capacitor, the air-dielectric type is the most stable and is to be preferred for small capacitances and variable tuning elements. As a rule, the improvements in circuit design that permit of greater frequency stability necessitate additional circuit components, additional tuning adjustments, narrower operating frequency ranges, smaller voltage or power outputs, or some combination of the above.

Bandwidth and Selectivity Parameters of Crystal Unit

THE CAPACITANCE RATIO, $r = \frac{C_o}{C}$

1-246. The bandwidth of a crystal unit refers to the particular frequency range over which the crystal unit can be operated in a given oscillator, filter, or transducer circuit. In the case of a conventional oscillator circuit, the applicable frequency range is that in which the crystal can appear as an inductive impedance. In cycles-per-second, this range is $\Delta f_a = f_a - f_r$. Percentage-wise, the bandwidth is $\frac{100\Delta f_a}{f_r}$, which, as shown in paragraph 1-208, is approximately equal to $\frac{100}{2r} = \frac{50C}{C_o}$. Although the practicable operating range does not extend over the entire band, it can be seen that the relative merit of a crystal unit insofar as its range of frequency adjustment is concerned can be indicated inversely by the parameter r, whereas the relative selectivity is indicated directly by r. In figure 1-95, it can be seen that the smallest theoretical values of r (when the distributed capacitance is negligible, so that $r = r_o$) are obtained with the low-frequency, length-extensional-mode elements of the X group. The smallest capacitance ratios are provided by element E, which has values of r as low as 120 to 125. These are equivalent to a resonance-to-antiresonance bandwidth on the order of 0.4 per cent of the nominal frequency. For the high-frequency A and B elements, the bandwidths are approximately 0.2 and 0.083 per cent, respectively.

1-247. Insofar as frequency control is concerned, the resonance-to-antiresonance bandwidth is important primarily as a relative index of the frequency range through which a parallel-resonant oscillator can be made to operate by varying the load capacitance, C_x. For example, the tuning adjustments of an oscillator employing an A element can vary the frequency approximately two-and-a-half times as much as can the same adjustments if the oscillator employs a B element. Although small frequency adjustments are possible, the high selectivity of quartz crystals precludes their use in frequency-modulated oscillators. Eventually, it may be that crystal units mounting high-frequency EDT plates, which have capacitance ratios as low as 20, will find an application in this field, but at the present time EDT crystals are used almost exclusively in filter networks. As a filter element, the capacitance ratio of a crystal is of greater importance than in frequency-control circuits. Filter networks, composed of crystal units alone, can be designed for a maximum pass band of $\frac{2\Delta f_a}{f_r}$, which in the case of quartz means a maximum pass band of 0.8 per cent. For the low-frequency networks, such as are normal to telephone carrier systems, this is much too selective for passing voice channels. For this reason, quartz crystals employed in l-f telephone carrier filters must be used in conjunction with inductors and capacitors. The narrow bandwidths of quartz elements used alone are primarily applicable in filters when it is desired to pass a single frequency, such as the pilot signal of a carrier system.

ELECTROMECHANICAL COUPLING FACTOR, k

1-248. To the extent that the equivalent circuit of figure 1-91 is applicable it can be assumed that when a crystal unit is connected across the terminals of a battery the ratio of the energy stored in electrical form to the energy stored in mechanical form is equal to the capacitance ratio $r = C_o/C$. The electrical energy is that stored in the static capacitance, C_o, and is equal to $\frac{1}{2} V^2C_o$, where V is the applied d-c voltage. The mechanical energy is the energy that is stored because of the piezoelectric strain in the crystal, and is equal to $\frac{1}{2} V^2C$. In transducer applications, it is useful to rate a crystal according to the ratio of stored mechanical energy to total applied electrical energy under the conditions of d-c or very-low-frequency applied voltages. The parameter for this purpose is the electromechanical coupling factor, k, equal to the square root of the ratio of the stored mechanical to the total input energy. As such, k is an index of the crystal efficiency as a transducer. This factor is given by the formula

$$k = \frac{d}{\sqrt{\frac{\epsilon}{4\pi} s}} \qquad 1\text{—}248\ (1)$$

where ϵ is the dielectric constant, s is the elastic compliance, and d is the piezoelectric constant giving the ratio of strain to field. According to the energy ratio, k^2 should equal $\frac{C}{C_o + C}$, or approximately $\frac{1}{r}$. However, the capacitance ratio, $\frac{C_o}{C}$, at resonance can be shown to be $\frac{\pi^2}{8}$ times as large as the theoretical ratio at zero frequency. Actually, then the ratio is

$$\frac{\text{stored mechanical energy}}{\text{total stored electrical energy}} = k^2 = \frac{\pi^2}{8} \cdot \frac{C}{C_o} \qquad 1\text{—}248\ (2)$$

where C and C_o are the equivalent capacitances at resonance. Since the bandwidth is proportional to C/C_o, so also is it proportional to k^2. In transducer applications, when an inductor is shunted across the crystal to tune out the electrical capacitance, and the crystal is operated near resonance, up to 90 per cent efficiency is possible in the conversion of electrical to mechanical energy. Under these conditions, k is not a direct index of the transducer efficiency, but it does serve as a parameter for estimating the frequency range over which the efficiency is 50 per cent or greater. The ratio of the highest to the lowest frequency for greater than 50 per cent conversion is:

$$\frac{f_H}{f_L} = \sqrt{\frac{1 + k}{1 - k}} \qquad 1\text{—}248\ (3)$$

Crystal Quality Factor, Q

1-249. The quality factor of a crystal unit is the Q of the motional arm at resonance. Thus,

$$Q = \frac{\omega L}{R} = \frac{1}{\omega CR} = \frac{1}{R}\sqrt{\frac{L}{C}} \qquad 1\text{—}249\ (1)$$

Quartz crystal units are obtained with Q's ranging in value from 10,000 to more than 1,000,000. The Q is a performance parameter that provides an indication of the ratio of the stored mechanical energy of vibration to the energy dissipated in the crystal unit per cycle at resonance. If I_s is the r-m-s current through the series arm at resonance, then, at the instant the current is a maximum, the equivalent capacitance C can be assumed to be completely discharged and all the vibrational energy, E_v, is at that instant in kinetic form. This energy is equivalent to that stored in motional-arm inductance, L. Therefore,

$$E_v = \frac{(1.414 I_s)^2 L}{2} = I_s^2 L \qquad 1\text{—}249\ (1)$$

The energy dissipated per second, P_c, is I_s^2R. Thus, the ratio of the stored mechanical energy to the energy dissipated per second is

$$\frac{E_v}{P_c} = \frac{I_s^2 L}{I_s^2 R} = \frac{L}{R} \qquad 1\text{—}249\ (2)$$

It can be seen that for a given wattage, the greater the L/R ratio the greater will be the amplitude of vibration. Regardless of the wattage, for a given L the amplitude of vibration will vary approximately directly with the current. Theoretically, since there are no Military Standards setting a minimum limit for the series-resonance resistance, a crystal unit can be so excellently mounted that it would be vibrated near its elastic limit if attention were given only to the power dissipation rather than to the current. Such a situation is not likely to arise except possibly in the case of a crystal-controlled power oscillator, where space and cost limitations require a crystal drive level far in excess of the rated level. More important from the point of view of maintaining a sinusoidal wave shape of the excitation voltage and of improving the stability of the oscillator is the ratio of the stored energy to the energy dissipated per cycle, rather than per second. In terms of angular frequency, the dissipation per radian is I_s^2R/ω, so that

$$\frac{E_v}{P_c/\omega} = \frac{I_s^2\omega L}{I_s^2R} = Q \qquad 1\text{—}249\ (3)$$

In an actual series-resonant circuit, it is the Q of the entire circuit rather than of the crystal unit alone that must be considered, so that R should be replaced by the total circuit resistance. If a tuned, class-C-operated circuit is to be effective in maintaining a sinusoidal wave shape and in reducing harmonics, the energy stored in the circuit should be at least twice the amount that is dissipated over the entire cycle. That is,

$$(\min)\ \frac{E_v}{P_c/f} = \frac{Q}{2\pi} = 2$$

This requirement is met easily in quartz-crystal circuits, but it is an important consideration in the design of plate tank circuits that are to be fed in pulses not smoothed by the action of the crystal. The crystal Q is also an important parameter in crystal filters. In general, the higher the Q the sharper the pass band.

1-250. Since the Q of a crystal is equal to $\frac{1}{R}\sqrt{\frac{L}{C}}$, and since the $\frac{L}{C}$ ratio for a given frequency can be increased to almost any value desired by decreasing the electrode area and by orienting the crystal in a direction of weak piezoelectric effect, or by using twinned crystal blanks, it might be wondered why much larger values of Q are not in use. The reason is that the L/C ratio and the equivalent series-arm resistance of the crystal are not independent of each other. As $\sqrt{L/C}$ increases, so also does R. This can be intuitively seen if it is kept in mind that fundamentally the Q is the ratio of the energy stored to the energy dissipated per angular cycle. Suppose that we have two crystal plates, A and B, both of approximately the same size and normal frequency, and both mounted exactly alike in that the frictional losses of one are the same as those of the other for the same energy of vibration. We shall also assume that these mechanical losses account for most of the crystal driving power. In other words, we are assuming that the two crystals have approximately the same quality factor, Q. Now, suppose that crystal A has a much smaller electrode area than does crystal B, or that for some other reason the piezoelectric effect of A is very weak compared with that of B. Under these conditions, crystal A will have a much larger equivalent L/C ratio than does crystal B. But since the Q of A equals the Q of B, it can be seen that the series-arm R of A must be greater than the series-arm R of B in the same proportion as the square roots of the respective L/C ratios. It should be understood that the Q and the L/C ratio are comparatively independent variables as far as R is concerned. Where R could not be estimated without a knowledge of Q and L/C, the latter theoretically could be approximated separately and independently by an examination of the fabrication of a crystal unit. L and C, for example, are approximately predetermined by the electrode area and the type and size of the crystal element. The Q is also to a certain degree a function of the same variables, but for given internal frictional properties, is primarily determined by the quality of the crystal finishing and mounting. Thus it is that the Q is largely determined by the frictional losses and is not subject to control by varying the L/C ratio. Indeed, as the L/C ratio increases, the piezoelectric effect can become so weak and the resistance so high that the crystal cannot be shocked into oscillation unless very high voltages are employed. Once in oscillation, a high L/C crystal unit could presumably operate satisfactorily, except that only very small currents could be withstood without the crystal shattering or arcing. There is a hypothetical case where an exceptionally large L/C ratio could be practical. Such a situation would arise if for any reason the external circuit resistance faced by a crystal could not readily be reduced below some large minimum value. In this event, the use of a crystal unit of normal Q but large L/C ratio would prevent the over-all circuit Q from being excessively degraded by the external resistance. Ordinarily the selection of a crystal unit will be made on the basis of considerations other than the L/C ratio, but where all else is equal, including the average values of Q, it might be assumed that the crystal units having the somewhat higher values of L/C are generally more suitable for those oscillators which do not require a crystal to sustain oscillations, but only to stabilize them. Such a circuit oscillating at or near the crystal frequency can build up the crystal vibrations over a large number of cycles of small amplitude, thereby obviating the need of large voltage surges or abnormally high vacuum-tube amplification.

Stability of Crystal Parameters

1-251. Regardless of how well designed a crystal oscillator may be, or how high the degree to which the crystal stabilizes fluctuations in the external circuit, the over-all performance will depend upon the stability of the crystal parameters, themselves. Changes in the crystal parameters are primarily due to aging, changes in the ambient temperature, spurious modes, and to changes in the drive level. Aging, here, is used in its broadest sense to include practically any nonreversible change in the crystal characteristic from whatever cause. The principal causes and effects of aging are discussed in paragraphs 1-172 through 1-181.

EFFECT OF TEMPERATURE UPON CRYSTAL PARAMETERS

1-252. The temperature-frequency characteristics of quartz plates are covered in the description of the various elements, and will not be repeated here except to note that a change in the frequency means a change in the LC product of the motional arm. To what extent the frequency drift may be due to a change in L and to what extent to a change in C would require very precise measurements of f_r and f_p, and the approximate formula in paragraph 1-225 for computing C from the measured parameters would need to be replaced by a more rigorous equation. Although of theoretical value, such small changes in L or C are not, in themselves, of practical importance in circuit design—rather it is the change in the LC product (i.e., in the frequency) which is important, and which must be kept to a minimum. Low-temperature-coefficient crystals have been developed for this purpose, but only the GT, at low frequencies, and the AT, to a lesser extent, at high frequencies provide a near-zero coefficient over a wide temperature range. The more exacting the requirements, the more expensive the crystal unit will be. Fortunately, zero coefficients can be obtained at different temperatures by slight variations in the orientation angle of the cut. By mounting the crystal in an oven thermostatically controlled near the zero-coefficient point of the crystal, the temperature effected frequency deviation can be kept very small. Indeed, an ideal oven having a zero temperature fluctuation would permit any type of quartz cut to be stable provided the drive level remained constant. Nevertheless, the use of an oven is to be avoided where possible, because of the additional cost, space, weight, and power requirements, and also because the crystal pins of the oven increase the shunt capacitance across the crystal. The additional shunt capacitance proves increasingly objectionable at the higher frequencies, and makes it necessary that either the oven dimensions be as small as possible or that the entire oscillator be mounted within the oven. Either requirement serves to reduce the stability of the oven temperature, particularly if the ambient temperature varies between wide extremes. For ovens of practical size and construction, some frequency deviation is to be expected as a result of temperature changes. If this deviation is to be kept to an absolute minimum, precise temperature-coefficient characteristics must be specified in selecting a crystal unit, or a greater precision in temperature control than is now attainable in the average crystal oven must be sought. An ingenious method of obtaining practically a zero temperature coefficient for A elements over a span of 20°C and more is being developed by the Hunt Corporation. During a luncheon conversation several years ago between E. K. Morse, S. Ryesky, and D. Neidig (the former, a Government representative, the latter two of Hunt) concerning the possibilities of improving the frequency stability of Radio Set AN/ARC-1 in its first modification, the idea originated of operating two temperature-compensating equal-frequency A elements in series. The angles of cutting could be so selected as to provide equal temperature coefficients of opposing polarities which would cancel when both crystals were operating at the same temperature. However, little was attempted in this field until recently. Experimental models show that over room-temperature ranges the frequency deviation can be quite negligible. Aging data is still insufficient, but over a period of six months a stability of about one-half part per million has been achieved, with three-fourths of the drift occurring in the first three months. Probably the most significant recent activity in the development of fabrication processes designed to stabilize the crystal parameters against changes in temperature centers around the current investigations under the direction of Dr. E. A. Gerber of the Signal Corps Engineering Laboratories. As reported by Mr. D. L. Hammond in a modest paper, *Effects of Impurities on the Resonator and Lat-*

tice Properties of Quartz, presented at the 1955 Signal Corps Frequency Control Symposium, a systematic exploration is under way to discover and catalog the effects on the parameters of quartz crystals which have been synthetically grown to include controlled percentages of impurities. Impurity elements being experimented with include aluminum, boron, calcium, germanium, lead, selenium, tin, titanium, and zirconium. This work undoubtedly has revolutionary possibilities. The discoveries already made presage the probability that temperature effects, which are now so important a problem, can in the future be largely eliminated by growing crystals, for particular cuts, with controlled impurities of proper quantities and proportions.

1-253. A crystal operated at an overtone mode will have temperature-frequency characteristics different from those exhibited by the same crystal at its fundamental vibration. For the control of very high frequencies the A element is normally preferred to the B element, because of its stronger piezoelectric effect and because of the smaller frequency deviation possible for large variations in temperature. However, an AT cut ideally oriented for operation at the fundamental mode is not usually ideally oriented for the higher modes. A research team at Philco Corporation investigating the characteristics of harmonic-mode crystals found that by far the greatest change in the temperature-frequency characteristics of A elements occurs at the first operable harmonic jump, i.e., between the fundamental and the third harmonic. (See figure 1-112.) Since the subsequent changes at the higher harmonics are relatively small, a crystal suitably oriented for the fifth harmonic will usually be suitable for operation under the same temperature conditions at all other overtones. The sensitivity of the crystal to slight changes in the orientation angle is acute. Figure 1-113, for example, shows the degree by which the characteristic curve of an 11th-harmonic A element is rotated by successive changes in the orientation angle of only 3 minutes each. If this crystal were to be operated at room temperature, an orientation of approximately 35°27′ would appear to be preferred. For operation under temperature variations of —55° to +70°C, an orientation of 35°30′ permits the minimum total frequency deviation from a room-temperature mean. Finally, if the crystal is to be mounted in an 85° crystal oven, an orientation of 35°33′ would be optimum.

EFFECT OF SPURIOUS MODES UPON CRYSTAL PARAMETERS

1-254. Closely allied with the problem of temperature control is the problem of avoiding spurious modes. Spurious modes are most apt to occur in the case of thickness-shear crystals. Among these elements, the AC and BC cuts provide the purest frequency spectrum, but unless crystals of these types are provided with precise temperature control their larger temperature coefficients prevent their being preferred over A and B elements. Cutting the crystal blank to the proper face dimensions is the most important factor in avoiding unwanted modes, but even when due precautions are taken, sudden apparent variations in the motional-arm parameters of individual crystal units

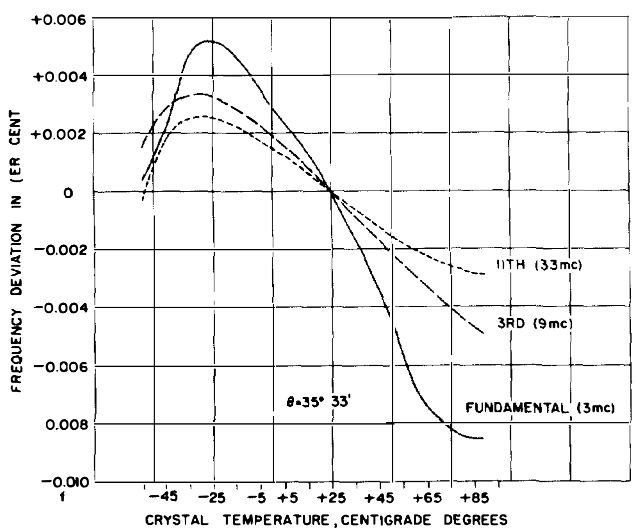


Figure 1-112. Typical variations in frequency-temperature characteristics of A element when operated at different harmonics

are not uncommon. These effects occur most often during variations in temperature, and are due to the fact that the temperature coefficients of nearby modes are quite high. The activity and frequency curves versus temperature of an erratic A element at series resonance in a tuned bridge circuit are shown in figure 1-114. The activity was measured by the grid current. No tuning adjustments were made during the temperature run. Note that the sudden jumps occur at some of the same frequencies, which, at the high-temperature portion of the curve, are apparently of a reasonably pure mode, indicating that the temperature coefficients of the desired and the unwanted modes are different. It can also be seen that the sudden dips in frequency are accompanied by abrupt changes in activity, the latter probably being due to higher motional resistances for the unwanted modes. Unwanted modes are not always accompanied by changes in the resistance. For example, a sudden jump from one frequency to another, but without the dipping effects shown in figure 1-114, where the temperature-frequency curve is effectively broken into two smooth curves, may have very little effect on the activity. This type of frequency jump, which was quite common in the old Y-cut crystals, seems to be due primarily to small defects in the finishing of the crystal blank. Where only one such jump occurs during the temperature cycle, it can usually be eliminated by a slight retuning of the oscillator circuit. However, retuning the oscillator circuit, particularly if the crystal is

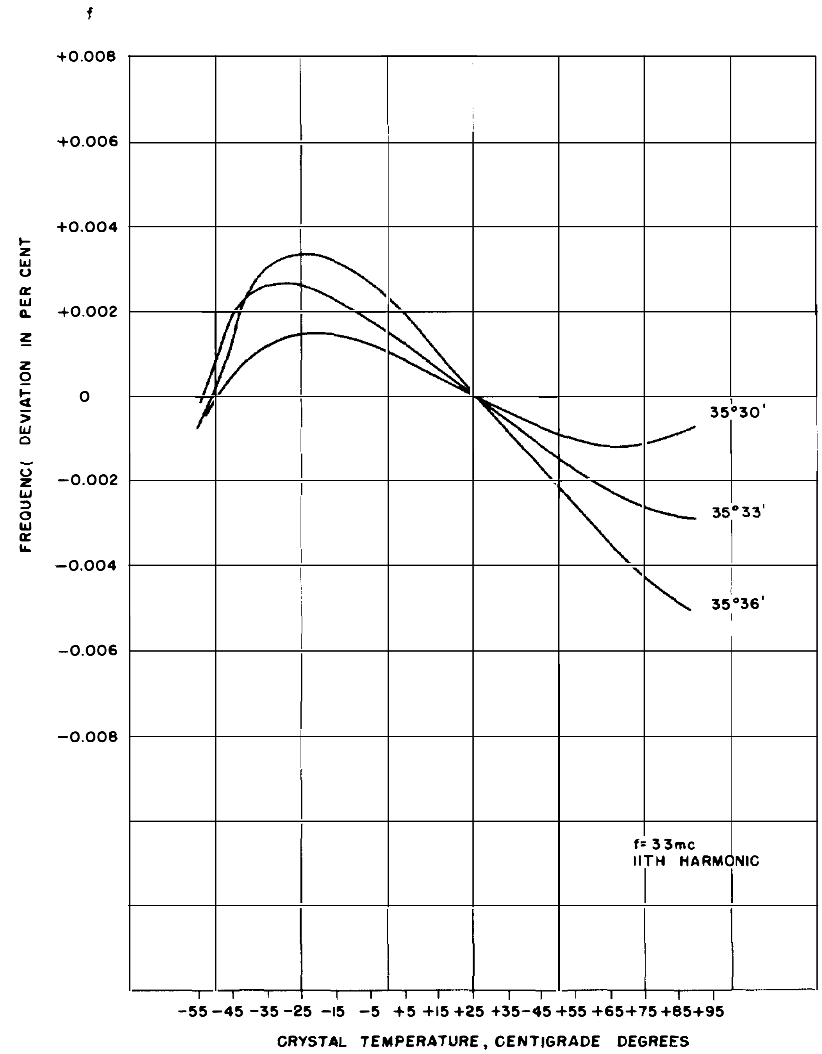


Figure 1-113. Large clockwise angle of rotation of frequency-temperature curve of harmonic-mode A element caused by small increments (3 minutes of arc) in the cutting orientation angle about the X axis

being operated near series resonance, will have little effect upon those temperature-frequency characteristics due to unwanted modes that are inherent functions of the major dimensions of the crystal blank. A crystal unit having characteristics similar to those shown in figure 1-114 should not be used where the operating temperature is expected to extend into the erratic region. Unfortunately, the specifications for most of the crystal units listed in Section II of this handbook are not rigorous enough to provide a guarantee against unwanted modes for every type of unit, if the effects upon the frequency and the effective resistance do not cause over-all deviations beyond the maximum allowed. On the other hand, "jumpy" crystals are the exception rather than the rule, but if particular precautions are necessary where wide temperature variations are to be encountered,

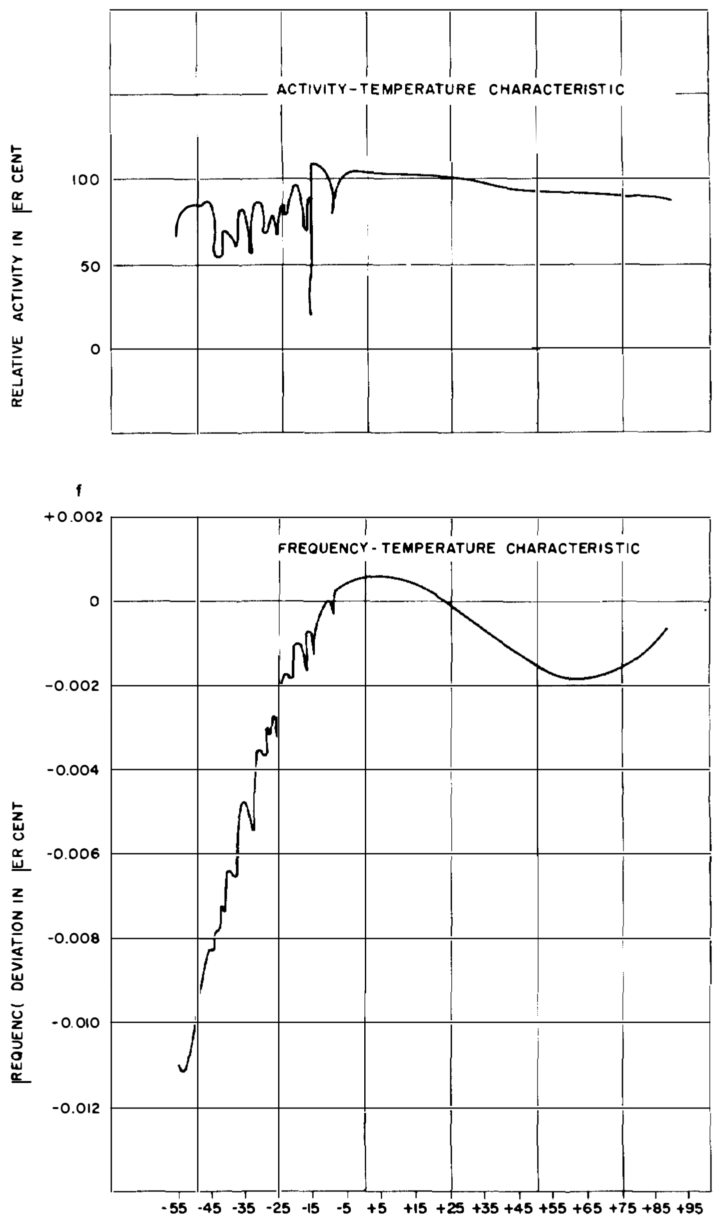


Figure 1-114. Activity-temperature and frequency-temperature characteristics of harmonic-mode A element, showing effects of unwanted modes

only those crystal units should be used which are specified by Military Standards to be free of unwanted modes over the desired temperature range.

1-255. The overtone modes of the thickness-shear elements are more likely to be troubled with spurious frequency dips of the type shown in figure 1-114 than are the fundamental modes, but a crystal that is erratic at its fundamental vibration usually exhibits a pure frequency spectrum at a high harmonic. Indeed, because the frequencies of the unwanted and the desired harmonics do not increase in the same proportion, one method of lessening the probability of interfering modes at the higher harmonics is to deliberately cut the crystal with edge dimensions which favor spurious responses at the fundamental frequency. Nevertheless, the overtone crystals have a tendency to oscillate at two or more thickness-shear frequencies. Usually, this seems to be due to slight differences in the thickness of the crystal from one point to another. For each order of the harmonic, n, the crystal can be imagined to be divided into n layers perpendicular to the thickness, with each layer being a separate crystal vibrating 180° out of phase with the neighboring layers on each side. If n is an even number, the separate sections tend to cancel each other's electrical effects at resonance. For this reason the even harmonics cannot easily be electrically excited. In the case of the odd harmonics, there is always effectively one vibrating layer whose alternating polarity is not neutralized. Most of the activity is more or less centered in one particular region of the crystal plate. If the thickness at an active point differs slightly from the thickness at a neighboring point, there will be a tendency to jump from one activity center to another, and small jumps in the frequency can result. In the case of crystal plates vibrating at high harmonic modes, a small variation in the thickness dimension is generally more likely to produce a sudden frequency jump than if the same crystal were vibrating at its fundamental mode. If there is little difference between the equilibrium conditions of two vibrating stages, the frequency may shift back and forth at an audio rate, thereby effectively modulating the oscillator output with an audio frequency. Such frequency jumps are best avoided by the use of ceramic-button holders, the design of which concentrates the excitation in a small area at the center of the crystal where the most uniform thickness is attainable. Occasionally, it is found that the small frequency jump occurs only at a particular adjustment of the oscillator, and therefore it can be avoided by slight changes in the oscillator tuning. Even so, unless the temperature is to remain reasonably constant, a crystal unit exhibiting any tuning jump at all should not be used. For although an unwanted mode that occurs during a temperature cycle may never appear during a tuning adjustment, the reverse situation is rarely found—a frequency jump that can be caused by a tuning adjustment is almost certain to appear during a temperature cycle.

EFFECT OF DRIVE LEVEL UPON CRYSTAL PARAMETERS

1-256. There is insufficient data and standardization at the present time to analyze or to predict exactly the effect a change in the drive level will have on a crystal unit of a given type. Not only do crystal units of the same type exhibit various reactions, depending on the nominal frequency, the method of fabrication, and the manufacturer's specifications, but even when all these factors are the same for a sample of crystal units, the individual reactions to changes in drive level are unpredictable. The frequency and series-arm resistance curves versus drive level in figures 1-115 and 1-116 are shown as examples. These curves were prepared from data obtained during a Signal Corps research project at New York University by a research team consisting of Messrs. Don J. R. Stock (Director), L. Silver, E. Strongin, A Yevlove, and A. Abajian. The curves in both figures were made from the same set of 9-mc crystal units —AT-cut, electrode-plated, wire-mounted types CR-18/U and CR-19/U, all made by the same manufacturer.

FREQUENCY VERSUS DRIVE

1-257. In figure 1-115, note that although the frequency of the average crystal unit tends to increase with drive level, this effect is by no means to be found at all drive levels for all crystal units. Unfortunately, the temperature-frequency curves for these same crystals are not available, so it is not possible to judge how much of the frequency deviation is due simply to the rise in temperature with drive level. However, the frequencies of other A elements have been tested for frequency deviation versus power, and even though the increases of temperature due to drive occur at points of negative slope on the frequency-temperature curve, the actual frequency-drive level curve generally reveals a positive slope. This increase in frequency with drive is apparently due to a relatively large temperature-gradient coefficient. The net effect on the frequency is due to the combined influences of the changes in both the temperature and the temperature gradient, which influences

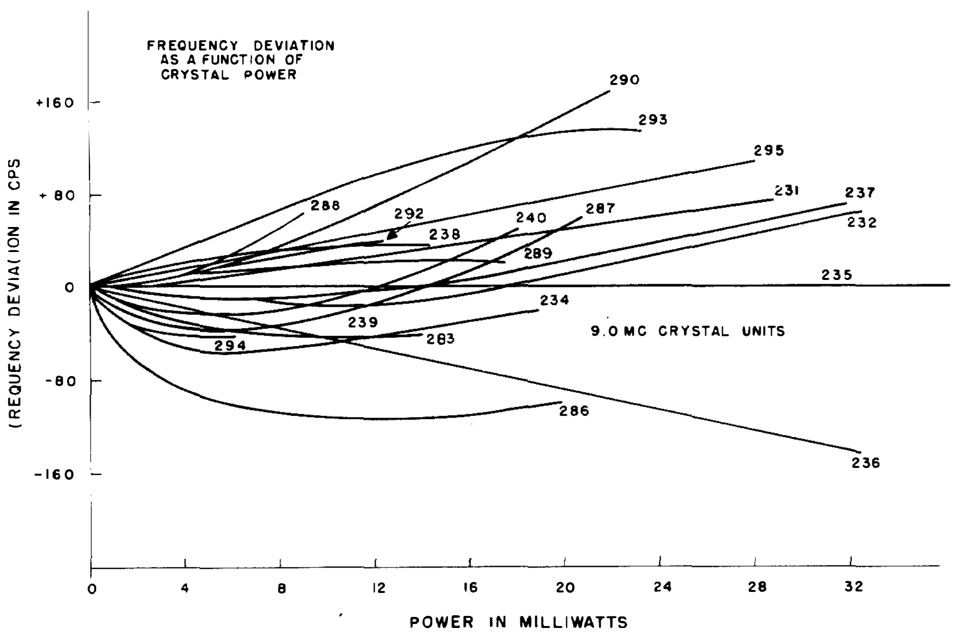


Figure 1-115. Frequency deviation versus drive for a random sample of 9-mc A elements. All crystal units are the products of the same manufacturer and are similarly fabricated and mounted in HC-6/U holders

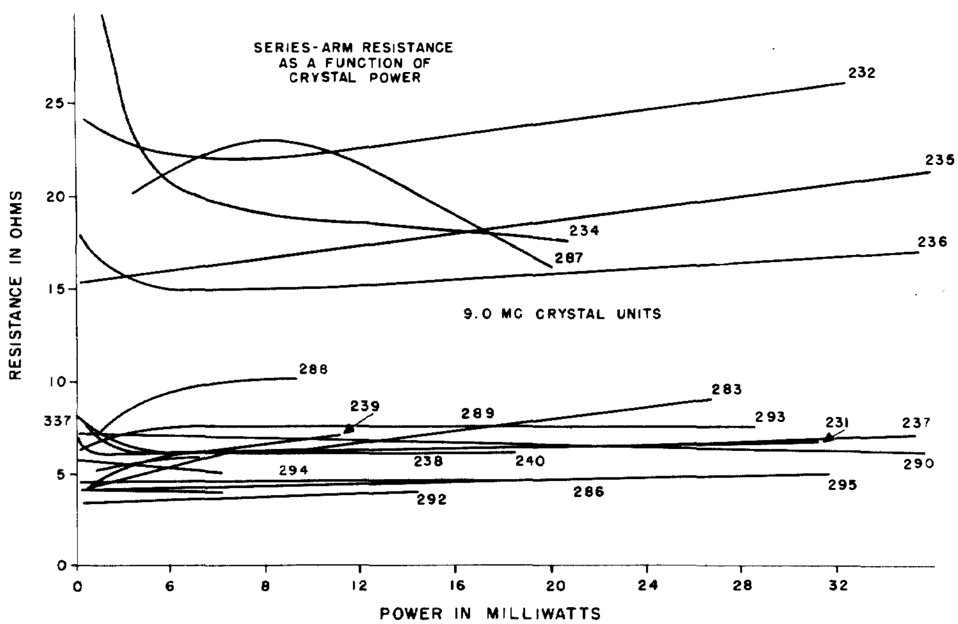


Figure 1-116. Resistance deviation versus drive for a random sample of 9-mc A elements. Curves are for the same sample of crystal units whose frequency-drive characteristics are shown in figure 1-115. Correspondingly numbered curves are those of the same crystal unit

may or may not be in opposition. From the appearance of the curves in figure 1-115, it is possible that those curves starting with a negative slope may be primarily responding according to the normal temperature coefficient. No data is available concerning the degree by which the orientation of the crystal plate relative to the mounting wires might influence the thermal-gradient effect.

1-258. The rise in temperature per milliwatt of drive varies widely with the types of mounting used and the sizes of the crystal plates. For wire-mounted units, most of the heat generated is due to friction at the points where the crystal is supported. With the heat source thus concentrated in a small region of the crystal surface, steep thermal gradients can be expected. The over-all rise in temperature is also greater in the case of wire-mounted units, since most of the thermal-leakage must be through the air, which, like all gasses, acts as a thermal insulator. If the crystal unit is vacuum-sealed, the temperature change per milliwatt may increase by a factor of from two to ten, depending upon the size of the supporting wires and how much of the crystal surface is metal-plated. With the air evacuated, the heat leakage is primarily through the supports and by radiation. The amount lost by radiation depends largely upon the emissivity of the crystal surface, which is approximately 40 times as great for unplated as for plated areas. If it can be assumed that the heat is evenly distributed over the volume of the crystal, the temperature rise of a one-centimeter-square crystal wire-mounted in an HC-6/U holder (not evacuated) can be expected to be approximately 0.3 to 0.4 centigrade degrees per milliwatt of drive. In practice, however, the temperature of the parts of the crystal where most of the heat is generated may increase as much as 10 times this amount. If high or variable drive levels are to be used, pressure-mounted crystal units should be employed. The relatively large contact area between the crystal and the supporting electrodes permits a more uniform distribution of the heat, thereby reducing the magnitudes of the thermal gradients. The pressure mounts also provide a much higher thermal conductivity away from the crystal, thus enabling a much smaller temperature rise per milliwatt of drive. Finally, the pressure mount provides better mechanical and aging protection for the crystal when operated at high-amplitude vibrations. Regardless of the type of mounting, it is never desirable from the point of view of stability or of long crystal life to use a higher crystal drive than absolutely necessary.

1-259. In tests made with GT-cut crystals, where

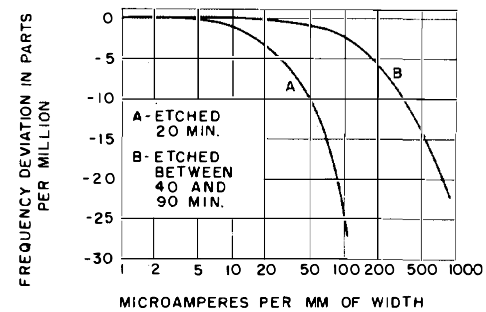


Figure 1-117. Frequency deviation versus crystal current density for two GT-cut crystals which were subjected to different periods of etching*

the frequency deviation with temperature is practically zero over a 100-degree centigrade range, the opportunity has been afforded to study the frequency deviation due to drive alone without the complication of temperature-coefficient effects. Experiments with G elements, as reported by A. R. D'Heedene, reveal a negative frequency deviation with drive, as shown in figure 1-117. Note that the GT plate given a deep etch maintained its stability during much higher drive levels than did the plate etched only 20 minutes. Since the effective resistance of the better finished crystal can be expected to be less than, and to be more stable with increasing drive than, the resistance of crystal A, the changes in frequency with changes in crystal power may have been much closer than the curves in figure 1-117 indicate if it was assumed that the resistances of the two crystals were equal. If the change in frequency is due primarily to changes in the thermal gradients, it is more directly a function of the crystal power. On the other hand, if the frequency deviation is due primarily to mechanical strains resulting from high amplitudes of vibrations, it is more directly a function of the crystal current. Although the evidence now suggests that it is the thermal gradients that are the primary factors, certainly a lowering of the frequency can be expected for any mode if the elastic limit is approached too closely. After crystal units are subjected to high amplitudes of vibration, they do not return immediately to their original frequencies when the drive is reduced to a low level. A period of days or weeks may ensue before the crystal unit regains its former characteristics, during which time the performance resembles that of a crystal rapidly aging.

1-260. Although the frequency-versus-drive characteristics of individual crystal units deviate considerably from the norm, the characteristics are generally similar enough to plot reasonably dependable average curves when the fabrication processes and the frequencies are the same. Such curves showing average frequency deviation versus power are illustrated in figure 1-118. Each curve represents the average of several samples from a representative manufacturer for a given frequency. The curves with the same letter correspond to crystal units of the same manufacturer. All the crystals are A elements, metal-plated and wire-mounted in HC-6/U holders. In every case, it can be seen that the average tendency is for the frequency to increase with power.

RESISTANCE VERSUS DRIVE

1-261. The resistance curves shown in figure 1-116 are more or less typical of the wide variations that must be considered in the design of an oscillator. A minimum performance level must be maintained regardless of the resistance of the crystal unit, as long as the resistance complies with the military specifications. Actually, the average series-arm resistance of the crystal units shown is quite low for 9-mc crystal units. As would be expected, the resistance generally increases as the amplitude of vibration increases. About one crystal unit in eight, however, exhibits a steady decrease in resistance as the drive increases. The initial resistance of such a unit is usually higher than the average. Note in figure 1-116 that a number of the curves have relatively sharp negative slopes at very low power levels. This characteristic is not uncommon, particularly in the case of harmonic-mode elements, where it has become a problem requiring special test procedures. Harmonic-mode

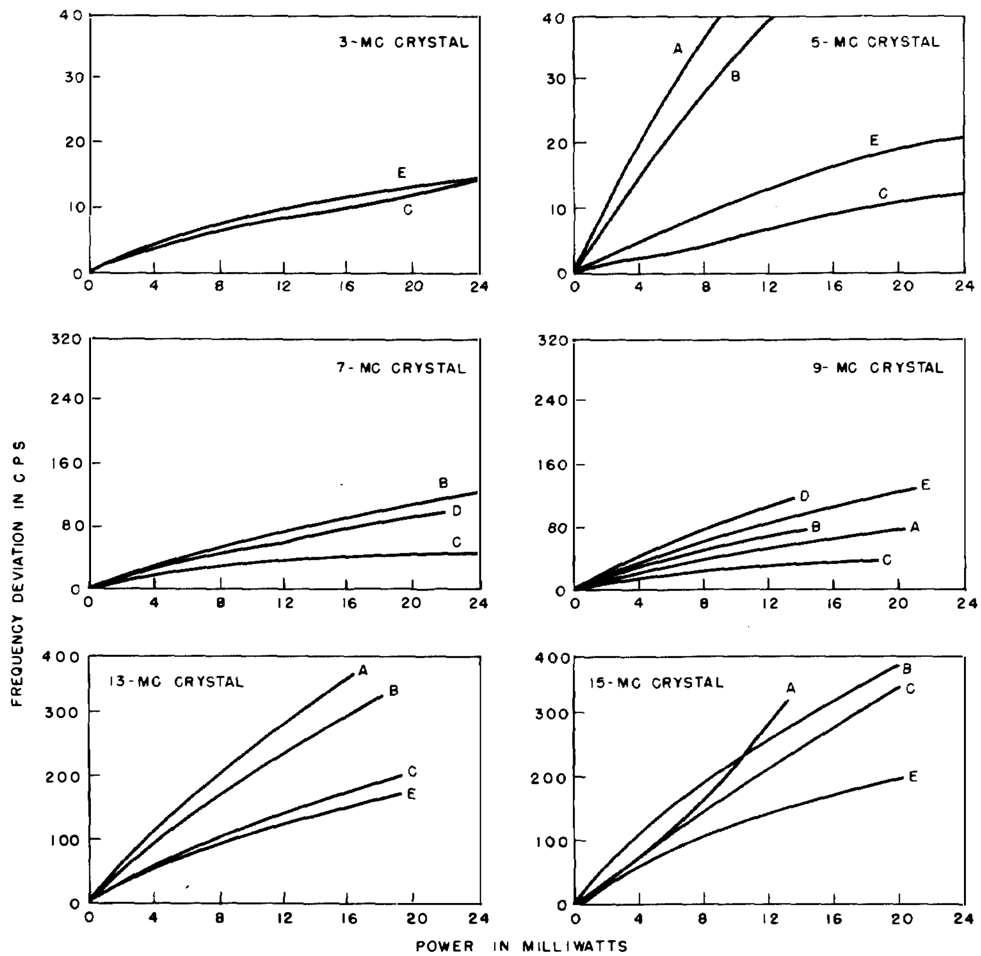


Figure 1-118. Average frequency deviation versus drive. Each curve represents the average of a random sample of several similarly constructed units of one manufacturer. Curves having the same letter represent the characteristics of crystal units of the same manufacturer

crystal units are now required to pass performance tests at two drive levels. The first is at the normal maximum recommended drive level; the second is to ensure that the resistance falls within specifications when the drive is at a minimum. In the case of fundamental thickness-shear elements, sharp negative slopes of the resistance-drive curves at low drive levels are not as common an occurrence percentage-wise as is suggested by the 9-mc samples in figure 1-116. Much more likely to be found are resistance curves with the slopes slightly more positive at very low drive levels.

1-262. In the design of an oscillator for military equipment a principal consideration is to ensure that the crystal drive does not exceed the recommended maximum when one crystal unit of the same standard type, but of perhaps a greatly different resistance, replaces another. If the drive is not kept to the lowest practicable level, the resistance of a borderline crystal may well be increased beyond the permissible limits, thereby excessively degrading the oscillator stability.

CRYSTAL OSCILLATORS

For a comprehensive cross-index of crystal-oscillator subjects, see end of Section I.

FUNDAMENTAL PRINCIPLES OF OSCILLATORS

1-263. An oscillator can be defined as any physical system having a periodic motion. If the motion is plotted as a function of time, a wave shape or a sequence of wave shapes that fairly accurately repeats itself would be considered the fundamental cycle of a stable oscillator. On the other hand, if there were a continuous change in the wave shape, the oscillator would be classified as being unstable. An oscillator that is unstable in the general sense, may, however, have a stable component of frequency, or amplitude, or some combination thereof. Of course, all oscillating systems are unstable to some degree, so that the terms *stable* and *unstable* define classifications that are somewhat arbitrary though none the less convenient.

1-264. Oscillators may also be classified according to the way in which the oscillations are controlled. A number of classifications are possible, but of those which consider the oscillating system alone, there are three general types: free, forced, and forced-free. Free oscillators are those whose oscillating energy is entirely self-contained in the oscillating state, and whose waveform and frequency are determined entirely by the properties of the system. The solar system, purely from the point of view of the planetary motions, is an example. A quartz crystal vibrating freely in space is another. Forced oscillators are those in which the energy, wave shape, and frequency are under the control of an external power source. An example would be the cone of a loudspeaker, or a quartz filter crystal, where the vibrations are controlled by the signal source. "Forced-free" oscillators can be described as those which are driven by the energy of an external source, but where the frequency is primarily determined by the properties of the system. Crystal oscillator circuits are of the forced-free type. Again, the classification is somewhat arbitrary, for in the final analysis there are no absolutely free nor absolutely forced oscillations, nor can two systems be rigorously considered as distinct when there is an exchange of energy between them. In fact, fourth and fifth categories are possible. In the one, the frequency control and drive are both inherent in the system, yet not in the same sense as that defined for free oscillators. By a stretch of the imagination, a good example is to be found in the hula dancer. In a fifth category, the energy is supplied by the oscillating system, but the frequency is controlled externally. An example is to be found by considering each limb of the hula dancer as a separate oscillating system. Still other categories are possible. Insofar as a crystal oscillator circuit is concerned, as distinct from its power source, we can consider it an independent controlling system in respect to the frequency, but only to the extent that the circuit can predetermine the periodic characteristics of both the input and the output energies.

FUNDAMENTAL REQUIREMENTS OF STABLE FORCED-FREE OSCILLATIONS

1-265. There are two fundamental conditions that are always met when a physical system is being maintained in a *stable* state of forced-free oscillation. First, the primary source of energy, or "prime mover," is supplying energy at the same average rate at which energy is being expended by the system. Second, all forces acting on the oscillations are, themselves, stable periodic functions which have frequencies relative to the frequency of the oscillator that can be expressed by rational numbers (e.g., $\frac{f_F}{f_o} = 1, \frac{1}{2}, \frac{8}{7}$, etc). In the practical case, this latter property cannot occur simply by coincidence between independent systems, so that the condition implies that the periods of all forces acting on the stable oscillations are controlled by the oscillator, itself. The most important of these forces are those exerted by the power source in driving the oscillator and those exerted by the output as a result of reaction with the load. The first condition for stable oscillations ensures that the average amplitude of oscillation is stable. The second condition ensures that the fundamental frequency is stable. Together, they ensure that the waveform is stable.

1-266. As a simple example of a forced-free oscillator, consider a system consisting of a swinging pendulum. If the frictional losses per cycle are small relative to the energy stored in the system, and if the amplitude is small, the oscillations are essentially those of simple harmonic motion, with the frequency being determined by the geometry of the system and the gravitational field. Assume that the pendulum, each time it reaches a certain point in its cycle, triggers a latch that releases a spurt of energy from a power source. If each kick received imparts the same amount of energy to the system, an equilibrium of stable oscillations will be reached when the input pulses are being completely transformed into a simple harmonic flow of energy to the surroundings. In order for the power source to transmit energy to the system, it must exert its force while the system is moving in the direction of the force. Otherwise, it will be the "power source," rather than the oscillator, that gains energy. If the system dissipates energy at the exact instantaneous rate at which it is received, the applied force will not, itself, produce momentary accelerations in the pendulum's swing each cycle. However, the losses from the system do not occur at simply a single interval during the cycle, but obey a sine-wave function extending over the entire period. Only at the instants of zero kinetic energy, at the end of each swing, can the instantaneous losses be considered zero. Thus, each pulse of energy must accelerate the pendulum in its direction of motion so that the waveform must deviate somewhat from a pure sine shape. The distortion is a minimum when the ratio of the total stored energy to the energy input per cycle is a maximum, and when, *during the input interval,* the ratio of the energy dissipated to the energy absorbed is a maximum. In other words, the distortion becomes smaller the higher the "Q" of the pendulum, the longer the interval over which the impulse is spread, and the more the impulse is centered at the middle of the swing where the instantaneous power dissipation is the greatest. Now, even though the impulses distort the waveform, the oscillations are stable if exactly the same pattern is repeated periodically. If instead of swinging back and forth, the pendulum is swinging through a complete circle, it is easier to see that if several impulses are transmitted to the system each cycle, the same pattern will continue to be repeated as long as the frequency, or frequencies, of the impulses are related to the frequency of the pendulum by a ratio of whole numbers. The fundamental of the pendulum cycle need not equal the fundamental of the *stable* wave, but it must be a harmonic thereof. For example, suppose that a pulse of energy is imparted to the system only once every four cycles. Then the fundamental period of the stable waveform is four times the period of the pendulum. If five impulses are delivered for each four cycles of the system before being repeated in the same phase as before, the period of the stable waveform will again equal four pendulum cycles. Only when there are one, two, three, etc, impulses repeated each cycle does the period of the stable wave equal the natural period of the oscillator. In the same way by which the forces exerted by the energy sources distort the sine-waveform, so do the forces exerted by the load into which the pendulum loses its energy. If the impedance during any interval of the swing changes from cycle to cycle without repetition, a stable waveform is not obtainable.

APPLICATION OF FUNDAMENTAL OSCILLATOR PRINCIPLES IN THE DESIGN OF ELECTRONIC OSCILLATORS

1-267. In the application of electronic oscillators, it is not usually a stable over-all waveform that is the first requirement, but a stable fundamental frequency. Practically, however, these two effects are not independent, and the generation of the one involves the generation of the other. The deviation from a pure sine wave in the a-c output of a stable crystal oscillator will be entirely due to the presence of harmonics of the fundamental. Such fractional components as $\frac{5}{4}$ ths of the fundamental or the harmonics thereof do not appear. With the load impedance constant, the conditions of stability are reached when energy is being supplied at the average rate of dissipation and at the same phase interval of each cycle. In the generalized crystal oscillator circuit of figure 1-108 (B), the first condition is met when ρ_s is equal to $-R_e$. The second condition is met when X_1 is equal to $-X_e$. In application, the two conditions are not independent of each other, for the build up in the energy of oscillation depends upon the power source not exerting its force in phase opposition to the oscillations. Indeed a common approach to the analysis of an oscillator circuit is to establish a single equation that expresses simultaneously the equilibrium requirements of both the rate and the phase of the energy supply. Equation 1—289 (1) for the Pierce and Miller circuits is an example. Since this type of equation, when fully developed, usually becomes quite cumbersome, such an approach is only occasionally followed in this manual, it being more convenient to treat the two basic equilibrium conditions separately. The first condition, that the rates of energy supply and of energy dissipation be equal, can be assumed to be satisfied if the magnitude of the rms voltage between any two points in the circuit is constant. This condition can be expressed by an equation which equates the *loop gain* to unity. By this we mean that, starting with the input circuit, or at any convenient point, the overall voltage gain around the oscillator loop back to the starting point is unity at equilibrium. If the loop gain is greater than unity, oscillations build up, if less than unity, they do not start, or, if already started, they die down. As an example, the loop-gain equation of a simple oscillator of the type shown in figure 1-177(D), where the feedback energy is transformer-coupled from the plate circuit to the grid circuit, can be expressed as follows:

$$G_1G_2G_3 = \frac{E_p}{E_g} \cdot \frac{E_s}{E_p} \cdot \frac{E_g}{E_s} = 1 \qquad 1\text{—}267\ (1)$$

where G_1 is the gain of the vacuum tube, G_2 is the gain of the transformer in the plate circuit, and G_3 is the gain of the feedback from the transformer secondary to the vacuum-tube input. Although the loop-gain equation may at first glance appear trivial since the product of the voltage ratios equals unity regardless of what voltage values are assigned, it should be remembered that a necessary qualitative implication requires that each voltage ratio represent the gain of an actual transfer of energy from one circuit to another. When the voltage ratios are expressed in terms of the circuit parameters an overall network formula is established that will serve to discipline the oscillator design. In a similar manner, the second condition of equilibrium can be expressed as an equation of the *loop phase rotation,* in which the total phase shift in the voltage around an oscillator loop is equal to zero, or to some integral multiple of 360 degrees. Continuing the example of the transformer-coupled oscillator above:

$$\theta_{pg} + \theta_{sp} + \theta_{gs} = 0 \text{ (or } 360^\circ) \qquad 1\text{—}267\ (2)$$

where θ_{pg} is the phase of E_p with respect to E_g, θ_{sp} is the phase of E_s with respect to E_p, and θ_{gs} is the phase of E_g with respect to E_s. In most cases approximately ideal conditions can be assumed so that the loop phase requirements need only be analyzed qualitatively. For instance, in the example given of the simple transformer-coupled oscillator, let it be imagined that the plate circuit is to be designed so that the vacuum tube faces a purely resistive load. Thus, θ_{pg} will represent the 180-degree phase reversal introduced by the vacuum tube, θ_{sp} will represent a counter 180-degree reversal by the plate transformer, so that the principal phase consideration is to design a grid circuit that will allow E_g to be in phase with E_s at the desired frequency. The loop-phase considerations are much more involved in the case of the conventional one-tube resonator circuit shown in figure 1-119. The loop phase rotation in this general type of circuit applies to such oscillators as the tuned-grid-tuned-plate, the Hartley, the Colpitts, the Pierce, and the Miller. It is discussed in detail in following paragraphs. The loop equations are the guides by which the design engineer approaches the basic oscillator problems of obtaining the desired amplitude, the desired frequency, the desired amplitude stability, and the desired frequency stability. As a general rule, these four fundamental design considerations are handled with the aid of the loop equations in the following ways:

a. An oscillator is designed to provide a certain amplitude of oscillation by ensuring that the parameters that vary with the amplitude (usually the vacuum-tube parameters) reach their limiting values, as defined by the loop-gain equation, when the desired amplitude is reached.

b. An oscillator is designed to oscillate at a given frequency by ensuring that the loop-phase equation holds at, and usually, only at, the desired frequency. Should the loop-phase equation also have a solution at some other frequency (e.g. the loop phase of the transformer-coupled oscillator mentioned above may well equal zero at more than one mode of the crystal's vibration), the design must ensure that the loop gain is less than unity at the unwanted frequency.

c. The amplitude stability is improved by counteracting or minimizing variations in those circuit parameters which, as indicated by the loop-gain equation, are most likely to cause changes in the amplitude.

d. The frequency stability is improved by counteracting or minimizing variations in those circuit parameters which, as indicated by the loop-phase equation, are most likely to cause changes in the frequency.

Before proceeding to a discussion of the particular types of oscillators, let us first examine in detail the phase relations of the conventional resonator circuit of figure 1-119(B). If a firm qualitative understanding of the operation of this type of circuit is had, the reader should be greatly aided in interpreting the physical meaning of equations later to be derived.

PHASE ROTATION IN VACUUM-TUBE OSCILLATORS

1-268. The conventional equivalent circuit of a vacuum-tube amplifier is shown in figure 1-119 (A). The equivalent generator voltage is equal to $-\mu E_g$, where μ is the amplification factor of the tube, and E_g is the excitation voltage on the grid. R_p is the plate resistance of the tube, and Z_L is the a-c load impedance. The minus sign of the generator voltage indicates a 180-degree phase difference between the equivalent emf and E_g. For oscillations to build up, energy must be fed back in the proper phase from the plate circuit, or from some circuit in a following stage. The control grid of the vacuum tube is effectively an escapement device for controlling the release of energy from the power source. Since this energy must be released each cycle so as not tô be in phase opposition to the oscillator, the grid voltage alternations must be "timed" by the activity in the rest of the circuit. This means that a sufficient and properly phased part of the energy released by the action of the grid must be fed back from the plate circuit each cycle, or from some other circuit of a following stage, in order to continue the periodic release of energy. The initial rush of plate current is to be sufficient to shock the circuit into oscillation, and the initial alternating voltage fed back to the grid circuit must be sufficient for the vacuum tube to generate more a-c energy than is lost during the first cycle. R_p increases with the amplitude of oscillations until equilibrium is reached.

1-269. The phase relation between the grid and plate voltages of an oscillator vacuum tube at equilibrium is the same as that which would occur if the grid were excited at the same frequency from an external a-c source and the tube were connected as a conventional amplifier, operating into the same equivalent load impedance it faces as an oscillator. However, only a certain impedance relationship among the components of a particular oscillator circuit can provide a feed-back producing the proper input phase. It is this necessary impedance relationship that determines the frequency. In the usual single-tube oscillator, the equivalent vacuum-tube generator, of voltage $-\mu E_g$, must drive a plate-coupled feed-back circuit that causes the voltage appearing across the grid to be rotated 180 degrees ahead of or behind the generator emf. The simplest method of reversing the phase is by transformer coupling. On the other hand, if two tubes are used, the reversal can be accomplished by the second tube alone. Either of these methods can

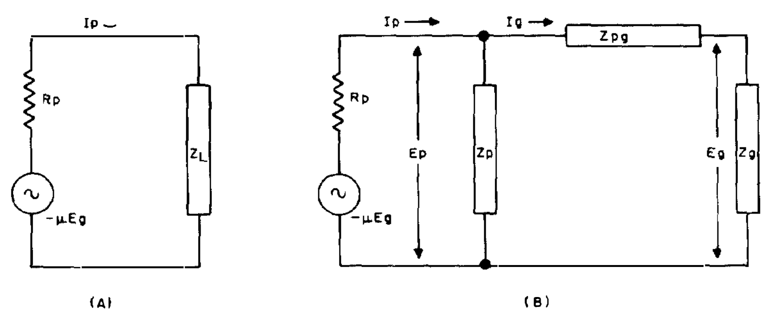


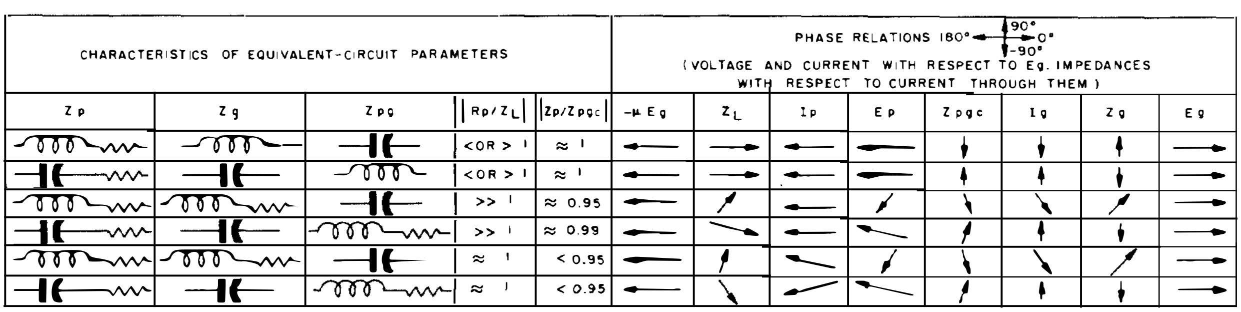
Figure 1-119. (A) Equivalent circuit of vacuum-tube amplifier. (B) Equivalent circuit of crystal oscillators of the Pierce and Miller types

enable a crystal oscillator to work into a more or less resistive load, so that fluctuations in the circuit parameters can have little effect on the feed-back phase, and, hence, upon the frequency. In the conventional parallel-resonant circuits, such as are illustrated in figure 1-109, the phase is rotated as shown in figure 1-119(C). First, assume an ideal case in which the resistive losses in the feed-back arm are zero. In this case, but only in this case, Z_L would need to be resistive. The frequency would be that at which the plate and feed-back arms operate as a parallel-resonant tank. There would be no phase shift in the voltage across Z_L, and E_p would be of the same sign as $-\mu E_g$. Z_p and Z_g, the impedances of the plate circuit from plate to cathode and of the grid circuit from grid to cathode, respectively, are reactive, and must always have the same sign. Z_{pg}, the plate-to-grid impedance is the dominant impedance in the feed-back circuit, and is always opposite in sign to Z_p and Z_g. In the ideal circuit, if Z_p and Z_g are positive, Z_{pg} is negative, so the current, I_g, leads E_p, and therefore $-\mu E_g$, by 90 degrees. If Z_p and Z_g are negative, Z_{pg} is positive, so I_g lags E_p 90 degrees. The voltage across Z_{pg}, of course, would be in phase with E_p in both instances. Since Z_g is opposite in sign to Z_{pg}, E_g thus is opposite in sign to E_p, and the required reversal takes place. Note that I_g is first rotated in phase with respect to E_p; next E_g is rotated in the same direction with respect to I_g.

1-270. In an actual circuit, the feed-back losses cannot be zero, so that a 180-degree reversal cannot be obtained in the conventional feed-back circuit alone. This means that E_p must first be rotated by an amount exactly sufficient to make up the difference. Assume first that R_p is much greater than Z_L, so that I_p can be assumed to be essentially in phase with the equivalent generator voltage. In this case, an inductive Z_L causes E_p to lead the emf, whereas a capacitive Z_L causes E_p to lag the emf. Unless the effective Q of the feed-back circuit is very low, Z_L, must be very nearly resistive, for the shift in the phase of E_p need not be large. In any event, the rotation of E_p must be in the same direction as that of I_g and E_g. For this to occur, the susceptance of Z_p must be greater in magnitude than the susceptance of $(Z_{pg} + Z_g)$. That is, the reactive component of the current through Z_p must more than cancel the reactive component of I_g. The smaller the value of R_p compared with the value of Z_L, the more nearly will Z_L control the phase of I_p, and the more detuned must the parallel circuit become in order to obtain the necessary rotation of E_p. If practically all the resistance in the feed-back arm is between the grid and the cathode, as is normally the case when E_g is developed directly across the crystal unit, E_p must be rotated through a larger angle than otherwise, thereby requiring the parallel circuit to be detuned to a greater degree. This is because E_p must be rotated by an amount effectively equal to the sum of two angles. One of the angles is the difference between the actual phase of I_g relative to E_p and the ideal phase of ± 90 degrees. The second angle is the difference between the actual phase of Z_g and its ideal phase of ± 90 degrees. If all the resistance is contained in the large impedance Z_{pg}, only the phase deficiency of I_g is reflected in the phase of E_p. On the other hand, if all the resistance is effectively contained in the small impedance Z_g, the effect on the phase of E_p by I_g is normally small by comparison with the effect due to the grid-to-cathode resistance. Expressed in polar form:

$$E_g = I_g Z_g\ (\theta_{Ig} + \theta_{Zg}) = I_g Z_g\ (0°)$$

where θ_{Ig} and θ_{Zg}, which must be equal in magnitude



(C)

Figure 1-119. (C) Chart showing ideal and typical phase relations necessary for forced-free oscillations of the circuit shown in (B)

but opposite in sign, are the phase angles of I_g with respect to E_g and of Z_g with respect to I_g. Now, letting $\theta_{Z_{pgc}}$ equal the phase of the total feed-back impedance, $(Z_{pg} + Z_g)$, with respect to the current through it, and θ_{E_p} equal to the phase of E_p with respect to E_g,

we have $\quad \theta_{E_g} = \theta_{I_g} + \theta_{Z_g} = 0$

and $\quad \theta_{E_p} = \theta_{I_g} + \theta_{Z_{pgc}}$

so that $\quad \theta_{E_p} = \theta_{Z_{pgc}} - \theta_{Z_g} \qquad$ 1—270 (1)

Since $\theta_{Z_{pgc}}$ is opposite in sign to θ_{Z_g}, these two phase angles add numerically. If it is assumed that Z_{pgc} is approximately six times the magnitude of Z_g, but that all the feedback-arm resistance is between the grid and cathode, the Q of Z_{pgc} will be approximately five times the Q of Z_g. Under these conditions the rotation of E_p from the ideal value of -180 degrees is approximately 20 per cent greater than the deviation of $-\theta_{Z_g}$ from its ideal value of ± 90 degrees. If Z_g represents the effective impedance of a crystal unit in parallel with the grid-to-cathode capacitance and resistance, the minimum rotation of θ_{E_p} occurs when the effective Q of the crystal and its shunt impedance is a maximum, provided $Z_{pg} >> Z_g$. Similarly, if Z_{pg} represents the Z_e of a crystal unit whose effective resistance is much greater than the equivalent series resistance of the grid-to-cathode impedance, the rotation of θ_{E_p} depends primarily upon the crystal unit Q_e, and is a minimum when Q_e is a maximum.

1-271. In a conventional parallel-resonant crystal oscillator having an ideal feed-back arm, the frequency would be determined entirely by the resonance of the tank circuit, so that fluctuations in R_p, although effective in changing the activity, would not affect the frequency. In practice, the equivalent resistance of the crystal unit is a parameter of the feed-back arm, so that the detuning of the tank becomes very nearly a direct function of the effective Q of the crystal and its shunt capacitance. Note that the phase of E_p with respect to E_g and $-\mu E_g$ is determined entirely by the parameters of the feed-back arm. As long as oscillations continue, variations in R_p or Z_p can only change the phase of E_p indirectly, i.e., by causing a change in the Q of the feed-back arm. Since the effective resistance can vary by as much as a factor of 10 between minimum and maximum values for the same standard type of crystal unit, an oscillator cannot be designed too closely upon the assumption that the load impedance Z_L will be essentially the same either in phase or magnitude when one crystal unit is replaced by another—even if it is of the same type and nominal frequency. This limitation might possibly be minimized by switching the crystal unit from the feed-back circuit to the plate circuit, and replacing it with a high-Q inductor in the feed-back arm. If the grid were operated with bias sufficient to prevent the flow of grid current, a very high and predictable Q could be obtained in the feed-back circuit, and an approximately resistive Z_L could be assumed for the tank regardless of the variations in Z_p due to variations from one crystal unit to the next.

1-272. An oscillator that can be represented by the equivalent circuit shown in figure 1-119 will show the following phase characteristics and related effects.

a. The phase of E_p is entirely determined by the over-all Q of the feed-back circuit and the Q of the grid-to-cathode impedance, Z_g. The phase angle, θ_{E_p}, is given by equation 1—270 (1) as being equal to $(\theta_{Z_{pgc}} - \theta_{Z_g})$. Let us now assume that $(\theta_{Z_{pgc}} - \theta_{Z_g})$ is determined by an imaginary over-all Q of the feed-back circuit. This we define to be

$$Q_f = |\cot \theta_{E_p}| = |\cot(\theta_{Z_{pgc}} - \theta_{Z_g})|$$

$$= \left| \frac{\tan \theta_{Z_{pgc}} \tan \theta_{Z_g} + 1}{\tan \theta_{Z_{pgc}} - \tan \theta_{Z_g}} \right|$$

$$= \frac{Q_{pgc} Q_g - 1}{Q_{pgc} + Q_g} \approx \frac{Q_{pgc} Q_g}{Q_{pgc} + Q_g} \qquad \text{1—272 (1)}$$

It can be seen that if either Q_{pgc}, the actual effective over-all Q of the feed-back circuit, or if Q_g, the Q of the grid-to-cathode impedance, is very large compared with the other, Q_f is approximately equal to the smaller Q.

b. For a given phase difference between E_p and $-\mu E_g$, the ratio of Z_p to the total feed-back impedance, Z_{pgc}, is less than 1 by an amount which increases as the ratio of R_p/Z_L decreases. In other words, the ratio of the r-f current in the plate circuit to the r-f current in the grid circuit increases as R_p decreases.

c. The value of R_p/Z_L is partly a function of the ratio $\frac{Z_{pgc}}{Z_g} = E_p/E_g$. Assume, for example, that $R_p >> Z_L$, so that $E_p/\mu E_g \left(= \frac{Z_L}{R_p + Z_L}\right) \approx \frac{Z_L}{R_p}$. Also assume that during oscillations Z_g is decreased, but that Z_{pg} remains essentially constant. The ratio E_p/E_g is thus increased, and likewise the ratio $E_p/\mu E_g \approx \frac{Z_L}{R_p}$. In other words, as E_g becomes a smaller component of the total voltage across the feed-back circuit, E_p cannot decrease in the same proportion, else each succeeding cycle would be weaker than the one before; so the ratio R_p/Z_L must decrease. Part of the change is due to the increase in Z_L, and part is due to a decrease in R_p. If it is assumed that a large percentage change in Z_g causes only a small percentage change in Z_{pgc}, then Z_L remains essentially constant in magnitude and R_p becomes the principal variable. In any event, as R_p/Z_L decreases, the effective Q of Z_L, as represented by an equivalent resistance and reactance in series, must increase in order to compensate for the increased phase shift of I_p.

1-273. From the qualitative discussion in the foregoing paragraphs it can be seen that in the conventional parallel-resonant crystal oscillators the state of an oscillator in operation is primarily determined by the impedance relations in the feed-back arm. Since the impedance of a crystal unit changes very rapidly with a small change in frequency, a crystal connected in the feed-back circuit makes the oscillator less critical in design than would otherwise be the case. Where maximum stability is required, the vacuum tube will be operated as nearly class B as possible. Under class-A conditions, R_p and μ are approximately given by the d-c, plate-characteristic data of the tube. In the case of power oscillators, amplifier operation will normally be class C, although class B or even class AB may be employed in particular circuits. In these cases, the effective tube parameters cannot be known beforehand, but reasonably accurate approximations can be made and optimum operating conditions can be reached by more or less trial-and-error final adjustments. The operation of conventional oscillators is made less critical, both in starting oscillations and in maintaining a constant amplitude, by the use of gridleak rather than fixed bias. From the point of view of phase rotation, the conductance of the gridleak somewhat decreases the Q of Z_g, and thereby necessitates increased detuning of the tank. Nevertheless, in low-power oscillations the gridleak losses can normally be considered negligible in comparison with the crystal losses. It is as a limiter and stabilizer of the amplitude that the gridleak bias is most important. Any changes in the circuit that tend to change E_g automatically change the bias in such a direction that R_p and g_m of the tube are readily adjusted to new equilibrium values, so that the tendency is one of immediate opposition to the change in so far as the activity is concerned.

TYPES OF CRYSTAL OSCILLATORS

1-274. Crystal oscillators are frequently classified as being either *crystal-controlled* or *crystal-stabilized*. A crystal-controlled oscillator is defined as an oscillator that cannot oscillate if the crystal is removed or is defective. A crystal-stabilized oscillator, on the other hand, operates as a "free-running" oscillator if the crystal is removed. When the crystal unit is properly inserted and the "free-running" frequency is made to approach the normal resonance of the crystal, the mechanical vibrations of the crystal sharply increase. At some point the piezoelectric effect will be sufficient to suddenly "capture" the oscillations and thereby synchronize them at the crystal-circuit frequency. Generally, the crystal-controlled oscillator is preferred, since it is not desirable that oscillations continue if the crystal unit suddenly or gradually becomes defective. Also, crystal-controlled oscillators are normally less critical to design and are less likely to jump suddenly from one frequency to another. The crystal-stabilized oscillator does have the possible advantage of being able to operate successfully with very-high-Q crystal units whose piezoelectric coupling, however, would be too weak for the crystal to build up oscillations from a single initial impulse. A number of oscillator circuits appear to be border-line cases, that can only arbitrarily be classified as crystal-controlled or crystal-stabilized. For example, the CI meter circuit shown in figure 1-106, if connected in the crystal position of S_1, would fail to oscillate if the crystal terminals were open, but not if they were shorted. Most of the oscillator circuits that are discussed in the following paragraphs are classified as crystal-controlled inasmuch as the oscillations do not occur if the crystal units are disconnected.

1-275. A more practical classification from the standpoint of circuit design and of selection of a crystal unit is that of series- and parallel-resonant crystal oscillators. In general, the series-resonant type provides the greater frequency stability and can generate the higher frequencies; whereas the parallel-resonant type is the more economical to construct, can operate over a wider frequency range by the substitution of different crystal units, and can generate the greater power output. There are, nevertheless, a number of exceptions to the general rule.

PARALLEL-RESONANT CRYSTAL OSCILLATORS

1-276. The first quartz oscillators to find general usage as frequency-control devices were of the parallel-resonant type. These oscillators are used primarily with fundamental-mode crystals at frequencies below 20 mc. In the conventional circuits of this type, the crystal must operate between its resonant and antiresonant frequencies, thereby behaving as an inductor. Under these conditions the circuit does not oscillate if the crystal unit becomes defective. Unless a frequency monitor is to be available, these oscillators should be permitted a tolerance of 0.002 per cent or greater, depending primarily upon the tolerance rating of the crystal unit to be used. Under extreme operating conditions, an oscillator error of approximately twice the crystal unit tolerance should be permitted. The conventional circuits are of the Pierce and Miller types, or their modifications. Maximum stability is achieved with low crystal drive and class-A to class-B operation of the tube. Maximum power efficiency is achieved with class-C operation. Since fundamental-mode crystals become too thin and fragile for operation above 20 mc, overtone crystals are necessary at these higher frequencies. However, the shunt capacitance, C_o, and the series-arm L remain the same whether the thickness-shear crystal is operated at the fundamental or the overtone mode, whereas the series-arm C varies inversely with the square of the harmonic. Thus, the capacitance ratio, C_o/C, increases with the square of the harmonic, so that the electromechanical coupling may be too weak to initiate oscillations at normal voltages if the crystal unit is to be operated at parallel resonance. For this reason and also because C_o, as well as the tube capacitances shunting the crystal, have larger susceptances at the higher frequencies, which greatly reduce the operating range of the crystal unit, the parallel-mode circuits are unsuitable for use at the higher frequencies. If the basic circuits are modified to employ series-mode crystals, or are used in conjunction with frequency-multiplying stages, they can provide stable control of very high frequencies. The introduction of a number of multiplier stages with the attendant problems of preventing unwanted frequencies is usually less to be preferred than the direct generation of the end frequency by the use of overtone-mode crystals in series-resonant circuits. Although frequency multiplication involving more than one multiplier stage is still widely used in conjunction with parallel-resonant master oscillators, this usage is found principally in medium- to high-frequency transmitters where various multiplying combinations can provide the maximum number of channels with a minimum number of crystal units. Control by parallel-mode circuits of frequencies above 30 mc is not very common. One example is to be found in Radio Set AN/ARC-1A. In this equipment, a fundamental frequency of 8 mc, for instance, would be doubled in the oscillator plate circuit and increased nine times more in the plate circuit of the following stage. Thus, with only two tubes, a parallel-resonant oscillator can control a frequency of 144 mc and higher. In the analysis of the particular oscillators to follow, the Pierce oscillator has been chosen as something of a reference circuit as well as a point of departure in the discussion of many design considerations to be encountered in crystal oscillators. For this reason, the reader will find the treatment of the Pierce circuit, both qualitatively and mathematically, considerably more detailed than that of the other types of circuits. Space forbids as comprehensive a treatment for the other circuits, but the design problems and methods illustrated in the particular case of the Pierce oscillator are applicable in principle to all oscillators.

The Pierce Oscillator

1-277. The Pierce oscillator is fundamentally a Colpitts oscillator in which the plate-to-grid tank inductance has been replaced by a crystal unit, as shown in figure 1-120. The design of the Pierce oscillator is simpler and less critical than that of any other crystal circuit. As long as R_l and R_g are large compared with the capacitive reactances shunting them, the Pierce circuit will oscillate with crystal units covering a wide band of frequencies. The use of load resistance, R_l, in figure 1-120 (A) aids in maintaining a reasonably flat response over a wide frequency range without the necessity of tuning adjustments other than the switching from one crystal unit to the next. Where a broad frequency range is not required, or where greater activity is necessary, an r-f choke should be used in place of R_l; otherwise, power approximately equal to $I_b^2R_l$ is simply wasted (I_b = average d-c plate current). But even with this economy, the Pierce oscillator cannot be used to generate as large an output as the Miller circuit. The principal reason is to be found in the fact that the impedance Z_{pg} of the equivalent circuit, which in this case is provided by the crystal unit in parallel with the plate-to-grid capacitance of the tube, must be approximately equal to, or greater than, Z_p and Z_g combined. Since the impedance of the crystal unit is fixed by its frequency and rated load capacitance, larger plate impedances are possible if the specified crystal impedance is Z_g instead of Z_{pg}. Thus, for the same crystal current, larger output voltages can be developed across the tank in the Miller than in the Pierce circuit. On the other hand, the effective feed-back phase Q_f is greater if the crystal is not connected between the grid and cathode. This permits the tank to appear more nearly resistive to the tube, so that fluctuations in R_p have less influence upon the frequency. Thus, a Pierce oscillator is generally more frequency stable than a Miller oscillator using the same crystal unit. The typical Pierce circuit employs a triode, although screen-grid tubes are usually to be preferred, because the higher R_p and the negligible plate-to-grid capacitance serve to improve the frequency stability. The oscillator is generally used at frequencies above 200 kc and below 15,000 kc. If an overtone mode is to be excited the oscillator must be made frequency-selective, by replacing R_1 with an inductor, L_1. The inductance of L_1 must be such that the antiresonant frequency of L_1 in parallel with C_1 is lower than the operating frequency of the crystal, in order that Z_p will appear capacitive. The use of the inductor-capacitor combination also reduces the harmonic content of the output waveform. If small L/C ratios are used, the effective plate-to-cathode capacitance will be much greater for the overtones than for the fundamental frequency, so that the former are more readily bypassed to ground than would be the case if no coil were used. In deciding upon the type of oscillator circuit to use, those rule-of-thumb factors most favorable to the selection of a Pierce circuit are:

a. The frequency lies between 200 and 15,000 kc.

b. The permitted frequency error is not less than 0.02 per cent, or 0.015 per cent if a regulated voltage supply is available.

c. The oscillator is to be capable of untuned operation over a wide frequency range, simply by switching from one crystal unit to another.

d. Only a small voltage output is required.

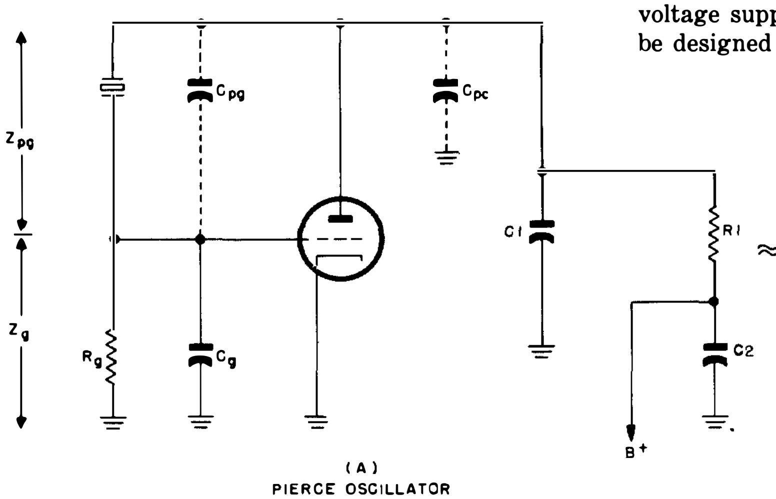
e. The oscillator must be of inexpensive design.

f. The oscillator must not be critical in operation, but able to oscillate readily with relatively large deviations in the parameters of the external circuit.

g. Wave shape is not critical.

h. Same as above, except that the permitted frequency error is 0.01 per cent and thermostatic control of the crystal temperature is feasible.

i. Same as above, except that the permitted frequency error is 0.005 per cent, thermostatic control of the temperature is feasible, a regulated voltage supply is available, and the oscillator can be designed to operate at one frequency only.



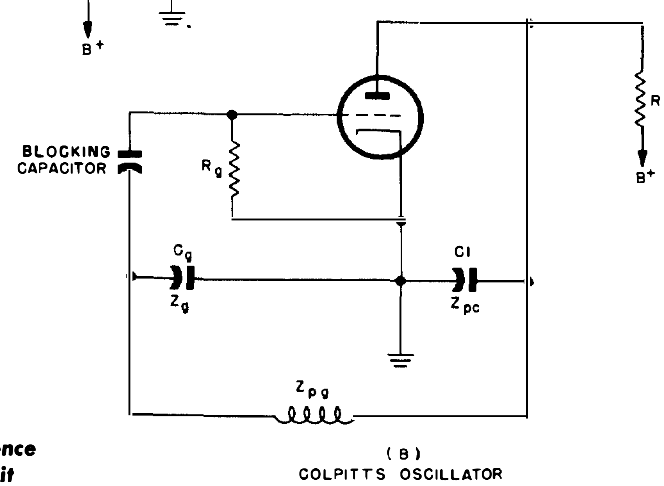


Figure 1-120. Diagrams illustrating the equivalence between the Pierce circuit and the Colpitts circuit

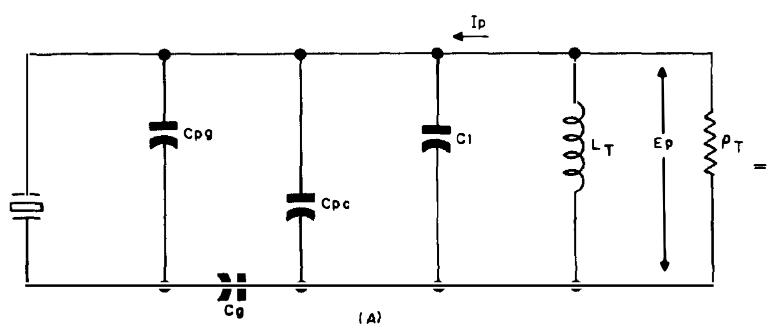
ANALYSIS OF LOAD CAPACITANCE, C_x, IN PIERCE CIRCUIT

1-278. Once that it has been decided to employ a Pierce type oscillator, the standard type of crystal unit is chosen which provides the desired frequency and frequency tolerance and which has been tested according to the Military Standards for parallel-resonance operation. One of the first design considerations is to ensure that the crystal unit will effectively operate into its rated load capacitance, C_x. Such operation is necessary, else there can be no assurance that one crystal unit of the same type can replace another and still fall within the drive-level and effective-resistance specifications. For most parallel-resonance crystal units the value of C_x is 32$\mu\mu$f, although at frequencies under 500 kc, values of 20 $\mu\mu$f are common. In particular instances, still other values of C_x are designated. To a first approximation, referring to figure 1-121 (A), the crystal unit operates into a load capacitance equal to C_{pg} plus the total of C_g in series with the parallel combination of C_l, C_{pc}, and the effective inductive impedance presented by the vacuum tube. Since the Q of the feed-back arm is not infinite, E_p, it will be recalled, must be rotated slightly away from $-\mu E_g$; the direction is such that for a particular frequency $C_{pc} + C_l$ must be slightly larger than would otherwise be the case. Even though the actual equivalent tank circuit is slightly detuned, mathematically the crystal unit is to be in resonance with an effective load capacitance C_x. (See figure 1-108 (D).) The vacuum tube appears to the tank circuit as a negative resistance having a positive reactive component sufficient to cancel the excess susceptance of Z_p. At equilibrium, the tube can be represented by an equivalent inductance, L_T, in parallel with a negative resistance, ρ_T, as in figure 1-121 (A). Note that ρ_T is smaller than ρ of figures 1-121 (C) and (D). This is because ρ_T is not connected directly across the crystal, but faces an impedance, approximately Z_L, that is less than the crystal PI. In figure 1-121 (B), L_T has been replaced by an equivalent negative capacitance, C_n. If C_{pg} can be considered negligible, X_e' and R_e' are equal to X_e and R_e, the equivalent parameters of the crystal unit alone; otherwise, the values of X_e' and R_e' are based upon the assumption that the shunt capacitance, C_o, of the crystal has been increased by an amount equal to C_{pg}. In figure 1-121 (C), C_p ($= C_{pc} + C_l$) and C_n are shown combined into a single plate-to-cathode capacitance, C_p'. In figure 1-121 (D), C_p' and C_g are represented by a single load capacitance, C_x'. If C_{pg} is negligible, C_x' becomes the equivalent load capacitance, C_x, into which the crystal unit operates, and it should be

$$C_p = C_{pc} + C_l \qquad C_x' = \frac{C_p' C_g}{C_p' + C_g}$$

$$C_n = \frac{1}{\omega^2 L_T} \qquad \rho_T = \left(\frac{X_e' + X_{C_g}}{X_e'}\right)^2 \rho$$

$$C_p' = C_p + C_n \qquad -\rho = \frac{(X_e')^2}{R_e'} = \frac{X_e^2}{R_e} = \text{PI}$$



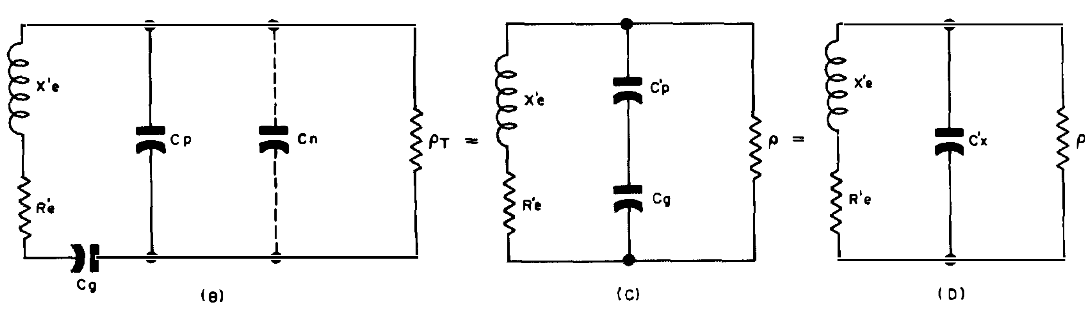


Figure 1-121. Generalized Pierce oscillator. X_e' and R_e' are the effective impedance parameters of the crystal unit based upon the assumption that C_{pg} is a part of C_o. C_n is an equivalent negative capacitance having a positive reactance equal to that of L_T. I_p is the r-f plate current of the vacuum tube.

equal to the value specified for the particular crystal. In any event $(C_x' + C_{pg})$ is the load capacitance that the crystal unit faces, and which should be equal to the rated value, C_x. The value of C_l includes not only the tuning capacitance in the plate circuit, but also the distributed capacitance of the output leads as well as the effective input capacitance of the next tube. C_g is the input capacitance of the oscillator tube. The losses due to R_g, the effective grid-to-cathode resistance, and to R_L, the effective load resistance, have been assumed to be negligible compared with those in the crystal unit. These assumptions can be made without appreciable error in a low-power Pierce circuit that requires only a minimum of loading. From an inspection of figure 1-121 it can be seen that in order for a crystal unit to operate into its rated load capacitance, the design must be such that

$$C_x = C_{pg} + C_x' = C_{pg} + \frac{C_p' \, C_g}{C_p' + C_g} \qquad \text{1—278 (1)}$$

Effect of C_{pg} in Pierce Circuit

1-279. C_{pg} effectively increases both the reactance, X_e, and the resistance, R_e, of the crystal unit. This does not mean that the true effective X_e and R_e are changed, for these are fixed by the fact that X_e must resonate with the rated load capacitance, C_x. Still, insofar as the impedance from plate-to-grid is concerned, X_e and R_e effectively have increased values which make it appear that the crystal shunt capacitance is increased by an amount equal to C_{pg}. This effect is not desirable, since R_e is effectively increased by a greater percentage than X_e. (See equations (1) and (2) in figure 1-98 for the effect on X_e and R_e if X_s is held constant but X_{Co} is decreased.) Thus, the larger the value of C_{pg}, the smaller the Q of the feed-back arm becomes the more the tank circuit must be detuned, the greater must be the value of the negative capacitance, C_n, and hence the greater the frequency instability due to changes in the tube parameters. The plate-to-grid capacitance needs to be considered only if a triode is used or if the crystal unit is oven-mounted. In the average triode C_{pg} is on the order of 1.5 to 2.5 $\mu\mu$f, sufficient to increase the effective value of C_o by as much as 50 per cent in some cases. When the second grid of a tube is used as the oscillator anode, as in the case of pentagrid converters, C_{pg} is usually on the order of 1 $\mu\mu$f. The pin-to-pin capacitance introduced by ovens may be as high as 5 $\mu\mu$f. The C_{pg} of screen-grid tubes can all but be neglected, since the increase in capacitance across the crystal is only about one-thousandth of the total. Because of its negligible C_{pg}, a pentode is preferred when the frequency deviation must be kept to a minimum. In the remaining discussion of the Pierce circuit, we shall assume that C_{pg} is negligible, so that X_e' and R_e' will represent the actual effective impedances of the crystal unit, and C_x' will equal the rated capacitance, C_x. Although we shall employ the unprimed symbols X_e and R_e to designate the plate-to-grid impedances, it should be remembered that this is only a convenience, for where X_e is predetermined by the rated load capacitance and the frequency, X_e' necessarily increases or decreases, respectively, with increases and decreases in C_{pg}. Similarly, C_x' varies negatively with C_{pg}, but we shall assume that it is a constant equal to C_x.

Determination of the Effective Negative Capacitance, C_n, Introduced by Vacuum Tube in Pierce Circuit

1-280. First, in order to avoid possible confusion, it should be pointed out that the selected reference or zero phase angle of the equivalent circuit in figure 1-119 is not the same as that implicitly assumed in the negative-resistance circuit of figure 1-121 (A). In figure 1-119, the reference phase has been taken as the phase of E_g, whereas in figure 1-121 (A) it is the phase of the current through the negative resistance ρ_T (not I_p), which in turn is the same as the phase of the r-f plate voltage, E_p. Now, I_p, in the negative-resistance circuit, is physically the same as the I_p of the vacuum-tube generator circuit. The equivalent current through ρ_T represents that component of I_p in phase with E_p—not that part of I_p in phase with $-\mu E_g$. The current through the negative resistance is thus smaller in magnitude than the total r-f plate current. The imaginary current through L_T, or C_n, is equal and opposite to that component of I_p which is 90 degrees out of phase with E_p. In the phasor chart in figure 1-119 (C), the bottom line shows the phase relations that are approached in a Pierce circuit if R_p of the tube approaches Z_L in magnitude. Note particularly that E_p and I_p rotate in opposite directions. E_p must lag I_p by an angle whose tangent is at least as great as $\frac{R_e}{X_e + X_{Cg}}$; that is, the tangent of the angle cannot be less than the reciprocal of the Q_f of the feed-back arm. The minimum angle occurs when R_p is very much greater than the load impedance Z_L and the gridleak losses are negligible. On the other hand, the phase difference between E_p

and I_p cannot be greater than 90 degrees, for the simple reason that E_p is a counter emf produced by I_p flowing through an equivalent impedance, Z_L, which has no component of negative resistance. There is another limitation in that the rotation of I_p with respect to $-\mu E_g$ cannot exceed 90 degrees minus the necessary rotation of E_p with respect to $-\mu E_g$. Otherwise, the necessary rotation of E_p cannot occur. As this extreme is approached, Z_L approaches a pure reactance approximately equal to X_{Cp}, I_p approaches a 90-degree phase lead over E_p, and the apparent Q of the entire plate circuit (approximately Z_L/R_p) approaches the Q_f of the feed-back circuit. These maximum and minimum phase angles are summarized in the following table. The phase angles are defined by the absolute values of their tangent expressed in terms of the phase Q of the feed-back arm, Q_f. Also shown are the limiting Q's of Z_L and $(Z_L + R_p)$; which respectively determine the phases of I_p with respect to E_p and $-\mu E_g$. In the last column are shown the limiting values of C_n which are discussed in the following paragraph.

	With Respect to:		Q of:		C_n (Approx)
	E_p	$-\mu E_g$	Z_L	$R_p + Z_L$	
Angle of E_p at all times	0°	$\tan^{-1}(1/Q_f)$			$-C_p\left(\frac{1}{Q_f^2} + \frac{R_e}{R_p + R_e}\right)$
Minimum angle of I_p	$\tan^{-1}(1/Q_f)$	0°	$1/Q_f$	0	$-C_p/Q_f^2$
Maximum angle of I_p	90°	$\tan^{-1}(Q_f)$	∞	Q_f	$-\frac{C_pR_e}{R_p + R_e}$

1-281. The approximate expressions given in the table above for the limiting values of C_n are derived upon the assumption that the unsigned phase angle between I_p and E_p is given by the equation

$$\tan^{-1}\left(\frac{X_{ZL}}{R_{ZL}}\right) = \tan^{-1}\left(\frac{1}{Q_f}\right) + \tan^{-1}\left(\frac{X_{ZL}}{R_p + R_{ZL}}\right) \qquad 1\text{—}281\ (1)$$

where X_{ZL} and R_{ZL} are the *absolute* values of the equivalent series reactance and resistance whose vector sum is equal to Z_L; $\tan^{-1}\left(\frac{1}{Q_f}\right)$ is the unsigned phase difference between E_p and $-\mu E_g$, and $\tan^{-1}\left(\frac{X_{ZL}}{R_p + R_{ZL}}\right)$ is the unsigned phase difference in the opposite direction between I_p and $-\mu E_g$. From equation (1) it follows that

$$\frac{X_{ZL}}{R_{ZL}} = \tan\left[\tan^{-1}\left(\frac{1}{Q_f}\right) + \tan^{-1}\left(\frac{X_{ZL}}{R_p + R_{ZL}}\right)\right] \qquad 1\text{—}281\ (2)$$

On applying the general trigonometric equation for the tangent of the sum of two angles

$$\tan(x + y) = \frac{\tan x + \tan y}{1 - \tan x \tan y}$$

Letting $\tan x = 1/Q_f$ and $\tan y = \frac{X_{ZL}}{R_p + R_{ZL}}$, equation (2) becomes

$$\frac{X_{ZL}}{R_{ZL}} = \frac{\frac{1}{Q_f} + \frac{X_{ZL}}{R_p + R_{ZL}}}{1 - \frac{X_{ZL}}{Q_f(R_p + R_{ZL})}} = \frac{R_p + R_{ZL} + Q_f X_{ZL}}{Q_f R_p + Q_f R_{ZL} - X_{ZL}} \qquad 1\text{—}281\ (3)$$

On rearranging, equation (3) can be expressed as an equation for R_p:

$$R_p = \frac{R_{ZL}^2 + X_{ZL}^2}{Q_f X_{ZL} - R_{ZL}} = \frac{Z_L^2}{Q_f X_{ZL} - R_{ZL}} \qquad 1\text{—}281\ (4)$$

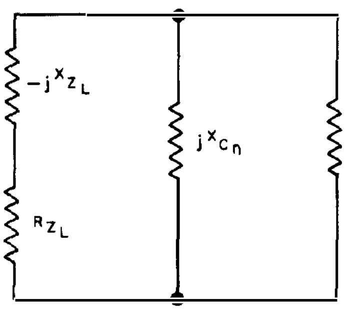


Figure 1-122. Generalized Pierce oscillator. X_{Z_L} and X_{C_n} represent the positive magnitudes of the equivalent reactances of the load and the dynamic effects of the tube, respectively. ρ_T is the same as the ρ_T in figure 1-121

Note that for oscillations to be maintained, $Q_f X_{Z_L}$ must be greater than R_{Z_L}, else R_p becomes negative. Referring to figure 1-122, the effective admittance of the Z_L and X_{C_n} branches in parallel is equal to $\frac{1}{R_{Z_L} - jX_{Z_L}} + \frac{1}{jX_{C_n}}$. From this expression it can be shown by straightforward manipulation that if the reactive component of the admittance is to be zero

$$X_{C_n} X_{Z_L} = R_{Z_L}^2 + X_{Z_L}^2 = Z_L^2 \quad 1\text{—}281\ (5)$$

On substitution of equation (5) in equation (4) and rearranging

$$X_{C_n} = R_p \left(Q_f - \frac{R_{Z_L}}{X_{Z_L}} \right) \quad 1\text{—}281\ (6)$$

The values of C_n as listed at the end of paragraph 1-280 are obtained from equation (6) by substituting the values of Q_f and R_{Z_L}/X_{Z_L} when these are expressed in terms of the basic circuit parameters. In paragraph 1-289, it is shown that at equilibrium:

$$X_{Cp} + X_{Cg} + X_e + \frac{R_e X_{Cp}}{R_p} = 0 \quad 1\text{—}289\ (3)$$

The term, $\frac{R_e X_{Cp}}{R_p}$, accounts for that part of the negative capacitance which is necessary to compensate for the phase shift in I_p, but not for that part which compensates for the phase shift of E_p. For example, if R_p were infinite, the phase shift of I_p, would be zero, and likewise the term, $\frac{R_e X_{Cp}}{R_p}$, in equation 1—289 (3). Nevertheless, E_p must still be rotated by an angle equal to $\tan^{-1}\left(\frac{1}{Q_f}\right)$, so that the tank cannot actually be parallel-resonant. Some value must be assigned to the negative capacitance, for the apparent resonance to hold in the generalized negative-resistance circuit. Now,

$$Z_L = \frac{Z_p Z_{pgc}}{Z_p + Z_{pgc}}$$

Letting

$$Z_p = jX_{Cp}$$

and

$$Z_{pgc} = R_e + j(X_e + X_{Cg})$$
$$= R_e - j\left(X_{Cp} + \frac{R_e X_{Cp}}{R_p}\right)$$

we can express Z_L as a complex function equal to $R_{Z_L} - jX_{Z_L}$, where X_{Z_L} is still assumed to be unsigned. Thus,

$$(\text{complex})\ Z_L = \frac{jX_{Cp}\left[R_e - jX_{Cp}\left(\frac{R_p + R_e}{R_p}\right)\right]}{R_e - \frac{jX_{Cp}R_e}{R_p}}$$

On multiplying both numerator and denominator by $\left(R_e + \frac{jX_{Cp}R_e}{R_p}\right)$, we find that

$$R_{Z_L} = \frac{X_{Cp}^2}{R_e\left(1 + \frac{X_{Cp}^2}{R_p^2}\right)} \approx \frac{X_{Cp}^2}{R_e} \quad 1\text{—}281\ (7)$$

and

$$X_{Z_L} = \frac{|X_{Cp}|\left[R_e + \frac{X_{Cp}^2}{R_p^2}(R_p + R_e)\right]}{R_e\left(1 + \frac{X_{Cp}^2}{R_p^2}\right)}$$
$$\approx \frac{|X_{Cp}|(R_e R_p + X_{Cp}^2)}{R_e R_p} \quad 1\text{—}281\ (8)$$

So

$$\frac{R_{Z_L}}{X_{Z_L}} = \frac{|X_{Cp}| R_p^2}{R_e R_p^2 + X_{Cp}^2 (R_p + R_e)}$$
$$\approx \frac{|X_{Cp}| R_p}{R_e R_p + X_{Cp}^2} \quad 1\text{—}281\ (9)$$

Also, when assuming that the grid losses are negligible,

$$Q_f = \frac{X_e + X_{Cg}}{R_e} = \frac{|X_{Cp}|(R_p + R_e)}{R_e R_p} \quad 1\text{—}281\ (10)$$

The expression on the right in equation (10) is obtained by substitution from equation 1—289(3). On substituting equations (9) and (10) in equation (6), we have

$$X_{C_n} = |X_{C_p}| \left[\frac{R_p + R_e}{R_e} - \frac{R_p^2}{R_e R_p + X_{C_p}^2 (1 + R_e/R_p)}\right]$$

$$X_{C_n} = |X_{C_p}| \left[\frac{R_p X_{C_p}^2 (1+R_e/R_p) + R_e^2 R_p + R_e X_{C_p}^2 (1+R_e/R_p)}{R_e^2 R_p + R_e X_{C_p}^2 (1 + R_e/R_p)}\right] \quad 1—281\ (11)$$

Equation (11) is obtained by using the exact values of R_{Z_L}/X_{Z_L} and Q_f as given by equations (9) and (10). Although the equation for X_{C_n} involves the difference between two nearly equal terms, the error introduced by using the approximate values of Q_f and R_{Z_L}/X_{Z_L} is negligible for all practical purposes as long as $R_p >> R_e$.

Now,

$$C_n = -\frac{1}{\omega X_{C_n}}$$

so, on substitution of equation (11),

$$C_n = -C_p \left[\frac{R_e^2 R_p + R_e X_{C_p}^2 (1 + R_e/R_p)}{R_p X_{C_p}^2 (1+R_e/R_p) + R_e^2 R_p + R_e X_{C_p}^2 (1+R_e/R_p)}\right] \quad 1—281\ (12)$$

In the practical case, R_p is much greater than R_e and $X_{C_p}^2$ is much greater than R_e^2, so that the approximate equation for C_n becomes

$$C_n = -C_p \left[\frac{R_e^2 R_p + R_e X_{C_p}^2}{R_p (X_{C_p}^2 + R_e^2)}\right]$$

$$= -C_p \left(\frac{1}{Q_f^2} + \frac{R_e}{R_p + R_e}\right) \quad 1—281\ (13)$$

The last term on the right inside the parentheses is obtained by assuming that $\frac{X_{C_p}^2}{X_{C_p}^2 + R_e^2}$ is more nearly equal to $\frac{R_p}{R_p + R_e}$ than to 1. We make this assumption arbitrarily for the convenience in remembering the limiting values of C_n. That part of C_n which is necessary so as to compensate for the phase shift of E_p with respect to $-\mu E_g$ is approximately equal to $-C_p/Q_f^2$; whereas the part necessary to compensate for the phase shift of I_p with respect to $-\mu E_g$ is approximately equal to $\frac{-C_p R_e}{R_p + R_e}$. It is this latter component that is accounted for by the term, $\frac{X_{C_p} R_e}{R_p}$, in equation 1—289 (3). By equation (12), when R_p/Z_L approaches infinity, the limiting value of C_n is found to be

$$\underset{(R_p/Z_L \to \infty)}{C_n} = \frac{-C_p R_e^2}{X_{C_p}^2 + R_e^2} \approx \frac{-C_p}{Q_f^2} \quad 1—281\ (14)$$

When R_p/Z_L approaches its smallest possible value, the ratio R_{Z_L}/X_{Z_L} becomes negligible compared with Q_f, so that X_{C_n} by equation (6) is approximately equal to $Q_f R_p$. Substituting equation (10) for Q_f, we have, when $\left(\frac{X_{Z_L}}{R_p + R_{Z_L}} \to Q_f\right)$

$$C_n = \frac{-1}{\omega X_{C_n}} \approx \frac{-1}{\omega Q_f R_p}$$

$$= -\frac{R_e}{\omega |X_{C_p}| (R_p + R_e)}$$

$$= \frac{-C_p R_e}{R_p + R_e} \quad 1—281\ (15)$$

As indicated in figure 1-121, the total effective plate capacitance is

$$C_p' = C_p + C_n = C_p \left(1 - \frac{1}{Q_f^2} - \frac{R_e}{R_p + R_e}\right) \quad 1—281\ (16)$$

When $R_p >> Z_L$, the Q of Z_L approaches $1/Q_f$ in value and

$$C_p' \approx C_p \left(1 - \frac{1}{Q_f^2}\right) \quad 1—281\ (17)$$

When R_p is small compared with $X_{C_p}^2/R_e$, then R_{Z_L}/X_{Z_L} becomes small compared with Q_f and

$$C_p' \approx \frac{C_p R_p}{R_p + R_e} \quad 1—281\ (18)$$

R_p is normally much greater than R_e. Only low-frequency crystals have effective resistances which approach in value the plate resistances of low-power vacuum tubes. For all practical purposes in the average Pierce circuit, C_p' can be assumed to equal C_p, the static plate capacitance, except when considering problems of frequency stability. What is important to note in the limiting equations for C_n is the fact that if the tank is to be operated well off resonance, R_p becomes quite an important factor in determining the frequency. In this case, because C_n is relatively large, any variation in the tube R_p has a great effect upon the frequency. It should be remembered that the parameter Q_f has been used to account for the required rotation of E_p with respect to $-\mu E_g$. Insofar as the gridleak is effective in increasing the necessary phase shift, $Q_f = \dfrac{X_e + X_{Cg}}{R_e}$ cannot be assumed, and the complete equation 1—272 (1) must be used.

THE EFFECT OF R_p UPON THE FREQUENCY OF PIERCE CIRCUIT

1-282. It can be seen that for large values of the R_p/Z_L ratio, C_n is small and its percentage variation with changes in R_p is smaller still. C_n is, effectively, a frequency-determining parameter, but more exactly it is a mathematical function that indirectly expresses the effect of R_p and Q_f upon the frequency. The smaller the Q_f and R_p, the larger is C_n; and the larger C_n, the greater is the effect of R_p. Since R_p is subject to change with changes in the tube voltages, tube aging, and the like, it is important to keep C_n as small as possible. This can be done by designing the circuit to operate with as high of value of tube R_p as is practicable. For a given tube, the higher values of an effective R_p are to be obtained when the tube is conducting during only a small fraction of a cycle. This in turn requires that the oscillator tube be operated class C, so that a larger grid bias than otherwise is required. However, if the crystal drive level is to be kept low and if the gridleak is to have a negligible effect on the effective Q_f, and hence upon C_n, the gridleak resistance must be as large as practicable without running the risk that the tube will block or operate intermittently. The limiting value of R_p occurs when $-\mu E_g$ is just sufficient to maintain oscillations. If the vacuum tube could operate into a pure resistance, I_p would be in phase with $-\mu E_g$, and R_p would be eliminated as a frequency-determining element. In the conventional Pierce circuit this could occur only if Q_f were infinite.

Phase-Stabilized Pierce Circuit

1-283. If an inductor is inserted in the plate circuit of the oscillator, as indicated in figure 1-123, having a reactance equal and opposite to the effective reactance X_{ZL}, then I_p undergoes no phase rotation, and changes in R_p, although affecting the activity, will have little effect upon the frequency. With I_p in phase with $-\mu E_g$, the Q of Z_L must equal $1/Q_f$, and the operation of the tank is the same as it would be if R_p were infinite. If Q_f is reasonably large and is approximately equal to $|X_{Cp}/R_e|$,

$$R_{ZL} \approx Z_L \approx X_{Cp}^2/R_e \approx |X_{Cp}| Q_f$$

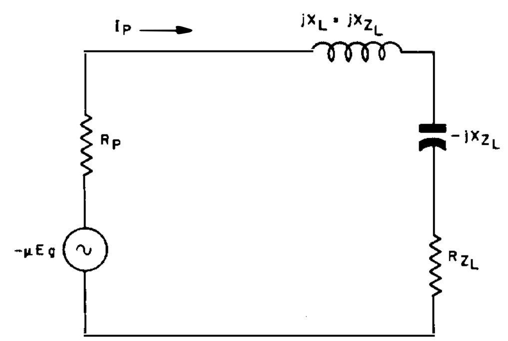


Figure 1-123. Coil inserted in plate circuit of Pierce oscillator to prevent phase of I_p from being influenced by changes in R_p

Since

$$X_{ZL}/R_{ZL} = 1/Q_f$$

X_{ZL} is approximately equal in magnitude to X_{Cp}. Thus, for the vacuum tube to look into a resistive load, the inductor should have a reactance approximately equal in magnitude to X_{Cp}. This value assumes that the gridleak and output losses are negligible. When such assumptions cannot be made, the value of the series plate reactance becomes a more involved function. Llewellyn analyzed this type of circuit and eliminated R_p from the frequency-determining equation (phase rotation equation) by equating the sum of the factors of R_p to zero. Although the approach is different and the grid losses are assumed to be predominant, Llewellyn's mathematical elimination of the effects of R_p upon the frequency by the introduction of a plate inductor in series with the tank appears to be equivalent to the qualitative condition that I_p must be held in phase with $-\mu E_g$. The experimenter, nevertheless, should be warned that the theory of this type of stabilization has been analyzed above, and also by Llewellyn, only in terms of the phase relations. Difficulty will probably be experienced in obtaining stable oscillations without additional modifications to ensure that the limiting characteristics are changed from a voltage- to a current-controlled nature. This feature of oscillator theory has not been fully explored, but see paragraphs 1-585 to 1-598 for a general discussion, and paragraph 1-323 for a particular example of an attempt, which was not entirely successful, to phase-stabilize a Pierce circuit.

Conditions for Maximum R_p in Pierce Circuit

1-284. Referring to figures 1-119 (A) and (B), we shall begin with the assumption that the tank is operating near resonance so that $Z_L \approx \dfrac{Z_p^2}{R_e}$, where R_e (not shown) is the effective resistance of the crystal unit whose total impedance is represented by Z_{pg}. Z_L, therefore, is very nearly resistive, and I_g is approximately equal in magnitude to the current through Z_p.

Thus, $I_g \approx E_p/Z_p$

$$\text{also, } E_g = I_g Z_g = \frac{E_p Z_g}{Z_p}$$

$$\text{Rearranging, } E_p/E_g = \frac{Z_p}{Z_g} = C_g/C_p$$

In the interest of maxium stability it is desirable for R_p to be a maximum. The problem is to find what capacitance ratio, C_g/C_p, permits the largest possible value of R_p consistent with the rated drive level and load capacitance of the crystal unit. The phase-rotation equations do not enter the problem —only those equations that concern the magnitude of the equilibrium voltages and currents are of concern now. The crystal specifications indirectly set an upper limit for the tank current, I_g. Thus, the output voltage, $E_p \approx I_g Z_p$, also has an upper limit, since Z_p $(= Z_{pg} - Z_g)$ has a theoretical maximum equal to $\mu Z_g = \frac{\mu Z_{pg}}{\mu + 1} \approx \frac{\mu X_e}{\mu + 1}$, which is approached as R_p approaches zero. At the other ideal extreme, Z_p and Z_L approach zero and the R_p/Z_L ratio becomes very large. Now, a large R_p/Z_L is desired, but some compromise must be made, since the Q of the feed-back circuit becomes increasingly small as Z_g approaches Z_{pg} in magnitude. A rigorous treatment of the problem to find that relation between R_p/Z_L and Q_f that provides an optimum frequency stability would require that a complete equation of frequency stability be established and that those impedance relations be determined that produce a minimum frequency deviation for small changes in the circuit parameters. Equation (2) in paragraph 1-288, which is a first order expression for the fractional change in frequency for a change in R_p, indicates that the percentage deviation increases directly with the first power of C_p, and inversely with the second power of R_p. This suggests that the stability increases as long as C_p/R_p^2 decreases with an increase in C_p, and is a maximum at the value of C_p, if existent, at which this ratio begins to increase. Such an approach will not be attempted here. Unless all the characteristic curves of a vacuum tube are available, so that either μ, R_p, or g_m can be used as an independent variable to eliminate the other two from the equations, concrete conclusions cannot be reached concerning the optimum design of an oscillator using that particular tube. A more qualitative analysis is presented below, and although the indicated optimum relations cannot be considered conclusive, they can serve as first approximations. All impedance, current, and voltage symbols given below are considered positive and undirected.

$$\text{Now, } E_p = I_p Z_L$$

$$\text{And } I_p = \frac{\mu E_g}{R_p + Z_L} = \frac{\mu E_p Z_g}{Z_p(R_p + Z_L)}$$

$$\text{So } E_p = \frac{\mu E_p Z_g Z_L}{Z_p(R_p + Z_L)}$$

$$\text{or } \mu Z_g Z_L - R_p Z_p = Z_p Z_L$$

On substituting $(R_p g_m)$ for μ, where g_m is the transconductance of the tube,

$$R_p g_m Z_g Z_L - R_p Z_p = Z_p Z_L$$

$$\text{or } R_p = \frac{Z_p Z_L}{g_m Z_g Z_L - Z_p} \qquad 1\text{—}284\ (1)$$

Dividing both sides by Z_L, we have

$$\frac{R_p}{Z_L} = \frac{Z_p}{g_m Z_g Z_L - Z_p} = \frac{1}{\frac{g_m Z_g Z_L}{Z_p} - 1}$$

Our present concern is to seek the largest practical value of R_p/Z_L, so that the phase of I_p will be least affected by small changes in R_p. Now, $Z_g \approx Z_{pg} - Z_p$, where Z_{pg} represents the predetermined crystal impedance, which is approximately equal to X_e. Also, $Z_L \approx Z_p^2/R_e$. On substitution of these values in the equation for R_p/Z_L, it is found that

$$\frac{R_p}{Z_L} = \frac{R_e}{g_m Z_p Z_g - R_e} = \frac{R_e}{g_m(Z_{pg}Z_p - Z_p^2) - R_e} \qquad 1\text{—}284\ (2)$$

It can be seen from equation (2) that for oscillations to be maintained g_m $(Z_{pg}Z_p - Z_p^2)$ must be greater than R_e. A maximum R_p/Z_L ratio is approached as the product $g_m(Z_{pg} Z_p - Z_p^2)$ approaches the value of R_e. Of course, it is impossible for the denominator in equation (2) to be actually equal to zero, for then R_p would be infinite; but it is plausible to assume that a denominator much smaller than the value of R_e can be realized. Thus, we can write

$$(\text{optimum})\ g_m \approx \frac{R_e}{Z_{pg}Z_p - Z_p^2} = \frac{R_e}{Z_p Z_g} \qquad 1\text{—}284\ (3)$$

The more nearly this equality is approached, the greater will be the frequency stability. The question arises, is it preferable to seek this equality with a small or a large value of g_m? Assuming that

$(R_p/Z_L) >> 1$, the equation, $I_p = \frac{\mu E_g}{R_p + Z_L}$, can be written approximately $I_p = \mu E_g/R_p = g_m E_g$, or $g_m = I_p/E_g$. A large g_m means a large r-f plate current for a given excitation voltage. This would be desirable from the point of view of maximum output, but an examination of the denominator in equation (3) shows that a large transconductance means that the plate impedance, Z_p, or the grid impedance, Z_g, must be made small if the equation is to hold. A small Z_g (large Z_p) means a large Z_L and also a large E_p/E_g ratio. Both consequences are incompatible with a large R_p/Z_L ratio. The former is obviously so, and the latter is implicitly so because the ratio of E_p/E_g times $\frac{R_p + Z_L}{Z_L}$ $\left(= \frac{\mu E_g}{E_p}\right)$ must equal μ. Every increase in E_p/E_g therefore requires an approximately proportional decrease in the R_p/Z_L ratio, insofar as μ can be assumed to remain constant. On the other hand, a large Z_g and small Z_p permits a large R_p/Z_L ratio and has the additional advantage of permitting a given excitation voltage with a minimum crystal current. It is under these conditions that equation (3) will be most nearly exact. There are serious disadvantages, however, when operating at a maximum R_p/Z_L ratio; the most important of which is that the Q of the feed-back circuit rapidly decreases as Z_g is increased, since $Q_f \approx \frac{Z_{pg} - Z_g}{R_e}$. Also, the voltage output is weak, and has a tendency to instability. This will be discussed more fully later. Since the excitation voltage is stronger for a given crystal current, the grid losses increase proportionately and may no longer be negligible. Furthermore, unless the tube is operated class C, the power efficiency is very low. These last mentioned disadvantages, nevertheless, are minor compared with the effect on Q_f. The minimum effective Q of the average crystal unit when operating into its rated load capacitance is not unduly large. A grid-to-cathode reactance equal in magnitude to three-fourths X_e reduces Q_f to one-fourth Q_e. Since the purpose of a large R_p/Z_L ratio is to permit the entire plate-circuit impedance $(R_p + Z_L)$ to appear as nearly resistive as possible, the better stability risk is to operate the parallel-resonant oscillator with small rather than large I_p and g_m. Since we are assuming that $g_m \approx I_p/E_g$, it can be seen that the smaller the value of g_m, the smaller is the r-f plate current for a given excitation voltage, or, for a given plate current, the smaller the value of g_m, the greater the excitation voltage. The problem becomes one of determining what capacitance ratio, C_g/C_p, permits the smallest possible g_m. By equation (3), g_m is a minimum when the denominator of the right-hand term is a maximum. Since the impedance of the crystal unit, Z_{pg}, is to be held constant, $(Z_p + Z_g)$ is also a constant. Thus, the product $Z_p Z_g$ can easily be shown to be a maximum when

$$Z_p = Z_g = Z_{pg}/2 \qquad 1\text{—}284\ (4)$$

A maximum operating R_p and a minimum I_p with a given excitation voltage can thus be obtained when the capacitance and voltage ratios are

$$E_p/E_g = C_g/C_p = 1$$

It is quite fortunate that g_m has a minimum value. At all other operating values a small change in the $\frac{C_g}{C_p}$ ratio causes g_m and R_p, and hence the frequency, to change. At the minimum g_m the rate of change in the tube parameters is necessarily zero, so that the stability in this respect is a maximum. When the more exact equation 1—289 (2) is used instead of equation (1) above, and when μ/R_p is substituted for g_m, it can be shown that

$$R_p = \frac{X_{Cp} X_e + X_{Cp} X_{Cg} + \mu X_{Cp} X_{Cg}}{R_e} \qquad 1\text{—}284\ (5)$$

Note that R_p, as long as μ is constant, is inversely proportional to R_e. Now, approximately

$$X_{Cp} = -(X_e + X_{Cg})$$

Substituting in equation (5), R_p becomes a function of X_e, R_e, μ, and X_{Cg}. Assuming that the first three parameters are constant, it can be shown that R_p is a maximum when

$$X_{Cg} = -\frac{X_e(\mu + 2)}{2(\mu + 1)} = -\frac{X_e}{2}\left(1 + \frac{1}{\mu + 1}\right) \qquad 1\text{—}284\ (6)$$

If $(\mu + 1) >> 1$, equation (6) states approximately the same conditions as does equation (4). If μ is small, equation (6) should be accepted as the more accurate in computing the optimum C_g/C_p ratio, since a minimum g_m coincides with a maximum R_p if the d-c plate voltage is kept constant. The capacitance ratio and values corresponding to equation (6) are

$$C_g/C_p = \frac{\mu}{\mu + 2} \qquad 1\text{—}284\ (7)$$

or

$$C_g = \frac{2C_x\ (\mu + 1)}{(\mu + 2)} \qquad 1\text{—}284\ (8)$$

and

$$C_p = \frac{2C_x\ (\mu + 1)}{\mu} \qquad 1\text{—}284\ (9)$$

Under these conditions the excitation voltage becomes greater than the voltage across the plate load by a factor of $\frac{\mu + 2}{\mu}$, and the following additional relations hold:

$$(\text{max})\ R_p = (\mu + 1)\ Z_L \qquad 1\text{—}284\ (10)$$

$$(\text{min})\ g_m = \frac{1}{Z_L} - \frac{1}{R_p} \qquad 1\text{—}284\ (11)$$

As a practical consideration in design as well as for the sake of simplicity in discussion it is convenient to assume that the optimum C_g/C_p ratio is equal to one rather than $\frac{\mu}{\mu + 2}$. However, in interpreting the equations above, a word of caution is necessary. Returning to equation (2), it will be seen that the maximum to be sought for R_p/Z_L is a "practical," not a "mathematical" maximum in the sense that a curve of R_p/Z_L rises to a peak and then decreases. The curve of equation (2) plotted against Z_p passes from positive to negative infinity as the denominator passes through zero and thus is discontinuous at that point. However, for any given value of g_m sufficiently large for R_p/Z_L to be positive, the curve does have a true minimum, not a maximum, at the point where Z_pZ_g is a maximum. To avoid confusion as a result of this apparent contradiction, it is important to recall that the "practical" maximum is to be sought by making equation (3) as nearly true as possible, and not by the process of making $Z_p = Z_g$. This latter consideration is in the interest of over-all stability and maximum activity (if measured by the d-c grid current) for a given d-c plate voltage and load capacitance. Another point that should be well understood is that the minimum g_m, minimum μ, and maximum R_p, are all coincidental. From the point of view of frequency stability the real interest is in the maximum R_p. From equation (10) it can be seen that the magnitude of the maximum R_p will increase with μ; but remember that this value of μ is the minimum obtainable with a given tube and E_b. As will be discussed more thoroughly in paragraphs 1-294 and 1-295, an oscillator vacuum tube cannot be operated so that μ is the maximum possible without the risk of amplitude instability. Thus, class-A operation where the tube is operated only along the straight portion of the E_cI_b curve is not feasible in gridleak oscillators. Understand that if equation (7) holds, equations (10) and (11) automatically hold. μ in each equation is the effective μ when equilibrium is reached and is not the starting μ. It is the minimum μ that can be obtained as long as the crystal resistance, R_e, and the total load capacitance, C_x, remain constant. Because R_p, g_m, and μ all pass through extremes at the optimum capacitance ratio, it might be thought that the operating conditions are more ideally unique than they actually are, because the instantaneous rate of change for all the tube parameters with the capacitance ratio is zero under these conditions. Remember, however, that these maximum and minimum values apply only in the event that the total C_x remains constant. An independent variation in C_g or C_p will cause the frequency to change, and the tube parameters will vary. For instance, g_m will tend to vary directly with both C_g or C_p. Only when C_g and C_p are adjusted simultaneously so as always to maintain the same total load capacitance will the instantaneous changes in the tube parameters be zero as the capacitance ratio is varied through its optimum value.

1-285. If equation 1—284 (3) is expressed as a function of Z_{pg}, by substituting from equation 1—284 (4), the minimum value of g_m becomes, approximately,

$$(\text{min})\ g_m = \frac{4\,R_e}{Z_{pg}^{\ 2}} \qquad 1\text{—}285\ (1)$$

Since Z_{pg} is the crystal impedance, approximately equal to X_e, the minimum value of g_m can be expressed as

$$(\text{min})\ g_m = \frac{4}{PI} \qquad 1\text{—}285\ (2)$$

where PI is the performance index. Using a one-to-one capacitance ratio and a vacuum tube of high R_p, equilibrium will be reached at the value of g_m defined in equation (2). Such operation generally provides the maximum frequency stability in a Pierce oscillator. In estimating the value of PI, X_e is numerically equal to the reactance of the rated load capacitance, and R_e must be assumed to be the maximum permissible effective resistance according to the military specifications of the crystal unit being used.

CAPACITANCE RATIO, C_g/C_p, FOR GREATER OUTPUT IN PIERCE CIRCUIT

1-286. Where a maximum output consistent with the minimum frequency-stability requirements is desired in a Pierce oscillator, the C_g/C_p ratio can be increased and a vacuum tube providing a large transconductance and a large amplification factor should be used. The first consideration is that I_g must not exceed a value that would cause the power dissipation in the crystal unit to exceed the specified drive level. If the output of the oscillator is capacitively coupled to the grid of a buffer amplifier, the output power becomes a minor consideration compared with the output voltage. If this voltage is to be a maximum for a given tank current, the plate impedance Z_p must be a maximum. This means that the capacitance ratio, C_g/C_p, must be as large as practicable. The larger this ratio, however, the smaller will be the excitation voltage for a given I_g. The smaller the excitation voltage, the smaller will be the gridleak bias, and consequently R_p will be less whereas g_m and μ will be greater. An examination of equation 1—284 (3) reveals that the required magnitude of g_m becomes very large as Z_p approaches the value of Z_{pg} of the crystal unit. Where a relatively large frequency deviation can be tolerated a large R_p may not be necessary, so that increased voltage outputs can be obtained with tubes of high transconductance at low d-c plate voltages. In any event the C_g/C_p ratio can never exceed the amplification factor of the tube, nor can the r-f plate voltage be greater than, nor equal to, the voltage across the crystal unit. It should be understood that the higher voltage outputs are only to be had with a large C_g/C_p ratio because of the limitations on the crystal drive level and the load capacitance, C_x, and are not due to the fact that the Pierce or Colpitts type of circuit is inherently more active when the C_g/C_p ratio is a maximum. To the contrary, with a fixed C_x, maximum amplitude of oscillations is to be obtained when $C_p = C_g = 2C_x$. Where a larger than minimum power rather than voltage output is required, this can probably best be achieved when C_g/C_p lies between 1 and 2, and in this case a larger g_m is necessary to maintain oscillations. When a relatively large power output is required, the Pierce circuit should not be used.

HOW TO ESTIMATE THE FREQUENCY VARIATION AND STABILITY OF A PIERCE OSCILLATOR

1-287. In paragraph 1-243 it was shown that the frequency-stability coefficient, F_{Xe}, of the crystal unit is defined as the percentage change in X_e per percentage change in frequency. By equation 1—243 (1), $F_{Xe} = \frac{2C_T^2}{C\,C_x}$. The reciprocal, $1/F_{Xe}$, is thus equal to the percentage change in frequency per percentage change in reactance. Since X_e is equal numerically to the reactance of the total load capacitance C_x, the fractional change in the load reactance multiplied by $1/F_{Xe}$ will give the fractional change in frequency. Thus,

$$\frac{df}{f} = \frac{d\omega}{\omega} = \frac{1}{F_{Xe}} \cdot \frac{dX_x}{X_x} = -\frac{1}{F_{Xe}} \cdot \frac{dC_x}{C_x} \qquad 1\text{—}287\ (1)$$

In the Pierce circuit, if it can be assumed that the interelectrode plate-to-grid capacitance and the grid and output losses are negligible, C_x will equal $\frac{C_p' C_g}{C_p' + C_g}$ where $C_p' = C_p + C_n$. (See figure 1-121 (C).) If it is desired to find the fractional change in frequency for a small change in, say C_p', equation (1) can be used by expressing dC_x as a function of dC_p', thus: $dC_x = dC_p' \cdot \frac{dC_x}{dC_p'}$

With

$$C_x = \frac{C_p' C_g}{C_p' + C_g}$$

then

$$\frac{dC_x}{dC_p'} = \frac{C_g^2}{(C_p' + C_g)^2}$$

and

$$\frac{dC_x}{C_x} = \frac{C_g^2\, dC_p'}{(C_p' + C_g)^2} \Big/ \frac{C_p' C_g}{C_p' + C_g}$$

$$= \frac{C_g\, dC_p'}{C_p'\,(C_p' + C_g)} = \frac{C_x}{(C_p')^2} \cdot dC_p'$$

On substituting in equation (1):

$$\frac{df}{f} = -\frac{1}{F_{Xe}} \cdot \frac{C_x}{(C_p')^2} \cdot dC_p' \qquad 1\text{—}287\ (2)$$

Likewise,

$$\frac{df}{f} = -\frac{1}{F_{Xe}} \cdot \frac{C_x}{C_g^2} \cdot dC_g \qquad 1\text{—}287\ (3)$$

In the event that $C_p' = C_g = 2C_x$
Equations (2) and (3) become

$$\frac{df}{f} = -\frac{dC_p'}{4\,F_{Xe}} \qquad 1\text{—}287\ (4)$$

and

$$\frac{df}{f} = \frac{-\,dC_g}{4\,F_{Xe}} \qquad 1\text{—}287\ (5)$$

1-288. When a more detailed expression of the frequency deviation is desired C_p' can be replaced by $(C_p + C_n)$, and C_n, in turn, can be expressed as a function of its variables. A rigorous analysis of the effect of each parameter upon the frequency would be quite involved. Probably the simplest approach for determining the frequency deviation due to a change in some particular circuit parameter would be to begin with an appropriate equation in paragraph 1-287, and express the differential element as a function of the differential of the particular circuit parameter. This is the method that was used when dC_x in equation 1—287 (1) was expressed as a function of dC_p' and of dC_g in equations 1—287 (2) and 1—287 (3), respectively. As an additional example, suppose that it is desired to determine approximately the frequency deviation due to a change in R_p of the vacuum tube. Let it be assumed that $C_g = 2C_x$. The most appropriate equation to begin with is 1—287 (4), since dC_p' can be expressed as a function of dR_p. The problem is to determine the function that gives the change in C_p' due to an infinitesimally small change in R_p, and to substitute that function for dC_p' in equation 1—287 (4). Since $dC_p' = \frac{dC_p'}{dR_p} \cdot dR_p$, the first step is to determine dC_p'/dR_p, and then simply to multiply this by dR_p. Now, by equation 1—281 (16),

$$C_p' = C_p - C_p/Q_f^2 - \frac{C_p\,R_e}{R_p + R_e}$$

so

$$dC_p'/dR_p = C_pR_e/(R_p + R_e)^2$$

or

$$dC_p' = \frac{C_p\,R_e}{(R_p + R_e)^2} \cdot dR_p \qquad 1\text{—}288\ (1)$$

Substituting this function for dC_p' in equation 1—287 (4):

$$\frac{df}{f} = -\,\frac{C_p\,R_e}{4(R_p + R_e)^2} \cdot \frac{dR_p}{F_{Xe}} \qquad 1\text{—}288\ (2)$$

Note than an increase in R_p causes a decrease in the frequency. It should be well understood that the equations above are only rough approximations. For example, one of the approximations in equation (2) is the assumption that R_p and R_e are independent variables, which, of course, is not true. However, the direct effect of a change in R_p upon the frequency can be considered much greater than the indirect effect due to a change in R_e resulting from the initial change in frequency. The differential element, dR_p, in equation (2) can, in turn, be expressed as a function of an infinitesimal change, dE_b, in the d-c plate voltage. The general equation for this function is $dR_p = \frac{dR_p}{dE_b} \cdot dE_b$. However, the derivative term, dR_p/dE_b, will be quite difficult to determine, except by experiment, since it depends upon both the tube and circuit characteristics at the operating voltages. For its mathematical expression the principal considerations would be the change in R_p with a change in E_b, assuming a constant grid bias, the change in grid bias due to the change in plate voltage, and the change in R_p due to a change in grid bias, assuming a constant plate voltage. In using the equations above it is only necessary to substitute small finite changes in the independent variable for its differential. For example, if the input capacitance were to decrease an amount $\Delta C_g = 1\mu\mu f$, the fractional change in frequency would be given approximately by equation 1—287 (3) if we were to substitute $-1\mu\mu f$ for dC_g.

ENERGY AND FREQUENCY EQUATIONS OF PIERCE CIRCUIT AS COMPLEX FUNCTIONS OF LINEAR PARAMETERS

1-289. It is beyond the scope of this handbook to present the more rigorous analyses of the various oscillator circuits. These can be obtained from the various reference sources listed in the index. Actually, even when following the more explicit equations, so many approximations must be made for the sake of simplicity, and so many unknowns are involved, such as stray circuit capacitance, that the final solutions can rarely be considered more than general indices of the actual circuit conditions. If maximum mathematical exactitude is desired in determining the frequency and activity characteristics of an oscillator, the analysis should be performed by differential equations assuming nonlinear parameters. Such equations are quite involved and are rather difficult to interpret qualitatively. If a moderately rigorous analysis is desired, the equivalent circuit in figure 1-119 (B) may be assumed to have linear parameters, as has been assumed in our previous discussion, but instead of handling the impedances as real numbers, to represent them by complex functions. For the Pierce circuit

$$Z_p = jX_{Cp}$$
$$Z_g = jX_{Cg}$$
$$Z_{pg} = R_e + jX_e$$

An equation expressing the conditions for oscillation is derived very similarly to the method followed in paragraph 1-284, except that I_g is expressed by the exact equation

$$I_g = \frac{E_p}{Z_{pg} + Z_g}$$

rather than by the simplifying equation

$$I_g \approx E_p/Z_p$$

For either the Pierce or Miller circuit, the conditions for stable oscillation are expressed by the equation

$$-R_p = \frac{Z_p(Z_{pg} + Z_g)}{g_m Z_p Z_g + Z_s} \qquad 1\text{—}289\ (1)$$

where $Z_s = Z_p + Z_{pg} + Z_g$. Equation (1) is similar to equation 1—284 (1) except that in the derivation of equation (1) I_g is expressed by its exact function and the impedances represent complex quantities. When the impedances are expressed as complex functions, the right-hand side of equation (1) can be reduced to the sum of a real quantity and an imaginary quantity, each with the dimensions of impedance.

Thus, $R_p = R + jX$

However, since R_p is not, itself, reactive, the imaginary term, jX, must equal zero, and the real term, R, must equal R_p. In this manner two equations, $X = 0$ and $R = R_p$, involving the same variables are obtained, both of which must hold if stable oscillations are to be maintained. A minimum of two equations is necessary, since there are two independent functions to perform. One function is to fix the frequency so that the excitation voltage is properly phased, and the other function is to ensure that the feed-back energy per cycle is exactly equal to its dissipation per cycle. If both sides of equation (1) are divided by R_p, the right-hand side again reduces to the sum of real and imaginary parts. The real part must be equal to 1, and the imaginary part must again be zero. It can be shown that the real part becomes, after multiplying through by R_p:

$$R_p = \frac{X_{Cp}(X_e + X_{Cg})}{R_e - g_m X_{Cp} X_{Cg}} \qquad 1\text{—}289\ (2)$$

This equation, rather than equation (1) is the real equivalent of equation 1—284 (1) and serves the same purpose in that it defines the conditions necessary for the feed-back energy to be sufficient and stable—in other words, for the loop gain to equal unity. The equation for the imaginary part when made equal to zero can be expressed as:

$$X_{Cp} + X_{Cg} + X_e + \frac{R_e X_{Cp}}{R_p} = 0 \qquad 1\text{—}289\ (3)$$

When this equation holds, the loop phase shift is zero. Whereas equation (2) is said to define the feed-back energy requirements, equation (3) is said to define the frequency requirements, Note that in equation (3), the term $\frac{R_e X_{Cp}}{R_p}$ is equivalent to the reactance of a dynamic capacitance

$$C_d = \frac{C_p R_p}{R_e} \qquad 1\text{—}289\ (4)$$

which can be imagined to be in series with C_p. The capacitance of the combination becomes

$$C_p'' = \frac{C_p C_d}{C_p + C_d} = \frac{C_p R_p}{R_p + R_e} \qquad 1\text{—}289\ (5)$$

which is exactly the same as the small-R_p value of C_p' in equation 1—281 (18). C_d is thus a positive series dynamic capacitance equivalent to part of the negative parallel dynamic capacitance C_n. Equation (3) indicates that as R_p increases indefinitely $X_{Cp} + X_{Cg} + X_e \rightarrow 0$. This is not to be interpreted as meaning that the tank circuit approaches a parallel-resonant state as a limit or that the total dynamic capacitance approaches zero. Actually, even if the sum of the first three reactances did equal zero, the tank would not be at resonance because of the presence of R_e in the feed-back arm, and a dynamic capacitance would need to be effectively present. What equation (3) does show is that, as R_p increases indefinitely, the frequency becomes entirely determined by the tank-circuit parameters. In the limit, $X_{Cg} + X_e = -X_{Cp}$. As this state is approached, the impedance of the feed-back arm can be represented as

$$Z_{pgc} = R_e + jX_{pgc} = R_e - jX_{Cp}$$

The impedance of the tank circuit is

$$Z_L = \frac{Z_{pgc} Z_p}{Z_{pgc} + Z_p} = \frac{(R_e - jX_{Cp})\, jX_{Cp}}{R_e - jX_{Cp} + jX_{Cp}}$$
$$= \frac{X_{Cp}^2}{R_e} + jX_{Cp}$$

The real component, X_{Cp}^2/R_e, is equivalent to R_{ZL}, and the imaginary component, jX_{Cp}, is equivalent to X_{ZL}. Thus, the tank circuit does not appear as a pure resistance in the limit, but approaches an effective

$$Q = \frac{X_{ZL}}{R_{ZL}} = \frac{|X_{Cp}|}{X_{Cp}^2/R_e} = \frac{R_e}{|X_{Cp}|}$$

$$= \frac{R_e}{|X_e - X_{Cg}|} = \frac{1}{Q_f}$$

This is the same effective Q as was determined qualitatively from the viewpoint of phase angles. The term R_eX_{Cp}/R_p in equation (3) is therefore not to be interpreted rigorously as the total dynamic reactance, but as that part of the dynamic reactance that is a function of the tube parameters. At all times, the total load reactance across the crystal terminals is $X_x = -1/\omega C_x = X_{Cg} + X_{Cp}'$ which is always slightly greater than $(X_{Cg} + X_{Cp})$.

CAPACITIVE ELEMENTS IN DESIGN OF PIERCE OSCILLATOR

1-290. It was declared that equation 1—289 (3) defined the frequency requirements of a Pierce oscillator. Since the frequency and the value of X_e and X_x are effectively predetermined constants, the primary problem involved in the solution of equation 1—289 (3) lies in determining the values of the lumped capacitances that must be inserted in the circuit to provide the correct value of C_x. In the average Pierce circuit the dynamic capacitance can be considered negligible when compared with the total load capacitance, so that C_x can be assumed to equal C_p and C_g in series to a first approximation. With C_p and C_g decided upon, approximate values of g_m can be had by equating the denominator of equation 1—289 (2) to zero and solving for the transconductance. The maximum and minimum equilibrium values of g_m coincide with the maximum and minimum expected values of R_e, respectively. Next, assuming that $g_m \approx I_p/E_g$, and that $E_g = I_gX_{Cg}$, where I_g is the crystal current, a vacuum-tube and plate voltage are chosen which will provide a maximum R_p, but which will not cause a crystal unit of any expected R_e to be overdriven. To determine the lumped capacitances that must be added to provide the correct values of C_p and C_g, it is first necessary to

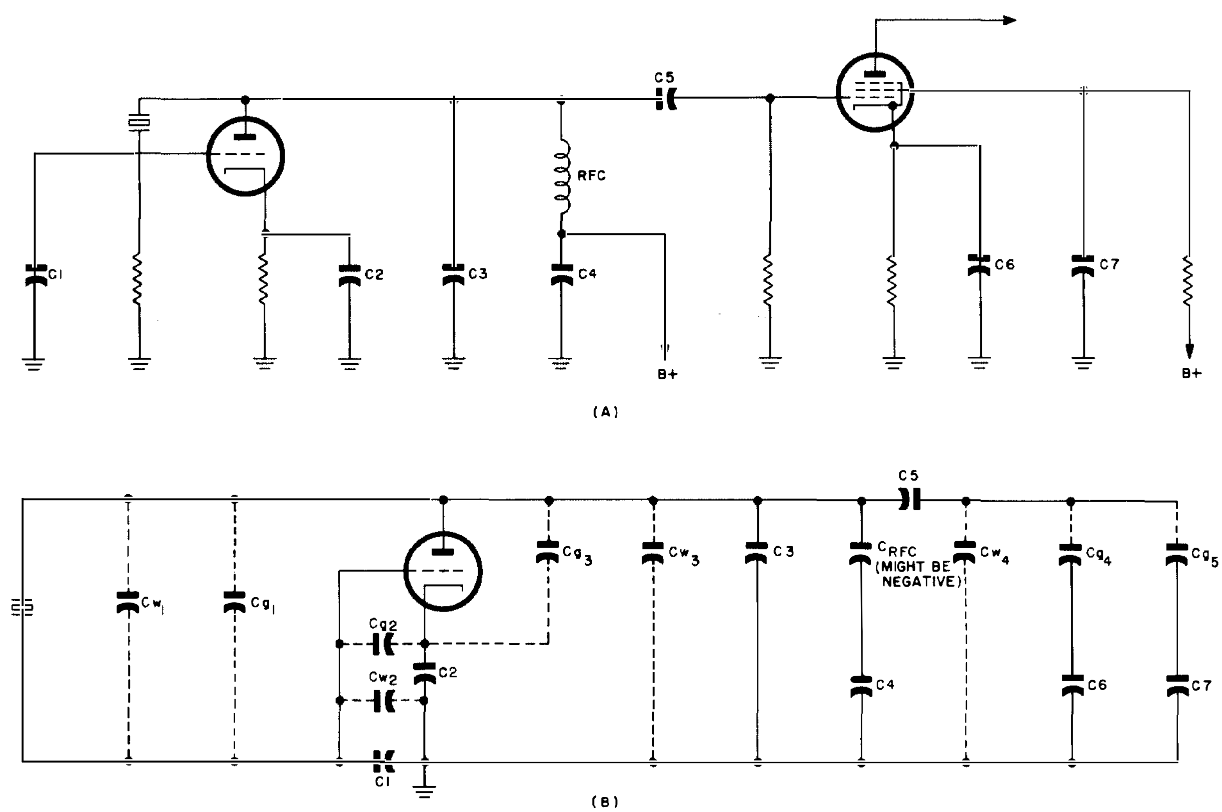


Figure 1-124. ***(A) Conventional Pierce oscillator and buffer-amplifier circuit. (B) Static capacitances of circuit (A). C_g = interelectrode capacitances. C_w = distributed capacitances of wires***

know the values of the stray static capacitances in the circuit. These stray elements effectively create a lower limit to the capacitance across the crystal unit. Obviously, they must not be allowed to exceed the total specified load capacitance. Preferably, they should be as small as possible and not be so distributed that an optimum capacitance ratio, C_g/C_p, cannot be achieved. The static capacitances to be considered in a conventional Pierce oscillator are illustrated in figure 1-124. Before the optimum values of C_1 and C_3 can be determined, an experimental circuit should be constructed with the various leads and circuit components reasonable facsimiles of those intended in the final production models. With C_1 and C_3 omitted, the static capacitances between plate and grid, plate and cathode, and grid and cathode can be measured. C_1 and C_3 can then be computed to provide the total required load capacitance.

1-291. In the early days of crystal oscillators, and even today where the oscillator is not required to meet a frequency tolerance less the 0.02 percent, the choice of grid and plate circuit capacitances was largely a matter of trial and error. Usually, the final choice was based upon the combination that provided the maximum activity for a given d-c plate voltage. For example, if the grid-to-cathode capacitance, C_g, of a Pierce oscillator is held constant while the plate load capacitance is varied from a minimum to a maximum, an activity curve, as measured by the gridleak current, is obtained similar to the one illustrated in figure 1-125 (A). Formerly, it was not unusual for the optimum capacitance to be considered a value slightly greater than that at which the activity is a maximum. In military equipment the principal consideration now is to ensure a given total load capacitance. As shown in figure 1-125 (B), increasing the plate-circuit capacitance causes the frequency to decrease. As the frequency decreases, so does X_e of the crystal unit, and at only one point along the curve will the crystal unit be operating into its rated capacitance. As stated previously, the circuit must provide the specified capacitance if there is to be an assurance that the required frequency tolerance is met when one crystal unit is replaced by another of the same type and nominal frequency. In an exceptional case, the most important consideration may be to maintain a fixed frequency relative to some frequency standard. For this purpose small variations in the load capacitance that can be made manually should be possible, but care should be taken that an operator is not to be able to vary the total more than is just sufficient to allow for a frequency variation equal to the bandwidth of the tolerance range. Otherwise, there can be a risk of overdriving a crystal unit, or of continuing in operation a defective crystal or other circuit component that should be replaced before a complete breakdown is threatened.

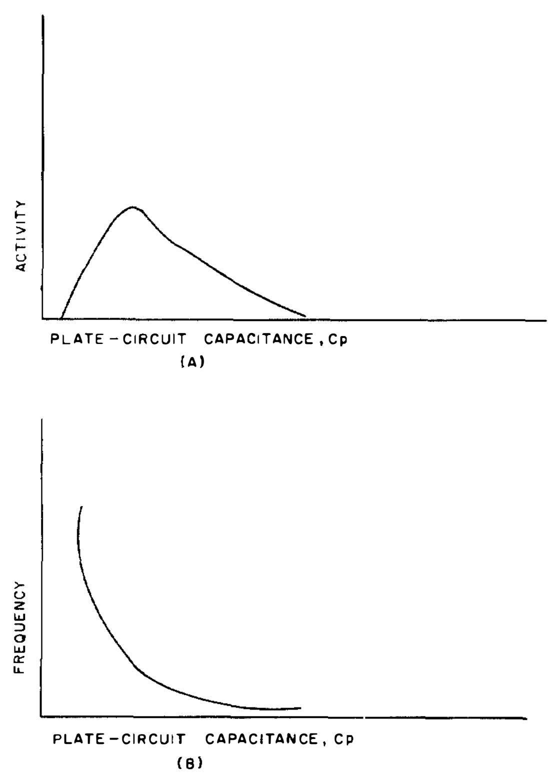
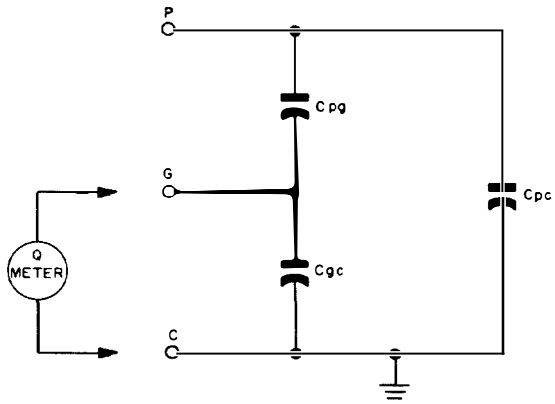


Figure 1-125. Variation of activity and frequency in Pierce oscillator as plate-circuit capacitance is increased

Measurement of Stray Capacitances In Pierce Circuit

1-292. In order to measure the stray static capacitances in a circuit such as that shown in figure 1-124, the lumped grid and plate capacitances, C_1 and C_3, as well as the crystal, should first be removed. The remaining elements can be assumed to form a three-element network as shown in figure 1-126. If an r-f choke is connected in the circuit, the frequency of the Q meter should be approximately the operating frequency of the oscillator. The measurements are made with all vacuum-tube voltages off. The three capacitances in figure 1-126 represent three independent variables, so that a minimum of three measurements is required to determine their values. A fourth measurement is desirable as a check on the accuracy of the first three. Any combination of measurements can be



(1) $C_{PG} = C_{pg} + \dfrac{C_{gc}\,C_{pc}}{C_{gc}+C_{pc}}$ (3) $C_{PC} = C_{pc} + \dfrac{C_{pg}\,C_{gc}}{C_{pg}+C_{gc}}$

(2) $C_{GC} = C_{gc} + \dfrac{C_{pg}\,C_{pc}}{C_{pg}+C_{pc}}$ * (4-9) $C_{XY} = C_{xy} + (C_{yz} \text{ OR } C_{xz})$

Figure 1-126. Equivalent network formed by the stray static capacitances of a vacuum-tube circuit. A minimum of three different measurements is required. The individual capacitances can be determined by solving simultaneously any three of the nine equations above when the respective terminal capacitances are known

**** Generalized equation for a measured terminal capacitance ($C_{XY} = C_{PG}$, C_{GC}, or C_{PC}) when either C_{xy} or C_{yz}, one of the two respective series-branch capacitances, is shorted out.***

made, the only restriction is that no two measurements are the same and that each of the capacitances is involved in at least one of the measurements. Three different measurements are possible between any one pair of electrodes. Thus, in measuring the capacitance between grid and cathode, the circuit is unchanged for the first measurement, the plate can be grounded for the second measurement, and shorted to the grid for the third. For the three measurements:

$$\text{(1st)}\quad C_{GC} = C_{gc} + \frac{C_{pg}\,C_{pc}}{C_{pg} + C_{pc}}$$

$$\text{(2nd)}\quad C_{GC} = C_{gc} + C_{pg}$$

$$\text{(3rd)}\quad C_{GC} = C_{gc} + C_{pc}$$

Theoretically, these three measurements could be sufficient, but one or more additional measurements are needed as a check in the event that the Q-meter leads or the shorting wires have significantly affected the readings. With C_{pg}, C_{pc}, and C_{gc} determined, the lumped capacitances for both grid and plate circuits can readily be determined.

BIAS VOLTAGE OF PIERCE CIRCUIT

1-293. Oscillator bias voltages are not as critical as are those of other types of vacuum-tube circuits, for usually it is neither wave shape, maximum output, nor maximum power efficiency that is of most importance, but simply a constant fundamental frequency. For this purpose, the Pierce circuit can be designed to obtain a maximum R_p with a minimum drive level, and the optimum bias voltage will be the one that most nearly answers the need. As a general rule, in the case of a vacuum tube that is not being driven to saturation, the effective R_p increases as the bias becomes more negative. If the bias is sufficient for the tube to be operated class B or class C, so that the plate current is cut off for an appreciable part of each cycle, such operation greatly improves the power efficiency. However, a bias developed across a gridleak resistor can never exceed the peak excitation voltage during stable oscillations. The excitation voltage, in turn, is limited by the current I_g, which must not exceed a value that would cause the losses in the crystal to exceed the permissible limit. When R_e is maximum, the limiting crest value of I_g is $I_{gm} = \sqrt{\frac{2P_{em}}{R_{em}}}$, where P_{em} is the maximum permissible drive level in watts, and R_{em} is the specified maximum effective resistance when the crystal unit is operating into its rated load capacitance. The maximum grid bias is thus

$$(\text{max})\ E_c = -\ Z_g\sqrt{\frac{2\ P_{em}}{R_{em}}} \qquad 1\text{—}293\ (1)$$

where $Z_g \left(\approx \frac{1}{\omega C_g}\right)$ is the r-f impedance between grid and cathode. It was found earlier (paragraph 1-284, equation 1—284 (8)) that a maximum R_p is to be had when $C_g = \frac{2C_x\ (\mu+1)}{\mu+2}$, or approximately when $C_g = C_p \approx 2C_x$. Equation (1) under these conditions becomes

$$(\text{max})\ E_c = -\ \frac{\sqrt{2\ P_{em}}}{2\omega\ C_x\sqrt{R_{em}}} \qquad 1\text{—}293\ (2)$$

Equation (2) is quite important in that it shows that the maximum grid bias obtainable with a maximum R_e is fixed by the crystal specifications. The maximum E_c for any given crystal unit may be obtained from equation (2) by substituting the actual R_e for R_{em}. The proper anode voltages for a given vacuum tube, or vice versa, to provide a required output are consequently also effectively predetermined by the crystal specifications. For maximum stability with any given vacuum tube in a conventional Pierce circuit, C_g and C_p should each be made equal to $2C_x$, or if the amplification factor is small, C_g/C_p should be made equal to $\frac{\mu}{\mu + 2}$, with the total capacitance in series made to equal C_x. With the capacitance so determined, the plate and screen voltages can be adjusted to give the desired output and excitation voltages. Since $E_p \approx E_g$, the output voltage is also effectively limited by the crystal specifications. With $C_g = C_p = 2C_x$, (max) $E_p = .707$ (max) E_c, where (max) E_p is the maximum r-m-s value of E_p, and (max) E_c is given by equation (2).

FIXED BIAS FOR PIERCE OSCILLATOR

1-294. It is not conventional to employ a fixed bias in a crystal oscillator, although it can be done—even to advantage in some cases. An r-f choke can be substituted for the gridleak resistance, thereby reducing the grid losses to an absolute minimum as long as the excitation is insufficient to overcome the bias and cause grid current to flow. In order for stable oscillations to be maintained, an increase in excitation must cause a decrease in amplification, and a decrease in excitation must cause an increase in amplification. When using a fixed bias, the choice of operating voltages is much more restricted than when employing gridleak limiting. Because of the more critical operating conditions, the replacement of one crystal unit with another having a different resistance may require additional circuit adjustments. If a fixed, class-B or class-C bias is used, a slight decrease in the amplitude of oscillations normally leads to the oscillations dying out all together. This is because the average amplification of the positive alternation of each cycle increases and decreases directly with the amplitude instead of inversely. For instance, with a class-C fixed bias, a decrease in the amplitude of one cycle would mean that the tube is cut off during a larger fraction of the succeeding cycle, thereby further decreasing the average amplification. On the other hand, if oscillations were once started, the tendency would be for the amplitude to build up until limited by grid and plate saturation. Only if limiting is provided by nonlinear elements, such as thermistors or varistors, in the external circuits is class-B or class-C fixed-bias operation possible if the tube itself is not to provide the limiting action. A fixed bias can be used if the tube is driven to saturation each cycle, but such operation is not practicable unless the utmost power is required from the oscillator, and in any event should not be attempted with the Pierce circuit. The only feasible application of a fixed bias

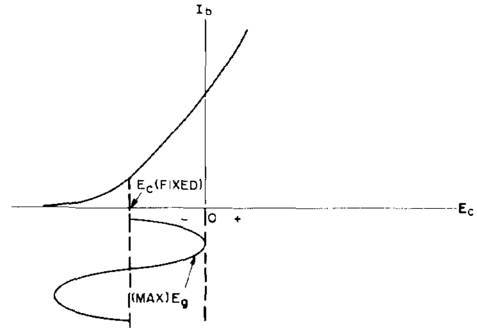


Figure 1-127. Point on E_c I_b curve at which stable oscillations can be achieved with fixed bias. For minimum grid losses, the peak of the maximum excitation voltage must not exceed the fixed E_c

in the Pierce oscillator is to operate the tube just above the knee of the E_cI_b curve (see figure 1-127). (E_c and I_b represent no-signal, d-c values of grid voltage and plate current respectively.) At the indicated operating point a slight decrease in the activity results in the average amplification of the negative alternation of the excitation cycle being greater, whereas the amplification of the positive alternation remains essentially constant. Thus, the over-all amplification of the weaker cycle is greater than that of the stronger, and a stable equilibrium is possible. An r-f choke would normally replace the gridleak resistor. Such operation practically eliminates the grid losses as long as the peak excitation voltage does not exceed the bias. Theoretically, then, the fixed bias permits a maximum Q in the feed-back circuit, and in this respect aids the frequency stability. There are, however, practical difficulties involved. If the C_g/C_p ratio of the external circuit is such that the oscillator is to operate at the minimum g_m, it may be difficult to find a vacuum tube that provides the desired transconductance when operated at suitable voltages just above the knee of the E_cI_b curve. The low transconductance can readily be achieved by using a remote-cutoff tube, but the amplitude stability will be more critical since the amplification of the positive alternations increases with amplitude, thereby tending to annul the limiting action of the negative alternations. Variations in the circuit capacitances have the following effects, which are essentially the same as those that occur with gridleak bias except that the amplitude variations are more pronounced in the fixed-bias circuit. A small decrease in the capacitance ratio, say by an increase in C_p, would mean that the voltage ratio,

$$E_p/E_g = C_g/C_p = \frac{g_m R_p Z_L}{R_p + Z_L} \approx g_m Z_L,$$

must become smaller. Since E_p/E_g varies inversely with C_p whereas Z_L varies inversely with C_p^2, an increase in C_p, as long as $R_p >> Z_L$, requires g_m to increase proportionately in order that equilibrium may be re-established. In other words, an increase in C_p must cause the excitation voltage to decrease. Conversely, a decrease in C_p causes g_m to decrease and the excitation voltage to increase. R_p varies in a direction opposite to that of g_m, but the percentage change in R_p is not as large as the percentage change in g_m. The change in g_m is approximately proportional to the change in C_p. If C_p is constant, but C_g is varied slightly, Z_L will be approximately constant, so that g_m must also vary directly with C_g in order for equilibrium to be maintained. Therefore, the changes in the excitation voltage with changes in C_g are similar to those with changes in C_p. When C_p and C_g are varied independently the total load capacitance C_x changes, and hence the frequency and X_e also change. If, during tuning adjustments, C_p and C_g are varied so that the same total capacitance is always maintained, then, in the region where $C_p \approx C_g$ the change in C_p is approximately equal but opposite to the change in C_g. In this case g_m, which has reached its minimum value, tends to remain constant, as do also the excitation voltage and the plate resistance of the tube. It should be understood that this optimum condition holds only for variations in capacitance that leave the total load capacitance unchanged. With C_x fixed, g_m must increase as the capacitance ratio is varied to either side of its optimum value. When C_g/C_p is made greater than one the change in g_m and the excitation is greater than when C_g/C_p is made less than one by an equivalent proportion. (e.g., Δg_m is greater when C_g/C_p is changed from 1 to 2 than when it is changed from 1 to ½.) An increase in g_m can occur only by virtue of a decrease in excitation voltage. Thus, if the grid impedance, Z_g, is made larger than the plate impedance, Z_p, the crystal current, I_g, must necessarily become smaller. Oscillations can thus be maintained with a smaller drive level. Nevertheless, optimum stability generally requires a maximum R_p, which, in turn, coincides with a maximum excitation voltage and minimum g_m.

1-295. According to equation 1-284(10), the optimum capacitance ratio will automatically cause the oscillator to seek an equilibrium when $R_p = (\mu + 1)\, Z_L$, or approximately when $R_p = \frac{\mu PI}{4}$, where PI is the performance index. Herein lies the principal limitation of fixed-bias operation of a Pierce circuit. The maximum to minimum values of the PI of a given type of crystal unit can be as much as 9 to 1 for the same frequency—and greater still if the oscillator is to operate over a wide frequency range. It may be difficult to find a tube to provide equivalent variations in the effective R_p unless the excitation voltage is to be so large that it drives the grid positive on the positive excitation peaks. In this case, the principal advantage of the fixed-bias—to maintain a maximum Q_f and to minimize the variations in the input impedance—is lost. Since an increase in R_p must be accompanied by an increase in the r-f plate current operating into a proportionally increased load impedance, the tendency will be for the crystal power (approximately equal to $I_p^2\, Z_L$) to vary directly with $(PI)^3$ or $\left(\frac{1}{R_e}\right)^3$. Unless there is some guarantee that the maximum R_e is not to be greater than twice the minimum expected value, some additional form of limiting must be used in the fixed-bias circuit, such as connecting a varistor across the r-f load, to ensure that the low-resistance crystal units are not over-driven. The fixed bias should not be less than three times that given by equation 1—293(2). The vacuum tube (preferably a pentode, because of its low plate-to-grid capacitance and high R_p) should be chosen and the anode voltage determined that permits operation at the lower end of the straight portion of the $E_c I_b$ curve. The optimum bias and plate voltages are best established by experiment. The principal problem is to ensure a sufficient output for crystal units of maximum R_e, without overdriving those crystals of minimum R_e. The average crystal unit has an R_e approximately one-third the maximum. An occasional crystal unit may have a value of R_e perhaps as small as one-tenth the maximum. It can be seen that a serious obstacle to the use of a fixed bias is that manual adjustments of the operating voltages are necessary when replacing crystal units, unless the plate circuit is to be rather heavily loaded. The output will tend to vary by a large factor from one crystal unit to the next. The same problem is encountered with the use of a gridleak bias, but voltage adjustments are not absolutely necessary, even under no-load conditions. A familiarity with fixed-bias operation is helpful, however, in that it aids the understanding of gridleak operation.

GRIDLEAK BIAS FOR PIERCE OSCILLATOR

1-296. The importance of having gridleak instead of fixed bias is two-fold: First, it permits a large initial surge of plate current, so that oscillations will build up quickly. If the tube were being operated class B or class C with a fixed bias, the bias would have to be removed before oscillations could build up at all. Second, it ensures a maximum stability in the output. If for any reason the excitation should increase or decrease, the d-c grid current and hence the bias follows the change, always acting in a direction that tends to annul the original change. When the oscillator is first turned on, the starting bias is zero regardless of the value of the gridleak resistance. Thus, insofar as the initial surge of current is concerned, the value of the gridleak resistance, R_g, is not a first-order factor. However, the value of R_g is significant in its effect upon the total build-up time. This effect is considered in paragraphs 1-304 and 1-305, where the conditions most favorable for oscillator keying are discussed. The present discussion considers only the effects of R_g upon the oscillator stability after the oscillations have reached a maximum amplitude. First, it is desirable that the grid losses be as small as possible, and that they at least can be considered negligible by comparison with the losses in the crystal unit. For a continuous flow of d-c grid current to be maintained, the grid must be positive with respect to the cathode at the positive peak of each excitation cycle. The amount of grid power that is dissipated, the extent to which the grid becomes positive, and the length of the period during which the grid is positive and electrons are flowing from cathode-to-grid depend upon how great a percentage of the total charge escapes through R_g during the remainder of the cycle. This, in turn, depends upon the ratio of the period of one cycle to the RC time constant of the grid circuit. This ratio, $1/R_gC_gf$, is seen to be equal to $\frac{2\pi\,|X_{Cg}|}{R_g}$. The smaller this ratio can be made, the smaller will be the percentage leakage of charge during one cycle, and the more nearly will the bias remain constant and equal to the peak excitation voltage. At high frequencies, ratios on the order of 1/50 and smaller are quite easily obtained. With the period of one cycle so short compared with the time it would take 63 per cent of the accumulated charge to leak off, it can be assumed that the bias voltage equals the peak excitation voltage in magnitude. Should the excitation voltage increase, the bias also increases. The peak excitation voltage is

$$E_{gm} = I_{gm}\,Z_g = 1.414\,I_g\,Z_g \approx |\,E_c\,| \qquad 1\text{—}296\ (1)$$

where E_c is the grid bias when $R_gC_g >> 1/f$, and I_{gm} is the peak r-f grid-circuit current. The bias voltage is also given by the equation

$$E_c = I_c\,R_g \qquad 1\text{—}296\ (2)$$

where I_c is the d-c grid current. If an r-f choke having an impedance that is large compared with R_g is connected in series with R_g, the r-f voltage across R_g becomes small compared with the d-c voltage. With this arrangement it is readily seen that the approximate grid power expenditure is

$$P_g = I_c\,E_c = E_c^2/R_g \qquad 1\text{—}296\ (3)$$

If the r-f choke is not present, so that the voltage across R_g varies sinusoidally from a peak of $-2E_{gm}$ to a peak of 0 on the positive alternation, the average squared voltage across R_g, equal to

$\frac{E_c^2}{2\pi}\int_0^{2\pi}(1 + \sin\omega t)^2\,d(\omega t)$, is found to be $1.5\,E_c^2$.

Thus, in the absence of an r-f choke,

$$P_g = 1.5\,E_c^2/R_g \qquad 1\text{—}296\ (4)$$

Clearly, if the grid losses are to be held to a minimum, R_g must be as large as possible. If an examination is made of several representative crystal oscillator circuits in actual production, it will be discovered that very few employ gridleak resistances higher than 100 kilohms, and only an occasional value of R_g is found higher than 0.5 megohms. The answer is principally to be found in the fact that the oscillator design is usually a compromise among several factors: (a) frequency stability, (b) output-voltage stability, (c) output control, (d) operating efficiency, (e) maximum economy in production costs, (f) minimum over-all weight and space requirements, (g) whether or not oscillator is to be keyed, (h) frequency range (i), value of C_g, (j) whether or not circuit is to permit switching from crystal to tuned circuit, and (k) the suppression of parasitic frequencies. Either a high or a low value of R_g can improve the performance in respect to any one of the factors above, depending upon what is required concerning the other factors. For example, a very large R_g can improve the frequency stability by reducing the grid losses, when only a small output is required. On the other hand, the same value of R_g could lead to both frequency and output voltage instability if maximum output or maximum operating efficiency were required. The effects of R_g relative to various factors listed above are discussed briefly in the following paragraphs.

Gridleak Resistance and Frequency Stability of Pierce Oscillator

1-297. The grid R_g can lead to frequency instability in two ways. As R_g is decreased, the grid losses load the feed-back circuit and require that the tank be operated farther away from resonance. R_p of the vacuum tube is reduced and its effect upon the frequency becomes more pronounced. On the other hand, if R_g is increased indefinitely, the oscillator can become self-modulated. This latter effect is more properly classed as a problem of output voltage stability, and is discussed in the paragraph 1-299. The principal frequency-stability importance of R_g is the degree by which it reduces the effective Q_f of the feed-back arm. (Q_f is the feed-back quality factor only from the point of view of phase rotation.) As defined by equation 1—272 (1)

$$Q_f \approx \frac{Q_{pgc}\,Q_g}{Q_{pgc} + Q_g}$$

where Q_{pgc} is the actual over-all effective Q of the feed-back arm, and Q_g is the effective Q of the input circuit. In order to analyze the effect of R_g in terms of the equivalent Pierce r-f circuit, we represent Z_g by an equivalent reactance, X_{Cg}' in series with an equivalent resistance, R_g', as shown in figure 1-128. The problem now is to determine R_g', such that

$$I_g^2\,R_g' = P_g \qquad 1\text{—}297\ (1)$$

where P_g is the power expended in the gridleak resistance. The power expended during the short current pulse when the grid is positive with respect to the cathode will be considered negligible compared with the power losses in R_g per cycle. Assuming that no r-f choke is used and that R_g is sufficiently large so that equation 1—296 (4) is approximately correct, equation (1) on substitution and rearrangement becomes

$$R_g' = \frac{1.5\,E_c^2}{I_g^2\,R_g} \qquad 1\text{—}297\ (2)$$

Since it has been assumed that R_g is large compared with X_{Cg}, then $X_{Cg}' \approx X_{Cg}$ and $E_c{}^2 \approx E_{gm}{}^2 = 2(I_gX_{Cg})^2$. On substitution in equation (2),

$$R_g' = \frac{3\,X_{Cg}{}^2}{R_g} \qquad 1\text{—}297\ (3)$$

If an r-f choke were used in series with R_g, the approximate value of R_g' would be

$$R_g' = \frac{2\,X_{Cg}{}^2}{R_g} \qquad 1\text{—}297\ (4)$$

The values of Q_g become

(without choke) $$Q_g = \frac{|X_{Cg}'|}{R_g'} \approx \left|\frac{R_g}{3\,X_{Cg}}\right| \qquad 1\text{—}297\ (5)$$

(with choke) $$Q_g \approx \left|\frac{R_g}{2\,X_{Cg}}\right| \qquad 1\text{—}297\ (6)$$

The equations for Q_{pgc} are

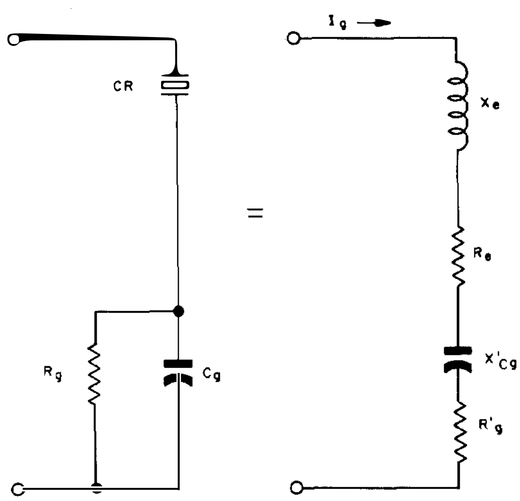
(without choke) $$Q_{pgc} = \frac{(X_e + X_{Cg})\,R_g}{R_g\,R_e + 3\,X_{Cg}{}^2} \qquad 1\text{—}297\ (7)$$

(with choke) $$Q_{pgc} = \frac{(X_e + X_{Cg})R_g}{R_g\,R_e + 2\,X_{Cg}{}^2} \qquad 1\text{—}297\ (8)$$

Assuming that $Q_{pgc}\,Q_g >> 1$, Q_f is given by the approximate equations

(without choke) $$Q_f = \frac{(X_e + X_{Cg})}{R_e - \dfrac{3\,X_e\,X_{Cg}}{R_g}} \qquad 1\text{—}297\ (9)$$

(with choke) $$Q_f = \frac{(X_e + X_{Cg})}{R_e - \dfrac{2\,X_e\,X_{Cg}}{R_g}} \qquad 1\text{—}297\ (10)$$



① $Q_{pgc} = \dfrac{X_e + X'_{Cg}}{R_e + R'_g}$ ③ $Q_f = \dfrac{Q_{pgc}\,Q_g - 1}{Q_{pgc} + Q_g}$

② $Q_g = \left|\dfrac{X'_{Cg}}{R'_g}\right|$ $\approx \dfrac{Q_{pgc}\,Q_g}{Q_{pgc} + Q_g}$

Figure 1-128. Equivalent Pierce feed-back circuit

In the previous discussions it has been supposed that the grid losses are kept negligible compared with the losses in the crystal, so that a Q_f equal to $\frac{X_e + X_{Cg}}{R_e}$ is approximately correct. Where this assumption cannot be made, those equations that define the frequency of the oscillator, such as equation 1—289 (3), can still be used for approximately correct answers if R_e is replaced by the appropriate denominator in equation (9) or (10). Thus, the effective frequency-determining resistance of the feed-back arm, can be defined as

$$R_{fe} = R_e + |3\,X_e\,X_{Cg}/R_g| \quad \text{or} \quad R_e + |2\,X_e\,X_{Cg}/R_g| \qquad 1\text{—}297\ (11)$$

On the other hand, in those equations that concern the equilibrium between the energy input and output of the feed-back arm, such as equation 1—289 (2), the effective feedback-circuit resistance to substitute in the place of R_e is the value

$$R_f = R_e + R_g' \qquad 1\text{—}297\ (12)$$

On multiplying equation (11) by R_g/R_e, it is apparent that if R_g is to have a negligible effect upon the frequency, it must be much greater than $\left|\frac{3\,X_e\,X_{Cg}}{R_e}\right|$. Similarly, if R_g is to have a negligible effect on the total feed-back power requirements, according to equations (3) and (12), it must be much greater than $3X_{Cg}^2/R_e$. If the oscillator is to be operated in the region of maximum R_p of the tube, $|X_{Cg}|$ will approximately equal $X_e/2$, and $\left|\frac{3\,X_e\,X_{Cg}}{R_e}\right|$ will approximately equal $\frac{3}{2}$ PI. Under these conditions, a good rule of thumb, if the other operating requirements permit, is to employ a grid resistance equal to 15 times the minimum permissible PI $\left(= \frac{1}{\omega^2 C_x^2\,R_{em}}\right)$, or greater, where R_{em} = (max) R_e.

Grid-Resistance Effects and "Class-D" Operation of Pierce Circuit

1-298. Typical curves showing the effect of the grid resistance upon the frequency and the frequency stability of a Pierce oscillator are shown in figure 1-129. The curves in figure 1-129 (A) were obtained for plate voltages (E_b) of 50, 60, and 70 volts under no-load conditions; the curves in figure 1-129 (B) were obtained for plate voltages of 75, 100, and 125 volts when a load resistance of 5000 ohms was connected across the plate circuit. The curves were plotted from measurements of an experimental Pierce oscillator during a USAF research project under the direction of E. Roberts at the Armour Research Foundation, Illinois Institute of Technology. The frequency deviation due to a change in R_p with a change in plate voltage is indicated by the frequency difference between the points on the curves that correspond to the same values of grid resistance. The four sets of curves for each of the two load conditions were obtained by maintaining the grid-to-cathode capacitance constant and varying the plate-to-cathode capacitance. The bottom set of curves in each graph represents the closest approach of the four sets to a C_g/C_p ratio of 1, and the top set represents approximately the largest C_g/C_p ratio (maximum E_p) at which oscillations can be maintained. As is to be expected the frequency deviation due to changes in voltage is greater for the larger than for the smaller values of C_g/C_p. However, part of this greater deviation is due to the fact that in the circuit from which these curves were plotted, the plate capacitance was obtained from a capacitor paralleled by an inductor. At the minimum effective value of C_p, the capacitor and inductor approach a state of parallel resonance, so that the rate of change in the equivalent plate reactance with a change in R_p is larger than would be the case if no inductor were present. Also, the fact that the total load capacitance facing the crystal is smaller at the higher frequencies contributes to the frequency deviation. Partially counteracting this latter condition is the increase in the effective Q_e of the crystal and in the Q_f of the feed-back circuit as a whole because of the rise in frequency. There are three variables influencing the frequency that are affected by a change in the grid resistance. The first of these is the Q_g of the grid-to-cathode impedance. As R_g becomes relatively small, Q_g becomes the dominant factor determining the phase of E_p with respect to E_g. As Q_g decreases, the tank circuit must appear more capacitive. Thus, the reactance of the crystal unit in the feed-back arm, and hence the frequency, must increase. This is indicated by the sharply rising tails of the curves in figure 1-129. The second frequency-determining factor affected by R_g is the effective grid-to-cathode capacitance, C_g. As R_g decreases, C_g effectively increases. That is, if R_g and C_g are represented by an equivalent R_g' and C_g' in series, it can be shown that C_g' increases with a decrease in R_g. As a result of this effect, as R_g becomes small, the frequency tends to decrease; but, as indicated in the upper sets of frequency stability curves of figure 1-129, when C_p is small compared with C_g, changes in C_g have little effect since the total load capacitance is approximately equal to the smaller capacitance. On the other hand, when the C_g/C_p ratio is small, as indicated by the lower-frequency curves in the figure, the changes in the effective C_g have a measurable effect upon the frequency. It can be seen at the low values of grid resistance that the effective increase in capacitance greatly diminishes the rise in frequency that would otherwise occur because of the decrease of the grid-to-cathode, Q_g. Indeed, the bottom curves in figure 1-129 (A) show that for the particular circuit and crystal unit the two opposing frequency effects of R_g apparently cancel each other when R_g is in the neighborhood of 100,000 ohms. This same set of curves indicates that a minimum frequency deviation with plate voltage occurs when the grid resistance is approximately 200,000 ohms. It is with some diffidence that we attempt to explain the reason why this particular value of R_g should provide a point of maximum frequency stability. Rather than interpret the effect as due to a possible optimum ratio, or as due to a possible variation of R_g with plate voltage that tends to cancel the effect of the variation in R_p, it seems more likely that the optimum results are due to a coincidence between the third frequency factor mentioned above and the characteristics of the tube, a 6C4 (triode), that was used in the test circuit. This third frequency factor is the average grid bias, which tends to increase and decrease with R_g, although the variations are not pronounced when R_g is large. As the bias increases, so also does R_p, which in turn causes the frequency to decrease. This can best be seen from an examination of the curves at the top of figure 1-129, where the effects of R_g on C_g' are negligible as they affect

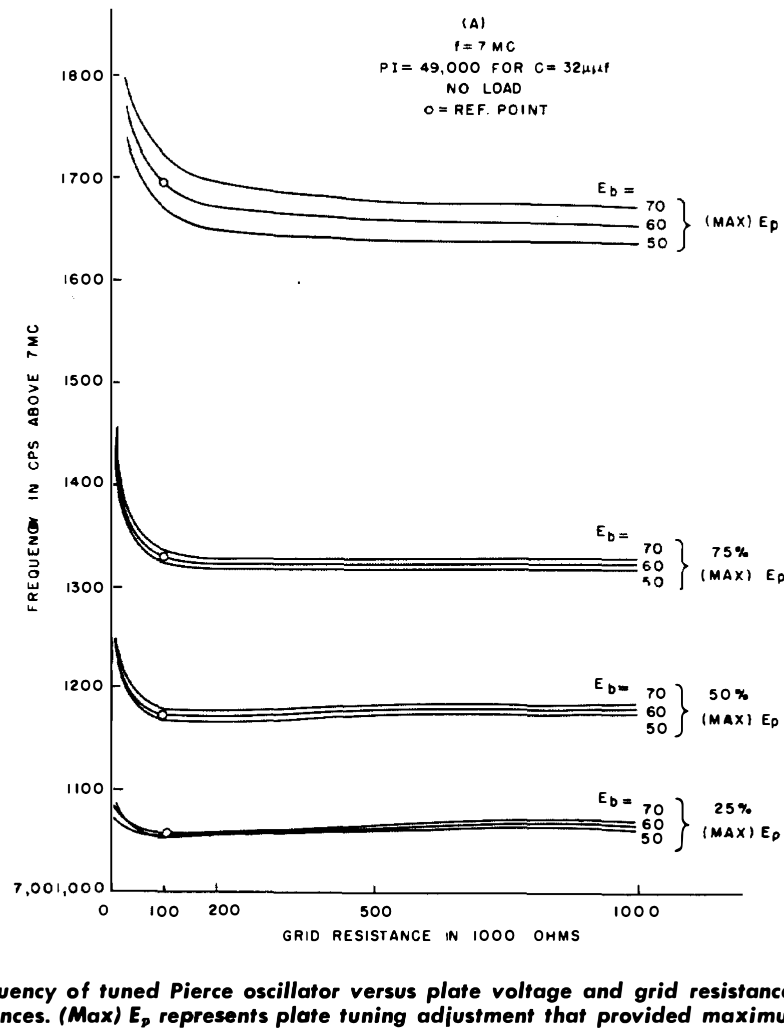


Figure 1-129. Frequency of tuned Pierce oscillator versus plate voltage and grid resistance for various plate-tuned load capacitances. (Max) E_p represents plate tuning adjustment that provided maximum r-f plate voltage when grid resistance and d-c plate voltage were values indicated by zero reference point. A 7-mc CR-18/U crystal unit was used, having a PI of 49,000 ohms when operating into a rated load capacitance of 32 $\mu\mu f$

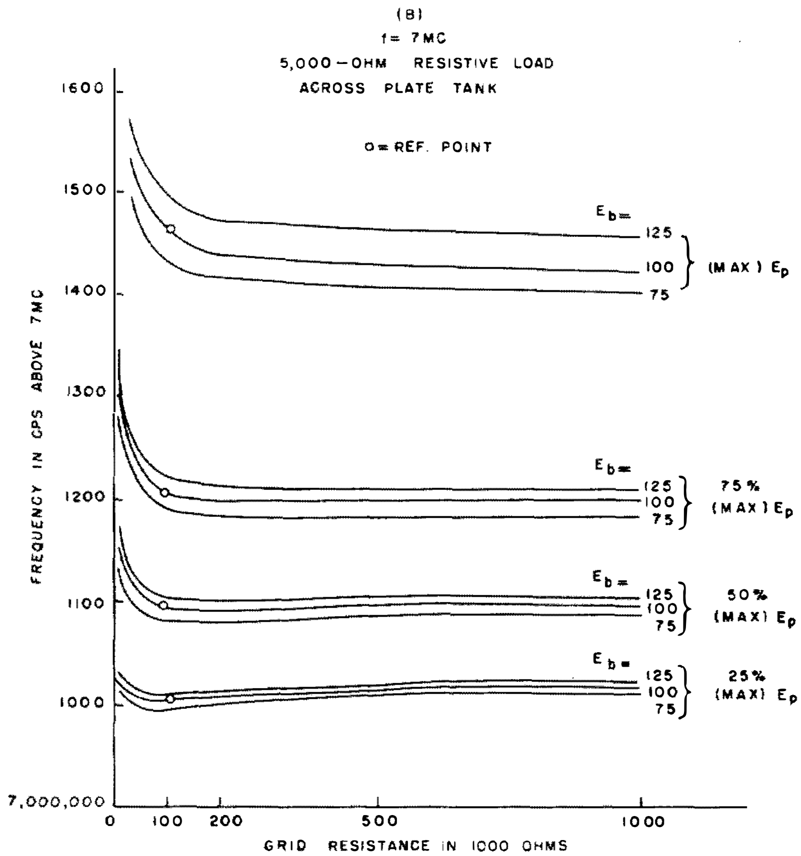
the frequency. As R_g increases, the bias approaches as a limit the magnitude of the peak excitation voltage. Thus, R_p also rises to some limiting value, causing the frequency to level off to some minimum value. If the plate voltage is increased when the tube is biased below the straight portion of the E_cI_b curve, one result is a decrease in R_p, which, in the curves of figure 1-129, clearly causes the frequency to increase. However, an increase in the plate voltage also causes an increase in the excitation voltage, and hence in the grid bias. Thus, it may well be that the point of maximum frequency stability as indicated in the bottom curves of figure 1-129 (A) is the result of an increase in bias with the increase in plate voltage just sufficient to hold R_p constant. A class of operation such that the percentage change in R_p due to a change in E_b is annulled by a percentage change in R_p due to a change in E_c, or vice versa, suggests interesting possibilities in stabilizing the plate resistance by methods other than plate-supply regulation. There is no evidence that this type of operation has been investigated, but on the strength that possibilities exist for practical application in oscillator circuits not employing a fixed bias, the name "class-D" operation is proposed. Three subclasses are possible: "D_1," where changes in E_b can occur more or less independently and E_c is the dependent variable; "D_2," where E_c can be considered independent and E_b dependent; and "$D_{1,2}$" where E_b and E_c are mutually dependent. Mathematically, this class of operation can be defined by the following equations:

(Class D)
$$\frac{\Delta R_p}{R_p} = \frac{1}{R_p}\left(\frac{\partial R_p}{\partial E_b}\right)_{E_c = \text{const.}} \Delta E_b + \frac{1}{R_p}\left(\frac{\partial R_p}{\partial E_c}\right)_{E_b = \text{const.}} \Delta E_c = 0 \qquad 1\text{—}298\ (1)$$

(Class D_1)
$$\Delta E_c = \frac{\partial E_b}{\partial E_c} \Delta E_b \qquad (\Delta E_b \text{ independent}) \qquad 1\text{—}298(2)$$

(Class D_2)
$$\Delta E_b = \frac{\partial E_b}{\partial E_c} \Delta E_c \qquad (\Delta E_c \text{ independent}) \qquad 1\text{—}298\ (3)$$

(Class $D_{1,2}$) Equations (2) and (3) both apply.



Equations (1), (2), and (3) sufficiently define "class-D" operation, but the parameter, R_p, represents a with-signal average plate resistance, and not the instantaneous or static resistances represented by conventional R_p curves; although the static values would apply if the signal amplitude were small relative to the bias, as would be the case if agc were being used. An increase in plate voltage can generally be expected to cause an increase in excitation and grid bias in the conventional oscillator circuits. This action, in turn, can be expected to cause an increase in the average R_p. Thus, if "class D" is to be in effect, it is necessary that an original increase in E_b cause R_p to decrease by the same amount as the change in E_c will cause it to increase. For this to occur, the plate characteristic curves must show positive slopes that increase with plate voltage; that is, the E_bI_b curves must be curving upward in the direction of increasing E_b at the operating voltage, as is quite characteristic of triode curves. Pentodes do not show this characteristic if the screen voltage is held constant, since the plate resistance tends to increase with increasing plate voltage. However, if the screen voltage varies with, and in the same direction as, the plate voltage, as can be the case when the two voltages are obtained from the same source, plate characteristics can be achieved similar to those of triodes, but with the advantages of larger values of R_p and an independent variable (E_{c2}) by which the rate of change of R_p with E_b, $(\partial R_p/\partial E_b)$, can be adjusted. Now there is an additional implied condition that must be met if class-D effects are to be achieved in conventional oscillator circuits. This is the requirement that g_m also remain constant. For example, by equation 1—289 (2) which is repeated here

$$R_p = \frac{X_{Cp}\,(X_e + X_{Cr})}{R_e - g_m X_{Cp}\,X_{Cg}}$$

it can be seen that feed-back equilibrium in a Pierce circuit requires that as long as R_p and the external circuit parameters remain constant, so also must g_m. An analogy here is to be found in class-A operation, which is defined by the operation of the tube along the straight portion of the E_cI_b curve, i.e., in a region of constant g_m. Since the principal purpose of class-A amplification is a distortionless output, by implication a necessary requirement is that the operation also be in a region of constant R_p. In "class D," on the other hand, a constant effective R_p is the sufficient definitive condition, but in application a constant effective g_m is a necessary implication. If g_m is to remain constant, equations for Δg_m similar to (1), (2), and (3) for ΔR_p must hold simultaneously. This can be achieved in one of two ways, or a combination thereof. Assume that the plate voltage increases. If R_g is sufficiently small for the positive excitation peak to drive the grid reasonably far above the zero point, an increase in excitation, although decreasing the average g_m on the negative alternation, can annul this effect by increasing the average g_m in the region above the zero grid voltage. For this to occur, the tube would have to be operated at plate voltages low enough for the zero grid point to lie well within the bend of the E_cI_b curve. This, indeed, was the state of the 6C4 tube when the curves of figure 1-129 (A) were determined. Presumably, under no-load conditions and a small C_g/C_p ratio, as the grid resistance was decreased, the changes in the positive excitation peaks with changes in plate voltage were just sufficient to maintain g_m approximately constant for values of R_g between 100K and 200K. If R_g is very large, the effect above is negligible, since any increase in the positive excitation above the zero grid voltage point becomes minor compared with the total increase of the negative alternation. For g_m to be stabilized when R_g is large, the change in plate voltage must make a change in the cutoff voltage comparable to the change occurring in the excitation voltage. Plate characteristics most probably favorable to "class-D" operation appear to be had with low plate voltages. Unlike the R_p and g_m in the conventional classes of amplifier operation, where R_p and g_m can be varied independently, this condition cannot exist in class-D operation of conventional oscillators, since R_p and g_m are tied together by the feed-back energy requirements at equilibrium. Any condition that would stabilize the one, would automatically stabilize the other. It is only R_p, however, that directly affects the phase of the feedback. Once R_p becomes large relative to the impedance across the tube, the percentage variations in g_m become very small, so that g_m can be assumed to be a constant for all practical purposes. It is R_p that requires critical attention if it is to be held constant. As discussed in paragraph 1-342, the curves shown in figure 1-146 strongly suggest the possibilities of "class-D" operation in the case of a Miller circuit. The solution of the "class-D" equations for a given vacuum tube and circuit can probably be approximated graphically, using families of R_p curves versus E_b and E_c, or curves of the deviations of the R_pE_b and R_pE_c curves. If rates of change of E_c with E_b can be obtained that provide a solution for equation (1) when the values of E_c and E_b are practicable, the possibility exists that an oscillator of any type, parallel- or series-resonant, using a gridleak bias (or agc) can be designed so that for all practical purposes it is independent of small fluctuations in the plate-supply voltage. Empirically, "class-D" operation is indicated at the point or points where the frequency deviation curves of an oscillator change in sign, or where frequency curves, such as those in figure 1-129, cross or touch each other. A full analysis of this class of vacuum-tube operation is beyond the scope of this handbook. It is suggested here only as a possible line of inquiry.

Gridleak Resistance and Output Voltage Stability of Pierce Circuit

1-299. The stability of the output voltage depends largely upon how readily the gridleak bias can follow small fluctuations in the excitation voltage. Imagine, for example, that the vacuum tube is being operated class C, and that after equilibrium is reached the positive peak of a certain excitation cycle happens to be slightly higher than the average. The average bias during the succeeding cycle will thus be slightly more negative than is normal, so that during this period the tube is conducting a smaller fraction of the time, and the peak excitation voltage will drive the grid less positive than before. This means that the amplification during this cycle will be less than the amplification during the preceding cycle. If the R_gC_g time constant is extremely large compared with the period of a cycle, the bias remains relatively fixed for the duration of several cycles. In which case the peak of several succeeding cycles must rise to progressively lower points on the E_cI_b curve. The oscillations will continue to decay until a sufficient amount of the bias charge of C_g has leaked through R_g to permit oscillations to again build-up. For this reason R_g cannot be increased indefinitely without the risk of the oscillator becoming self-amplitude-modulated. As R_g is gradually increased, the amplitude sooner or later begins rising and falling at a radio-frequency rate. If R_g is further increased, the modulation of the output can fall within the audio range. Finally, with extremely large values of R_g, the circuit behaves as a damped-wave blocking oscillator. Now, assume that R_g is infinite. As oscillations build up, the bias for each cycle is essentially the same as the peak excitation voltage of the preceding cycle. Eventually a peak excitation voltage is attained which causes the bias for the next cycle to be too great for the circuit to be resupplied with all the energy that will be lost during the period of the cycle. If the peak bias is equal to or greater than the cutoff bias, the oscillations will die out completely, since any decrease in excitation with class-C bias means a decrease in average amplification. To avoid this possibility, it is important that sufficient electrons escape from the grid so that, at the beginning of the cycle immediately following the first peak cycle, the bias will have returned to approximately the same starting point. Expressed in another way, to avoid the intermittent activity, there must be an assurance that the positive peak of every cycle will drive the grid positive. This assurance is to be had for all operating conditions if the bias voltage decreases at a greater rate than would the positive excitation voltage peaks if the tube were cut off for a complete cycle. In practice, the vacuum tube can be conducting in polar opposition to E_p and hence effectively supplying energy to the circuit during the entire negative alternation of an E_p cycle. Nevertheless, if there is an assurance that the bias voltage drops as fast as the peak excitation voltage when no energy is being supplied to the circuit, the bias reduction is certainly sufficient if the net rate of energy-loss is reduced by virtue of a variable release of energy by the tube throughout a large part of each cycle. The problem, then, becomes one of first determining the percentage change in the peak excitation voltage that would occur during the period of one cycle if the tube were suddenly cut off.

1-300. At the instant that I_g is a maximum, the voltages across the reactances in the tank circuit are zero, and none of the circuit energy is stored in the capacitances. All the stored energy at that instant is in mechanical form, and is equal to the kinetic energy of the crystal as it swings through its position of zero potential energy. As discussed in paragraph 1-249, this stored energy is equal to I_s^2L, where I_s is the series-arm current, and L is the equivalent series-arm inductance. Now, when the crystal appears as an inductance, I_s is approximately equal to I_g plus the current, I_{Co}, which flows through the shunt capacitance, C_o, of the crystal unit. (Only the unsigned magnitudes of I_g and I_{Co} are considered here.) We can say, approximately, that the stored energy is equal to $(I_g + I_{Co})^2L$. As is also discussed in paragraph 1-249, the ratio of stored energy to the energy dissipated per radian, is equal to the Q of the circuit, which in this case is effectively $\frac{\omega L(I_g + I_{Co})^2}{(R_e + R_g' + R_L')I_g^2}$, or the equivalent value, $\frac{\omega L(I_g + I_{Co})^2}{R(I_g + I_{Co})^2 + (R_g' + R_L')I_g^2}$, where R_L' is the equivalent load resistance when represented as in series with the plate-circuit capacitance, and R is the series-arm resistance of the crystal. If R_g' and R_L' can be considered negligible, the circuit Q will be the actual crystal $Q\left(=\frac{\omega L}{R}\right)$. Assume that the plate current is cut off for an entire cycle, which is a period of 2π radians. The fraction of the energy dissipated during this time is approximately equal to $2\pi/Q = R/fL$, if Q is of sufficient magnitude that the percentage decrease in current is not large. Since the energy is proportional to the square of the voltage, the equivalent decay in peak excitation voltage is $E_{gm}\sqrt{R/fL} \approx |E_c|\sqrt{R/fL}$. If E_c, the grid bias, whose magnitude only we shall consider, is to decrease at the same rate, the bias charge, equal to C_gE_c, must leak through R_g at an average rate of $C_gE_c\sqrt{R/fL}$ during the period of one cycle. Thus

$$\text{(min) } I_c = \frac{E_c}{\text{(max) } R_g} = fC_gE_c\sqrt{R/fL} \qquad 1\text{—}300\ (1)$$

The maximum safe value of R_g for all operating conditions, if R_g' and R_L' are negligible compared with R_e, according to equation (1) is

$$\text{(max) } R_g = \sqrt{L}\Big/C_g\sqrt{fR} \qquad 1\text{—}300\ (2)$$

Since $\sqrt{L} = 1/\omega\sqrt{C}$, where C is the equivalent series-arm capacitance, and since $C \approx \frac{C_o}{r}$, in the case of partially plated elements, where r is approximately equal to the theoretical capacitance ratio, r_e, given in figure 1-95, then, on substitution in equation (2)

$$\text{(max) } R_g = \frac{\sqrt{r}}{\omega C_g\sqrt{fRC_o}} \qquad 1\text{—}300\ (3)$$

In paragraph 1-297 it was shown that when $C_g = 2C_x$, if R_g is to be considered negligible it should be at least 15 times the minimum permissible PI. Assume that $R_g/\text{(min) PI} = k$, (k is not to be interpreted here as a symbol for any quantity other than the ratio defined) and that it is desired that $k = k_m$, its maximum value consistent with equation (3), above. Let it also be assumed that R is approximately equal to R_{em}, the maximum permissible value of R_e, that $C_g = 2C_x$, and that $C_o = C_{om}$, the maximum permissible shunt capacitance specified for the crystal unit. Then,

$$\text{(max) } R_g = k_m \text{ (min) PI} = k_m/\omega^2C_x^2 R_{em}$$

also

$$\text{(max) } R_g = \sqrt{2\pi r}\Big/2\omega\, C_x\sqrt{\omega C_{om}\, R_{em}}$$

Thus,

$$\text{(max) } R_g/\text{(min) PI} = k_m = \frac{\omega C_x\sqrt{2\pi\, r\, R_{em}}}{2\sqrt{\omega\, C_{om}}} \qquad 1\text{—}300\ (4)$$

or

$$k_m^{\,2} = \pi\, r \mid \text{(min) } X_{C_o} \mid /2 \text{ (min) PI} \qquad 1\text{—}300\ (5)$$

It will be found in practice that the limiting values of k_m given by equations (4) and (5) are normally smaller than the minimum desired value of 15. If this should be the case in an actual circuit, the assumption that the power losses in the grid circuit are negligible can no longer be made, and the actual value of k_m would be even less than that given above. The factor $\sqrt{R/fL}$ in equation (1) is derived upon the assumption that only the crystal losses are significant. If this is not to be the case, this factor should be replaced by one equal to $\sqrt{\frac{\text{total energy expended per cycle}}{\text{energy stored}}}$. Now, even though it would seem from equations (4) and (5) that R_g cannot be safely made more than 5 to 10 times larger than the minimum PI, particularly if an AT cut is employed, since it has a value of r of only 250, and since C_x is normally no greater than 4 or 5 times C_{om}, it should be remembered that the value of k_m above is based upon the assumption that no energy is being fed to the circuit during an entire cycle, so that the net loss is equal to the gross loss. This condition is only approached in high-efficiency class-C circuits where the operating bias is several times the cutoff bias. Except in the case of power oscillators, such operation is not feasible because of the high operating voltages that are required. The larger the fraction of the cycle during which the tube is conducting, the larger the ratio of the usable R_g to that given by equation (3). If the tube is conducting one-half the time, class-B operation, the maximum safe R_g is more than twice that given by equation (3). For class-B and class-C operation, the output stability is almost entirely dependent upon the automatic adjustment of the bias, for any decrease in signal strength will mean a decrease in over-all amplification unless the bias can drop immediately to allow more energy to be fed to the circuit. On the other hand, it was found in paragraph 1-294 that if an oscillator tube is operated at a bias immediately above the knee of its E_cI_b curve, the bias can remain fixed and the variations in excitation directly produce a change in amplification that tends to annul the original variation. If the tube voltages are so selected that a gridleak bias at equilibrium is also at the optimum fixed-bias point, then limiting can be achieved both from the grid-leak action and the excitation swings. Under these conditions, R_g can be safely increased to values beyond one megohm, even at high frequencies. As a design consideration, however, the gap between the theoretical and the practical solution can prove quite wide. Among the optimum-bias bugs that resist extermination, there is the difficulty of finding a vacuum tube having the desired operating characteristics, and once found, there is the additional problem of maintaining an optimum operating state with crystal units having different values of effective resistance. These problems are discussed in some detail in succeeding paragraphs. The main problem is to reduce the grid losses to negligible proportions without endangering the output voltage stability. This can normally be done with any parallel-resonant crystal oscillator if the tube is conducting throughout most of each cycle.

Gridleak Resistance and Output Control in Pierce Circuit

1-301. If it is necessary for a Pierce oscillator to provide a higher voltage output than can be obtained under the conditions of maximum frequency stability, the C_g/C_p ratio can be increased. If the total load capacitance is to remain constant, C_g will necessarily be larger, and the excitation voltage smaller, so a smaller value of R_g can be used without the grid losses becoming significant. If the capacitance ratio is to be adjustable in order to permit an operator or technician to control the output voltage, R_g cannot be made larger than that value which would permit a stable output with the largest operable value of C_g at the highest frequency at which the oscillator is to be used. If such an adjustment is to be provided in a Pierce circuit, C_g and C_p should be so ganged as to always provide a constant load capacitance. This problem is discussed in paragraph 1-318. Insofar as the grid-to-cathode resistance is concerned, the maximum safe value of R_g becomes less if C_g is to be variable than otherwise. Without changing the C_g/C_p ratio, larger outputs can be achieved by reducing the value of R_g to a point where the grid leakage is so great that the average bias is considerably smaller than the peak excitation voltage. With this the case, the oscillations must build up to higher amplitude levels before equilibrium can be reached. Although the maximum excitation is still fixed by the rated drive level of the crystal unit, the output can be controlled somewhat within this restriction by a variable R_g. At the higher frequencies, this method of output adjustment requires such low values of R_g that the grid losses seriously affect the frequency stability. However, at very low frequencies, a variable R_g could be feasible as a means of adjusting the output of a Pierce circuit to a desired level when one crystal unit is replaced by another of different effective resistance. Although such a design feature has no particular recommendation, it could be preferred over those methods of output control that require adjustment of the C_g/C_p ratio, which risk changes being made in the total load capacitance. With grid control, the lowest adjusted value of R_g could be designed to provide the desired output when a crystal unit of maximum effective resistance (minimum PI) is connected in the circuit; whereas the larger values of R_g could ensure the same output with some theoretical minimum value of effective resistance. Since the crystal current, I_g, is practically constant as long as the output voltage E_p is constant, the power losses in the crystal, equal to $I_g{}^2R_e$, tend to vary directly with R_e, as long as E_p is held constant by adjustments of R_g. Under those conditions where the capacitance ratio does not change, a maximum crystal drive level is required for the crystal unit of maximum R_e, and a minimum crystal drive level when R_e is a minimum—the reverse of those conditions discussed in paragraph 1-294 when a fixed bias instead of a fixed output is assumed.

1-302. As applied to crystal oscillators in general it cannot be said that a variable gridleak resistance is advisable except for test purposes or unless its purpose is to obtain the minimum possible grid losses when changing from one crystal unit to another. As an output-voltage control device other methods are generally to be preferred. Except at very low frequencies, the resistance values necessary to appreciably lower the average bias are too small to prevent the grid losses from becoming a significant frequency-determining factor. This statement, of course, only expresses a general rule, and in specific instances the inter-relations among the circuit variables may be such as to annul the effects upon the frequency. For example, the bottom set of curves in figure 1-129(A) is to be expected theoretically to indicate a greater frequency stability when R_g is 1 megohm rather than when it is 0.2 megohm, but this effect was not observed. Figure 1-130 shows curves of output voltage obtained from the same experimental oscillator that was used in plotting the curves of figure 1-129. Although the curves are plotted as output-voltage versus crystal driving power, it should be understood that the actual independent variable for each curve is the plate voltage. Each curve represents a particular value of grid resistance. The cross lines intersect the curves at points corresponding to the same values of plate voltage. From figure 1-130 it can be seen that large percentage changes in the grid resistance can cause changes in the output voltage on the order of 30 per cent or so, but which increase sharply as R_g becomes small.

1-303. If an adjusable output voltage is desired, probably the best solution to the problem is to use a screen-grid tube having an r-f-bypassed, variable, voltage-dropping resistor in series with the screen supply voltage. Varying this resistance will control the output of the tube and the crystal driving power. The maximum permissible output voltage must be determined on the assumption that the crystal unit has the maximum permissible R_e. Since $I_g \approx \frac{E_p}{X_e + X_{Cg}}$ and the maximum permissible $I_g = \sqrt{P_{cm}/R_{em}}$, where P_{cm} and R_{em} are the maximum crystal driving power and effective resistance, respectively, then the maximum permissible *constant* E_p is

$$(\max)\ E_p = (X_e + X_{Cg})\sqrt{P_{cm}/R_{cm}} \qquad 1\text{—}303\ (1)$$

If it is assumed that $X_{Cg} \approx X_{Cp} \approx \frac{X_x}{2}$, where $X_x (= -1/\omega C_x)$ is the total load reactance equal and opposite to X_e, equation (1) becomes

$$(\max)\ E_p = \sqrt{\frac{P_{cm}}{R_{em}}} \Big/ 2\omega C_x \qquad 1\text{—}303\ (2)$$

According to equation (2), the maximum permissible constant E_p varies inversely with the crystal frequency. If the oscillator is to be used at more than one frequency, and at the same time is to provide the same output voltage regardless of the frequency, the maximum E_p is that value given by equation (2) for the crystal unit of highest frequency, assuming the crystal specifications are the same for all frequencies. With the ratio of C_g/C_p approximately equal to one, equation (2) also gives the value of E_g, which obviously will also remain constant. R_g can be made quite large, so that $|E_c|$ will approximately equal $E_p\sqrt{2}$. With E_c constant, and with the plate voltage E_b also assumed to be constant, the operating position of the tube on the E_cI_b curve largely becomes the function of the screen voltage. As the screen voltage is increased, g_m increases, which means that the slope of the

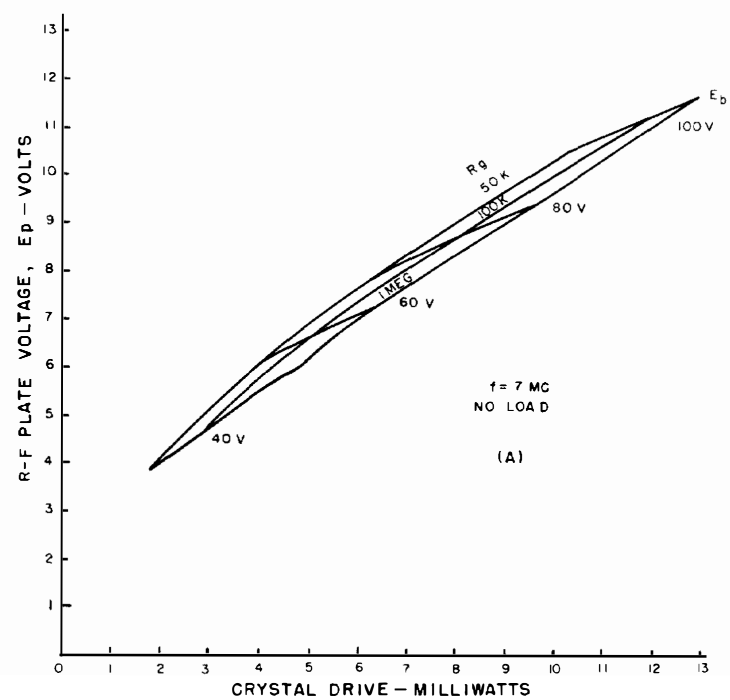


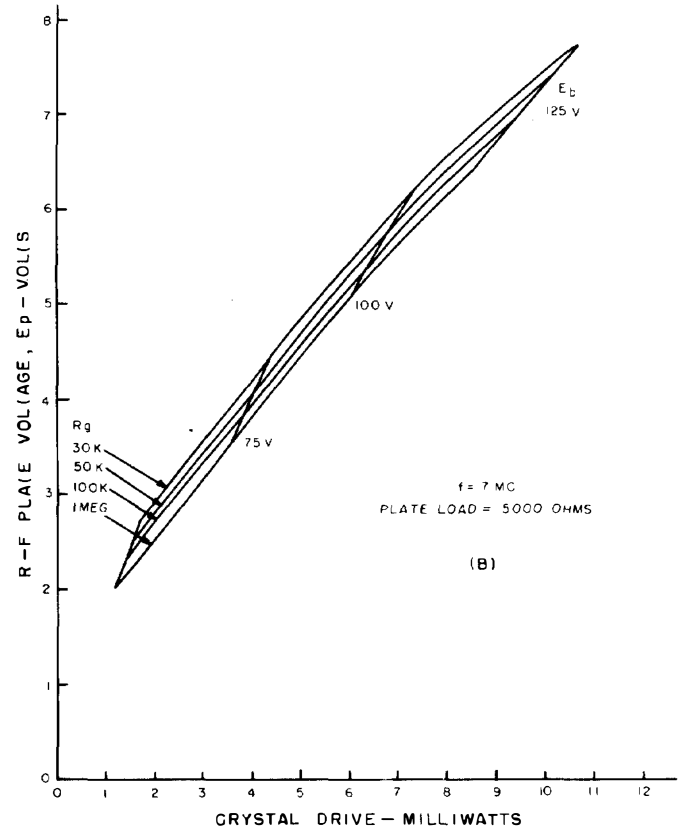
Figure 1-130. Output curves of tuned Pierce oscillator for different values of grid resistance when reactance of tuned plate circuit is adjusted for output voltages equal to 50 percent of the maximum attainable. A 7-mc CR-18/U crystal unit was used, having a PI of 49,000 ohms when operating into its rated load capacitance of 32 $\mu\mu f$

E_cI_b curve becomes steeper. Also, the cutoff bias is increased. Since E_c is being held constant, the effect is one of shifting the operating bias, percentage-wise, closer to or farther up the straight portion of the E_cI_b curve. From the point of view of using as large a value of R_g as is possible, it is desirable that the operating position be just above the knee of the E_cI_b curve when the screen voltage is to be a maximum, i.e., when R_e of the highest-frequency crystal unit is a maximum.

Gridleak Resistance and Oscillator Keying of Pierce Circuit

1-304. If avoidable, a crystal oscillator should not, itself, be keyed. For one reason, the oscillation build-up time is not negligible if rapid telegraph keying is desired. As the operable speed limit is approached the wave shape becomes distorted and the harmonic output is considerably increased. Even the keying of a crystal oscillator in a push-to-talk voice transmitter is not desirable if frequency stability is important, since on-and-off operation constantly raises and lowers the crystal temperature. Thus, the frequency is kept in a state of constant variation to a degree dependent upon the magnitude of frequency-temperature coefficient of the crystal unit at the average operating temperature. Unless necessary for reasons of economy in space, cost, or the like, the oscillator should be designed for continuous operation and the keying performed in one or more of the succeeding amplifier stages. Usually the keying circuit is designed to remove and apply by one means or another, a cutoff bias in the buffer-amplifier stage. During the time that the buffer amplifier is cut off, the crystal circuit continues to oscillate, but the signal cannot be amplified and applied to the succeeding stages.

1-305. When it is necessary to key the oscillator, itself, the reason is normally that the space and weight requirements are so limited that no more than one or two vacuum-tube stages can be allowed. For this same reason, the oscillator is probably required to develop as much output power as possible, so that a Miller, rather than a Pierce circuit is generally employed if crystal control of the frequency is required. Nevertheless, the factors affecting the build-up time are approximately the same in either circuit. Fundamentally, the reason that a crystal oscillator requires a relatively much longer build-up time than does a conventional inductor-capacitor tuned circuit of the same reso-



nant frequency, is because the energy to be stored in the crystal is much greater than that which would be stored in an inductor-capacitor circuit. For a given tank current, the stored energy is proportional to the inductance, so, to a first approximation, we can suppose that the build-up time of, say, a Pierce oscillator as compared with that of a Colpitts oscillator of the same frequency, is directly proportional to the inductance ratio. On the other hand, it can be imagined that the build-up time tends to vary inversely with the total effective resistance in the tank circuit. The greater this resistance, the more quickly do the losses in the circuit rise to equilibrium with the rate of energy supply. The build-up time also tends to vary inversely with the frequency. Clearly, if the frequency were one cycle per second, equilibrium could not be reached in a shorter period. Finally, the build-up time is a function of the electromechanical coupling of the crystal to the circuit. The larger the C_o/C ratio of the crystal unit, the weaker is the coupling and the longer is the period before equilibrium can be reached. The exact relations of all the circuit variables in an equation expressing the time required for the amplitude to rise to within one per cent or so of its equilibrium limit would, indeed, be quite involved. Insofar as the crystal is concerned, the build-up time can be expected to vary positively if plotted against L, C, C_o, and 1/R of the crystal unit. The percentage variation of the build-up time with a given percentage variation in L can be expected to be greater than with the same percentage variation in C, because of the fact that, say, an increase in C, although increasing the build-up time by lowering the frequency, will also tend to decrease the build-up time by improving the electromechanical ratio. Thus, if the frequency remains constant, a crystal oscillator can be keyed at a faster rate if the L/C ratio is kept to a minimum, provided C_o is not increased. In other words, a crystal element should be chosen that has as large a piezoelectric effect as possible, provided the frequency-temperature coefficient is small. For example, for high-frequency circuits, an AT-cut crystal which has a capacitance ratio $\frac{C_o}{C} \approx 250$ is to be expected to provide better keying characteristics than a BT-cut crystal, which has capacitance ratio of 650. Preferably, from the point of view of a maximum keying speed for a given output voltage, the gridleak resistance should be kept small, not only to load the circuit and to provide quick-action limiting, but also to keep the positive swings of the grid and the transconductance high. The oscillator will almost certainly be designed for maximum power output, so that the tank circuit will be well loaded, for which reason the grid resistance must be kept relatively small as a safeguard against intermittent oscillations. It is questionable as to just how much the effective tank resistance limits the build-up time. Of course, if the resistance were zero, the oscillations would theoretically continue to rise indefinitely. On the other hand, the time required for the amplitude to reach any given value is least when the energy being lost from the circuit is least. In this respect, the build-up time tends to vary directly, not inversely, with the tank resistance. It would seem, that to obtain a maximum keying speed it might be preferable to use a fixed bias or a cathode bias, instead of the gridleak action. Using a sharp-cutoff tube biased for class-A operation, a grid, plate, or output circuit limiting arrangement could permit the oscillations to build up to a given level under conditions of a maximum ratio of input to dissipated power. Above this amplitude level the ratio would sharply decrease. Such a circuit could raise the permissible keying speed, but since this is accomplished by virtue of sudden changes in the circuit parameters, which changes always accompany to some extent any limiting action, an increased frequency instability and harmonic output are almost certain to result. Although a crystal oscillator should not be designed to be keyed unless absolutely necessary, experimental circuits have obtained keying speeds approaching 400 words per minute. The higher the keying speed, however, the greater must be the frequency tolerance.

Gridleak Resistance When Pierce Circuit Permits Switching from Crystal to Variable LC Control

1-306. It is often necessary to provide a variable-tuned, inductor-capacitor auxiliary circuit to permit emergency operation at frequencies other than those provided by the available crystals, or in the event of crystal failure. For this purpose it is often possible and is usually desirable to use the same vacuum tube that is used during crystal control. For example, a Pierce circuit could be readily converted to a Colpitts circuit simply by switching from the crystal to a tuning inductor, or to an inductor shunted by a variable capacitor. However, when such a conversion is made, the ratio of the stored energy to the power dissipation becomes much smaller than that during crystal control. For this reason, the maximum safe value of gridleak resistance is much smaller than during crystal operation. For output voltages comparable to those obtained with crystal control, the LC circuit employs gridleak resistances ranging from 20,000 to occasionally 100,000 ohms. If the LC circuit is intended to furnish a much greater output than the crystal circuit, lower values of R_g may be necessary. Rather than require the crystal circuit to operate with small values of R_g, it would be preferable to connect an additional shunt resistor in the grid circuit when switching to variable-tuning control.

Gridleak Resistance When Used with Cathode Biasing Resistor in Pierce Circuit

1-307. In addition to the voltage across the gridleak resistance, part of the bias voltage can be furnished by an r-f-bypassed resistance in the cathode circuit. The cathode resistor protects the tube from excessive plate current should oscillations cease, and has the additional advantage of reducing the grid current and, hence, the grid losses. The power expended in the grid circuit will be approximately equal to E_cI_c where E_c is the *total* bias and I_c is the grid current. Actually, unless an r-f choke is used in the grid circuit, the grid losses will be somewhat greater than E_cI_c because of the a-c component of voltage across R_g. As the cathode component of the bias becomes small, the grid losses approach 1.5 E_cI_c as a limit. See paragraph 1-296. The values of the cathode resistance, R_k, usually range from 100 to 1000 ohms. The reactance of the bypass capacitor should be at least as small as $R_k/10$ at the lowest operating frequency. With R_k connected between cathode and ground, the d-c voltage developed equals $(I_b + I_c) R_k$; or approximately, I_bR_k. The total bias, E_c, is still approximately equal to $\sqrt{2}\, E_g$. The d-c grid current is given by the equation

$$I_c = \frac{|E_c| - E_k}{R_g} = \frac{\sqrt{2}\, E_g - I_b R_k}{R_g} \qquad 1\text{—}307\ (1)$$

where E_k is the voltage across the cathode resistor.

AGC USED WITH PIERCE OSCILLATOR

1-308. Where space and cost permit, optimum output stability can be had when the oscillator bias is provided through an automatic-gain-control circuit. Gridleak action can be effective in initiating oscillations, but the bias furnished through AGC should be of much greater magnitude in order to be of maximum effectiveness. A small increase in output voltage must cause a large increase in bias. The use of AGC reduces the grid losses to a minimum and maintains a constant amplitude of oscillation. It is this latter feature that is, of course, of most importance—particularly so when the same oscillator is to be switched from one crystal unit to another. The voltage requirements for constant output without risking the overdrive of any of the crystals are the same as those that apply in the case of manual adjustment of the output. (See paragraph 1-302.) An A-G-C circuit applicable for use with a Pierce, or Miller, type oscillator, is shown in figure 1-131. The oscillator output is amplified by V_2. The output of V_2 is then rectified by V_3. The oscillator bias equals the average rectified voltage across R_3. C_1 bypasses the r-f component to ground. If R_3 were increased indefinitely the

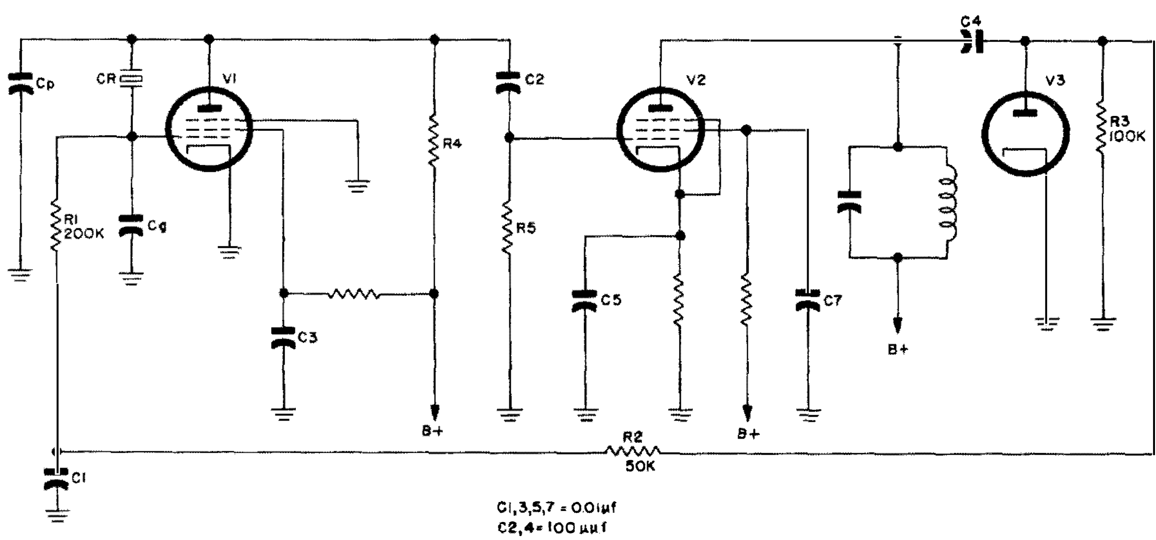


Figure 1-131. Pierce oscillator with automatic gain control

bias voltage would approach in magnitude the peak value of the V_2 output voltage. R_1, R_2, and R_3 are not critical—each can be made equal to 50K if a faster-acting gain control is required. However, R_1 should be kept as large as possible. Assume that the r-f losses in R_1, R_4, and R_5 are negligible and that C_p and C_g are approximately equal, so that V_1 is operating into a load impedance approximately equal to PI/4. Under these conditions g_m will be the minimum and R_p the maximum possible for sustained oscillations as long as the load capacitance across the crystal unit remains constant. The actual values of g_m and R_p are fixed by the vacuum-tube characteristics. Although the effective parameters of the tube are directly dependent upon the peak-to-peak magnitude of the excitation, as well as indirectly through the bias, it can be said that to a first approximation the equilibrium R_p and g_m are associated with a bias of more or less definite magnitude if the plate voltage is constant, and that approximately the same bias must exist regardless of whether it is developed by gridleak action or by AGC. Thus, the difference between AGC and gridleak control is not primarily in the magnitude of the bias, but in the amplitude of oscillations. Gridleak action requires that the peak excitation voltage of V_1 be slightly greater than the required bias; AGC requires that the peak excitation voltage of V_2 times the voltage amplification of the V_2 stage be slightly greater than the required bias of V_1. If the peak excitation voltage of V_2 is assumed to equal E_{pm}, which, in turn, is assumed to equal E_{gm}, $\left(\frac{C_g}{C_p} \approx 1\right)$, and k_2 is the effective amplification of the V_2 stage, then

$$|E_c| \approx k_2 E_{gm} \qquad 1\text{—}308\ (1)$$

or

$$E_{gm} \approx |E_c/k_2|$$

Since E_c is approximately fixed, it can be seen that the amplitude of oscillations is only $\frac{1}{k_2}$ as large as those that would exist by the gridleak method employing the same plate voltage. This is not a desirable feature where large output is required, but from the point of view of ensuring a low crystal drive and maximum stability, an A-G-C circuit has great advantages. Although the equilibrium amplitude is low, oscillations start as readily as with gridleak bias. AGC permits class-A operation with remote-cutoff tubes, and, since the limiting is very slow-acting, very pure sine-wave outputs and excellent frequency stability as well as amplitude stability is obtainable.

PLATE-SUPPLY CIRCUIT OF PIERCE OSCILLATOR

1-309. For optimum frequency stability it is important that the r-f impedance of the B^+ circuit be as high as possible relative to the impedance of the tank. If $C_g/C_p = 1$, the tank impedance equals PI/4. If the oscillator is intended to oscillate at only one frequency, or within a narrow frequency, range, it is generally preferable that the B^+ voltage be fed through an r-f choke. This method affords a maximum impedance with minimum loss and minimum voltage at the B^+ source. The inconvenience of an r-f choke is that its impedance changes with frequency, being inductive below its effective parallel-resonant point, and capacitive above. As long as this effect does not change the effective value of C_p by more than ± 10 per cent, the total load capacitance will not change by more than 5 per cent, if $C_g/C_p = 1$. Within these limits the use of a choke is to be preferred. For wide frequency ranges, a resistor should be used in the plate circuit, such as R_4 in figure 1-131. It is desirable for this resistance to be as high as 50K, or higher, from the point of view of frequency stability. On the other hand, the larger the resistance the higher the B^+ voltage source must be to provide a given plate voltage. Plate-supply resistances on the order of 5000 to 10,000 ohms have one other important advantage besides permitting lower B^+ sources. They load the oscillator tank so that differences in the resistance of the crystal from one unit to the next have very little effect upon the output impedance of the tube. Hence, when a change is made from one crystal unit to the next, the output voltage remains approximately the same.

1-310. The proper compromise in selecting a plate-circuit resistance depends upon the frequency-tolerance limits. The plate-circuit resistance does afford a certain frequency-stabilizing effect that is not provided by an r-f choke, particularly so when agc is used. The effect is one of reducing the change in R_p of the vacuum tube caused by a change in grid bias. For example, if the bias becomes more negative R_p increases, and I_b, the average plate current, decreases. There is then less voltage drop across the plate-supply resistor, and the resulting increase in plate voltage tends to decrease R_p, thereby annulling part of the increase in R_p due to the change in bias. The plate-voltage source should be regulated, if good stability is required. Where the frequency deviation must be kept to a minimum, the oscillator may require a separate rectifier unit, filter circuit, and voltage-regulator circuit.

CHOOSING A VACUUM TUBE FOR THE PIERCE CIRCUIT

1-311. It is no problem to find a vacuum tube that will permit a Pierce circuit to oscillate. Indeed, one of the major problems in tube circuit design is to prevent oscillations from occurring. With a crystal connected between the plate and grid of any vacuum-tube amplifier, the stray capacitance in the circuit is usually sufficient to cause oscillations to build up. If the plate voltage is not so high that the crystal is over-driven, the frequency stability of a stray-capacitance circuit may even be satisfactory for general-purpose use. Thus, the problem is not to find a vacuum tube that will work, but one that will be most satisfactory from the point of view of output stability and cost. First, a large tube is not necessary, since the Pierce circuit is not suited for large output. The choice of tube will depend somewhat upon the exact purpose of the oscillator and of the equipment of which it is a component. If the frequency tolerance is to be large, little thought need be given to fine points in the design, for the principal problem will be to keep the production costs to a minimum. A triode would be satisfactory, a 5K to 50K resistance in the plate circuit, a C_g/C_p ratio between 1 and 2, and a plate voltage sufficiently low so that the driving power of the crystal does not exceed the rated level for any effective crystal resistance meeting the specifications. A high-mu triode generally provides the better frequency stability because of its larger effective R_p, but it will have a higher plate dissipation for the same output voltage. Of the high-mu triodes, probably the 6AB4 is to be preferred as a simple unit, and the 12AX7 and the 12AT7 as twin triodes. It is the medium-mu tube that has been the most favored by design engineers when a triode has been chosen. Of these the 6C4, 6J4, and 7A4 single units, and the 6SN7-GTA twin unit are among the more popular. The 6C4 and the 6J4 are to be preferred for high-frequency operation. Since the 7A4 and the 6SN7-GTA have approximately 4 $\mu\mu$f capacitance between grid and plate, the 6C4, 6J4, 6J6, or the 12AU7, each with 1.5 $\mu\mu$f capacitance grid to plate, should provide the better frequency stability — particularly at high frequencies. The 7A4 and the 6SN7-GTA are generally more satisfactory for use in a Miller circuit. Where greater frequency stability is required, a pentode should be used. A pentode has the advantages of low plate-to-grid capacitance, greater R_p, and a screen grid whose voltage can be adjusted independently of the control-grid bias and plate voltage, thereby permitting a greater range of adjustments in the plate characteristics. Conventional pentodes must be operated at reduced voltages, to avoid overdriving the crystal, unless rather high C_g/C_p ratios are used. Subminiature pentodes have operating characteristics at their normal operating voltages ideally suited for crystal drive levels. The 1U4 is one such type having a sharp cutoff. Among the miniature pentodes having a sharp cutoff, the 6AU6, 6BC5, and 6AH6 are tubes generally recommended for wide-band, h-f circuits. The 6CB6, although designed principally for television use at 40 mc, should also be quite appropriate in crystal oscillator circuits. Remote-cutoff tubes are generally used only in special circuits. For example, if a low harmonic output is required, such tubes could be employed in conjunction with AGC. When the harmonic content is not of first importance, AGC is more effective if used with sharp-cutoff tubes, where a slight change in grid bias can make a much larger change in small-signal outputs than is possible if the slope of the E_cI_b curve changes very gradually. Actually, remote-cutoff tubes, when used, are usually found in doubler circuits, because of the large second-harmonic component that is produced. Although class-B and class-C operation with sharp-cutoff tubes can produce even greater harmonic outputs, there is the problem of ensuring that a crystal of large R_e will not be overdriven if it is to be operated in a class-B or class-C circuit. The output voltages of remote-cutoff tubes tend to vary more with crystals of different resistances than is the case when sharp-cutoff tubes are used. The reason is that in the former case the effective I_p continues to increase as R_e becomes small, since very large excitation voltages are required to override the cutoff point. On the other hand, the effective I_p begins to decrease when class-B operation is approached and such operation can be had with relatively small excitation voltages when sharp-cutoff tubes are used. If a remote-cutoff tube is desired, recommended types are the subminiature 1T4, the miniature 6BA6 and 12BA6, the lock-in 7A7, and the conventional-sized tubes such as the 6SK7 and 12SK7. The mention of particular vacuum tubes here should not be construed as official recommendation; they are named simply because they are the tubes commonly found in new equipment. The design engineer may very well find that the characteristics of other tubes are more appropriate for his needs.

Pierce-Oscillator Design Considerations When Vacuum Tube with Very Sharp Cutoff is Used

1-312. In making a preliminary approximation as to what the performance of a particular tube will be if used in a Pierce circuit, it should first be kept in mind that the ratio

$$E_p/E_g = C_g/C_p = \frac{\mu Z_L}{R_p + Z_L} = \frac{g_m R_p Z_L}{R_p + Z_L} \qquad 1\text{—}312\ (1)$$

is the gain of the tube. If the gain = k, and if $\frac{R_p}{Z_L}$ is 10 or greater, then

$$g_m = k/Z_L \qquad 1\text{—}312\ (2)$$

or

$$R_p = \mu Z_L/k \qquad 1\text{—}312\ (3)$$

Either equation (2) or (3) can be used to estimate approximately the grid bias for a given plate voltage, and vice versa, that can be expected if a particular tube is used. Assume, for example, that k = 1, that gridleak bias is to be used, and that the grid and load losses are negligible compared with the crystal driving power. In this case, the minimum expected Z_L will equal (min) PI/4, which occurs when a crystal unit has the maximum allowable R_e and is operated at the rated load capacitance, C_x. Under these conditions, the maximum permissible bias, as given by equation 1-293 (2), is

$$(\max)\ E_c = -\sqrt{2\,P_{cm}}\ /2\omega C_x \sqrt{R_{em}}$$

This maximum value of E_c is to be interpreted as a maximum that can be allowed *only* if R_e is a maximum *or* if the output voltage is to be the same magnitude regardless of the value of R_e. In this latter case, P_{cm} and R_{em} fix the output and bias limits for all crystal units of a given type. The constant output can be obtained in several ways: by the use of an actual or equivalent, parallel, plate load resistance that is small compared with the minimum nonloaded crystal tank impedance; by the use of AGC, by the use of manual voltage adjustments; or by other methods. The present discussion concerns only the noncontrolled nonloaded circuit. If a crystal unit of maximum R_e, being driven at the maximum drive level, is replaced by a crystal unit of smaller R_e, Z_L increases, and E_p and I_g tend to increase proportionately, so that the crystal driving power, equal to $I_g^2 R_e$, is greater than when R_e is a maximum. According to equation (3), insofar as it can be assumed that μ remains approximately constant (in practice, μ decreases somewhat) R_p increases proportionately with Z_L, so that although the equivalent generator voltage, $-\mu E_g$, increases, I_p remains constant. Thus, $I_p \approx g_m E_g \approx kE_g/Z_L \approx$ constant. In an actual circuit where $R_p >> Z_L$ and the vacuum tube has a very sharp cutoff, the effective I_p increases up to the point that the tube is cutoff for approximately three-fifths of the negative alternation (three-tenths of the entire cycle). As the excitation voltage increases beyond that point, I_p progressively decreases, although the total power supplied to the tank circuit continues to increase as long as the excitation voltage continues to increase. The conclusions above are derived in the special case of a C_g/C_p ratio of unity, by assuming that for all practical purposes the plate-current pulses are in phase with E_p, and that $E_b >> E_p$. Figure 1-132 illustrates different states of operation of the same oscillator circuit that can occur if crystal units of the same frequency but different values of R_e are inserted in the circuit. A change from the class-A to the class-C state could readily occur if the crystal R_e were reduced by more than one-half. The effective I_{pm} is defined by the equation

$$P_{ZL} = I_{pm}\, E_{pm}/2 \qquad 1\text{—}312\ (4)$$

where P_{ZL} is the power expended in the tank circuit. Since Z_L is very small compared with R_p, it can be assumed that the sinusoidal component, E_p, of the with-signal d-c plate voltage, e_b, is negligible by comparison with the average value, E_b; that is, $E_b \pm E_{pm} \approx E_b$. With this assumption we can treat I_{bm}, the value of the with-signal, d-c plate current, at the positive peak of excitation ($e_c \approx o$) as a constant. The assumptions above also imply that very little grid current exists; otherwise, the larger excitation voltages would drive the grid considerably above zero at the positive peaks. With the peak instantaneous d-c plate current a constant, the *total* energy supplied by the power source progressively decreases as Z_L and the excitation increase, since I_b, the average i_b, becomes progressively smaller, whereas E_b remains constant. (Actually, if the plate current is supplied through a resistor, a decrease in I_b causes E_b to increase somewhat. For the problem at hand, assume that a regulated B^+ is fed through an r-f choke.) Thus, it can be seen that as R_e becomes small the plate efficiency increases considerably. However, the efficiency of a crystal oscillator does not approach the high ratios of input to tank power that are obtained with conventional class-C power amplifiers and oscillators. The latter circuits can operate at efficiencies of 60 to 90 per cent because E_{pm} approaches E_b in magnitude. The instantaneous power being dissipated in the tube is the instantaneous value of $i_b e_b$, and the instantaneous power being delivered to the tank is $i_b e_p$. When i_b is a maximum, $e_b = E_b - E_{pm} << e_p =$

E_{pm}, so that most of power goes to the tank circuit. In the conventional Pierce oscillator such high efficiency is not to be approached unless the C_g/C_p ratio is to be made very large and E_b approaches in magnitude the voltage specifications of the crystal unit. Now, to obtain a maximum output without the risk of overdriving a randomly selected crystal unit, it will be useful to derive approximate equations concerning the change in crystal power with a change in R_e. The crystal power, we shall assume to equal the total tank power, P_{ZL}. In short, the problem is to be able to express P_{ZL} as a function of R_e. I_{bm}, E_b, and E_{co} (the cutoff voltage) will be considered constants, and i_p and e_p are to be assumed to be in phase. First, we express the effective I_{pm} for each class of operation in terms of the constants above and the angles ϕ and θ, where appropriate. (See figure 1-132.) As a safeguard against intermittent oscillations, which are most likely to occur when R_e is a maximum, assume that the bias for maximum R_e is to occur on the straight portion of the E_cI_b curve. If the oscillations are to build up at all, they must continue to do so until the negative excitation peak at least extends into the lower bend of the E_cI_b curve, for it is only beyond the straight portion of the curve that g_m can change in order to seek its equilibrium value—that is, unless R_g is so small that equilibrium is reached by virtue of the increase in grid losses alone. With a large R_g and a reasonably sharp cutoff, it is virtually impossible for oscillations to start if the amplification is not at least sufficient to increase the excitation to where the negative peak is very nearly equal to E_{co}. Assume, then, that with $R_e = R_{em}$, the oscillator is designed to operate approximately as shown in figure 1-132 (A). It can be seen intuitively that

$$\text{(Class A) } I_{pm} \approx I_{bm}/2 \qquad 1\text{—}312\ (5)$$

and with $C_g/C_p = 1$, considering only the unsigned magnitudes of the bias voltage,

$$\text{(Class A) } E_{pm} \approx E_c \approx E_{co}/2 \qquad 1\text{—}312\ (6)$$

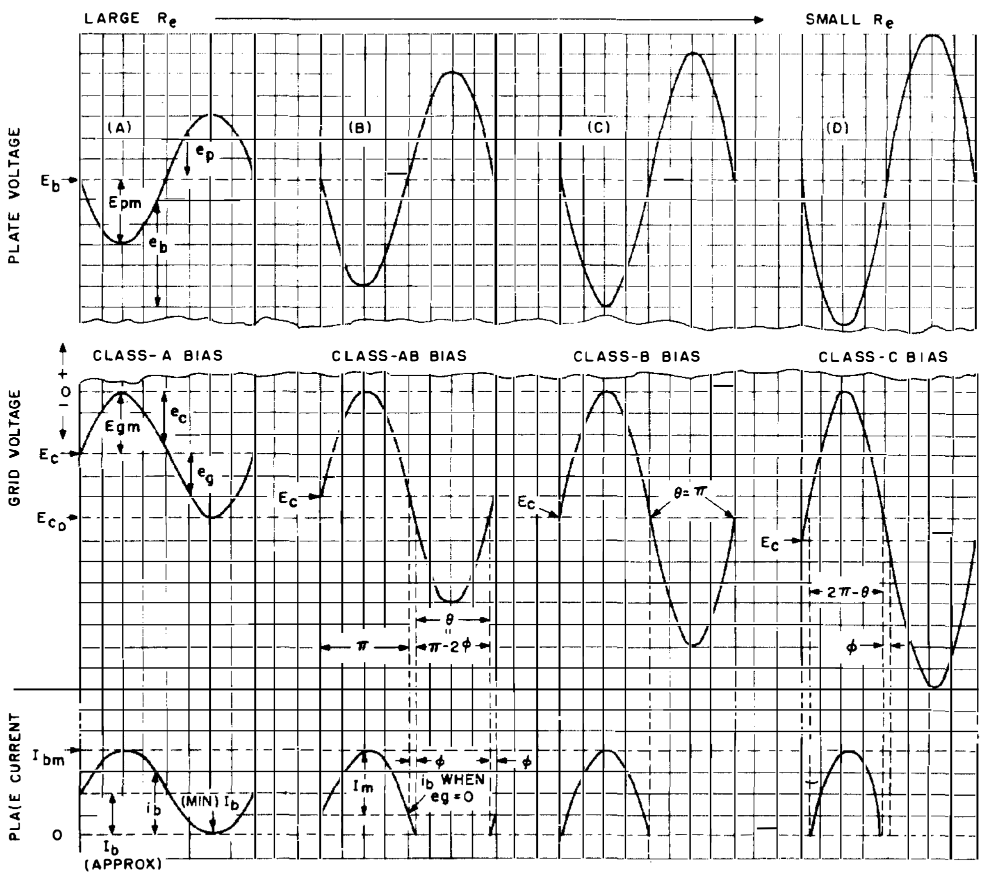


Figure 1-132. ***Change of state of Pierce oscillator with a C_g/C_p ratio of one, under no-load conditions when E_b is held constant and the effective resistance of the crystal changes***

so that

$$\text{(Class A)}\ P_{Z_L} = I_{pm} E_{pm}/2 \approx I_{bm} E_{co}/8 \qquad 1\text{—}312\ (7)$$

Equation (7) represents the maximum possible crystal power if a tube is not to be driven beyond cutoff. Referring now to figure 1-132 (B), we shall assume that a crystal unit with an R_e slightly less than the maximum is connected so that the bias is similar to that under AB operating conditions. I_m represents the *apparent* maximum I_p. It can be seen that except for the angle $(\pi - 2\phi)$, when the tube is cut off,

$$i_b \approx I_m (\sin \omega t + \sin \phi) \qquad 1\text{—}312\ (8)$$

where ϕ can be considered a constant. Now,

$$I_{bm} = I_m (1 + \sin \phi) \qquad 1\text{—}312\ (9)$$

so, on substitution in equation (8),

$$i_b = \frac{I_{bm}}{1 + \sin \phi} (\sin \omega t + \sin \phi) \qquad 1\text{—}312\ (10)$$

Similarly,

$$E_{co} = E_{gm} (1 + \sin \phi) = E_{pm} (1 + \sin \phi) \qquad 1\text{—}312\ (11)$$

so that

$$e_p = E_{pm} \sin \omega t = \frac{E_{co}}{1 + \sin \phi} (\sin \omega t) \qquad 1\text{—}312\ (12)$$

Since no energy is being supplied during the time that the tube is cut off, the energy delivered to the tank per cycle is

$$\int_{t\,(-\phi)}^{t\,(\pi + \phi)} e_p\, i_b\, dt = \frac{1}{\omega} \int_{-\phi}^{\pi + \phi} e_p\, i_b\, d\omega t \qquad 1\text{—}312\ (13)$$

where t = time in seconds.
Thus, the energy delivered per second, is

$$P_{Z_L} = \frac{f}{\omega} \int_{-\phi}^{\pi + \phi} e_p\, i_b\, d\omega t = \frac{1}{2\pi} \int_{-\phi}^{\pi + \phi} e_p\, i_b\, d\omega t \qquad 1\text{—}312\ (14)$$

On substitution of E_p and I_b from equations (10) and (12),

$$P_{Z_L} = \frac{E_{co}\, I_{bm}}{2\pi (1 + \sin \phi)^2} \int_{-\phi}^{\pi + \phi} (\sin^2 \omega t + \sin \phi \sin \omega t)\, d\omega t \qquad 1\text{—}312\ (15)$$

On integration,

$$\text{(Class AB)}\ P_{Z_L} = \frac{E_{co}\, I_{bm} (\pi + 2\phi + 2 \sin 2\phi)}{4\pi (1 + \sin \phi)^2} \qquad 1\text{—}312\ (16)$$

No maximum exists for equation (16) with values of ϕ between 0 and $\pi/2$. For class-A operation similar to that in figure 1-132 (A), $\phi = \pi/2$, so that equation (16) becomes

$$\text{(Class A)}\ P_{Z_L} = E_{co}\, I_{bm}/8$$

This checks, as is to be expected, with equation (7). For class-B operation, $\phi = 0$, so that equation (16) becomes

$$\text{(Class B)}\ P_{Z_L} = E_{co}\, I_{bm}/4 \qquad 1\text{—}312\ (17)$$

Note that the power expenditure in the crystal unit for class-B operation is exactly twice that found for class-A operation. Since E_{pm} under class-B conditions is equal to E_{co} (see figure 1-132 (C)), or twice the class-A value of E_{pm}, then, because $P_{Z_L} = I_{pm}E_{pm}/2$, the effective I_{pm} at class B must be equal to the same effective value as at class A. Thus,

$$\text{(Class B)}\ I_{pm} = I_{bm}/2 \qquad 1\text{—}312\ (18)$$

Also, since $E_{pm} = I_{pm}Z_L$, if E_{pm} has doubled but I_{pm} has not changed it can only mean that Z_L has doubled. In other words, if the oscillator is designed to operate class A when R_e is a maximum, it will operate class B when a crystal unit is inserted that has an effective resistance equal to $R_{em}/2$. Equation (16) can be generalized to apply for all operating states in which $2E_{gm}$ is equal to or greater than E_{co}. For greater simplicity, ϕ should be replaced by $(\pi - \theta)/2$, where $\theta = \pi - 2\phi$ is the angle during which the tube is cut off. θ is always positive, whereas ϕ would be negative in the case of class-C operation. Thus, equation (16) can be expressed

$$\text{(all classes)}\ P_{Z_L} = \frac{E_{co}\, I_{bm} (2\pi - \theta + \sin \theta)}{4\pi \left(1 + \cos \frac{\theta}{2}\right)^2} \qquad 1\text{—}312\ (19)$$

The slope of this equation is positive for all values of θ less than 2π, so that the power dissipated in a crystal unit always becomes greater as R_e becomes smaller. By substituting $\cos\frac{\theta}{2}$ for $\sin\phi$ in equation (11) and rearranging, we have

$$E_{pm} = E_{co} \Big/ \left(1 + \cos\frac{\theta}{2}\right)$$

so that

$$I_{pm} = \frac{2\,P_{Z_L}}{E_{pm}} = 2\,P_{Z_L}\left(1 + \cos\frac{\theta}{2}\right) \Big/ E_{co}$$

$$= \frac{I_{bm}\,(2\pi - \theta + \sin\theta)}{2\pi\left(1 + \cos\frac{\theta}{2}\right)} \qquad 1\text{—}312\ (20)$$

Equation (20), unlike equation (19), has a maximum when θ is approximately $3\pi/5$. That a maximum (or a minimum) occurs between $\theta = 0$ and $\theta = \pi$ is to be expected, since I_{pm} has the same value for each of those values of θ. This maximum is

$$(\text{max})\ I_{pm} \approx 0.54\ I_{bm} \qquad 1\text{—}312\ (21)$$

Now,

$$Z_L = 2\,P_{Z_L} / (I_{pm})^2 \qquad 1\text{—}312\ (22)$$

On substituting equations (10) and (20) in (22)

$$Z_L = \frac{2\pi\,E_{co}}{I_{bm}\,(2\pi - \theta + \sin\theta)} \qquad 1\text{—}312\ (23)$$

Rearranging and substituting $1/4\omega^2 C_x^2 R_e$ for Z_L, where C_x is the specified load capacitance of the crystal unit,

$$\theta - \sin\theta = 2\pi\,(1 - 4\omega^2 C_x\,E_{co}\,R_e/I_{bm}) \qquad 1\text{—}312\ (24)$$

Equation (24) is quite significant in that it predicts the approximate angle during which a given tube will be cut off for a given value of R_e. A Pierce oscillator designed so that the tube is operating with a class-A bias equal to $E_{co}/2$ when R_e is a maximum will have a value of θ equal to zero. Thus, when $R_e = R_{em}$, each side of equation (24) must vanish. For the right-hand side to equal zero,

$$4\omega^2\,C_x\,E_{co}\,R_{em} = I_{bm}$$

This is equivalent to saying that

$$\frac{1}{(\text{min})\ Z_L} = \frac{4}{(\text{min})\ PI} = \frac{I_{bm}}{E_{co}} = (\text{average})\ g_m \qquad 1\text{—}312\ (25)$$

which could have been predicted on the basis of equation (2). Assume that equation (25) holds, what will be the value of θ when a crystal unit having a practical minimum value of R_e equal to $R_{em}/9$ is connected in the circuit? When $R_e = R_{em}$, the negative term within the parentheses of equation (24) is equal to -1; with $R_e = R_{em}/9$, the same term is reduced to $-1/9$. Thus, the maximum θ to be expected is defined by:

$$[(\text{max})\ \theta \text{ for } (\text{min})\ R_e] \text{ when: } \theta - \sin\theta = 16\pi/9 \qquad 1\text{—}312\ (26)$$

Figure 1-133 shows that equation (26) requires that

$$(\text{max})\ \theta \approx 16\pi/9 - 1 \qquad 1\text{—}312\ (27)$$

In other words, when a sharp-cutoff tube is used and the oscillator is designed for class-A operation with crystal units of maximum R_e, the oscillator will be operating class C, with the tube cut off approximately three-fourths of the time, when crystal units of minimum values of R_e are connected in the circuit. Equation (24) can be generalized to define θ with reference to any convenient value of R_e, simply by assuming that $\theta = 0$ when $R_e = (\text{ref})\ R_e$. Thus,

$$\theta - \sin\theta = 2\pi\,[1 - R_e/(\text{ref})R_e] \qquad 1\text{—}312\ (28)$$

or

$$\theta_N - \sin\theta_N = 2\pi\,(N - 1)/N \qquad 1\text{—}312\ (29)$$

where $N = (\text{ref})\ R_e/R_{e_N}$, $\theta = 0$ when $R_e = (\text{ref})\ R_e$, and θ_N is the value of θ for the particular value of R_e symbolized by R_{e_N}. The reference R_e need not be the maximum permissible R_e. For a given oscillator of C_g/C_p ratio equal to 1, (ref) R_e would be the value of R_e that would cause the peak-to-peak excitation voltage to equal E_{co} in magnitude. Assuming that (ref) $R_e = R_{em}$, what then will be the ratios of P_{Z_L} and I_{pm} corresponding to minimum and maximum values of R_e? When $\theta = \frac{16\pi - 9}{9}$, as given by equation (27), $\cos\theta/2$ is very nearly $-2/3$, so equation (19) becomes

$$(\text{Class C max})\ P_{Z_L} = \frac{E_{co}\,I_{bm}\left(2\pi - \frac{16\pi}{9}\right)}{4\pi\left(1 - \frac{2}{3}\right)^2} = \frac{E_{co}\,I_{bm}}{2} \qquad 1\text{—}312\ (30)$$

On comparison with equations (7) and (17), which give values of P_{Z_L} of $E_{co}I_{bm}/8$ and $E_{co}I_{bm}/4$

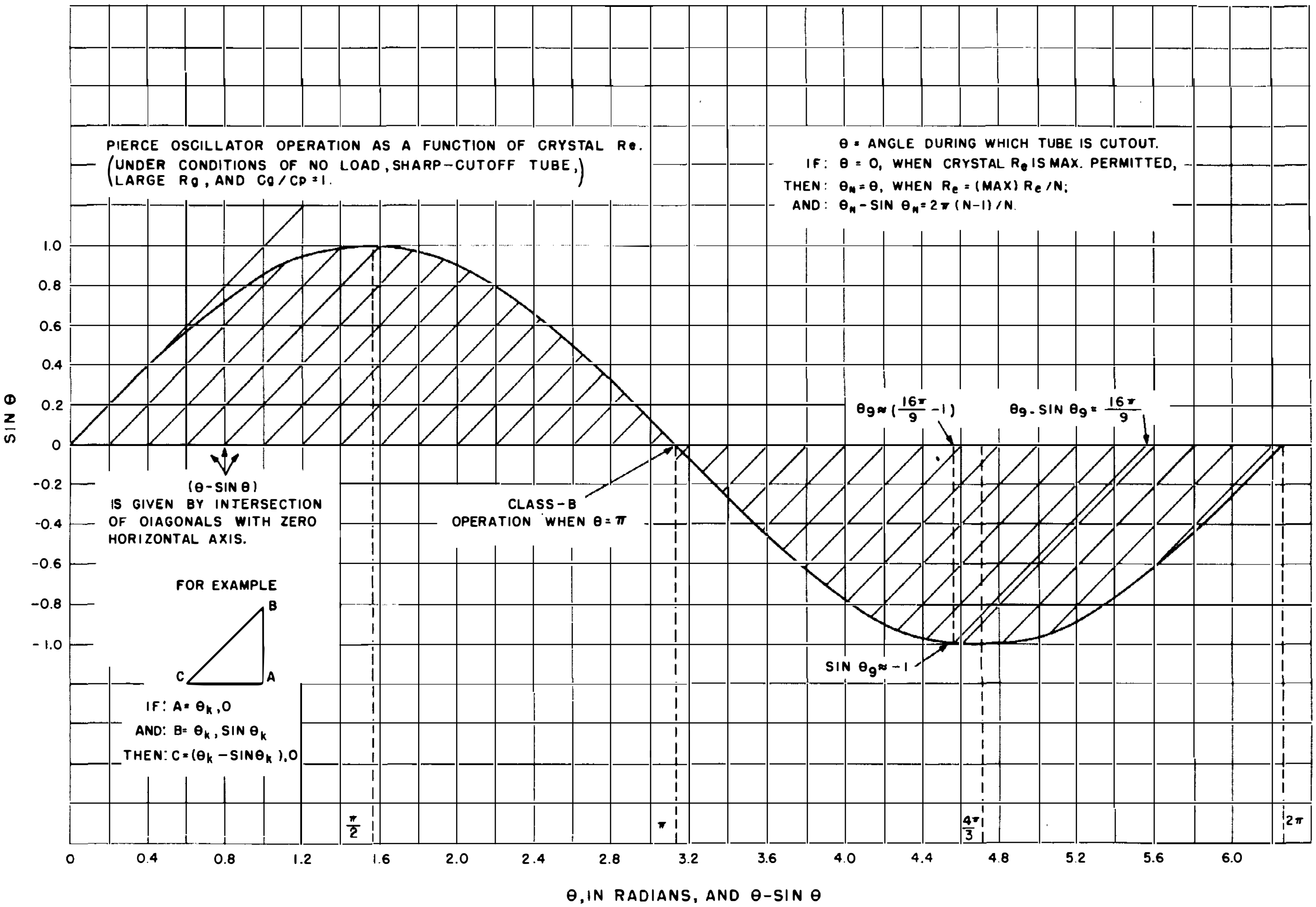


***Figure 1-133.** Sine curve plotted to same scale as the angle θ. 45-degree diagonal that intercepts sin θ_k on curve intercepts (θ_k—sin θ_k) on θ axis*

for class-A and class-B operation, respectively, we find that where there is to be no output control, the no-load tube voltage must be so chosen that a crystal unit of maximum R_e is not driven at more than one-fourth the rated drive level, otherwise crystals of small R_e will be overdriven. With a power ratio of 4 when the Z_L ratio is 9, it can be shown quite simply that the I_{pm} ratio is 2/3 and the E_{pm} ratio is 6. Thus,

$$\text{(Class-C min) } I_{pm} = I_{bm}/3 \qquad 1\text{—}312\ (31)$$

and

$$\text{(Class-C max) } E_{pm} = 3\,E_{co} \qquad 1\text{—}312\ (32)$$

The plate dissipation in the tube should be of little concern unless subminiature tubes are used. In any event the plate dissipation is a maximum when R_e is a maximum, so no thought need be taken for other than class-A operation. Approximately,

$$\text{(Class-A) plate power} = E_b I_b \approx E_b I_{bm}/2 \qquad 1\text{—}312\ (33)$$

Finally, the foregoing equations suggest that a Pierce oscillator employing a sharp-cutoff tube be designed for class-A operation on the assumption that R_e will be a maximum and that the maximum permissible drive level is one-fourth its actual rating. Under these assumptions, equation 1-293(2) should be changed to

$$\text{(max) } E_c = -\frac{\sqrt{2\,P_{cm}/4}}{2\omega C_x\sqrt{R_{em}}} = -\frac{\sqrt{P_{cm}}}{2\omega C_x\sqrt{2\,R_{em}}} \qquad 1\text{—}312\ (34)$$

where P_{cm} is the true drive-level rating. Since (max) E_c will also be equal to $E_{co}/2$, approximately, then (no longer continuing to treat E_{co} as a magnitude only)

$$\text{(max) } E_{co} = -\sqrt{P_{cm}}\Big/\omega C_x\sqrt{2\,R_{em}} \qquad 1\text{—}312\ (35)$$

At the same time, $P_{ZL}\left(=\frac{I_{bm}E_{co}}{8}\right)$ must not exceed $P_{cm}/4$. Consequently,

$$\text{(max) } I_{bm} = \frac{2\,P_{cm}}{\text{(max) }|E_{co}|} = 2\omega C_x\sqrt{2\,P_{cm}R_{em}} \qquad 1\text{—}312\ (36)$$

Equations (35) and (36) define the operating characteristics to be sought if a sharp-cutoff tube is to be used under conditions of maximum output for maximum stability. Remember, that equation (34) actually is an expression of the limitation on I_g, the crystal current, and therefore upon E_p and E_g. As far as the self-excitation voltage of a truly sharp-cutoff tube is concerned, it will be difficult to keep this voltage from building up until it reaches into the bend near the cutoff point. For this reason, the first consideration is that I_{bm} is not exceeded. As a safety measure, I_{bm} should not be greater than the value given by equation (36), even if the actual E_{co} is less than (max) E_{co}. The conclusions reached in the foregoing discussion are summarized in the following table.

PIERCE-CIRCUIT OPERATING LIMITATIONS DUE TO CRYSTAL SPECIFICATIONS OF LOAD CAPACITANCE, C_x, MAXIMUM PERMISSIBLE EFFECTIVE RESISTANCE, R_{em}, AND DRIVE LEVEL, P_{cm}

Plate Dissipation (max) $= E_b I_{bm}/2$
$E_{co} = -\sqrt{P_{cm}}/\omega C_x\sqrt{2\,R_{em}}$
$I_{bm} = 2\,P_{cm}/|E_{co}|$

Conditions are those for sharp-cutoff tube, negligible load gridleak bias with large R_g, $C_g/C_p = 1$, $E_b >> E_p$, $R_l >>$ (max) PI/4 = (max) Z_L, Class-A operation when R_e is maximum, and maximum permissible output.

R_e =	R_{em}	$R_{em}/2$	$R_{em}/9$
E_{pm}, E_{gm}, $\lvert E_c\rvert$ =	$\lvert E_{co}/2\rvert$	$\lvert E_{co}\rvert$	$3\lvert E_{co}\rvert$
P_c ($= P_{ZL}$) =	$P_{cm}/4$	$P_{cm}/2$	P_{cm}
I_{pm} =	$I_{bm}/2$	$I_{bm}/2$	$I_{bm}/3$
Z_L =	(min) PI/4	(min) PI/2	9 (min) PI/4
g_m =	4/(min) PI	2/(min) PI	4/9 (min) PI
θ =	0	π	$\frac{16\pi}{9} - 1$
Operation =	Class A	Class B	Class C

Pierce-Oscillator Design Considerations When Tube Cutoff Has Below-Average Sharpness

1-313. Unless a vacuum tube has plate characteristics resembling those of subminiature tubes when normal plate voltages are used, or unless by reducing the filament voltage such characteristics can be achieved, a Pierce oscillator tube must be operated at a plate voltage of from one-half to one-fourth normal. In so doing, it is very probable that the lower bend of the E_cI_b curve will become rather extended compared with the straight portion to the left of zero grid volts. In this event, the tube exhibits the characteristics of a remote-cutoff tube, except that the cutoff voltage is one-fifth or less that of a normal remote-cutoff tube operating at an equivalent reduced plate voltage. Where the cutoff is not sharp, it is quite easy for equilibrium to be reached with peak-to-peak excitation voltages much smaller in magnitude than E_{co}, and considerably greater ranges in R_e of the crystal unit can exist before cutoff is reached. Thus, in the more usual case, the assumptions used in paragraph 1-312 cannot be made unless greater care is taken in the oscillator design to ensure a peak-to-peak excitation voltage equal to $|E_{co}|$ when R_e is a maximum—an operating point much more difficult to locate and critical to maintain when a large steady decrease in the effective g_m occurs well before the cutoff point is reached, and which may require very low plate voltages if the maximum-R_e crystal unit is not to be overdriven. As a concrete example, suppose that the oscillator is to employ a 10-mc crystal unit of the CR-18/U type. At this frequency, $P_{cm} = 5$ mw, $R_{em} = 25$ ohms, and $C_x = 32\ \mu\mu f$. On substitution in equation 1—312 (35), we obtain a (max) E_{co} of approximately $-5V$. By equation 1—312(36), this value of E_{co} is to be obtained in a tube where the zero-bias, without-signal plate current is $I_{bm} = 2$ ma. Such characteristics are not easily obtained with conventional-sized vacuum tubes. It may be necessary to operate at the given value of I_{bm}, or slightly greater, and a cutoff voltage that is of a smaller magnitude than that indicated for (max) E_{co} in equation 1—312 (35), in which case all crystal units used will drive the tube beyond cutoff. An alternative approach is to operate at a larger than maximum E_{co}, but, if this be done, a safety factor should be allowed by assuming that I_p is to be the same for all values of R_e. Although this will not be strictly true, the assumption is a close approximation if the change in plate current between the values of $E_c = E_{co}/2$ and E_{co} is very small compared with the change in plate current between $E_c = 0$ and $E_c = E_{co}/2$. It can be seen that insofar as the effective I_p can be assumed to remain constant, E_p, and hence E_g, I_g, E_c, and the crystal driving power, $I_g^2R_e \approx I_p^2Z_L$, increase directly with Z_L, or inversely with R_e. The problem is to find the maximum permissible E_c, which, although applying to extended-cutoff operation only when R_e is a maximum, will not lead to a replacement crystal being over-driven if its resistance is less than the maximum. Again we assume a minimum R_e equal to $R_{em}/9$. In a manner similar to the derivation of equation 1—293 (2), we can say (max) E_c (with (min) R_e) =

$$(\text{max})\ E_c\ [\text{for (min)}\ R_e] = -\frac{\sqrt{2\,P_{cm}}}{2\omega C_x\sqrt{(\text{min})\ R_e}}$$

$$= -\frac{3\sqrt{P_{cm}}}{\omega C_x\sqrt{2\,R_{em}}} \qquad 1\text{—}313\ (1)$$

Now, if equation (1) gives the bias voltage when a crystal unit of minimum R_e is connected, *assuming that I_p is constant,* the bias that exists when a crystal unit of maximum R_e (= 9 (min) R_e) is substituted will be one-ninth the value above. Thus,

$$(\text{extended cutoff max})\ E_c\ [\text{for}\ R_e = R_{em}] =$$

$$-\frac{\sqrt{P_{cm}}}{3\omega C_x\sqrt{2\,R_{em}}} \qquad 1\text{—}313\ (2)$$

If a gridleak Pierce oscillator is not to have a loaded plate circuit, nor an adjustable nor controlled output voltage, nor a sharp cutoff, equation (2) gives the maximum bias that can be safely assumed when R_e is a maximum. The output voltage agreeing with equation (2) is two-thirds that given in paragraph 1-312 for a sharp-cutoff tube. If R_g is not large enough for the average E_c to approximate the peak excitation voltage, a maximum bias less than that given by equation (2) must be assumed. With large values of R_g, $|E_c|$ of equation (2) is the peak of the maximum excitation voltage when R_e is a maximum, and $|E_c|$ of equation (1) is the approximate peak when R_e is a minimum. If a C_g/C_p ratio other than 1 is used, equation (2) can be expressed more exactly

(extended cutoff max) E_c =

$$-\sqrt{2\,P_{cm}} \Big/ 3\omega C_g \sqrt{R_{em}} \qquad 1\text{—}313\ (3)$$

It should be understood that although equations (2) and (3) are derived from equation (1), it is wiser to select the vacuum tube and plate voltage upon the assumption that the resistance of the crystal unit is a maximum rather than a minimum. Since the effective amplification factor of the tube cannot be expected to be constant for all values of R_e, equations (1) and (2) will not both hold for the same circuit. If (1) is correct, (2) will indicate a value too low; if (2) is correct, (1) will indicate a value too high. Equation (2) therefore permits a safety factor in the event of an exceptionally low value of R_e. Also, if the oscillator performs properly with R_e a maximum, it will almost certainly operate when R_e is a minimum. The reverse is not necessarily true.

1-314. Equation 1—313 (1) is equivalent to a bias and output of the same magnitude as that obtained in paragraph 1-312 for sharp-cutoff conditions and minimum R_e; but the bias and output of equation 1—312 (2) for $R_e = R_{em}$, when E_{co} is assumed to be significantly greater than $2E_c$, are only two-thirds their equivalent sharp-cutoff values. In the case of the 10-mc CR-18/U crystal unit discussed in paragraph 1-313, the (practical max) E_c, as given by equation 1—313 (2) is $-\frac{5}{3}V \approx -1.7$ V. For the smallest values of R_e, the bias will approach -15 V. Assuming that $g_m \approx \frac{1}{Z_L}$ (according to equation 1—312 (2), when $C_g/C_p = 1$) and that

$$Z_L = \frac{\text{(min) PI}}{4} = \frac{1}{4\omega^2\ C_x^2\ R_{em}} =$$

$$\frac{10^{10}}{4 \times 6.28^2 \times 32^2 \times 25} \approx 2500 \text{ ohms}$$

then, $g_m \approx 10^6/2500 = 400$ μmhos when R_e is a maximum. This is a very small transconductance to be obtained with a bias of approximately -1.7 volts, and usually cannot be obtained at all with normal operating voltages except in the case of the small battery-operated tubes. The 1.7-volt maximum bias represents a peak-to-peak excitation maximum of 3.4 volts. With an average g_m of 400 μmhos, the limiting value of I_{bm} ($\approx 2I_{pm} \approx 2g_mE_{gm}$) becomes 1.4 ma, approximately. Only if agc is used to provide a much larger bias than can be obtained with a peak-to-peak excitation of 3.4 volts will it be possible to have such a small zero-signal plate current without operating conventional tubes at greatly reduced voltages. Generally, it is easier to operate with a small E_{co} and a larger I_{bm} and not attempt class-A operation. A large percentage of the crystal oscillators now in use drive the crystal units at a considerably higher level than is advantageous from the point of view of stability and long crystal life. Much of the care otherwise taken in the circuit design can be wasted if the first consideration is power output rather than frequency control. Where a vacuum-tube manual recommends a particular voltage of power amplifier for use as a class-C oscillator tube, the typical operating characteristics listed are rarely appropriate for military-standard crystal units, but apply more usually to LC circuits. The plate voltages must be considerably lower than the typical values indicated, in order to reach the small transconductances that must exist at equilibrium without overdriving the crystal unit.

1-315. Assume that a crystal unit is connected in a Pierce circuit using a conventional triode operating at its normal plate voltage, and that the C_g/C_p ratio is near unity. The equilibrium values of g_m and R_p cannot be reached until the amplitude is great enough for the tube to be operating class C, and the crystal unit will almost certainly be overdriven. There are four ways in which the circuit can be adjusted to prevent this overdrive: (a) the plate voltage can be reduced, (b) the filament voltage can be reduced, (c) the C_g/C_p ratio can be increased, or (d) the load losses can be increased. Of these methods, the first, reducing the plate voltage, seems to be the best from the point of view of frequency stability, although a reduction of the filament voltage may be worth consideration. Very possibly, if the filament voltage is decreased sufficiently to lower the zero-bias transconductance to as much as one-fifth its normal value, the operation of the circuit will become unduly sensitive to slight fluctuations in the filament power supply. The only data available at this writing is that reported by Messrs. Roberts, Novak, and Goldsmith of the Armour Research Foundation of Illinois Institute of Technology. Experimenting with a 6C4 tube and a 7-mc Miller circuit, it was found that a 30-percent decrease in filament voltage, which is equivalent to decreasing the filament power by approximately one-half or more, depending upon the temperature coefficient of the filament resistance, caused only a 2.5-cycle rise in frequency. (In a Pierce circuit the frequency would have decreased.) This effect on the frequency is very slight, but the exact decrease in the r-f plate current is not known. Nevertheless, the evidence is sufficient to suggest that if the tube characteristics are made suitable for a crystal circuit by reducing the filament voltage, any instability caused by further fluctuations in the filament voltage would appear primarily as variations in the output voltage, rather than as variations in the frequency. In view of the fact that a reduction in filament current permits a greater saving in power than does a reduction in plate voltage (and lengthens the tube life), this approach to the problem may well be worth experimentation. The conventional approach, however, is to operate with a low plate voltage. If a C_g/C_p ratio on the order of unity is to be used, the average tube will require plate or screen voltages of 40 to 50 volts, or less. The lower the voltage, the nearer class-A operation can be approached at equilibrium. A fair approximation of the operating conditions to be expected can be made from an inspection of a family of plate-characteristic curves. With $C_g/C_p = 1$, the peak-to-peak variations in plate voltage are the same as those of the excitation voltage, so for all practical purposes the plate voltage can be assumed to be constant. Thus, the load line can be assumed to be vertical, and the maximum and minimum amplitudes of I_b for a given plate voltage become the values, respectively, for grid voltages of 0 and $2E_{gm}$, where E_{gm} is the peak excitation voltage. For the 10-mc crystal unit taken as an example above, it was found that the peak-to-peak I_p for a maximum R_e was $2\,|g_m E_c| = 1.4$ ma. The correct plate voltage for a given tube is thus the value of E_b at which a change of grid voltage from 0 to -3.4 volts causes the plate current to decrease by 1.4 ma. This type of operation—class A to class AB—is generally more feasible when agc is used, if it is desired to apply for all values of R_e.

PIERCE-OSCILLATOR DESIGN CONSIDERATIONS FOR C_g/C_p RATIOS OTHER THAN ONE

1-316. When the C_g/C_p ratio is not approximately equal to one but the total load capacitance meets the crystal specifications, g_m is increased, and generally it will be easier to obtain desirable vacuum-tube characteristics at more convenient plate voltages. The first step, as before, is to theoretically limit the peak of the crystal current to $\sqrt{\frac{2P_{cm}}{R_{em}}}$ when R_e is a maximum. The peak excitation voltage, E_{gm}, equals $\frac{1}{\omega C_g}\sqrt{\frac{2P_{cm}}{R_{em}}}$ under these conditions. E_{pm} equals $\frac{1}{\omega C_p}\sqrt{\frac{2P_{cm}}{R_{em}}}$; g_m equals C_g/C_pZ_L ; I_{pm} equals g_mE_{gm}. With these values taken as a start we can retrace the steps taken in paragraphs 1-312 and 1-313, and determine the values of I_{bm} and E_{co} that do not permit the crystal to be overdriven for any value of R_e between R_{em} and $R_{em}/9$.

FINAL WORD ON CORRECT LOAD CAPACITANCE IN THE PIERCE CIRCUIT

1-317. A prime purpose of the military specifications regarding the load capacitance, effective resistance, drive level, and frequency tolerance of the different types of crystal units is to guarantee the replacement of a defective crystal unit in the field without special testing or other complications, and with the same ease that a defective vacuum tube can be replaced with a new tube of the same type. However, a crystal unit is more critical in its performance than a vacuum tube. As a result there can be no replacement guarantee unless the new crystal unit is inserted in a circuit where it will be operated under approximately the same load and drive conditions at which it has been tested. An inspection of the various types of oscillator circuits now in use, such as those illustrated in figures 1-135 to 1-138, most of which have been designed around the older types of crystal units, reveals a much greater versatility in operating conditions than is now desired in the design of new equipment. One of the requirements that is no longer within the jurisdiction of the design engineer is the effective load capacitance into which the crystal unit is to work. This means, that for a given nominal frequency and type of crystal unit, the crystal unit must exhibit a given inductive reactance, X_e, equal numerically to $1/\omega C_x$, where C_x is the rated load capacitance. Furthermore, it means that for each particular crystal unit there is but one frequency at which it is supposed to operate. This does not mean that all crystal units of the same type and nominal frequency have a single common operating frequency, rather that each has its own individual frequency, which, however, will not differ from the nominal frequency by more than the permitted tolerance. It is the effective operating *reactance* that the crystal units must have in common. Approximately,

$$X_e = \frac{4\pi L\Delta f}{1 + \frac{4\pi L\Delta f}{X_{Co}}} \quad \text{(Equation (1), figure 1—98)}$$

Now, $\Delta f = f_p - f_s$, where f_p is the operating parallel-resonant frequency and f_s is the series-resonant frequency of the motional arm. Assume that a 10-mc parallel-resonant crystal unit has a frequency tolerance of ± 0.02 per cent. This is equivalent to an absolute frequency tolerance of ± 2000 cps. Two crystal units at opposite extremes could be within specifications even though their operating frequencies, f_{p1} and f_{p2}, were 4000 cps apart.

If the rated load capacitance were 32 $\mu\mu$f and the crystals were A elements, Δf, itself, for each crystal would be on the order of 2000 cps. If the crystals were B elements of the same shunt capacitance, C_o, Δf for each crystal would be only in the neighborhood of 800 cps, because of the B element's larger series-arm inductance, L. It becomes obvious that there can be no expectation of "pulling" the frequencies together by making slight adjustments in the load capacitance. The lower-frequency crystal could not be raised to zero beat with a frequency 4000 cps higher without reducing C_x several-fold. The higher-frequency crystal could not even be "pulled" to the nominal frequency and oscillations still be maintained. For this reason, the design engineer should generally not attempt to provide an operator with frequency adjustments for the crystal oscillator. The only adjustments needed are those which can be factory preset, in order to compensate for slight differences in stray capacitance. If the frequencies to be generated must be in close agreement with some standard, or with the frequency of some controlling station, the task is to provide oven-controlled crystal units of smaller tolerance. Only when the desired operating tolerance is less than any provided by crystal-unit specifications alone, is it necessary to provide the operator with a frequency adjustment knob. Even then, the adjustment need not provide a tuning range greater than the specified crystal tolerance. Since the smaller tolerances are only 1/10 to 1/20 of the 0.02 per cent in the example above, the total variation in load capacitance may not need to be greater than ± 10 per cent of the specified capacitance.

1-318. It may be desirable to provide an operator with the means of controlling the output voltage

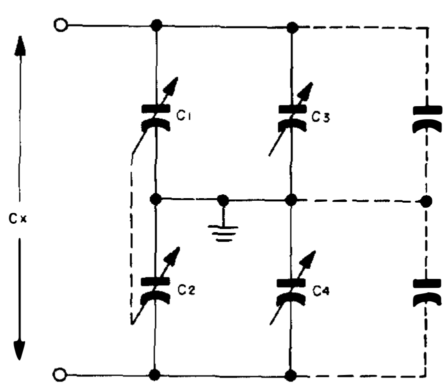


Figure 1-134. Ganged capacitances to enable adjustment of C_g/C_p ratio of Pierce circuit without changing total load capacitance

of a Pierce oscillator by varying the C_g/C_p ratio. In this case, care must be taken to ensure that the total load capacitance remains the same. If C_g and C_p are to be adjusted separately, some type of matching scales should be provided with the two tuning knobs, so that the correct load capacitance can always be had when, say, the two scales give the same reading. It is more desirable to have available ganged capacitors similar to those shown in figure 1-134 for each of the Military Standard capacitance ratings. The capacitors C_1 and C_2 in series are to be designed to always ensure a correct load capacitance when each is shunted by convenient predetermined fixed capacitance. The small variable capacitances C_3 and C_4 are adjusted until the sum of their values and the circuit stray capacitances, C_{s1} and C_{s2}, provide the correct fixed shunts for the ganged elements.

MODIFICATIONS OF PIERCE CIRCUIT

1-319. Figures 1-135, 1-136, 1-137, and 1-138 and their accompanying circuit-data charts reveal a great flexibility in the design of a Pierce oscillator. It would be very convenient to be able to put our finger on a single circuit and say that the design of this circuit is superior to all others. Unfortunately, this is not possible. One would first have to define what is meant by "superior design." The definition, at best would be a complex function of several physical and psychological variables. The very existence of a wide variety of circuit modifications suggests that no one circuit is superior to all others for all given tasks, although much of the variety can be attributed to the desire of the design engineer to create his own circuit and also to avoid the risk of possibly infringing upon the patent rights of another. Our space does not permit a detailed discussion of each of the circuits shown. Only a few of the highlights are to be mentioned. In general, most of the oscillators illustrated employ older-type crystal units; most of the circuits use B+ voltages on the order of 200 volts, and would overdrive the smaller-sized crystals currently recommended; and the load capacitances and the C_g/C_p ratios vary widely from one circuit to another. Those circuits that employ currently recommended Military Standard crystal units (crystal units having nomenclature type numbers, CR-15/U and higher) are designed to operate so that the crystal unit faces its rated parallel-mode load capacitance. In these circuits the plate supply is normally 100 to 120 volts, and the crystal unit of average resistance is operated at 4 to 5 milliwatts.

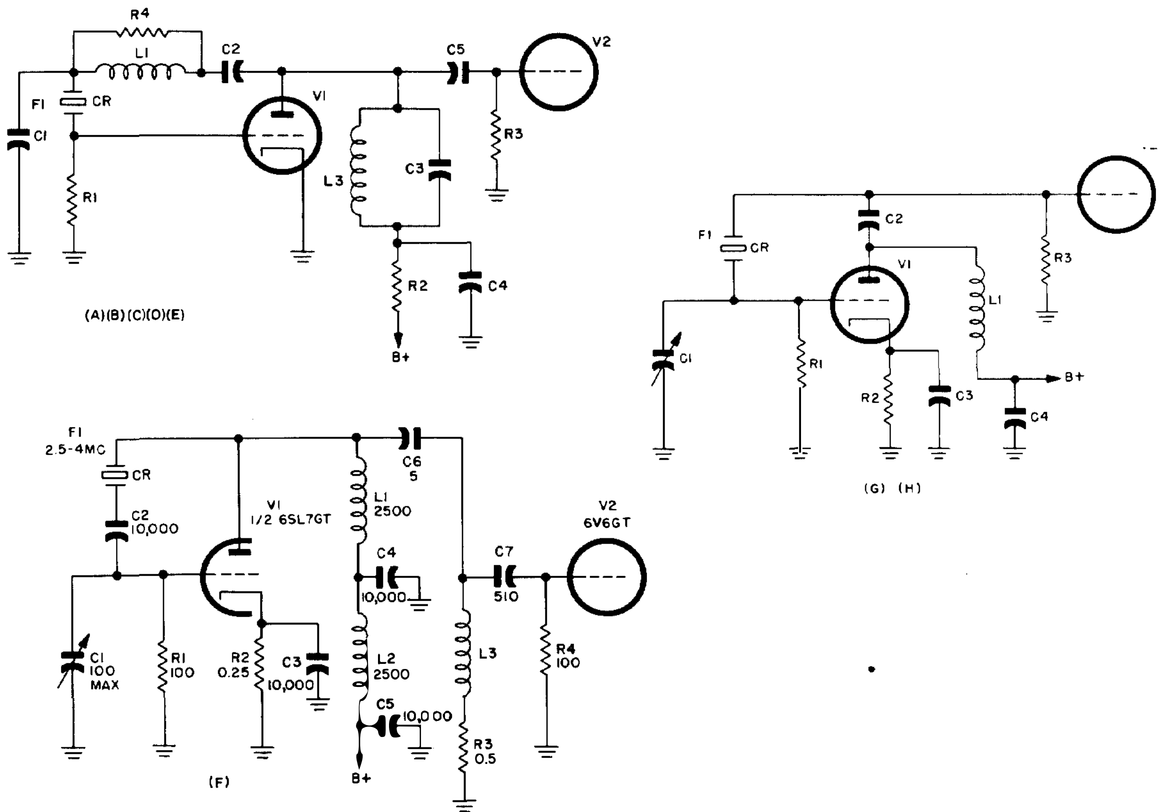
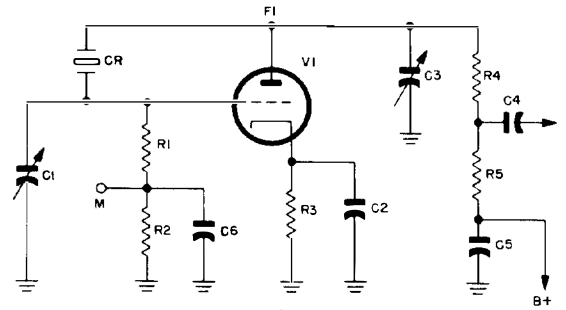


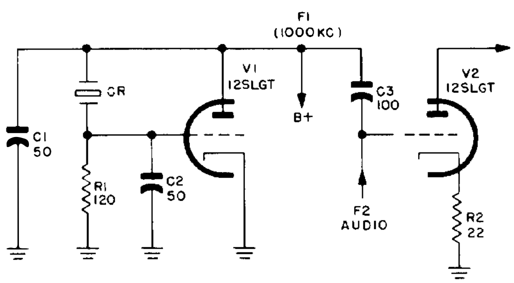
Figure 1-135. Modifications of Pierce oscillator using triodes

Fig.	Equipment	Purpose	F_1	F_2	F_3	CR	R_1	R_2	R_3
(A)	Radio Transmitter T-14(A/B/C)/TRC-1	M.O.	729-1041			CR-4/U (oven)	350	50	100
(B)	T-14(D/E)/TRC-1	M.O.	729-1041			CR-4/U (oven)	350	150	100
(C)	T-14H/TRC-1	M.O.	729-1041			CR-4/U (oven)	350	150	100
(D)	Test Oscillator TS-32(A/B)/TRC-1	Test oscillator	729-1041			CR-4/U (oven)	350	50	15
(E)	TS-32(C/D)/TRC-1	Test oscillator	729-1041			CR-4/U (oven)	350	47	15
(F)	Radio Transmitter T-177/FR	M.O.	2500-4000			FT-164 (oven)	100	0.25	0.5
(G)	Radio Transmitter Assembly OA-60B/FRT	M.O.	2000-4000			FT-164 (oven)	100	0.5	25
(H)	Radio Transmitter T-172/FR	M.O.	2000-4000			FT-164 (oven)	100	0.5	25
(I)	Radio Transmitter T-125A/ARW-34	M.O.	1000 (approx)	Audio for phase mod.		CANC 40138 Otis Elev. Co.	120	22	
(J)	Radio Receiver WE D-99945	2nd beating oscillator	3000			Entire circuit in oven			
(K)	Exciter Unit O-5/FR	M.O.	1800-5800			FT-249 (oven)	100	0.25	10

Circuit Data for Figure 1-135. F in kc. R in kilohms. C in $\mu\mu$f. L in μh.



(J)



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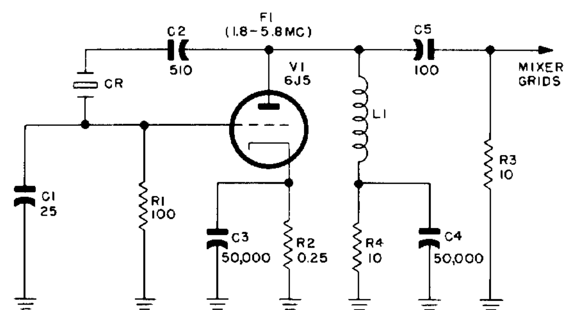
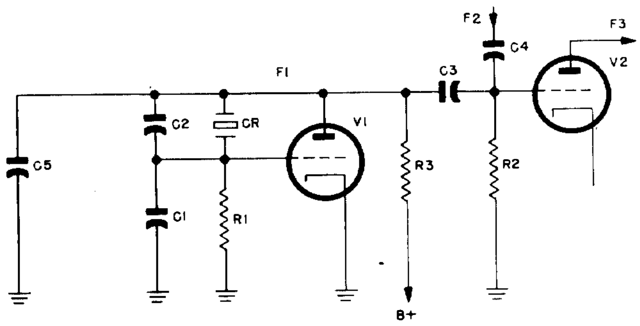


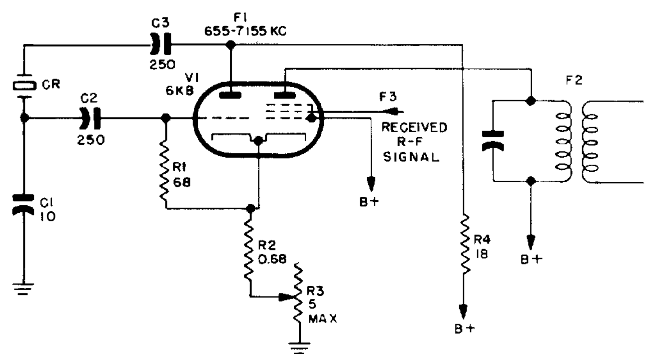
(K)

Figure 1-135. Continued

R_4	R_5	C_1	C_2	C_3	C_4	C_5	C_6	C_7	L_1	L_2	V_1	V_2
∞		40	∞	0	0	100			770	0	1/2 6SN7GT	6AC7
15		22	5000	40	5000	100			770	770	1/2 6SN7GT	6AC7
15		22	4700	39	4700	100			770	770	1/2 6SN7GT	6AC7
∞		40	∞	0	0	100			770	0	1/2 6SN7GT	1/2 6SN7GT
15		22	5100	39	5100	100			770	770	1/2 6SN7GT	1/2 6SN7GT
100		100	10,000	10,000	10,000	10,000	5	510	2500	2500	1/2 6SL7GT	6V6GT
		100	10,000	10,000	10,000				2500 (50Ω)		6J5GT/G	807
		100	10,000	10,000	10,000				2500 (50Ω)		6J5GT/G	807
		50	50	100							1/2 12SL7GT	1/2 12SL7GT
											WE272A	
10		25	510	50,000	50,000	100					6J5	



(L) (M)

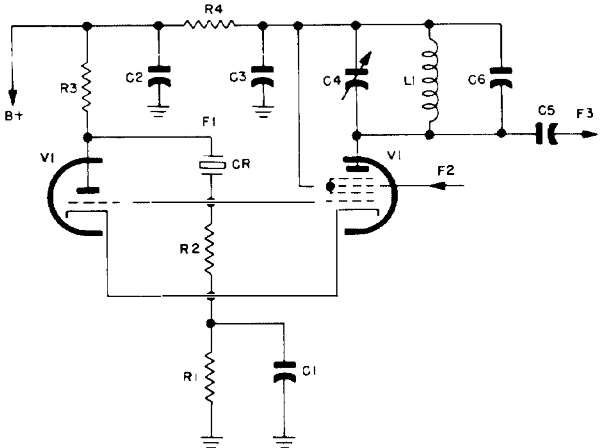


(N)

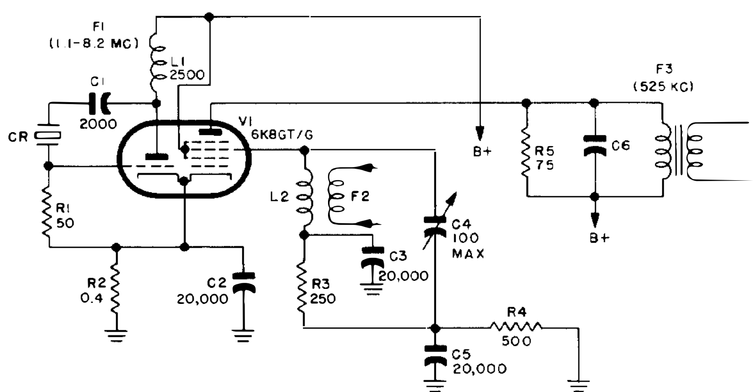
Figure 1-135. Continued

Fig.	Equipment	Purpose	F_1	F_2	F_3	CR	R_1	R_2	R_3
(L)	Receiver-Transmitter RT-173/ ARC-33	"Side-step" injector oscillator	7662.5	104,840-192,280	F_1+F_2	CR-18/U	47	68	68
(M)	Receiver-Transmitter RT-173/ ARC-33	Main channel local oscillator	12,517.8	15,325	F_2-F_1	CR-18/U	11	470	33
(N)	Lear Radio Set Model T-30AB-RCBBL-2	Local oscillator	655-7155	455 IF.			68	0.68	5
(O)	Communication Equipment AN/CRC-3	Local oscillator	4755-3845	4300 (1st I.F.)	F_1-F_2 or F_2-F_1 (455 I.F.)	FT-243	0 33	47	150
(P)	Radio Set AN/ VRC-2	Local oscillator	4755-3845	4300 (1st I.F.)	F_1-F_2 or F_2-F_1 (455 I.F.)	FT-243	0.33	47	150
(Q)	Radio Receiver R-114/VRC-4	Local oscillator	1175-8175	1700-8700	F_2-F_1 (525 I.F.)	Sig 2Z 3531B	50	0 4	250

Circuit Data for Figure 1-135. F in kc. R in kilohms. C in $\mu\mu$f. L in μh.



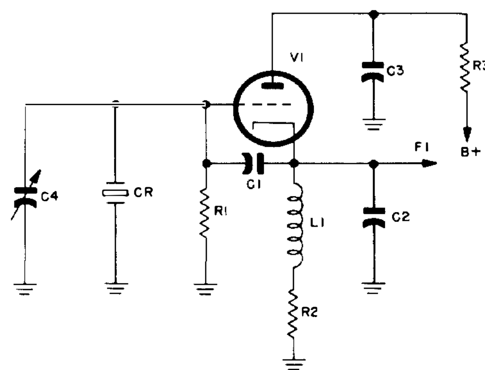
(O) (P)



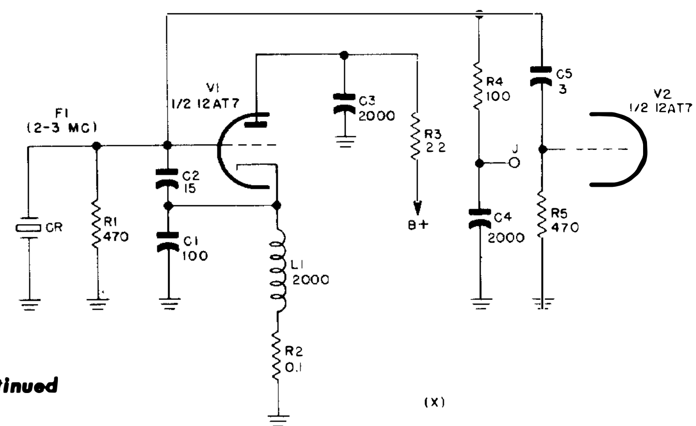
(Q)

Figure 1-135. Continued

R_4	R_5	C_1	C_2	C_3	C_4	C_5	C_6	C_7	L_1	L_2	V_1	V_2
		68	18	5	5	0					6AK5W	
		300	0	10	22	18					6AS6W	
18		10	250	250							6K8	
27		50,000	2,000	50,000	5-44	50	50				6K8GT	
27		50,000	2,000	50,000	5-14	50	50				6K8GT	
500	75	2,000	20,000	20,000	100	20,000			250u		6K8GT/G	



(R) (S) (T) (U) (V) (W)

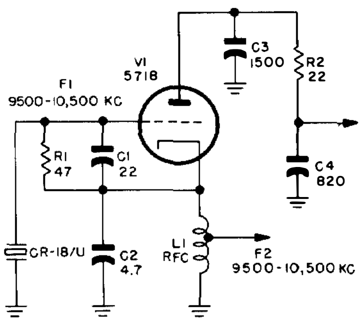


(X)

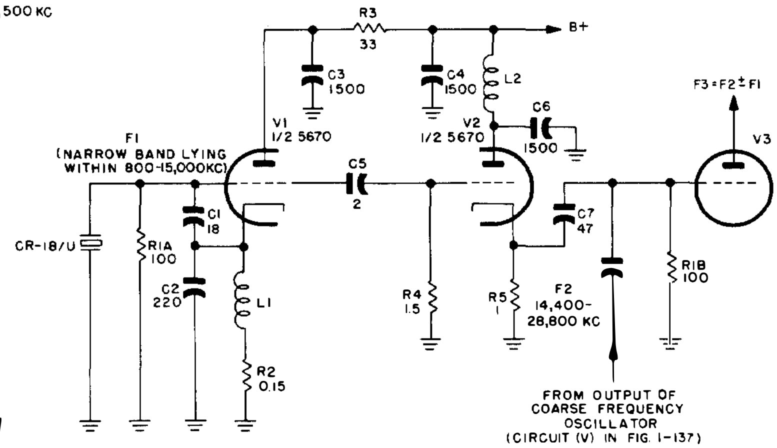
Figure 1-135. Continued

Fig.	Equipment	Purpose	F_1	F_2	F_3	CR	R_1	R_2	R_3
(R)	Receiver-Transmitter RT-178/ARC-27	2nd injector osc (rec guard channel)	15,950			CR-18/U	100	0.12	0.560+
(S)	Receiver-Transmitter RT-173/ARC-33	2nd monitor osc of trans. M.O.	3233.333–3900			CR-27/U	220	0	1
(T)	Receiver-Transmitter RT-173/ARC-33	3rd monitor oscillator of M.O.	3650–3800			CR-27/U	220	0	1
(U)	Receiver-Transmitter RT-173/ARC-33	4th monitor oscillator of M.O.	5172.917–5181.25			CR-27/U	220	0	1
(V)	Receiver-Transmitter RT-178/ARC-27	1st transmitter oscillator	3450			CR-18/U	100	0.56	100
(W)	Receiver-Transmitter RT-178/ARC-27	2nd trans. osc (heterodyned with 1st)	8250–9150			CR-18/U	100	0.56	33
(X)	Transmitter T-217/GR	1st i-f oscillator	1994–2894				470	0.1	2 2
(Y)	Radio Receiver-Transmitter RT-XA-101/ARC-22	Local osc in receiver	9500–10,500	F_1		CR-18/U	47	22	
(Z)	Signal Generator SG-13/ARN	Fine freq osc for mixing with output of coarse freq osc	Narrow band within 800–15,000 range (10 crystals)	14,400–28,800	$F_2 \pm F_1$	CR-18/U	100	0.15	33

Circuit Data for Figure 1-135. F in kc. R in kilohms. C in $\mu\mu$f. L in μh.



(Y)



(Z)

Figure 1-135. Continued

R_4	R_5	C_1	C_2	C_3	C_4	C_5	C_6	C_7	L_1	L_2	V_1	V_2
		24	100	3,000	0.3-3				500		1/2 12AT7	
		30	30	470					2000		1/2 5670	
		15	27	470					2000		1/2 5670	
		27	30	470					2000		1/2 5670	
		20	330	3000	0				500		1/2 12AT7	
		7	82	3000	1-8				500		1/2 12AT7	
100	470	100	15	2000	2000	3			2000		1/2 12AT7	1/2 12AT7
		22	4.7	1500	820				RFC		5718	
1.5	1	18	220	1500	1500	2	1500	47			½ 5670	½ 5670

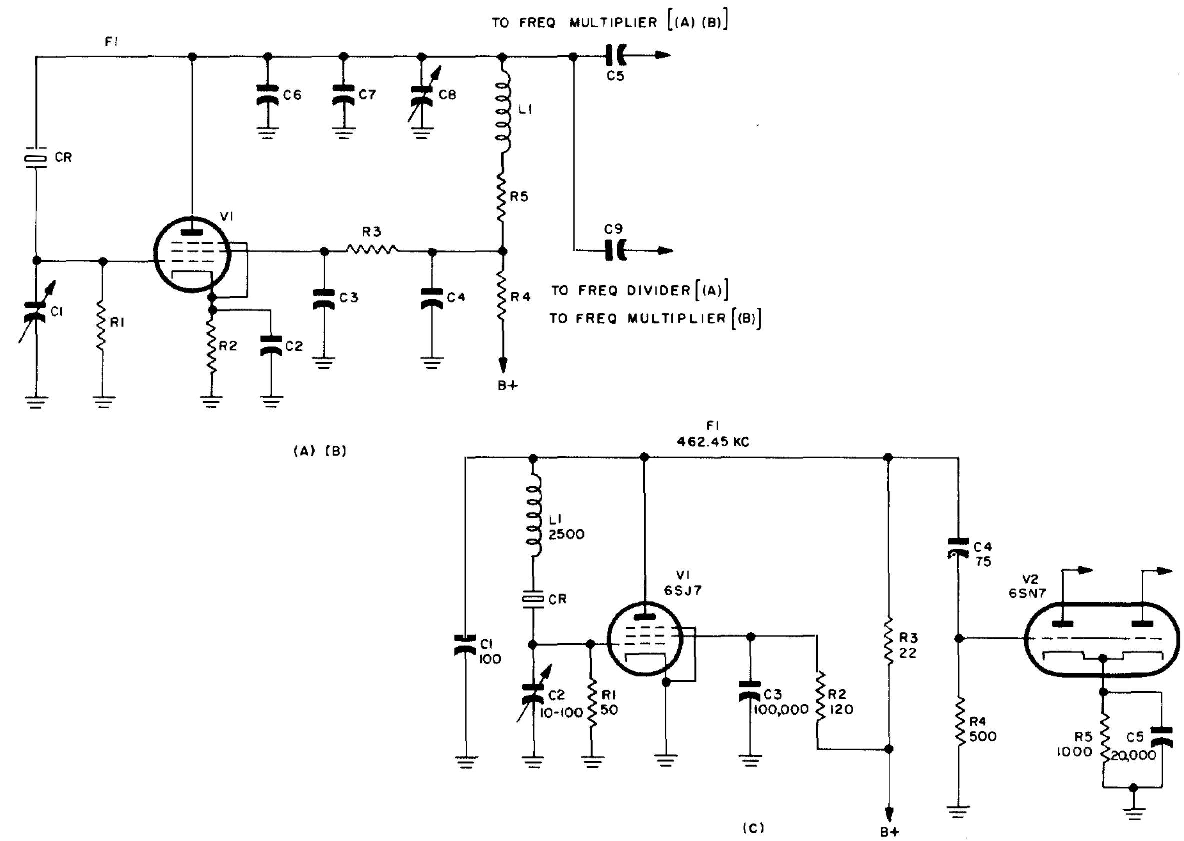
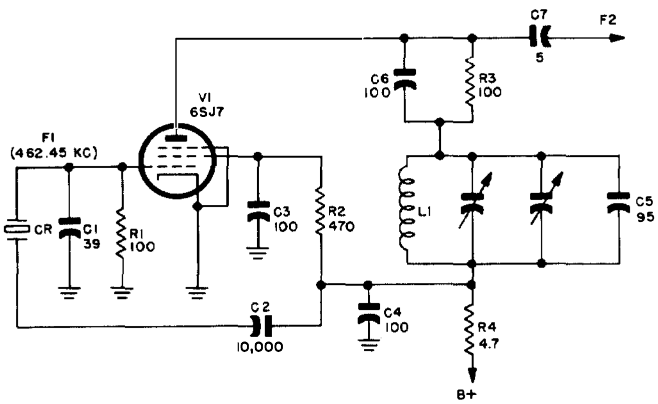


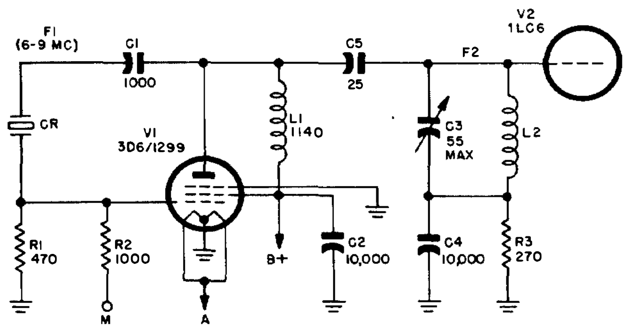
Figure 1-136. Modifications of Pierce oscillator using screen-grid tubes

Fig.	Equipment	Purpose	F_1	F_2	CR	R_1	R_2	R_3	R_4	R_5	R_6	R_7
(A)	Radio Set AN/FRC-10 (WECo Transmitter D-156000)	L-F osc for single-side-band operation	625		WECo 7B	250	1	1	20	0		
(B)	Same as (A)	H-F osc for single-side-band operation	940–5000		WECo 5AA	100	1	10		5		
(C)	Switching Unit SA-107 ( )/MRC-4	BFO for two diversity receivers	462.45			50	120	22	500	1000		
(D)	Radio Receiver R-270/FRR	BFO	462.45	F_1	Bliley SR-901 (FT-241A)	100	470	100	4.7			
(E)	Radio Receiver BC-659-( )	Local osc of receiver	5675–8650	$4F_1$	FT-243	470	1000	270				

Circuit Data for Figure 1-136. F in kc. R in kilohms. C in $\mu\mu$f. L in μh.



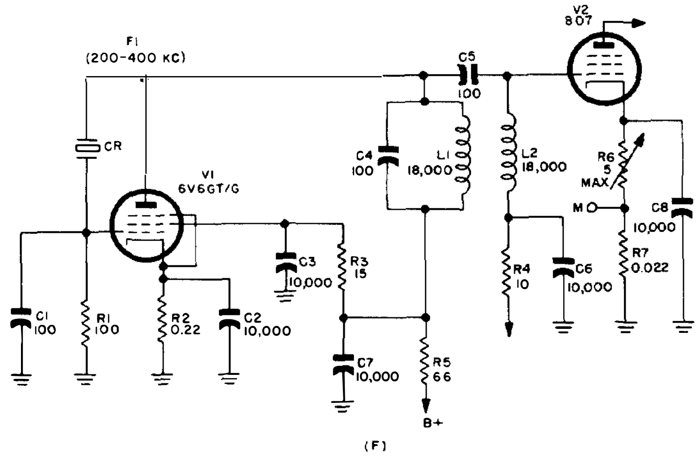
(D)



(E)

Figure 1-136. Continued

C_1	C_2	C_3	C_4	C_5	C_6	C_7	C_8	C_9	L_1	L_2	V_1	V_2
100	10,000	10,000	10,000	1000	250	50	100	10	200		RCA41	
Variable	10,000	10,000	10,000	1000	0	0	0	1000	0		RCA41	
100	10–100	100,000	75	20,000					2500 (25Ω)		6SJ7	6SN7
39	10,000	100	100	95	100	5					6SJ7	
1000	10,000	55	10,000	25					1140	8 turns 5/8-in. dia	3D6/1299	1LC6



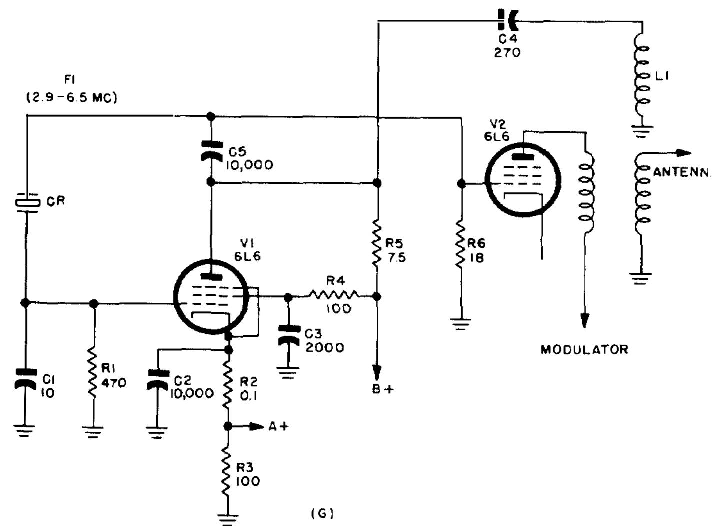
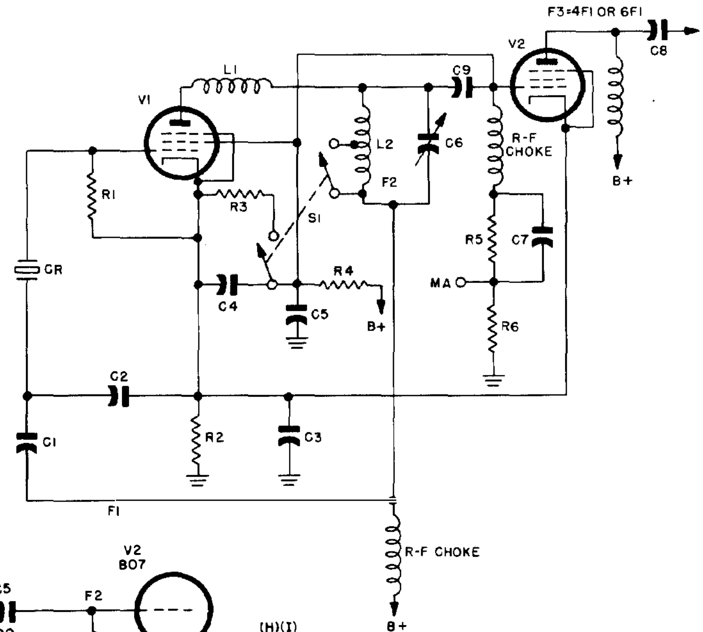


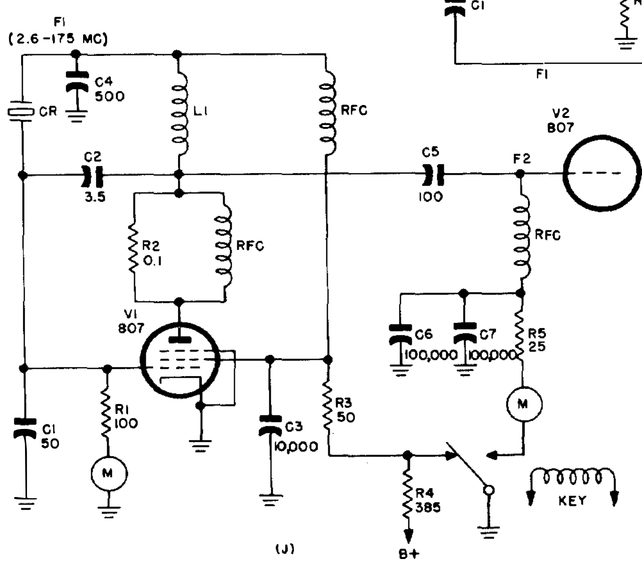
Figure 1-136. Continued

Fig.	Equipment	Purpose	F_1	F_2	CR	R_1	R_2	R_3	R_4	R_5	R_6	R_7
(F)	Radio Transmitter BC-329-N	M.O.	200-400		FT-249	100	0.22	15	10	66	5	0.022
(G)	Lear Radio Set T-30AB-RCBBL-2	M.O.	2900-6500			470	0.1	100	100	7.5	18	
(H)	Modulator-Transmitter T-233/URW-3	M.O. for remote-control transmitter	3966.6-5666.6	$2F_1$	CR-1A/AR	47	0.33	6.8	30	27	0.0051	
(I)	Modulator-Transmitter BC-1158	M.O.	3966.6-5666.6	$2F_1$	CR-1A/AR	47	0.33	6.8	30	27	0.0051	
(J)	Radio Transmitter 12GLX-2	M.O.	260-1750	F_1	Holder FT-249	100	0.1	50	385	25		

Circuit Data for Figure 1-136. F in kc. R in kilohms. C in $\mu\mu$f. L in μh.



(H)(I)



(J)

Figure 1-136. Continued

C_1	C_2	C_3	C_4	C_5	C_6	C_7	C_8	C_9	L_1	L_2	V_1	V_2
100	10,000	10,000	100	100	10,000	10,000	10,000		18,000	18,000	6V6 GT/G	807
10	10,000	2000	270	10,000							6L6	6L6
5100	510	10,000	2400	2400	45	470	470				1/2 815	1/2 815
5100	510	10,000	2400	2400	45	470	470				1/2 815	1/2 815
50	3.5	10,000	500	100	100,000	100,000			Per F_1		807	807

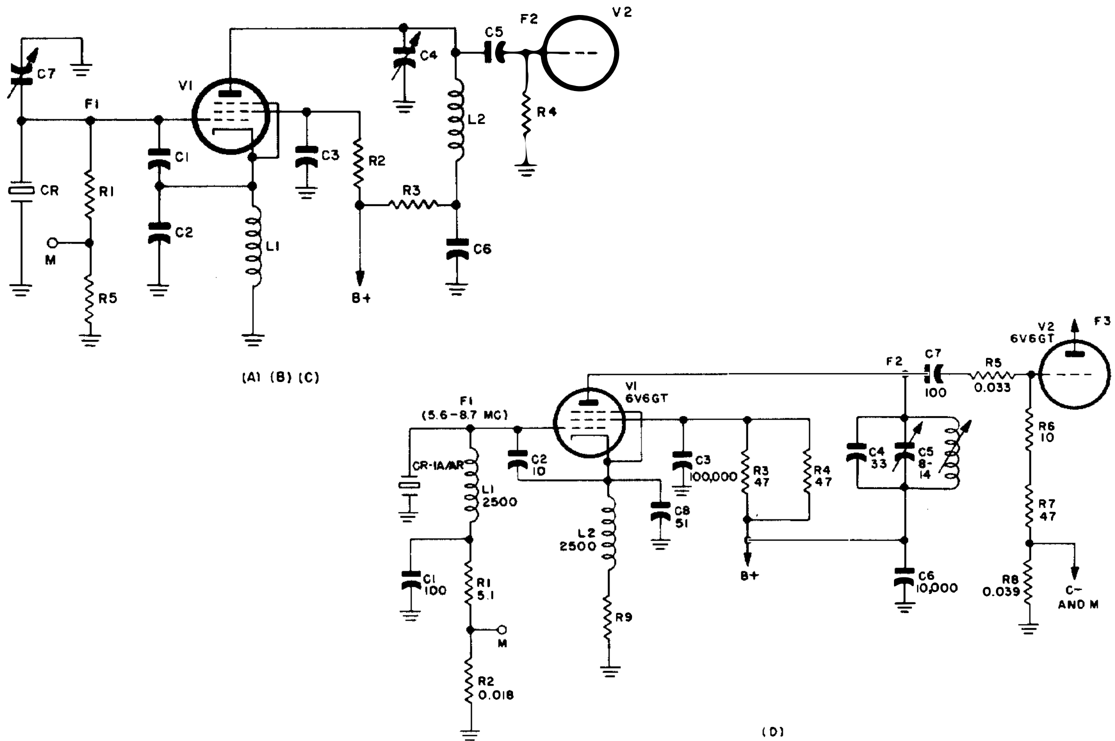
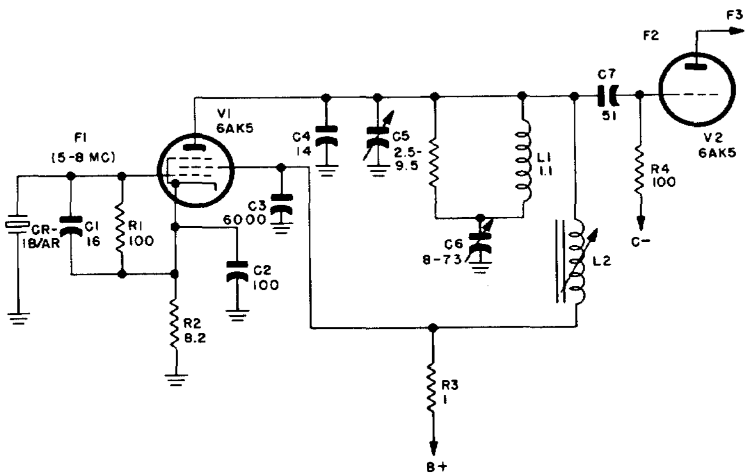


Figure 1-137. Electron-coupled Pierce oscillator modifications. All circuits except circuit (A) provide frequency multiplication

Fig.	Equipment	Purpose	F_1	F_2	F_3	CR	R_1	R_2	R_3	R_4	R_5
(A)	Receiver-Transmitter RT-178/ARC-27	Spectrum oscillator	10,000	F_1		CR-27/U	100	56	12	100	0
(B)	Radio Receiver R-252/ARN-14	2nd monitor osc of receiver vfo	11,275–11,725	F_1		CR-33/U	8.2	68	0		2.2
(C)	Radio Transmitter BC-400-(B,C,D,E)	M.O. for marker beacon	4166.67	$2F_1$		FT-164	200	35	5	20	0
(D)	Radio Transmitter T-67/ARC-33	M.O.	5555–8666	$2F_1$	$6F_1$	CR-1A/AR	5.1	0.018	47	47	0.033
(E)	Radio Set AN/ARC-1A	Main-channel heterodyne freq osc	5020–8120	$2F_1$	$18F_1$	CR-1A/AR or CR-18/AR	100	8.2	1	100	
(F)	Frequency Meter TS-186/UP	Crystal calibrator	5000	$2F_1$		Navy CG-40210; GE #G31 Thermocell. 6L6 tube envelope; heater; ± 0.002 per cent, −20 to 75°C	1.5	100	100	10	100
(G)	Frequency Meter TS-186(B/C)/UP	Crystal calibrator	5000	$2F_1$		CR-18/U (oven, 60°C)	1.5	100	100	10	39

Circuit Data for Figure 1-137. F in kc. R in kilohms. C in $\mu\mu$f. L in μh.



(E)

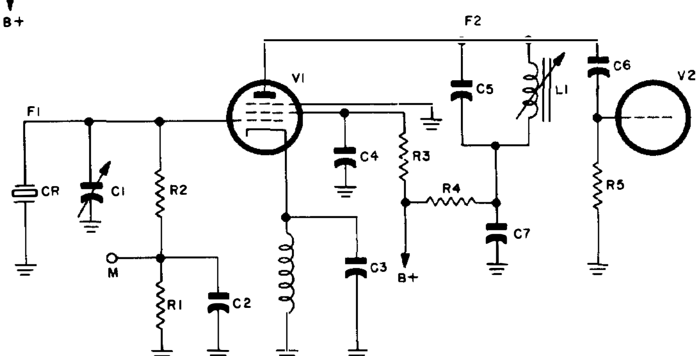
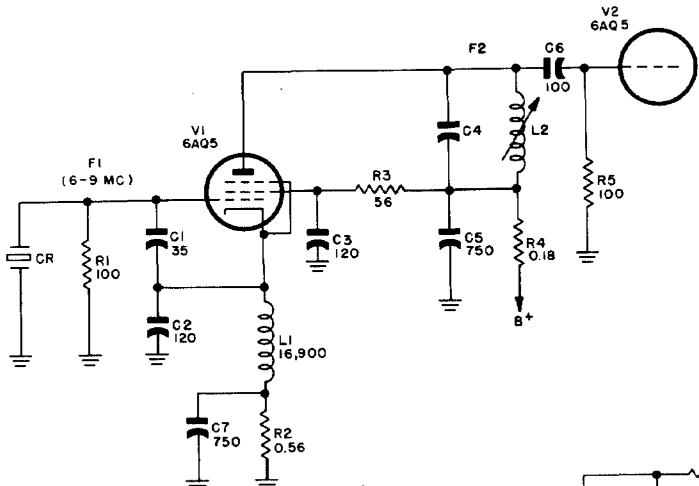


Figure 1-137. Continued

(F) (G)

R_6	R_7	R_8	C_1	C_2	C_3	C_4	C_5	C_6	C_7	C_8	L_1	L_2	V_1	V_2
			20	220	2000	0	100	0	1.5-7		500	0	6AG5	6AG5
			22	150	1800	0	470	1800	0				5654/ 6AK5	Discrim-inator rectifier
			22	100	1500	35	1500	1500	0		1000		6L6	6L6
10	47	0.039	100	10	100,000	33	8-114	10,000	100	51	2500	2500	6V6GT	6V6GT
			16	100	6000	14	2.5-9.5	8-73	51		1.1	15 turns	6AK5	6AK5
			4-12	470	100	470	22	100	470		4-9	35 (7Ω)	6SJ7	6SJ7
			6-36	470	100	470	22	51			4-9	35 (7Ω)	6SJ7	6SJ7



(H)

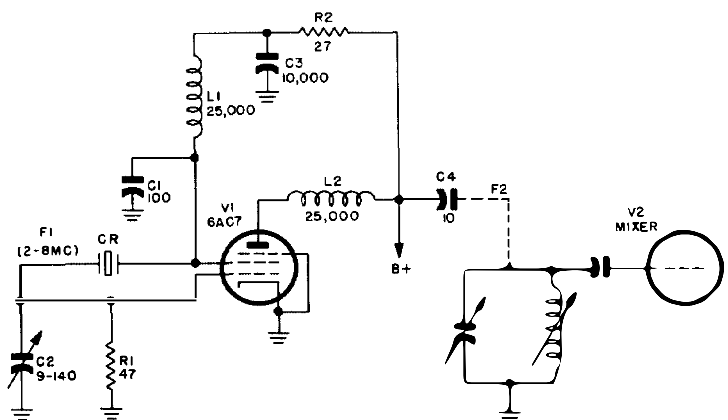
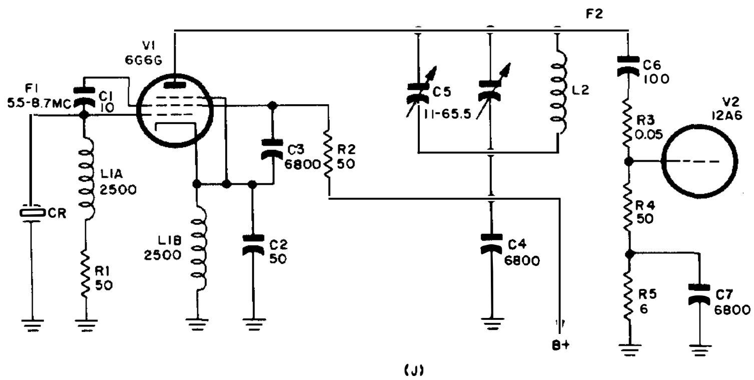


Figure 1-137. Continued

(I)

Fig.	Equipment	Purpose	F_1	F_2	F_3	CR	R_1	R_2	R_3	R_4	R_5
(H)	Aircraft Radio Corp. Radio Transmitter ARC Types T-13 and T-11A	M.O.	6000–9000	$3F_1$		CAATC #1081	100	0 56	56	0.18	100
(I)	Radio Receiver R-270/FRR	Local oscillator	1965–8511.6	F_1, $2F_1$, or $3F_1$		HC-1/U Holder	47	27			
(J)	Radio Transmitter BC-655-A,-AM	M.O.	5560–8660	$2F_1$		DC-11-( ), DC-16, DC-26, or CR-1( )/AR	50	50	0.05	50	6
(K)	Radio Transmitter BC-401-B	M.O.	1000–4575	$2F_1$			100	0.51			

Circuit Data for Figure 1-137. F in kc. R in kilohms. C in $\mu\mu$f. L in μh.



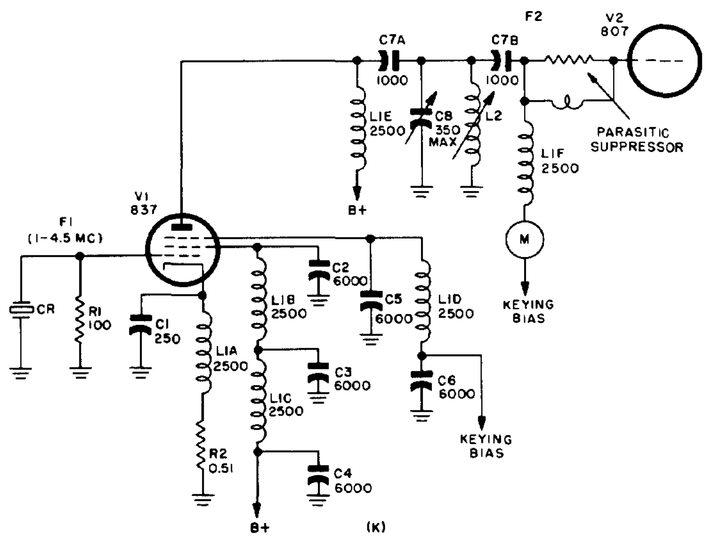
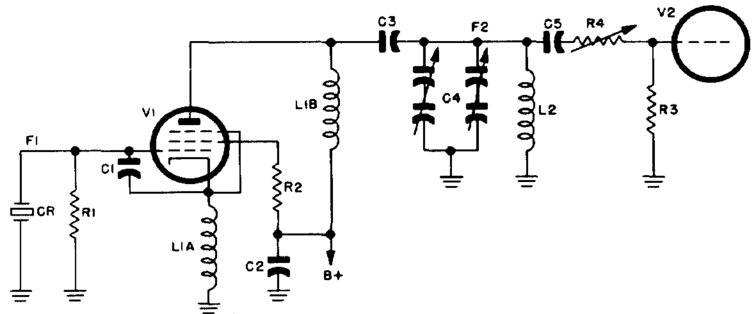


Figure 1-137. Continued

R_6	R_7	R_8	C_1	C_2	C_3	C_4	C_5	C_6	C_7	C_8	L_1	L_2	V_1	V_2
			35	120	120	40 (T-13) 50 (T-11)	750	100	750		16,900		6AQ5	6AQ5
			100	9-140	10,000	10					25,000 (160Ω)	25,000 (160Ω)	6AC7	Mixer
			10	50	6800	6800	11-65.5	100	6800		2500 (50Ω)	9-1/2 turns	6G6G	12A6
			250	6000	6000	6000	6000	6000	1000	350	2500		837	807



(L)(M)

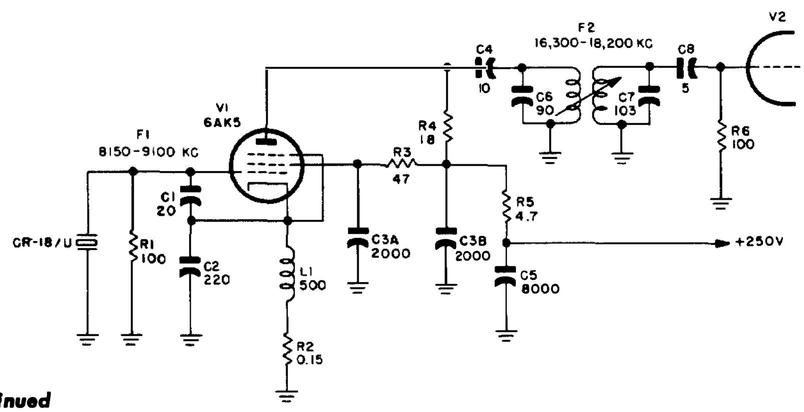
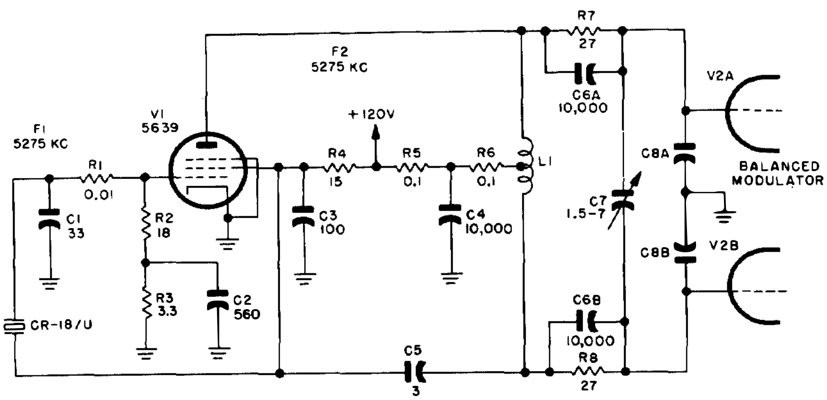


Figure 1-137. Continued

(N)

Fig.	Equipment	Purpose	F_1	F_2	F_3	CR	R_1	R_2	R_3	R_4	R_5
(L)	Frequency Meter BC-638-A	M.O. for signal generator	5555.55-8666.6	$2F_1$ (?)		Sig 2Z 3501-11A (Bendix)	100	40	150	0-2.2	
(M)	Frequency Meter BC-1420	M.O. for signal generator	5555.55-8666 6	$2F_1$ (?)			100	40	150	0-2.2	
(N)	Radio Receiver R-540/ARN-14C	Low freq injection osc	8150-9100 (20 crystals)	$2F_1$		CR-18/U	100	0 15	47	18	4 7
(O)	Radio Set AN/ARC-34(XA-1)	Sidestep osc	5275	F_1		CR-18/U	0 01	18	3 3	15	0.1
(P)	Radio Set AN/ARC-34(XA-1)	1st monitor osc of transmitter M.O.	833.333	F_1	F_1	CR-27/U	560	56	10	1000	

Circuit Data for Figure 1-137. F in kc. R in kilohms. C in $\mu\mu$f. L in μh.



(O)

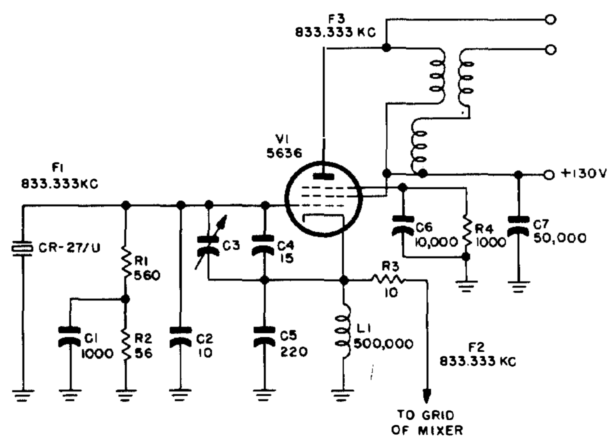
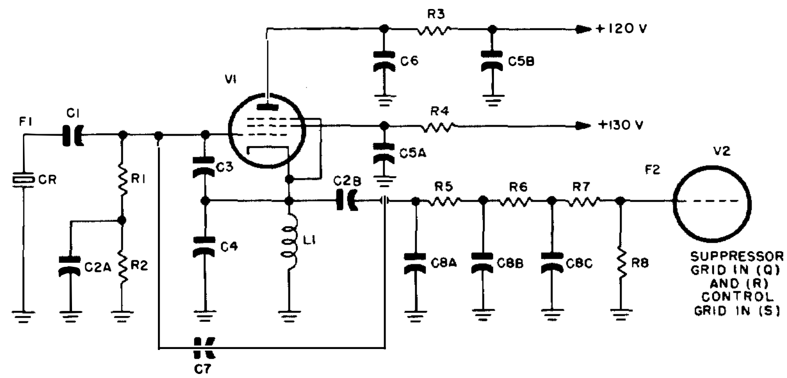


Figure 1-137. Continued

(P)

R_6	R_7	R_8	C_1	C_2	C_3	C_4	C_5	C_6	C_7	C_8	L_1	L_2	V_1	V_2
			25	10,000	3000	6–42 (each sect- ion)	1000				1700 (40Ω)	7 5	6SK7	9003
100			25	100	3000	6–42 (each sect- ion)	1000				1700 (40Ω)	9.3	6SK7	9003
			20	220	2000	10	8000	90	103	5	500		6AK5	
0 1	27	28	33	560	100	10,000	3	10,000	1.5–7				5639	Balanced modulator
			1000	10		15	220	10,000	50,000		500		5636	



(Q) (R) (S)

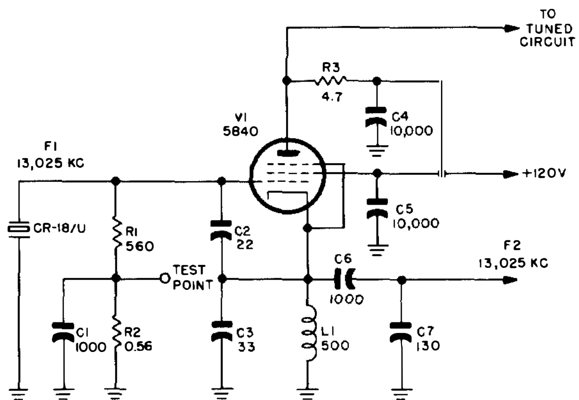
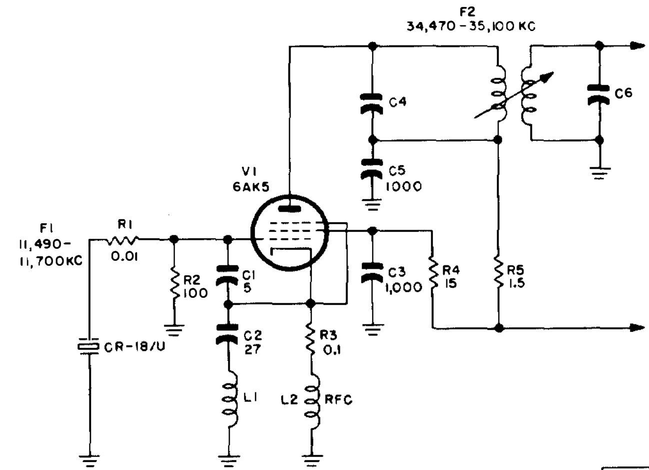


Figure 1-137. Continued

(T)

Fig.	Equipment	Purpose	F_1	F_2	F_3	CR	R_1	R_2	R_3	R_4	R_5
(Q)	Radio Set AN/ARC-34(XA-1)	2nd monitor osc of transmitter M.O.	3230–3900 (5 crystals)	F_1		CR-27/U	560	56	4.7	1.8	0.68
(R)	Radio Set AN/ARC-34(XA-1)	3rd monitor osc of transmitter M.O.	3650–3775 (4 crystals)	F_1		CR-27/U	560	56	4.7	1.8	0
(S)	Radio Set AN/ARC-34(XA-1)	4th monitor osc of transmitter M.O.	5130–5165 (5 crystals)	F_1		CR-27/U	560	56	4.7	1.8	0
(T)	Radio Set AN/ARC-34(XA-1)	2nd osc for guard channel	13,025	F_1		CR-18/U	560	0.56	4.7		
(U)	Radio Receiver R-322/ARN-18	Local osc	11,490–11,700 (20 crystals)	$3F_1$		CR-18/U	0.01	100	0.1	15	1.6
(V)	Signal Generator SG-13/ARN	Coarse freq osc for mixing with output of fine freq osc	4800–9600 (28 crystals)	$3F_1$		CR-18/U	100	0.15	100	18	

Circuit Data for Figure 1-137. F in kc. R in kilohms. C in $\mu\mu$f. L in μh.



(U)

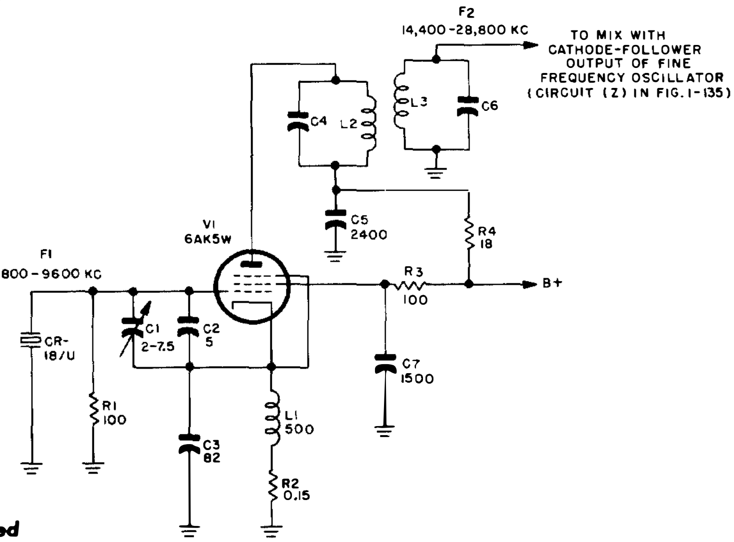


Figure 1-137. Continued

(V)

R_6	R_7	R_8	C_1	C_2	C_3	C_4	C_5	C_6	C_7	C_8	L_1	L_2	V_1	V_2
1	0.47	10	47	1000	10	33	10,000	470	0	22	5000		5840	5636
0	0	10	47	1000	10	33	10,000	470	12	0	5000		5840	5636
0	0	2.2	47	1000	10	33	10,000	470	10	0	5000		5840	5840
			1000	22	33	10,000	10,000	1000	130		500		5840	
			5	27	1000		1000					RFC	6AK5	
			2-7.5	5	82		2400		1500		500		6AK5W	

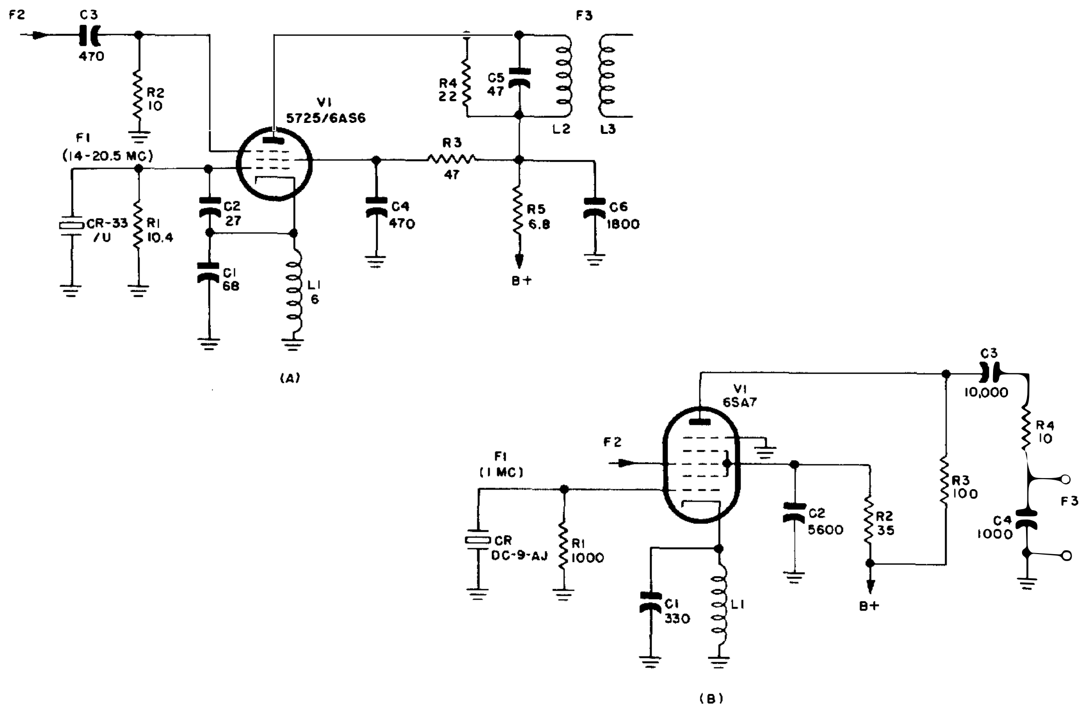


Figure 1-138. Electron-coupled Pierce oscillator modifications for heterodyne circuits

Fig.	Equipment	Purpose	F_1	F_2	F_3	CR	R_1	R_2
(A)	Radio Receiver R-252/ARN-14	1st monitor osc of receiver vfo.	14,000-20,500	From isolation amplifier of vfo.	F_2-F_1	CR-33/U	10 4	10
(B)	Signal Generator TS-413/U	Crystal harmonic heterodyne mixer	1000	Variable *rf*	$nF_1 \pm F_2$ (n = 1,2, . .)	DC-9-AJ	1000	35
(C)	Radio Receiver R-146A/ARW-35	2nd heterodyne osc	5456	5000	F_1-F_2	FT-243	47	47
(D)	R-F Signal Generator Set AN/URM-25C	Crystal calibrator harmonic generator and mixer	1000	10-50,000 (output from variable osc)	$nF_1 \pm F_2$	CR-18/U	1	270
(E)	Radio Receiver R-277 (XA-A)/ APN-70	Local osc	400 480 2850 2950 3000 3050	F_1-F_3	300 1100 (IF)	CR-25/U (LF_1) CR-18/U (HF_1)	22	0.068

Circuit Data for Figure 1-138. F in kc. R in kilohms. C in $\mu\mu$f. L in μh.

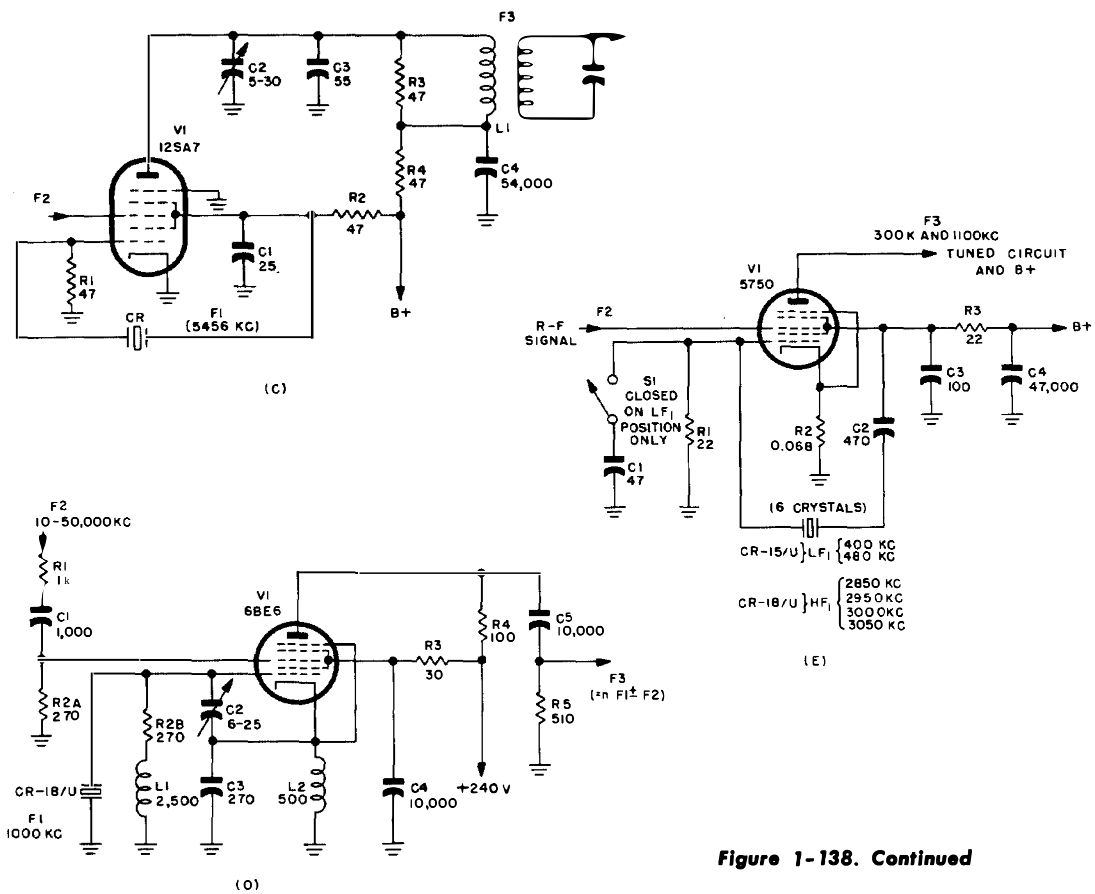


Figure 1-138. Continued

R_3	R_4	R_5	C_1	C_2	C_3	C_4	C_5	C_6	L_1	L_2	L_3	V_1
47	22	6 8	68	27	470	470	47	1800				5725/ 6AS6
100	10		330	5600	10,000	1000			2500			6SA7
47	47		25	5-30	55	50,000						12SA7
30	100	510	1000	6-25	270	10,000	10,000		2500	500		6BE6
22			47	470	100	47,000						5750

Preventing B^+ Voltage from Existing Across Crystal Unit in Pierce Circuit

1-320. There are two commonly used methods for preventing the application of the B^+ voltage across the crystal unit. One method is to connect a blocking capacitor in series with the crystal unit. The other method is to connect the crystal unit directly to ground, and the plate to ground through an r-f bypass capacitor. With this arrangement, the cathode must be operated above r-f ground. Except for the advantages of electron-coupling to the load to be had in the tri-tet circuit, which uses screen-grid tubes, the r-f-grounded-plate modification is not to be preferred since the plate-to-grid capacitance directly shunting the crystal unit is greatly increased by the addition of the grid-to-ground capacitance. When a capacitor is used to block the B^+ voltage from the crystal, a number of modifications are possible, three of which are shown in figures 1-135 (F), (G), and (K). In each figure, the blocking capacitor is the one labeled C_2, and has a reactance negligible with respect to that of the crystal unit. The more usual circuit arrangement is that shown in figure 1-135 (K). In figure 1-135 (G), C_2 plays a dual function in blocking the B^+ voltage from both the crystal and the grid circuit of the succeeding stage. The circuit in figure 1-135 (F), although not effective in blocking the B^+ voltage from the crystal unit, does, of course, reduce the d-c potential across the crystal equally as well as the other arrangements. It is desirable to keep the d-c potential across the crystal unit low; otherwise the crystal may be heavily strained in one direction, and the chance is increased that the elastic limit of the crystal will be approached on the alternation of the a-c voltage of the same polarity, or, at least, that the effect on the crystal will cause the performance characteristics to deviate from test specifications. Since the crystal unit, itself, is a capacitor of considerably greater dielectric thickness and smaller cross-sectional area than the usual blocking capacitor, it might seem questionable that the d-c voltage across the crystal unit could be expected to be significantly reduced. Certainly in the static state, an air-gap crystal unit should have at least as high a resistance to leakage currents as would the blocking capacitor. However, in the dynamic state some degree of ionization will occur if the voltage is excessive. The consequent leakage tends to charge the blocking capacitor to the full plate-to-grid d-c voltage. An unchecked B^+ voltage not only can increase the likelihood of corona effects, but also can lead to continuous discharge should corona losses once begin, and to an increase in the effective resistance of the crystal unit, a reduction of the grid bias, and perhaps even to arcing and puncturing of the crystal. A blocking capacitor, if it does not remove the d-c potential completely from across the crystal unit, at least can ensure that the potential is not sufficient to cause ionization. A blocking capacitor is generally more important in high-drive circuits employing air-gap crystal units.

1-321. Figure 1-135 (X) shows one example of a grounded-crystal, grounded-plate Pierce circuit. At first glance such an oscillator might very easily be mistaken for the Miller type. C_3 is a relatively large capacitance that causes the plate to be at r-f ground. C_1 is effectively C_p, the plate-to-cathode capacitance, and C_2 serves as the lumped part of C_g. By interchanging C_1 and C_3 and making the ground connection of the crystal a plate connection, essentially the same oscillator characteristics are obtained except that the B^+ is across the crystal unit. Note that the circuit is designed for the maximum possible output voltage, in that the output is taken across the crystal unit directly, instead of across C_1, the effective plate capacitance, alone. The fact that the plate is at r-f ground does not remove the grid-to-plate capacitance from shunting the crystal unit, but adds to this the grid-to-ground capacitance. In the circuit of figure 1-135 (R), a small tuning capacitance, C_4, is also shunted across the crystal unit. In addition to this there is the extra shunt capacitance contributed by the oven in which the crystal is mounted. In circuit (R), the C_1/C_2 ratio, which is approximately equal to the C_g/C_p ratio, is on the order of ¼. The gridleak losses are increased somewhat, since the grid resistance is connected across the crystal instead of across the grid capacitance. All these factors tend to reduce the effective Q_f of the feedback circuit, so it would appear that with crystal units of greater than average resistance the circuit operates with the tank considerably detuned from resonance. The fact is, however, that connecting the grid resistance across the crystal serves to concentrate most of the grid losses in the plate-to-grid circuit and to eliminate them from the grid-to-cathode circuits, and this probably increases the effective feed-back Q_f more than the increased losses diminish it. The circuit in figure 1-135 (X) is similar to that in figure 1-135 (R) except that the tuning capacitance has been eliminated and the output is obtained directly across the crystal unit. It is claimed that this arrangement tends to smooth the output and to reduce the harmonics. What probably is meant is that for a given output voltage the harmonic content is less. This can readily be seen, for the voltage across the crystal unit is equal to the sum of the voltages across C_1 and C_2. If the same output is to be taken across either capacitance alone, the excitation must be increased and the circuit will generally be operated beyond tube cutoff a greater fraction of the time, thereby increasing the higher-harmonic content. It should also be noted that the output arm in figure 1-135 (X), since it shunts the crystal, is effectively part of the feed-back circuit. Should the load increase or decrease, so also will the excitation. The circuit is thus a tri-tet modification where the output voltage is somewhat stabilized against changes in the load, but only at the sacrifice of frequency stability. This feature is not important in the particular fixed-load circuit of figure 1-135 (X), since the load in this circuit appears to be reasonably constant.

Electron-Coupled Pierce Oscillator

1-322. The electron-coupled oscillator permits a remarkable freedom from coupling between the plate load circuit and the crystal circuit. Screen-grid tubes are required, with the screen grid serving as the plate of the oscillator circuit. When electron coupling is used in conjunction with Pierce oscillators, the tri-tet arrangement, where the plate load circuit is in series with the oscillator tank, is generally the most advantageous, and is used in all the circuits shown in figure 1-137 except in circuits (I) and (O). With the screen at r-f ground, the vacuum-tube plate circuit performs as a conventional pentode r-f amplifier, with the excitation of the control grid a function of both the screen and plate r-f currents. Variations in the plate impedances have much less effect upon the frequency than do similar variations in the oscillator tank impedances. For this reason, even if the oscillator is to operate over a wide range of crystal frequencies, a tunable coil and capacitor tank can be placed in the plate circuit to obtain a smoother sine-wave output without running the risk of greatly changing the load capacitance of the crystal circuit. In effect, the electron-coupled oscillator reduces by one the number of amplifier stages that are required, and hence is particularly applicable for small portable transmitters where the crystal circuit must perform as nearly as possible the function of a power oscillator. The widest application of the electron-coupled Pierce circuit is for the purpose of frequency multiplication. In figure 1-137 (D), for example, the plate tank circuit is tuned to twice the crystal frequency. The L/C ratio of the plate tank should be as small as practicable, in order to increase the output selectivity and to ensure a low-impedance bypass through the coil for the fundamental frequency and through the capacitor for all harmonics higher than the second. The larger the angle θ during which the tube is cut off, the larger will be the percentage of the higher-harmonic generation in the output. In general, the lower the order of the harmonic, the greater is its energy content. For optimum output, the tube should not be heavily conducting during a positive alternation of the plate harmonic voltage, E_h. During such intervals the plate tank would be losing energy to the circuit at an instantaneous rate of $i_b e_h$, where i_b is the instantaneous d-c plate current and e_h is the instantaneous harmonic voltage across the plate tank. To meet the requirements above, plate current should be allowed to flow only during the interval of approximately one alternation of a harmonic cycle. Since the tube is to be cut on and off at the fundamental frequency, the plate tank, after receiving a pulse of energy during one alternation of a harmonic cycle, must oscillate freely for the remaining part of the fundamental period. If the frequency is being doubled, plate current should flow approximately one-fourth the time; if the frequency is being tripled, plate current should flow approximately one-sixth of the time. To generalize, if the frequency is to be multiplied n times, optimum n'th harmonic output is approached if the oscillator is designed so that the tube conducts approximately 1/2n of each fundamental cycle. In paragraph 1-312, it was found that for a given peak value of i_b, (I_{bm}), the effective I_p was essentially constant for all bias voltages between class-A and class-B operations, although a small maximum occured when the tube was cut off during three-fifths of the negative alternation. See equation 1-312 (20). If the plate load is constant, as is the case in the electron-coupled circuit, this point of maximum I_p coincides with the conditions of maximum output. If we assume that approximately the same conditions hold in the case of frequency multiplication, maximum harmonic output is approached if the tube continuously conducts 7/10 of the period of one harmonic cycle, or 7/10n of the period of the fundamental cycle. Since the maximum is not at all sharp, the optimum operating conditions are not critical and can be assumed to extend over a range which permits the tube to conduct from 1/2n to 7/10n of the time. That is, for optimum output, the oscillator section can be designed so that the crystal unit of average resistance allows the tube to be cut off during a fundamental-cycle angle within the range given by

$$\text{(optimum)} \quad \theta = \frac{\pi(2n - 1)}{n} \text{ to } \frac{\pi(10n - 7)}{5n} \qquad 1\text{—}322\ (1)$$

The principal advantage of the tri-tet circuit is that the excitation voltage tends to increase with an increase in load. If the output is to be inductively coupled to the succeeding stage, the tri-tet arrangement tends to stabilize the output voltage when the coefficient of coupling is varied. This feature originally found its greatest popularity among radio amateurs, although it was used principally in conjunction with Miller rather than Pierce oscillators. Figure 1-139 shows the basic tri-tet circuit as applied to Pierce and to Miller oscillators. It is the adaptability to frequency multiplication rather than to variable load conditions, however, that is of greatest importance when considering the tri-tet circuit for use in military equipments. The tri-tet frequency stability is low compared with that of the conventional pentode circuit because of the large stray capacitance (approximately 8 $\mu\mu$f) that directly shunts the crystal unit. About 4 $\mu\mu$f is the $C_{g_1g_2}$ of the tube, and the remainder is the capacitance of the grid leads to ground, which otherwise would be part of C_g.

Miscellaneous Pierce Circuit Modifications

1-323. Circuits (A) and (B) in figure 1-135 are of interest because they indicate two stages in the development of a particular modification of the Pierce oscillator. Originally L_3, C_3, C_4, and R_4 were not present. Since no grid capacitance is employed other than that of the tube, the C_g/C_p ratio is very small. The tube has approximately 4 $\mu\mu$f capacitance between plate and grid, and it may be assumed that the crystal oven adds a comparable amount directly across the crystal. In all probability the phase-shifting Q_f of the feed-back circuit is rather low, so that the tank can be expected to be more reactive than resistive. L_1 is resonant with C_1 at the mid-point of the intended frequency range. If the feed-back circuit is operated with a high effective Q_f, this value of L_1 should provide a zero phase shift in I_p. In this event, the tube would operate into a resistive load, and a theoretical independence of the frequency with changes in R_p could be predicted. As it is, the low feed-back Q_f

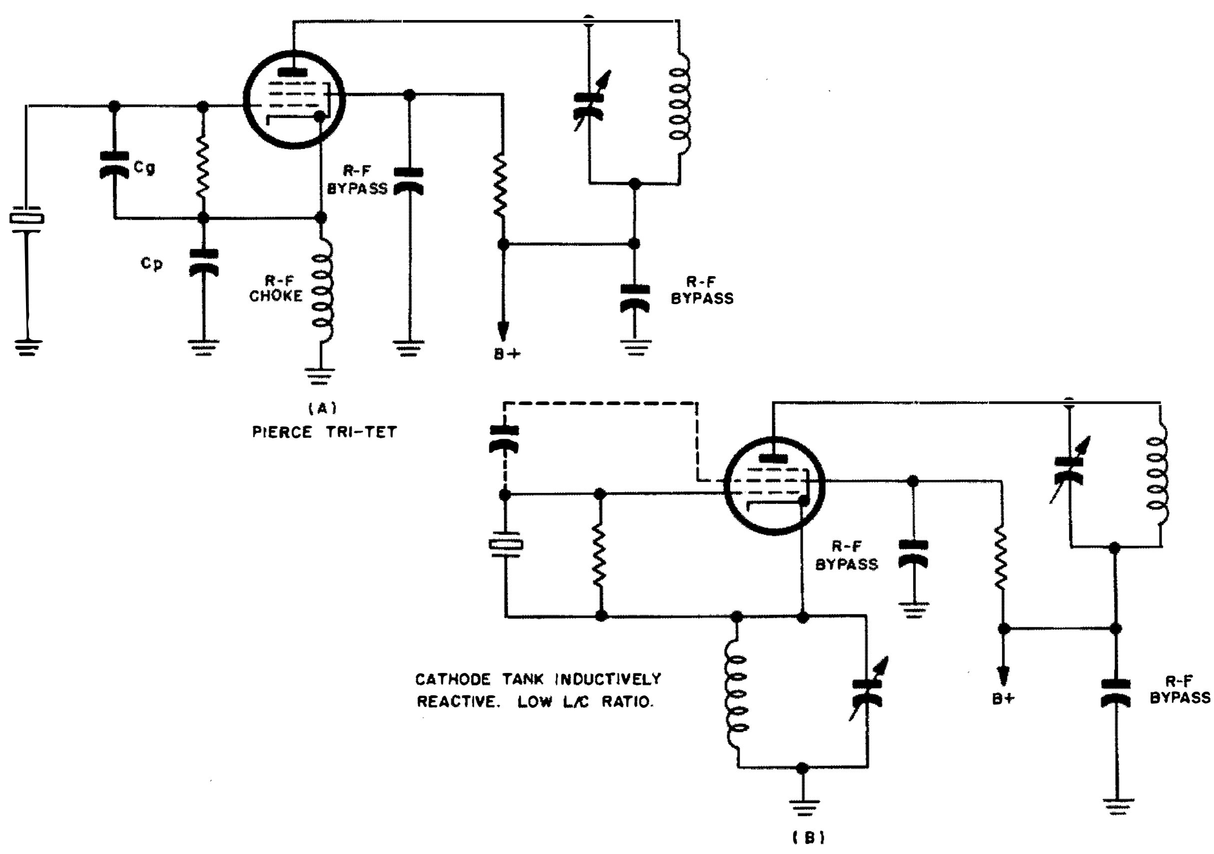


Figure 1-139. Basic tri-tet circuits where excitation voltage is a function of both screen- and plate-circuit r-f currents

probably requires the crystal tank circuit to be detuned to such a point that at equilibrim it appears either as a reactance much greater in magnitude than that of the coil L_1, or as a reactance approximately equal to that of C_1, in which case the lagging component of the current through the crystal unit is negligible compared with the current through C_1. Under these latter conditions, the tube would operate into a low-impedance, series-resonant circuit at the mid-point of the frequency range, where $X_{L1} + X_{C1} = 0$. Nevertheless, it was found that oscillations could not be maintained dependably at the mid-point of the frequency range. It was for this reason that the changes were made in the models represented by circuit (B), as indicated in the data chart for figure 1-135. The tank circuit L_3C_3 is resonant at the mid-point of the frequency range. C_1 has been changed so that the reactance of the coil L_1 is lower than the reactance of the crystal tank at all frequencies. R_4 has been added to dampen the effect of L_1 at the high end of the frequency range. It may be that the dead spot at the mid-point of the frequency range in circuit (A) was due only to transient effects in the crystal units before oscillations could build up, or it may have been due to the fact that the feed-back Q_f was insufficient to provide the necessary plate impedance to maintain equilibrium even if oscillations were once started.

1-324. An interesting circuit is that shown in figure 1-136 (D). The feed-back voltage is developed across C_4 by the r-f plate current. C_3, although of the same capacitance as C_4, maintains the screen at r-f ground, since R_2 is very large. The large value of R_2 keeps the screen voltage, and hence the out put, at very low values, so that the crystal is only weakly driven.

1-325. The circuit shown in figure 1-136 (E) is intended to supply a fourth-harmonic excitation of the V_2 stage. For this purpose the C_3L_2 tank is tuned to $4F_1$. A low L_2/C_3 ratio is provided, to ensure that the fundamental is effectively bypassed. The capacitance C_5 is kept small so as to present a high reactance to the fundamental, else the fundamental would be entirely bypassed around the crystal circuit.

1-326. The circuit shown in figure 1-136 (G) is something of a novelty in that a Pierce instead of a Miller oscillator is employed to directly excite the power amplifier of a small transmitter. The L_1C_4 arm is a neutralizing circuit which prevents the amplitude-modulated output stage from varying the effective impedance of the oscillator load. Normally, neutralizing networks are not necessary for crystal oscillators. Only when the oscillators drive power amplifiers directly is feed-back neutralization advisable. Even then, if the power amplifier is not modulated and performs as a frequency multiplier, neutralization is not necessary.

1-327. The electron-coupled converter circuits shown in figure 1-138 embody more or less the same features previously discussed. The basic methods illustrated for obtaining a heterodyne output are more or less self-explanatory, and will not be elaborated upon here.

The Miller Oscillator

1-328. The Miller oscillator is the crystal equivalent of a Hartley oscillator in which no mutual inductance exists between the plate-to-cathode and grid-to-cathode inductances. (See figure 1-140.) The Miller oscillator has an average frequency deviation of approximately 1.5 times that of the Pierce circuit. The plate circuit must appear inductive in order that the correct phase shift will be produced in E_p, the plate r-f voltage, to compensate for the

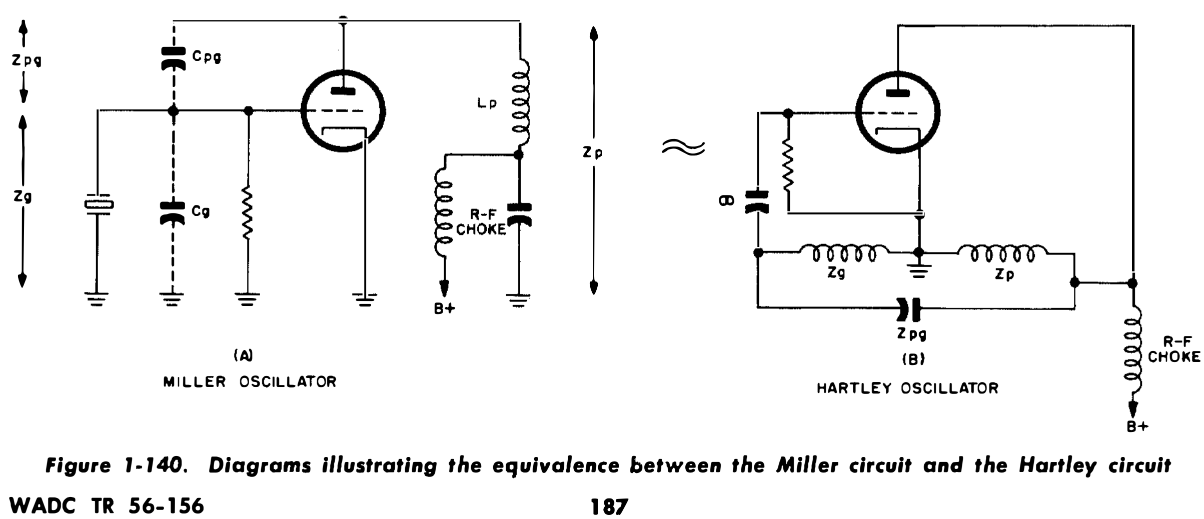


Figure 1-140. Diagrams illustrating the equivalence between the Miller circuit and the Hartley circuit

resistance in the feed-back arm, since this resistance prevents the necessary 180-degree phase rotation of the equivalent generator voltage of the amplifier from occurring entirely in the feed-back circuit. The effective load capacitance into which the crystal unit operates is, approximately,

$$C_x \approx C_g + \frac{1}{\omega}\left(\frac{1}{Z_{pg} - Z_p}\right)$$

$$= C_g + \frac{C_{pg}}{1 - \omega^2 L_p C_{pg}} \qquad 1\text{—}328\ (1)$$

where Z_{pg} and Z_p are both considered as unsigned magnitudes, and the various symbols correspond to those in figure 1-140. Since the load capacitance is a function of the frequency, a Miller oscillator cannot be operated at more than one frequency and still present the same load capacitance to each crystal unit except by providing for an adjustment of the circuit parameters. In spite of its greater frequency instability and lack of circuit simplicity as compared with the Pierce circuit, the Miller design is the one most widely used in crystal oscillators. The reason for this popularity is the greater output that can be obtained for the same crystal drive level. In either the Pierce or the Miller basic circuit, the output cannot exceed the voltage across Z_{pg}, the largest single impedance in the plate tank circuit. In the Pierce circuit, the maximum voltage is thus the maximum permissible across the crystal unit; in the Miller circuit the maximum voltage is $(k + 1)$ times the maximum permissible voltage across the crystal unit, where k is the gain of the stage, equal to E_p/E_g. This gain, theoretically (not practically), can approach the mu of the tube as a limit when the load impedance, Z_L, is large compared with R_p. Thus, the use of a Miller circuit permits a saving of one amplifier stage.

1-329. The feed-back capacitance of the Miller circuit is, normally, simply the plate-to-grid interelectrode capacitance of the tube. It cannot, of course, be less than this unless an inductive shunt is connected between the plate and grid. When a pentode is used, it is usually necessary to insert a small feed-back capacitance on the order of a few micromicrofarads. The waveform in the output is improved by the use of a tuned tank circuit having a low L/C ratio in place of L_p. The plate tank must be tuned to a frequency above the oscillator frequency, in order that the tank impedance will appear inductive. Such an arrangement also ensures a large effective L_p of high Q. A variable capacitance in the plate tank facilitates adjustments to obtain the correct load capacitance for the crystal unit.

MILLER-OSCILLATOR DESIGN CONSIDERATIONS

1-330. If it is decided to employ a Miller oscillator as a frequency generator, the choice should be dictated by the need of a greater output than can be obtained with a Pierce oscillator. An exception to this rule might be made if a tri-tet circuit is contemplated, in which case, the large capacitance that will directly shunt the Pierce-connected crystal may well prevent the stability from being as high as that of the Miller tri-tet circuit. The Miller circuit is the more critical to design insofar as maintaining the correct load capacitance is concerned, but the basic approach to the problem is the same as that which was followed in analyzing the equilibrium state of the Pierce circuit. Both oscillators are represented by the same basic circuit, shown in figure 1-119. We shall not repeat the steps involved in the derivation of the equilibrium equations in the particular case of the Miller oscillator. The basic equations given in the following paragraphs can be used as points of departure in the design of any Miller circuit. Also, by methods similar to those employed in the analysis of the Pierce circuit, the design limitations of a Miller oscillator in which the crystal unit is to be operated within specifications can be predetermined, approximately.

MILLER-OSCILLATOR EQUATION OF STATE

1-331. As in the case of the Pierce circuit, there are two equations that express the state of oscillation equilibrium in the Miller circuit. Originally derived by Koga, these two equations are the real and the imaginary parts of the general equation:

$$-\frac{\mu Z_p Z_g}{R_p Z_s + Z_p(Z_g + Z_{pg})} = 1 \qquad 1\text{—}331\ (1)$$

where

$$Z_s = Z_g + Z_p + Z_{pg}$$

and

$$Z_g = R_{cg} + jX_g; \quad Z_p = jX_p; \quad Z_{pg} = jX_{pg}.$$

R_{cg} is the effective resistance of the grid circuit, accounting for both the crystal and gridleak losses. The losses in the plate circuit are assumed to be negligible. On solving equation (1), the real part can be expressed as

$$-R_p = \frac{X_p\,[Z_g^2\,(\mu + 1) + X_g X_{pg}]}{R_{cg}\,(X_p + X_{pg})}$$

1—331 (2)

Equation (2) defines the conditions that exist when the feedback power input equals the power dissipated in the grid circuit. The imaginary part of the equation (1) defines the frequency, or, more exactly, the impedance relations that must exist if the feedback is to be of proper phase. This is given as

$$X_s = \frac{X_p X_{pg} - R_{cg} R_p}{R_p Q_g} \qquad 1\text{—}331\ (3)$$

Where Q_g is equal to X_g/R_{cg}, and $X_s = X_p + X_g + X_{pg}$. Equation (1) is the same as equation 1-289 (1) except that the terms are rearranged and μ is substituted for the product $R_p g_m$. Equations (2) and (3) correspond to equations 1-289 (2) and (3), respectively. If it is assumed that $Z_g \approx X_g$ and that $(X_p + X_{pg}) \approx -X_g$, equation (2) above can be simplified, thus:

$$R_p \approx \frac{X_p\,[X_g\,(\mu + 1) + X_{pg}]}{R_{cg}} \qquad 1\text{—}331\ (4)$$

Remember that X_p and X_g are positive, and that X_{pg} is negative. When equation (3) is rearranged as follows

$$X_g + X_p + R_{cg}/Q_g - \frac{X_p X_{pg}}{R_p Q_g} = -X_{pg} \qquad 1\text{—}331\ (5)$$

it can be seen that the effect of the tube R_p on the frequency is a function of the term $\left(\frac{X_p X_{pg}}{R_p Q_g}\right)$ only.

LOAD CAPACITANCE OF CRYSTAL UNIT IN MILLER OSCILLATOR

1-332. The load capacitance into which the crystal unit operates in a Miller circuit has been derived by Koga to be

$$C_x = C_g + C_{pg} + C_v \qquad 1\text{—}332\ (1)$$

where C_g and C_{pg} are as represented in figure 1-140, and

$$C_v = \frac{\mu\, C_{pg}}{1 + R_p^2/X_p^2} \qquad 1\text{—}332\ (2)$$

It appears that equation (2) gives a value to C_v that, for a Miller oscillator operated at the rated load capacitance of the crystal unit, is probably between three and four times too small for the average circuit. Equation (2) is derived from equation 1-331 (1), beginning by expressing the latter equation in the following form:

$$\frac{1}{Z_g} + \frac{1}{Z_{pg}} + \frac{\mu}{Z_{pg}\left(1 + R_p/Z_p + \dfrac{R_p}{Z_g + Z_{pg}}\right)} = 0 \qquad 1\text{—}332\ (3)$$

It is next assumed that the term in parentheses

$$1 + R_p/Z_p + \frac{R_p}{Z_g + Z_{pg}} \approx 1 + R_p/Z_p \qquad 1\text{—}332\ (4)$$

Such an assumption not only implies that the feed-back current is negligible compared with the r-f current through the plate coil, L_p, but that the feed-back impedance is so high relative to R_p that $R_p/(Z_g + Z_{pg})$ is negligible compared with 1. The former implication requires that Z_L, the effective load impedance across the tube, be approximately equal to jX_p; the latter implication requires that $|X_{pg}| - |X_g| >> R_p$. If the effective phase-determining Q_f of the feed-back circuit is 10 or more, as is very likely to be the case when standard crystal units are operating at their rated local load capacitance and C_g is not excessive, then E_p must be very nearly in phase with $-\mu E_g$. Such a condition cannot exist simultaneously with equation (4) unless $R_p << X_p$—an operating state that would be very undesirable from the point of view of frequency stability. If R_p is to have a reasonable value at the rated load capacitance of the crystal unit, the impedance of the feed-back arm cannot be greatly different from that of the plate circuit. Equation (4) would be sufficiently accurate for very low values of Q_g and very large transconductances for the tube; however, it would seem that for crystal units that are to be operated well above series resonance, the approximation of equation (4) should not be made. In this case

$$1 + R_p/Z_p + \frac{R_p}{Z_g + Z_{pg}} = 1 + R_p/Z_L \qquad 1\text{—}332\ (5$$

and Z_L approaches X_p^2/R_{cg} as R_p and Q_g increase. Since Z_L, as used above, represents an involved complex quantity, an exact expression of equation (2) will not be attempted here. As can be seen from equation 1-331 (5), if $1/Q_g$ and X_p/R_p are each on order of 1/10 or smaller, $X_g + X_p \approx |X_{pg}|$. On the other hand, if X_p/R_p is not small, the variable parameters of the vacuum tube and the variations to be expected in the effective resistance from one crystal unit to another will have such a large influence upon the effective load capacitance that there can never be an assurance that a crystal chosen at random will be operated according to specifications. In other words, a Miller oscillator cannot be designed to provide approximately a specified load capacitance unless $X_g + X_p \approx |X_{pg}|$. Under these conditions

$$C_x \approx C_g + \cfrac{1}{\omega\left(\cfrac{1}{\omega C_{pg}} - \omega L_p\right)} = C_g + C_{pg} + C_{Xp} \qquad 1\text{—}332\ (6)$$

where

$$C_{Xp} = C_{pg}\left(\frac{-X_p}{X_{pg} + X_p}\right) \qquad 1\text{—}332\ (7)$$

It must be understood that equations (6) and (7) assume that R_p is large compared with X_p, and that X_g is large compared with R_{cg}. For this latter condition to hold, the grid-to-cathode capacitance, C_g, must be kept as small as possible. If the assumptions above cannot be made, it is not feasible to expect a Miller oscillator to operate at approximately the same load capacitance for all crystal units, nor can good frequency stability be expected. A more comprehensive capacitance equation for the Miller circuit—one that holds approximately for all operating conditions — can be expressed as

$$C_x = C_g + C_{pg} + C_{Xp}' \qquad 1\text{—}332\ (8)$$

where C_{Xp}' is given by equation (7), except that X_p is replaced by X_p', where

$$X_p' = X_p + R_{cg}/Q_g - \frac{X_p\,X_{pg}}{R_p\,Q_g} \qquad 1\text{—}332\ (9)$$

It will be seen that X_p' has been so chosen that equation 1-331 (5) can be expressed in the form

$$X_g + X_p' + X_{pg} = 0 \qquad 1\text{—}332\ (10)$$

In the event that the plate circuit contains a capacitance shunting the coil, equations (6) and (7) still hold except that X_p refers to the total parallel reactance in the L_pC_p branches.

MAXIMUM R_p OF MILLER OSCILLATOR TUBE UNDER GIVEN LOAD CONDITIONS

1-333. Referring to figure 1-141, R_o and R_{ge} are defined as follows:

$$R_o = E_p^2/P_o \qquad 1\text{—}333\ (1)$$

where P_o is the power dissipated in the output circuit, and

$$R_{ge} = E_g^2/P_g \qquad 1\text{—}333\ (2)$$

where P_g is the power dissipated in the grid circuit. If the gridleak losses are negligible, P_g equals the crystal power and R_{ge} equals the PI of the crystal unit. As can be seen from the equations in figure 1-141, either PI must be small or R_g very large for this assumption to hold. With R_p large compared with Z_L, I_p is approximately equal to g_mE_g. We shall assume that $X_s = X_g + X_p + X_{pg} \approx 0$. Under these conditions it can be shown quite simply that in the circuit of figure 1-141, letting $k = E_p/E_g$,

$$g_m = \frac{k^2\,R_{ge} + R_o}{k\,R_{ge}\,R_o} \qquad 1\text{—}333\ (3)$$

For a given R_{ge} and R_o, equation (3) has a minimum g_m when

$$k^2 = R_o/R_{ge} \qquad 1\text{—}333\ (4)$$

Since a maximum R_p coincides with a minimum g_m, equation (4) also establishes the conditions for a maximum R_p. Now, R_{ge} is a function of R_e of the crystal unit, so that a circuit design using equation (4) should be based on a most probable value of R_{ge} (i.e., a most probable value of R_e), which will usually correspond to a value of R_e between one-third and one-fourth of the maximum R_e. Equation (4) should not be interpreted to mean that if R_o/R_{ge} is adjusted to equal a fixed value of k^2, the g_m of the tube will therefore be a minimum relative to its values for other R_o/R_{ge} ratios. Such an interpretation would only hold true if the product R_oR_{ge} were constant. Where equation (4) holds, it can be shown that the ratio of output power to crystal power is

$$P_o/P_g = 1 \qquad 1\text{—}333\ (5)$$

OPTIMUM VALUE OF $k = E_p/E_g$ FOR MILLER OSCILLATOR

1-334. Practical values of k, unless a pentode is used, are limited by the plate-to-grid and grid-to-cathode interelectrode capacitances of the tube and the specified load capacitance of the crystal unit. If $X_s = X_g + X_p + X_{pg} \approx 0$, then

$$X_p = kX_g \qquad 1\text{—}334\ (1)$$

and

$$-X_{pg} = (k + 1)\,X_g \qquad 1\text{—}334\ (2)$$

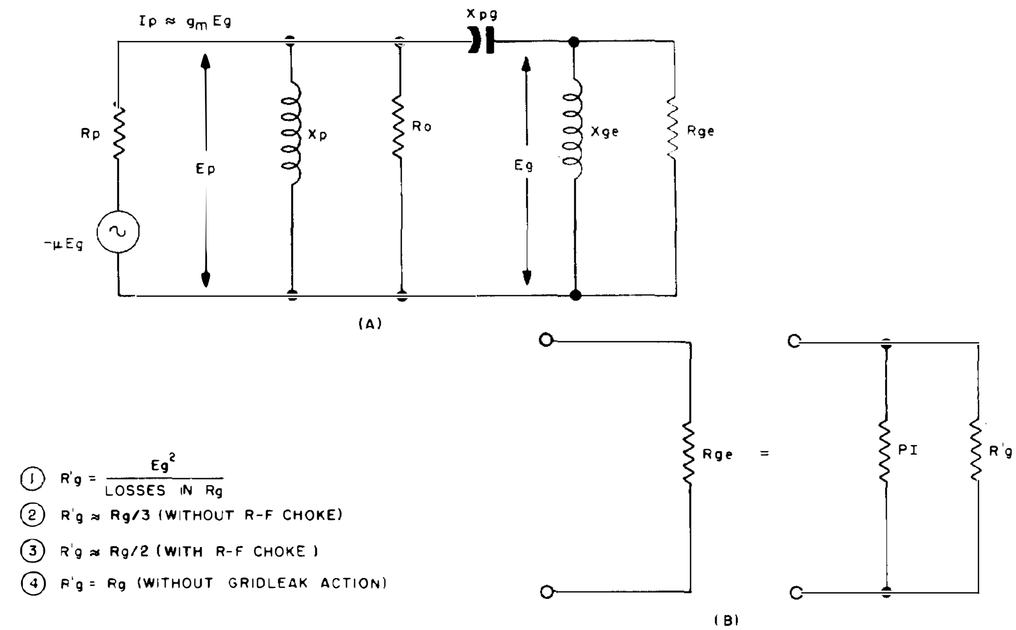


Figure 1-141. Equivalent circuit of Miller oscillator. R_p is assumed to be large compared with the total load impedance. R_o is an equivalent resistance accounting for the output losses. R_{ge} is an equivalent resistance accounting for the crystal and grid losses; it is approximately equal to the resistance of the parallel circuit shown in (B). PI is the performance index of the crystal unit; R_g' is the equivalent grid resistance; and R_g is the actual gridleak resistance

By equation (2)

$$k = \frac{-(X_{pg} + X_g)}{X_g} = \frac{C_x - C_g - C_{pg}}{C_{pg}} \qquad 1\text{—}334\ (3)$$

where

$$X_g = \frac{1}{\omega(C_x - C_g)}, \quad X_{pg} = -1/\omega C_{pg},$$

C_x = rated load capacitance, and C_g = grid-to-cathode capacitance. With triodes, values of k above 4 or 5 are difficult to obtain. If the oscillator is to be designed with no other feedback-circuit capacitances than those provided by the interelectrode capacitances, C_{pg} and C_g, of the tube, it can be assumed that k is a fixed parameter equal to the value given by equation (3). The output arm must thus be designed to provide a reactance, $X_p \approx kX_g$, if the crystal unit is to operate into its rated load capacitance.

1-335. From the point of view of frequency stability it is desired that the term (X_pX_{pg}/R_pQ_g) in equation 1—331 (5) be as small as possible relative to $(X_g + X_p)$, or, equivalently, to X_{pg}'. In other words,

$$X_{pg} \Big/ \frac{X_p\, X_{pg}}{R_p\, Q_g} = R_p/kR_{eg}$$

should be a maximum. With equation 1—331 (4), it can be shown that

$$R_p/k = Q_gX_g(\mu - 1) \qquad 1\text{—}335\ (1)$$

Equation (1) indicates that as the fraction of the loop reactance (kR_{eg}/R_p) dependent upon R_p becomes smaller, the effective amplification factor of the tube becomes greater. Intuitively from equation (1) it can be seen that with Q_gX_g constant for a given vacuum tube and plate voltage, E_b, R_p must increase as k is made larger, otherwise μ could not decrease. Nevertheless, the larger that k becomes the smaller the value of R_p/k. If k = 1, a frequency stability almost approaching that of the Pierce circuit can be achieved, but with twice the output voltage. Lower values of k would soon detune the oscillating tank to a point where the simplifying assumptions made regarding k would no longer hold.

1-336. Since the principal purpose of using a Miller instead of a Pierce circuit is to eliminate an amplifier stage, and since E_g is limited to the maximum voltage that can be placed across the crystal unit, the ratio $E_p/E_g = k$ can be chosen to give a desired gain over that which would be obtained with a Pierce oscillator operating at the same crystal drive level. E_g in the Miller circuit can be assumed to be twice the E_g of a Pierce circuit that has a $k = 1$. If an imaginary gain of 10 is desired, k for the Miller circuit should be equal to 5. For k to be 5, according to equation 1—334 (3)

$$C_{pg} = (C_x - C_g)/6$$

If $C_x = 32\ \mu\mu f$ and $C_g = 8\ \mu\mu f$, C_{pg} must be $4\ \mu\mu f$, which is a value quite representative of the average triode amplifier, or which could be obtained with a pentode by using a small external plate-to-grid capacitance. Equation 1—335 (1) can be rewritten in the form

$$1/R_p = g_m - R_{cg}/kX_g^2 \qquad 1\text{—}336\ (1)$$

For $k = 5$, the value of $1/R_p$, and hence the percentage effect of R_p on the loop reactance, will be a minimum the more nearly that g_m can be made to approach in value $R_{cg}/kX_g^2 = R_{cg}/5X_g^2$. If the effective Q of the crystal unit, $Q_e = \frac{X_e}{R_e}$, is equal to 10 or more and if the gridleak losses are negligible, it can be shown that

$$R_{cg} \approx \frac{R_e\, X_{Cg}^2}{(X_e + X_{Cg})^2} \qquad 1\text{—}336\ (2)$$

and that

$$X_g \approx \frac{X_e\, X_{Cg}}{X_e + X_{Cg}} \qquad 1\text{—}336\ (3)$$

Thus,

$$Q_g = X_g/R_{cg} = \frac{X_e(X_e + X_{Cg})}{R_e\, X_{Cg}} \qquad 1\text{—}336\ (4)$$

and

$$X_g^2/R_{cg} = X_e^2/R_e = PI \qquad 1\text{—}336\ (5)$$

where PI is the performance index of the crystal unit. If it is further assumed that the output losses are negligible, the impedance of the crystal tank is

$$Z_L = \frac{X_p^2}{R_{cg}} = \frac{k^2\, X_g^2}{R_{cg}} = k^2\, PI \qquad 1\text{—}336\ (6)$$

Equation (1) can thus be written

$$\frac{1}{R_p} = g_m - \frac{1}{k\, PI} = g_m - \frac{k}{Z_L} \qquad 1\text{—}336\ (7)$$

If k is fixed by output considerations, the percentage effect of R_p upon the loop reactance becomes a function of R_p alone, being a minimum when R_p is a maximum. If R_p is increased without limit, g_m approaches k/Z_L as a limit, and the greater the PI, the larger will R_p become. In the case of the Pierce oscillator, it will be recalled that a maximum R_p was obtained by a proper choice of k. This optimum k was the one that provided the maximum excitation voltage. In the Miller circuit, the excitation is the voltage developed across the crystal unit, and thus is limited by the crystal specifications regardless of the value of k. The smaller that k is made, the smaller will be the effective R_p, but, even so, the percentage effect of R_p upon the effective loop reactance will also be smaller. With k fixed by the requirement to eliminate an amplifier stage, the problem of obtaining a maximum R_p becomes one of keeping the load requirements to a minimum, selecting the vacuum tube, determining the proper operating voltages consistent with the crystal specifications, designing a test model accordingly, and experimenting for optimum results over the resistance range to be expected in the crystal units.

OPERATING CONDITIONS OF MILLER OSCILLATOR PROVIDING MAXIMUM R_p FOR GIVEN g_m.

1-337. If the bias of a tube is supplied by agc, the excitation voltage is small by comparison, so that the operating point of the tube can be theoretically estimated by consulting the R_p and g_m curves plotted against grid voltage. The operating bias for a given plate voltage would approximately be that giving values of R_p and g_m that obey equation 1—336 (7). Unfortunately, there are no curves available that indicate the effective R_p and g_m for large excitation voltages where the tube is cut off a large fraction of each cycle. Nor has a theoretical basis been established for estimating the probable rates of change in R_p and g_m as the excitation is increased under various circuit conditions. If time permits, experiments designed to furnish such data may gain for the engineer a valuable insight into the characteristics of his design models. Most probably the "dynamic" curves of R_p and g_m will correspond closely to the static curves. Yet the possibility exists that significant differences in the rates of change in the tube parameters may be discovered under certain operating conditions. In equation 1—336 (7) it can be seen that for any large value of R_p, g_m very nearly equals k/Z_L. For example, if $R_p = 0.5$ megohm, the difference between g_m and k/Z_L is only 2 micromhos. An R_p of 1 megohm corresponds to practically the same value of g_m, the difference being only on the order of 1 micromho. Thus, when R_p is large, g_m can be considered more or less a circuit constant. During the time that the tube is cut off, R_p is infinite and g_m

is zero. From the point of view of a large effective R_p it is desirable that the cutoff angle be a maximum. The larger the g_m of the tube above cutoff, the greater can be the cutoff angle. A sharp-cutoff tube would be preferred for this purpose. It is also desirable to have R_p as high as possible above cutoff. For this purpose, a high-mu tube is to be preferred. A theoretical estimate of the optimum relation between the values of R_p and g_m, for a tube of the same class-A mu, that provides a maximum over-all effective R_p cannot be attempted here. However, it would seem that the emphasis should be placed upon the larger g_m/R_p ratio. The effective R_p of a pentode can always be increased artificially by inserting a high resistance in the plate circuit in series with the oscillating tank, as illustrated in figure 1-142. In testing a given tube for those bias and excitation conditions which provide a maximum R_p, it may be preferable to control the bias independently of the oscillations, or by using an adjustable, r-f-bypassed cathode resistor. A crystal unit should be employed having parameters known to remain constant over the experimental drive-level range. After the tube has warmed up, if the cathode bias is used, the cathode resistance can be decreased until oscillations begin. The cathode resistance can then be increased until the frequency is a maximum. The maximum frequency would be an indication of an equilibrium point of maximum R_p. In order for oscillations to be maintained in the event that all the bias is developed across the cathode resistance and the excitation is insufficient to drive the grid positive, a small percentage decrease in the excitation amplitude must cause at least an equal percentage decrease in the average plate current, and hence in the bias. Such operation will require that the tube be cut off for a large fraction of each cycle. An adjustable cathode resistance cannot be considered a particularly practical design feature, but it may prove advantageous in an experimental circuit for finding the operating conditions that provide a maximum R_p for a given g_m.

FREQUENCY-STABILITY EQUATIONS FOR MILLER CIRCUIT

1-338. Regardless of whether the circuit conditions are such that the effective load capacitance of the crystal unit is assumed to be given by equation 1—332 (1), by equation 1—332 (6), or by equation 1—332 (8), the fractional change in frequency for a small change in any one of the equivalent component capacitances is given by the general equation

$$\frac{df}{f} = -\frac{1}{F_{Xe}} \cdot \frac{dC_x}{C_x} \qquad 1\text{—}338\ (1)$$

where dC_x is equal to dC_g, dC_{pg}, dC_v, dC_{Xp}, or dC_{Xp}', and represents an incremental change in any of the component capacitances, and F_{Xe} is the frequency-stability coefficient of the crystal unit, equal to $2C_T^2/CC_x$. (See equation 1—243(1).) If a tuning capacitor, C_p, is connected across L_p in the plate circuit, and if equations 1—332 (1) and (2) are assumed approximately correct, it can be shown

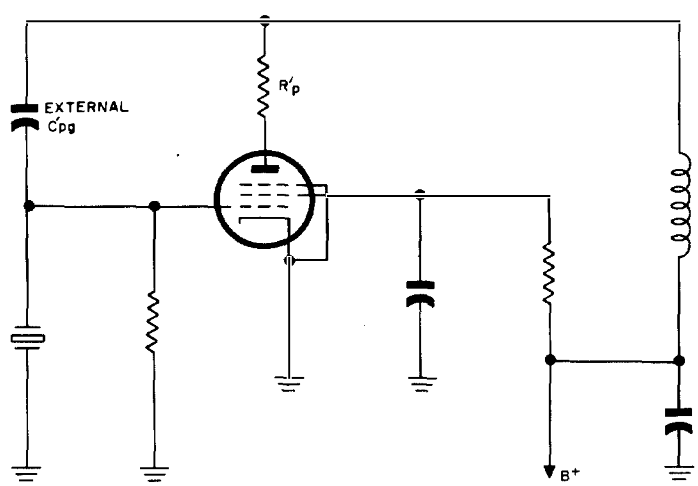


Figure 1-142. The large resistance, R_p', connected in plate circuit effectively increases R_p of the tube. This method can be used to improve the frequency stability of a Miller oscillator employing a screen-grid tube and an externally connected feed-back capacitance, C_{pg}'

that for variations in C_p,

$$\frac{df}{f} = -\frac{C_{pg}}{C_x} \cdot \frac{2\,\mu\,\omega\,C_p\,R_p^2}{F_{Xe}\,X_p\,(1 + R_p^2/X_p^2)} \cdot \frac{dC_p}{C_p}$$

1—338 (2)

where ω = angular frequency. If the circuit is operating at maximum activity, in which case $Z_L \approx X_p = R_p$, equation (2) becomes

$$\frac{df}{f} = -\frac{C_{pg}}{C_x} \cdot \frac{\mu\,R_p\,\omega C_p}{2\,F_{Xe}} \cdot \frac{dC_p}{C_p}$$

1—338 (3)

Where equations 1—332 (6) and (7) can be assumed to be approximately correct, it can be shown that a fractional change in the plate reactance, X_p, causes a fractional frequency deviation of

$$\frac{df}{f} = -\frac{(C_x - C_{pg})\,dX_p}{C_{pg}\,F_{Xe}\,X_p} \qquad 1\text{—}338\ (4)$$

If no tuning capacitor is provided to shunt the plate coil L_p, equation (4) can be expressed in terms of a fractional change in L_p, thus:

$$\frac{df}{f} = -\frac{(C_x - C_{pg})\,dL_p}{C_{pg}\,F_{Xe}\,L_p} \qquad 1\text{—}338\ (5)$$

If desired, equations indicating the frequency stability when other parameters are varied can be derived by following a procedure similar to that employed in the analysis of the Pierce circuit.

MILLER CIRCUIT AS A SMALL POWER OSCILLATOR

1-339. The ratio of the output power to the input power is given by the equation

$$P_o/P_g = k^2\,R_{ge}/R_o$$

where $k = E_p/E_g$, and R_{ge} and R_o are the resistances represented in figure 1-141. In practice, ratios of R_{ge} to R_o can be obtained on the order of 4 for crystal units of maximum effective resistance. If k = 5, this would mean a power ratio of 100. A 10-mw crystal unit could thus be used to develop a 1-watt output. Much higher power outputs, of course, can be obtained with crystal units of small values of R_e or of higher power ratings. It cannot be recomended that a crystal be driven beyond its rated power level, but if an exception should ever arise, the Miller circuit will require the least overdrive. If a larger drive level is necessary than can be obtained with Military Standard crystal units, the cognizant military agency should first be consulted. It may be that one or more of the crystal manufacturers has available a nonstandard crystal unit with crystal dimensions and mounting sufficient to withstand the required drive—perhaps by operating with an overtone mode — without the risk of significant parameter variations. As a final resort, it will be found that most of the Military Standard crystal units can withstand, without shattering, drive levels from 10 to more than 20 times the rated drive. If need be, power outputs greater than 35 watts can be obtained with the Miller circuit, using a beam power tube or a power pentode. It is much easier to obtain a large output from a high-mu than from a low-mu tube for the same crystal drive. Also, it is easier to obtain a large output from, say, a 50-watt tube operated at low efficiency, than from a smaller tube operated at high efficiency. An r-f choke must be used in the grid circuit if large output is to be developed. Furthermore, fixed bias that is sufficient to prevent the grid from drawing current must be used, so as to reduce the grid losses to a minimum. The voltage gain of the oscillator, $k = E_p/E_g$, should be as high as possible. With proper design, except that the crystal unit is operating at tolerances greater than those specified for low drive levels, the Miller circuit can be made to drive a power amplifier of 300 watts or more. Some crystal units can withstand as much as 120 ma r-f current and still be within the safe-operating range as far as shattering is concerned. A pressure-mounted unit is generally to be preferred at high drive levels, because of the added protection it offers, and because its greater thermal conductivity permits the generated heat to escape more rapidly. For maximum output, the oscillator must operate into an impedance matching the R_p of the tube. If a Miller oscillator is to drive a power amplifier, great care must be taken in neutralizing the feedback from the amplifier, or the crystal may easily be overdriven to the point of shattering—that is, unless the power amplifier is to serve as a multiplier stage, or if a screen-grid tube is used as the amplifier tube. The plate supply voltage for the oscillator can be as high as 350 to 650 volts, and that of the power amplifier, 1500 to 2000 volts. If a low-power (7.5 watts, approximately) oscillator tube is used, a fixed bias of 40 to 60 volts will be required for high efficiency. A 3.5 to 4.5-ampere current in the plate L_pC_p tank can be obtained under these conditions. A fixed bias is usually not necessary when a 50-watt tube operated at low efficiency is used. With the same plate voltage as for the low-power tube, a plate tank current of 4.5 to 7.5 amperes can be had. When a fixed bias is used, some arrangement must be provided to cut it in after oscillations build up and the negative peaks of E_p must be sufficient for plate limiting to occur at the positive peaks of E_g. When used as a power oscillator, the Miller circuit is often required to operate as a variable-tuned circuit, with a coil or tank-circuit in place of the crystal, the circuit thereby being converted into a tuned-plate-tuned-grid or a Hartley type oscillator. Because of this, the various tuning adjustments and meters that are needed in the variable circuit are also available in the crystal circuit. In this event, the rated load capacitance of the crystal unit will usually exist more in theory than in application. Since the crystal is intended to be operated at high drive, the risk is greatly increased that a chance adjustment may overload the crystal to the shattering point. This risk can be minimized by the use of an r-f milliammeter in series with the crystal, with the danger zone well marked. Besides excessive plate voltage and stray feedback due to poor shielding or neutralization, a poorly bypassed screen-grid circuit can lead to an overloaded crystal, as also can an excessive control-grid bias. Now, no attempt should be made to design a circuit in which a crystal unit is to be operated above its rated power level unless weight, space, or expense requirements demand the elimination of every possible amplifier stage; unless greater frequency stability is required than can be obtained with a conventional inductor-capacitor network; or unless a long operating lifetime is not a primary consideration. Even so, if a Military Standard unit is operated beyond specifications, it should be well understood that it is no longer effectively a standard type, and no guarantee exists concerning the replacement of one crystal unit by another.

TYPICAL CHARACTERISTICS OF MILLER OSCILLATOR

1-340. Figure 1-143 shows an experimental Miller circuit, the performance characteristics of which were investigated by Messrs. E. A. Roberts, Paul Goldsmith, E. K. Novak, and J. Kurinsky of the Armour Research Foundation at the Illinois Institute of Technology. The crystal units used are of the type CR-18/U. The crystal PI indicated for each of the characteristic curves (figures 1-144 to 1-148) of the oscillator in figure 1-143 is the value observed when the crystal was operating into its rated load capacitance. The PI at the rated load capacitance indicates the relative activity of the crystal unit, but is not intended to imply that the same PI is in effect for all variations of the load capacitance. The curves in figure 1-144 indicate (excitation voltage)^2 and the output voltage as the plate tuning capacitance is varied. An increase in the plate capacitance means an increase in the effective value of L_p, so that the frequency decreases. Thus, as C_p increases (C_2 in figure 1-144), the reactance, X_e, of the crystal unit decreases. Oscillations cease whenever the Q_g of the grid circuit becomes too small for the proper phase rotation to take place, or the ratio of X_p/X_g becomes too high for the feed-back voltage to be of sufficient amplitude, or the plate tank approaches the parallel-resonant state, so that E_p can no longer assume its proper phase, which requires the plate arm to be an inductive reactance smaller in magnitude than the capacitive-feedback reactance. The percentage points in figure 1-144 refer to percentages of the maximum output voltage that was obtained through variations of the plate tank capacitance alone. The six curves shown represent values for

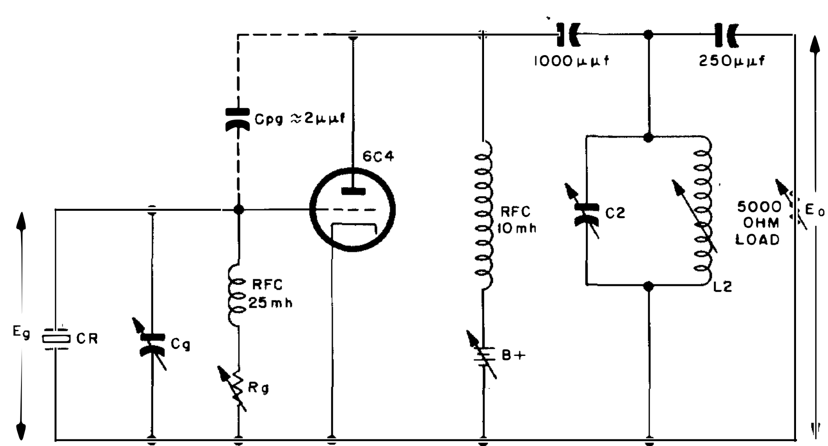
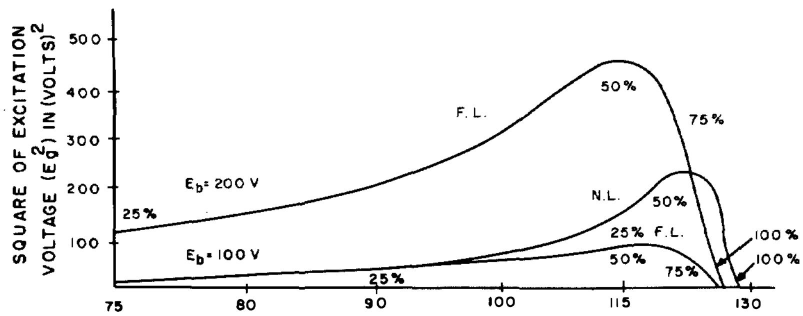


Figure 1-143. Experimental Miller oscillator whose characteristic curves are plotted in figures 1-144 to 1-148

an E_b of 200 volts with a full load of 5000 ohms, and for an E_b of 100 volts with full load and with no load. It is interesting to note that in each of the three pairs of curves the maximum grid voltage occurs at a smaller load capacitance than that at which the output voltage is a maximum. For each curve where the plate tuning capacitance is the same, it can be approximately assumed that the load capacitance is the same. Also, between the values of $C_p = 75$ and $C_p = 90$ $\mu\mu$f, it can be assumed that the percentage change in C_x is small. Since the PI of the crystal unit is the same where the load capacitance is the same, the excitation-voltage-squared curves indicate the relative crystal drive for the different E_b and load conditions. Also note that the two pairs of curves representing full-load conditions coincide fairly closely at their points of equal percentages. This is important in interpreting the curves in figure 1-145, each of which represents 50 per cent output, and hence approximately the same load capacitance and frequency. Exceptions are the 10K curves in figure 1-145, as can be checked by figure 1-146. The performances curves in figure 1-145 are the Miller equivalents of the Pierce curves in figure 1-130. It can be seen that the Miller output is much



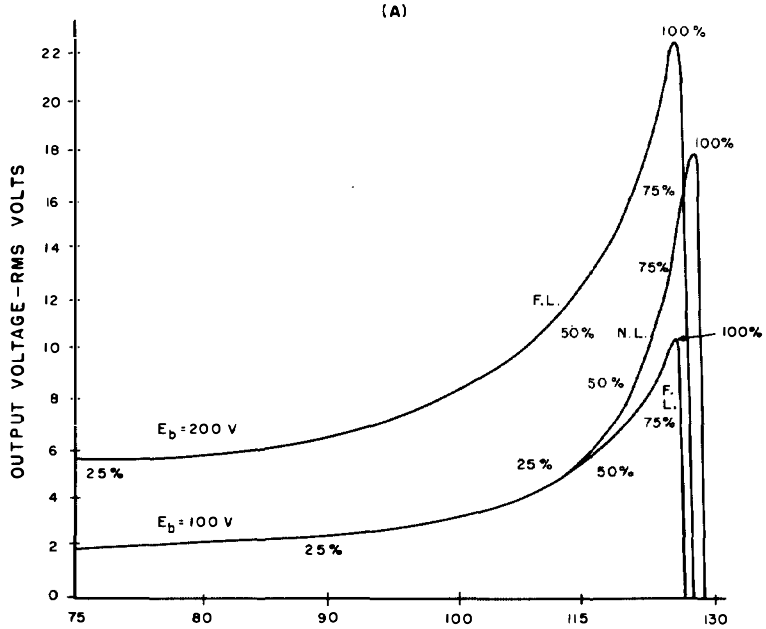


***Figure 1-144.** (A) Square of excitation voltage and (B) rms value of output voltage versus plate tuning capacitance of experimental Miller oscillator. Frequency = 7 mc; gridleak resistance = 1 megohm; and PI of CR-18/U crystal unit (with load capacitance of 32 $\mu\mu$f) = 49 kilohms. F.L. = full load conditions (5,000 ohms across plate tank) and N.L. = no load conditions. Percentage points refer to percentages of maximum output voltage obtainable under given load and d-c plate voltage conditions*

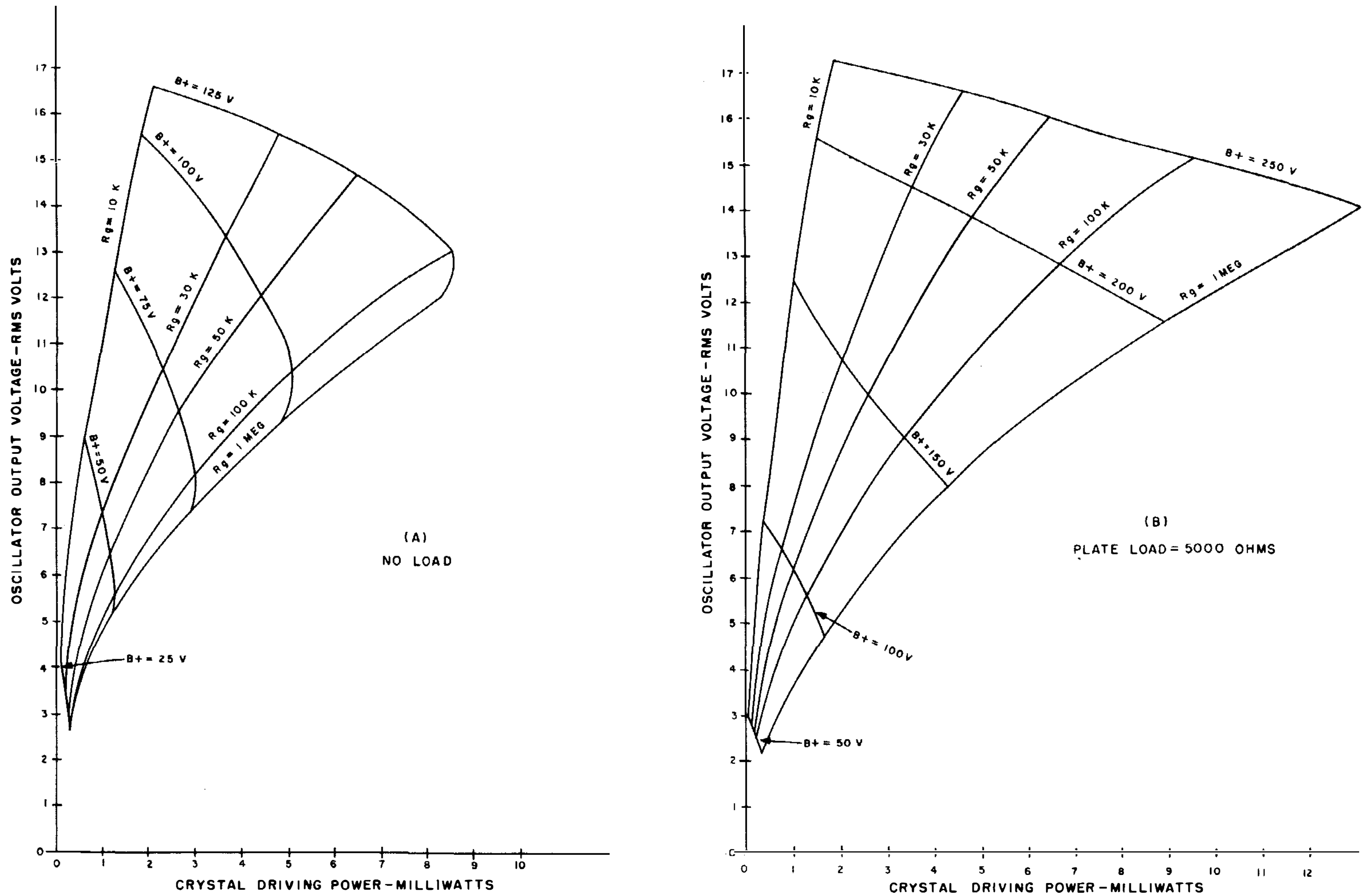


***Figure 1-145.** Output voltage versus crystal power for various values of gridleak resistance as the d-c plate voltage is varied in experimental Miller oscillator. The circuit is tuned for output voltages 50 percent of the maximum possible by variations in plate tank alone. Frequency = 7 mc; plate tank L/C = 0.04; and PI of CR-18/U crystal unit (with load capacitance of 32 $\mu\mu$f) = 49 kilohms*

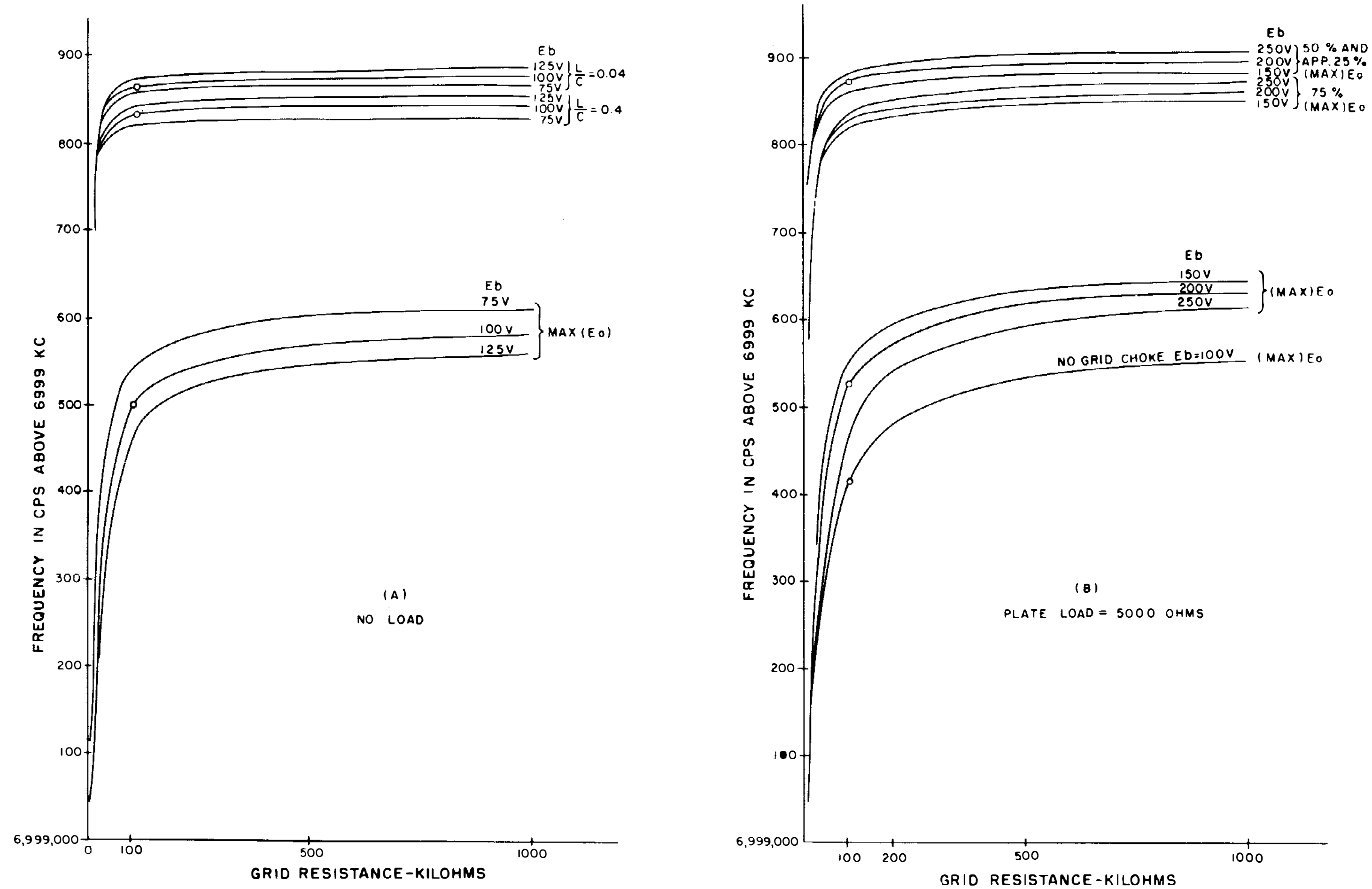


Figure 1-146. Frequency of experimental Miller oscillator versus d-c plate voltage and grid resistance for various plate-tuned load capacitances. (Max) E_o represents the plate tuning adjustment that provided maximum output voltage when grid resistance and d-c plate voltage were values indicated by zero reference point. A 7-mc CR-18/U crystal unit was used, having a PI of 49,000 ohms when operating into a rated load capacitance of 32 $\mu\mu f$

more sensitive to changes in the grid resistance. This is to be expected, since the grid-to-cathode r-f impedance and excitation voltage is much greater in the Miller circuit. A crystal r-f voltage of 2 volts represents an excitation of 2 volts in the Miller circuit, but usually of only 1 volt or less in the Pierce circuit. If the curves in figures 1-130 and 1-145 were plotted against excitation voltage instead of crystal driving power they would be much more similar in appearance.

1-341. The frequency curves in figure 1-146 are the Miller equivalents of the Pierce curves in figure 1-129. Note that as R_g is decreased the frequency falls, whereas in the Pierce circuit the frequency increases. This is one reason why the Miller circuit becomes so much more frequency sensitive to changes in R_g when R_g is small. As the Q_g of the grid circuit is decreased because of a decrease in R_g, the effective Z_L across the tube must appear more inductive in order for E_p to shift in the correct direction to compensate for the decreased phase rotation between grid and cathode. For this to occur, the net capacitive reactance of the feedback arm must increase, which can only come about if the inductive reactance of the crystal unit decreases. Hence, the frequency falls, and in so doing, the Q_g of the crystal and grid circuit becomes smaller still, so that an additional drop in the frequency is necessary to compensate for the decrease in the crystal Q_e. In the meantime, the bias decreases and the grid goes positive a larger fraction of the time. This tends to decrease R_p, which contributes even more to the drop in frequency. With all these effects adding in the same direction, the large frequency sensitivity of the Miller with changes in the grid resistance is ex-

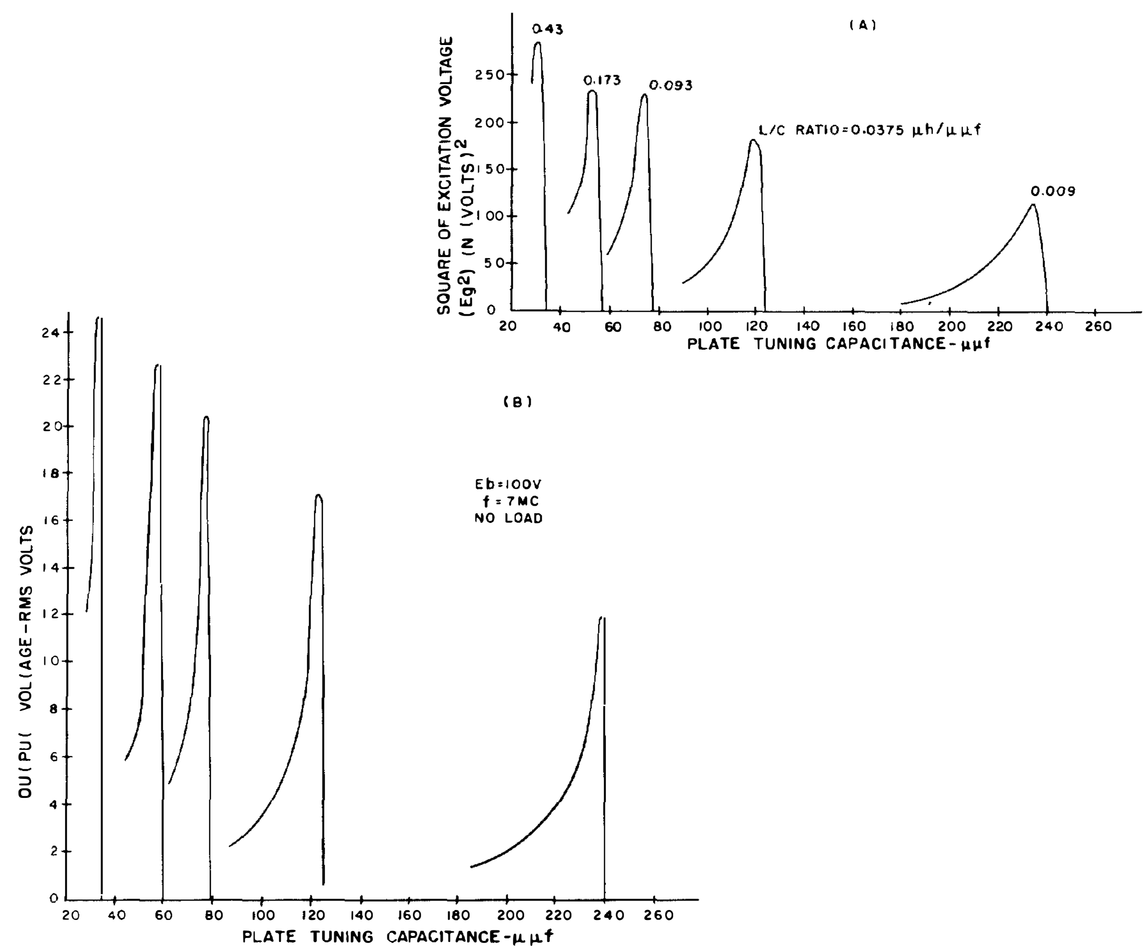


Figure 1-147. (A) Square of excitation voltage and (B) rms value of output voltage of experimental Miller oscillator versus plate tuning capacitance for various L/C ratios of plate tank. Same crystal unit as was used for curves in figure 1-144

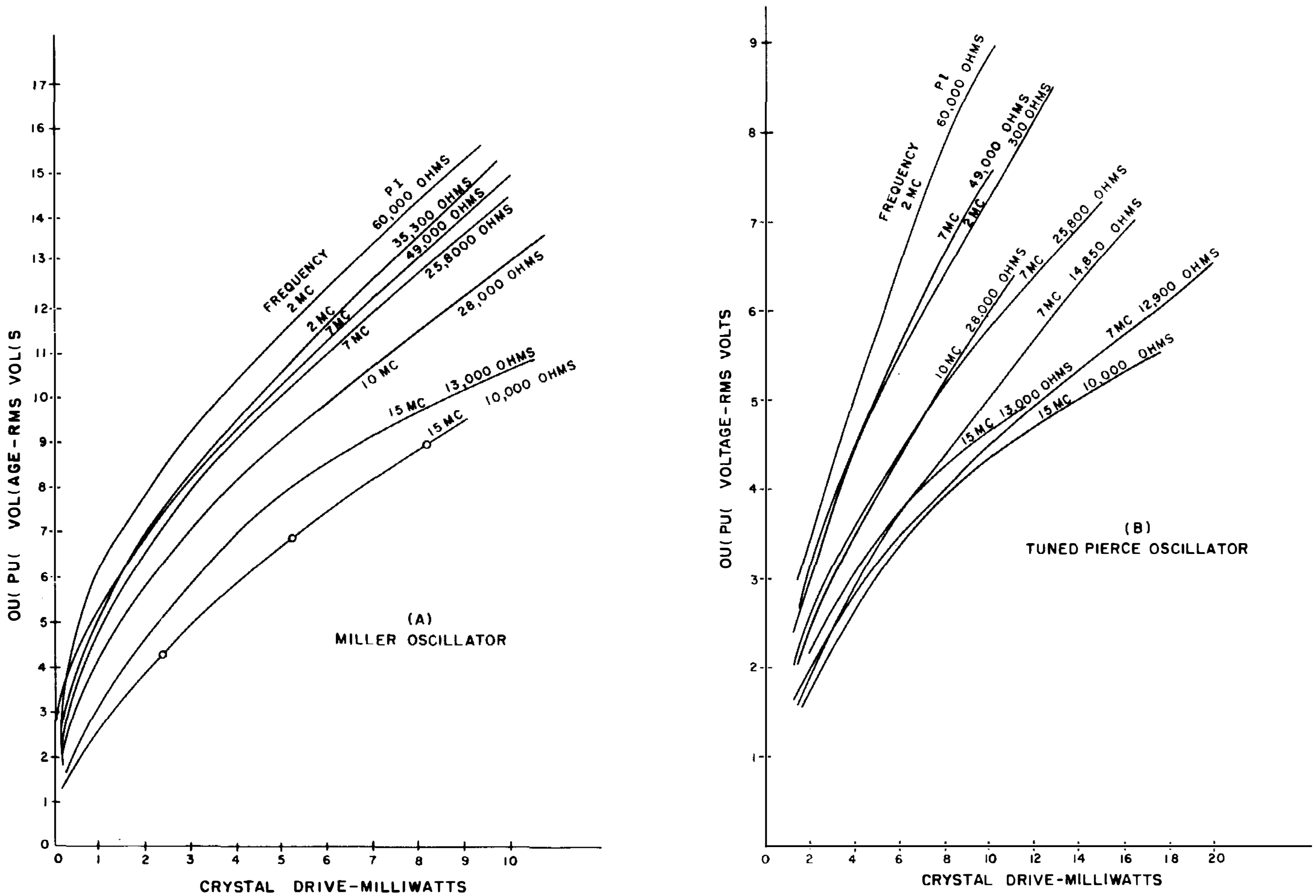


Figure 1-148. Output and crystal drive of experimental Miller and Pierce oscillators as d-c plate voltage is varied for crystal units of various frequencies and PI's. The values of PI assume a rated load capacitance of 32 $\mu\mu f$. Both oscillators were tuned to provide 50 percent (max) E_0 and to operate into a 5000-ohm plate load

plained. The sensitivity is a maximum when L_p is a maximum, for then the frequency is a minimum and the crystal is operating nearest its series-resonant state. This fact makes an exception to the rule that the larger the effective C_x, the greater the stability.

1-342. Of special interest in the curves of figure 1-146 is the fact that those representing the 50-per-cent-maximum-output adjustment show an increase in frequency with an increase in plate voltage, whereas the curves representing a maximum output voltage show a decrease in frequency when the plate voltage is increased, even though the same plate voltages are applied in each case. Now, in the Miller oscillator, the frequency increases and decreases in the same direction with R_p. The plate characteristics of the 6C4 tube, the tube being used when the curves in figure 1-146 were plotted, indicate a decrease in R_p as the plate voltage increases. Thus, we should expect the change in frequency of the (max) E_o curves to be due to the change in R_p caused by the change in plate voltage. On the other hand, the oppositely directed change in frequency of the lower-percentage-E_o curves must be due to an oppositely directed change in R_p brought about by a change in the bias. A re-examination of the crystal voltage curves does indeed show at the 50-per-cent-E_o adjustment that the grid excitation, and hence the bias, is near the maximum. In figure 1-145, it can be seen that for large values of grid resistance the changes in output voltage due to changes in the plate voltage cause a maximum variation in the crystal drive. The evidence is quite strong that there is an operating region between the oppositely changing frequency curves where the changes in R_p due to changes in E_b and E_c will annul each other. From an inspection of the crystal voltage curves in figure 1-144 we would guess that such operating points will lie on both sides of the maximum-E_g region. Such a state of operation would be an example of "class-D" operation described in paragraph 1-298.

1-343. The curves shown in figure 1-147 indicate the effect of variations in the L/C ratio of the plate tank circuit obtained by increasing the value of L_2 in figure 1-143. Increasing L_2 increases the impedance into which the tube operates, and thus increases the r-f plate voltage. This also has the effect of decreasing the frequency and the reactance of the crystal unit. For this reason, the crystal voltage does not increase in the same proportion as the plate voltage. Much greater stability is obtained with low L/C ratios, but much greater values of E_p/E_g, and hence of power gain, are to be obtained with large L/C ratios. Figure 1-148 compares output-vs-drive curves for several different frequencies and values of PI when the same crystal units are used in both Miller and Pierce oscillators. Note that in the Pierce circuit the slopes of the curves consistently increase with an increase in PI. In the Miller circuit the tendency is for the slopes to increase with decreasing frequency primarily, and secondarily with the PI. This may be due to the fact that the gridleak resistance used in the Miller circuit was smaller than that in the Pierce circuit. In any event, the average bias will tend to be less at the lower frequencies, since the grid charge has more time during a cycle to leak off.

MODIFICATIONS IN DESIGN OF MILLER OSCILLATOR

1-344. A number of Miller oscillators currently being used in military equipment are illustrated in figures 1-149, 1-150, 1-151, 1-152, and 1-153. The values of the circuit parameters, where available, are given in the accompanying circuit-data charts. None of the crystal units employed in these circuits is now recommended for equipments of new design. Nevertheless, all the circuits shown can be modified in one way or another and used with currently recommended crystal units which have been tested for parallel resonance. The necessary modifications would be those that would ensure a correct load capacitance and would not permit a crystal to be overdriven within the expected range of effective resistance. The circuits illustrated suggest the wide adaptability of the Miller oscillator for different output requirements and uses. It is not possible to single out a particular circuit and declare this design to be preferred. The engineer will need to design and test his own circuit for the particular requirements of the equipment in which his oscillator is to be used. Quite often the type of vacuum tube or other circuit components most readily available influence the design. Unlike the Pierce, the Miller circuit must include a means of adjusting the plate impedance to ensure the correct load capacitance for the crystal if the oscillator is to operate to more than one frequency. In the circuits of figures 1-149 to 1-153, the switching arrangements of those circuits designed to operate over a wide frequency range are for the most part omitted. Most often, a separate plate coil is provided for each crystal position. Because of space limitations in the circuit-data charts, occasionally two different components in a circuit having the same value or being of the same type are assigned the same symbol number.

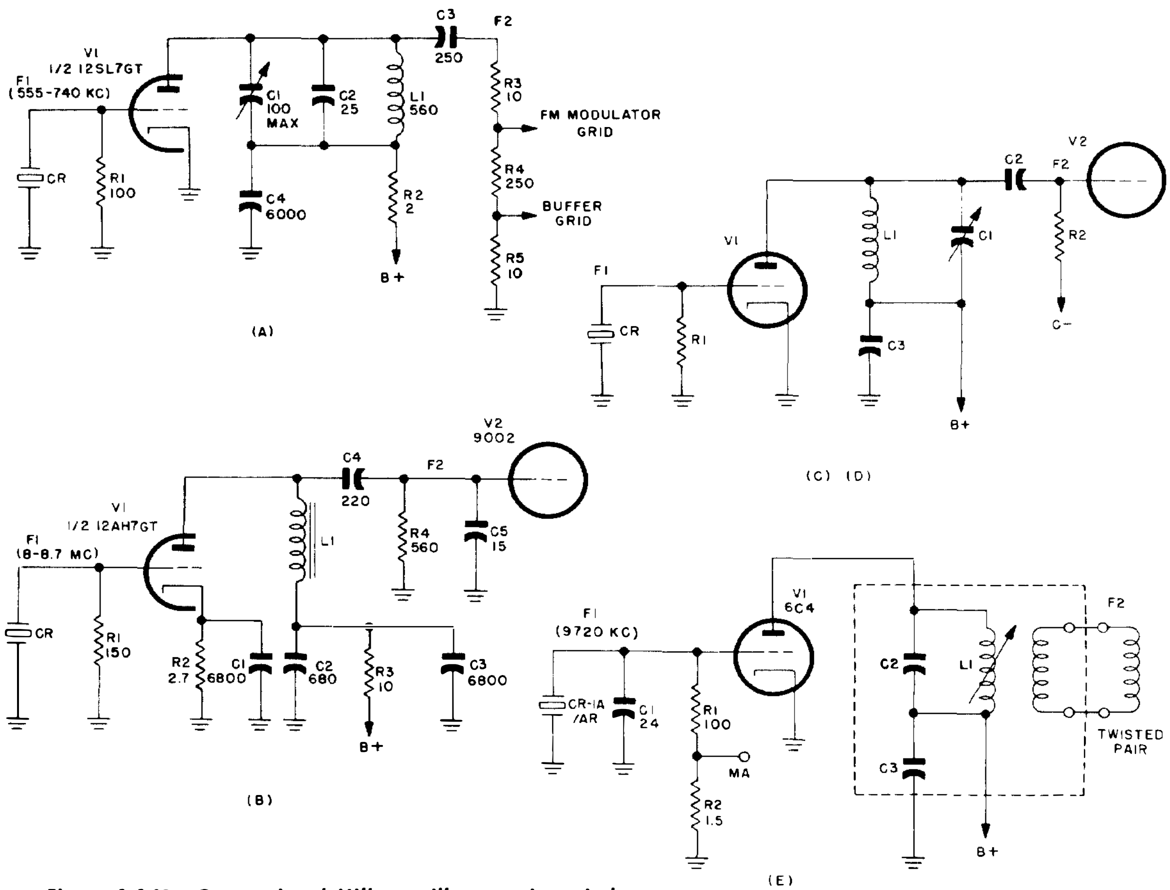


Figure 1-149. Conventional Miller oscillators using triodes

Fig.	Equipment	Purpose	F_1	F_2	CR	R_1	R_2
(A)	Radio Modulator and Transmitter BC-925	M.O.	555-740	F_1	Bliley AR-3	100	2
(B)	Radio Receiver BC-624-A,-AM,-C	Local oscillator	8000-8720	F_1	DC-11-( ), DC-16, DC-26, or CR-1( )/AR	150	2.7
(C)	Target Control Transmitting Equipment RC-56-A	M.O.	5583-6167	F_1	CR-1B/AR	100	100
(D)	Test Set TS-67/ARN-5	6.9-mc and 20.7-mc signals for testing receivers in ILS	6900	F_1	CR-1A/AR	50	70
(E)	Radio Set AN/ARC-1A	Local osc in receiver	9720	F_1	CR-1A/AR	100	1.5
(F)	Frequency Meter TS-323/UR	Oscillator for heterodyne freq meter and crystal calibrator	1000	nF_1	Dallon's Laboratories D-1000	100	15
(G)	Signal Generator I-222	Calibration for vfo of signal generator	5000	nF_1	Holder FT-243	22	22
(H)	Radio Transmitter BC-1332	M.O.	8333.33	F_1	CR-1A/AR	51	1

Circuit Data for Figure 1-149. F in kc. R in kilohms. C in $\mu\mu$f. L in μh.

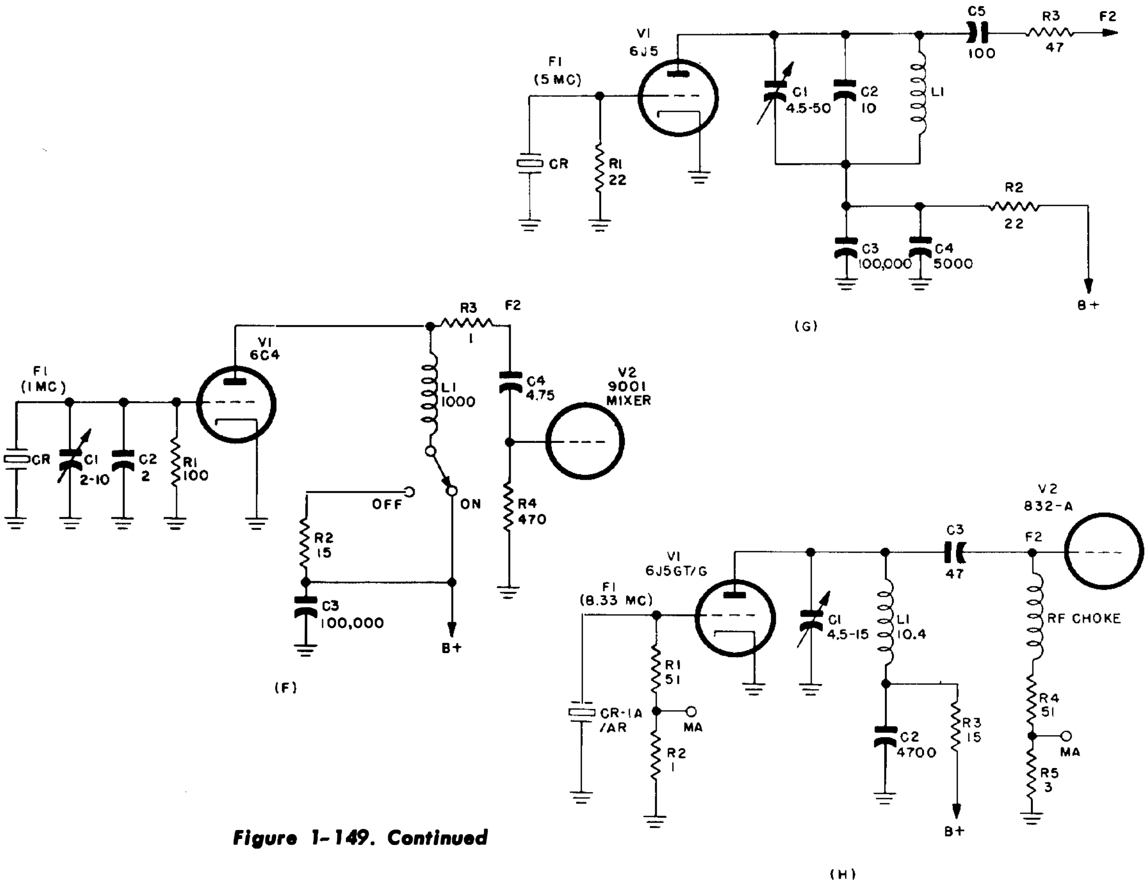
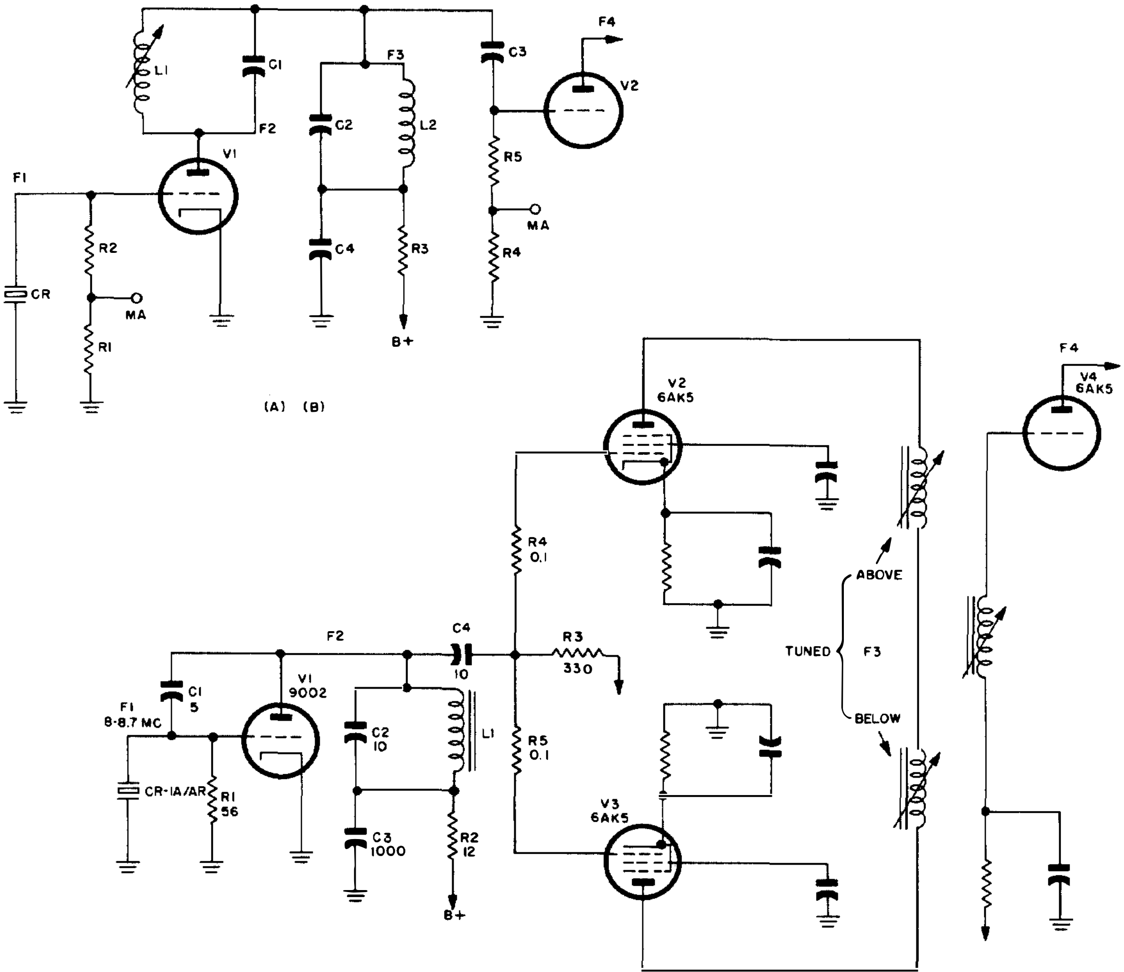


Figure 1-149. Continued

R_3	R_4	R_5	C_1	C_2	C_3	C_4	C_5	L_1	V_1	V_2
10	250	10	100	25	250	6000		560	1/2 12SL7GT	
10	560		6800	680	6800	220	15	26 turns	1/2 12AH7GT	9002
			10-100	100	4000				12J5GT	12J5GT
			5-50	50	4000				1/2 6SN7GT	1/2 6SN7GT
			24						6C4	
1	470		2-10	2	100,000	4.75		1000	6C4	9001
4.7			4.5-50	10	100,000	5000	100	44 turns No. 24 AWG; 0.77 x 1.94 in.	6J5	
15	51	3	4.5-15	4700	47			10.4	6J5GT/G	832-A



(C)

Figure 1-150. Modified Miller oscillators using triodes

Fig.	Equipment	Purpose	F_1	F_2	F_3	F_4	CR	R_1	R_2	R_3	R_4	R_5
(A)	Radio Receiver BC-733-DM	Local oscillator	5633–5745	F_1	$3F_1$	$9F_1$	Holder FT-243	1.5	180	8.2	1.5	220
(B)	Radio Receivers R-57/ARN-5 and R-89( )/ARN-5A	Local oscillator	6498–6548	F_1	$2F_1$	$4F_1$	Holder FT-243 (±0.02%)	1.5	8.2		1.5	39
(C)	Radio Receivers R-77/ARC-3 and R-77A/ARC-3	Local oscillator	8000–8727	F_1	F_1	$11F_1$-$18F_1$	CR-1A/AR	56	12	330	0.1	0.1
(D)	Test Oscillo-scope TS-100/AP	Nautical-mile range-synchronizing oscillator	80.86	F_1	F_1	F_1	±25 cps; −30° C to 50° C	2200	5.1	8.2	5.1	50

Circuit Data for Figure 1-150. F in kc. R in kilohms. C in $\mu\mu$f. L in μh.

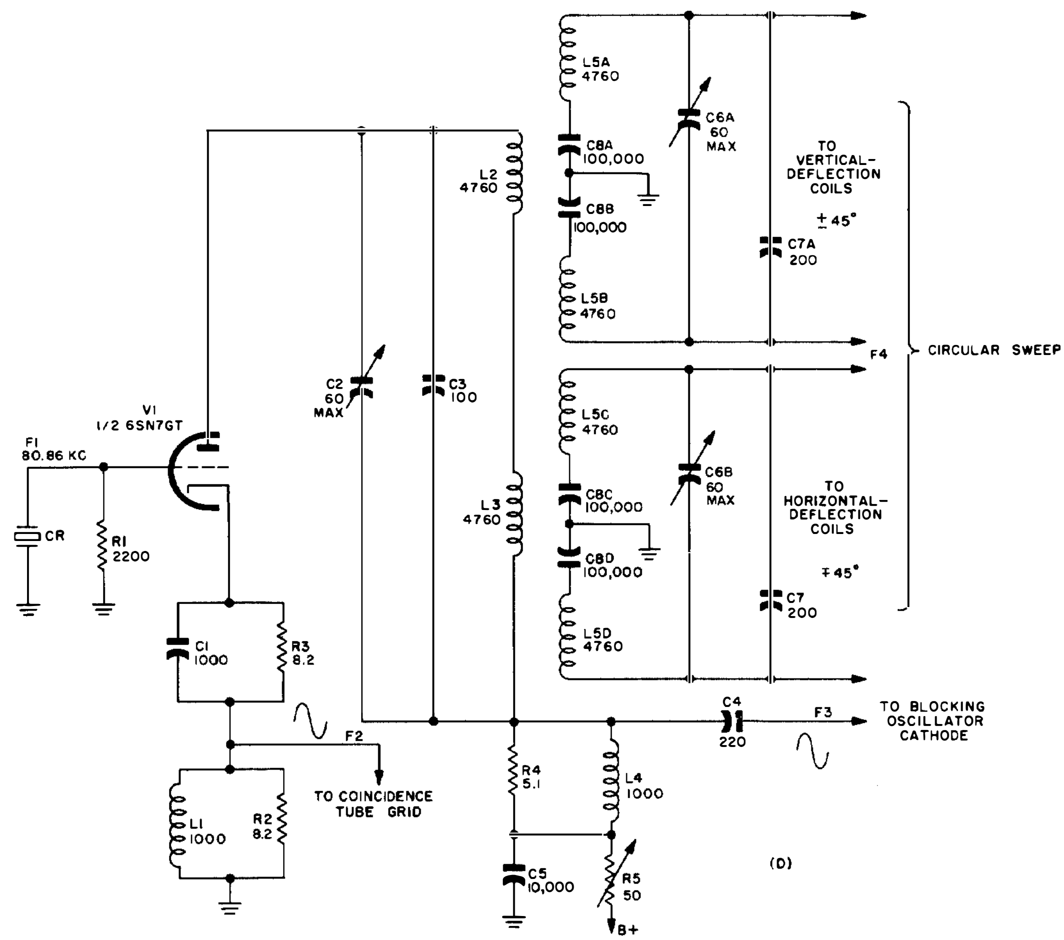
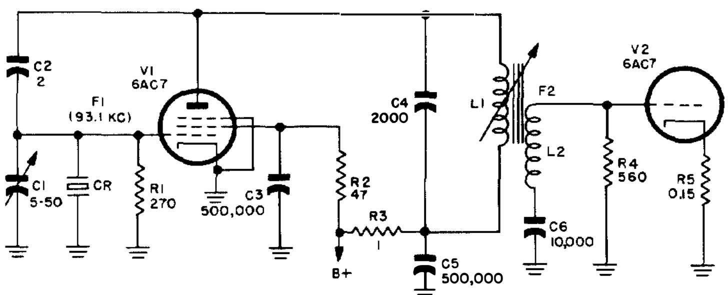
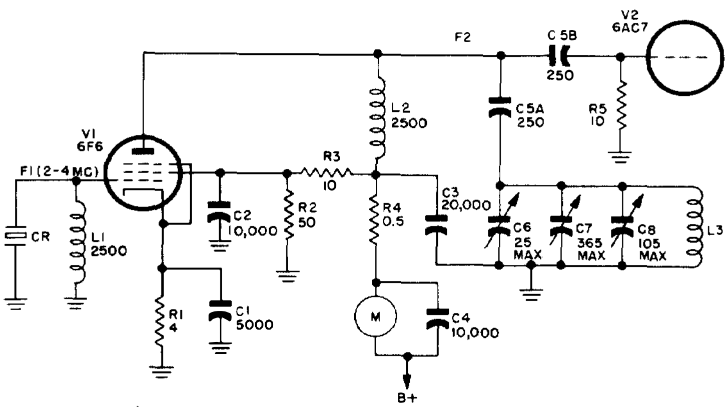


Figure 1-150. Continued

C_1	C_2	C_3	C_4	C_5	C_6	C_7	C_8	L_1	L_2	L_3	L_4	L_5	V_1	V_2	V_3	V_4
50	10	100	6000					44 turns, 3/8-in. dia	22 turns, 3/8-in. dia				1/2 12AH7 GT	1/2 12AH7 GT		
30	22	270											1/2 12SN7 GT	1/2 12SN7 GT		
5	10	1000	10					33 turns					9002	6AK5	6AK5	6AK5
1000	60	100	220	10,000	60	200	100,000	1000	4760	4760	1000	4760	1/2 6SN7 GT			



(A)



(B)

Figure 1-151. Miller oscillators using screen-grid tubes. (Effective suppressor in beam-power tubes is connected to cathode inside tube, although external connection may be indicated in diagram.)

Fig.	Equipment	Purpose	F_1	F_2	F_3	F_4	CR	R_1	R_2	R_3	R_4	R_5	R_6
(A)	Monitor ID-18/ CPN-2	Mile range synchro-nizing oscillator	93.109	F_1			75°C oven	270	47	1	560	0.15	
(B)	Radio Trans-mitters BC-339-E-to-M	M.O.	2000–4000	F_1			Holder FT-164	4	50	10	0.5	10	
(C)	Range Marker Generator TD-42/ FPS-3	Nauti-cal-mile range calibra-tor osc	80.86	F_1				100	0.15	68	15	470	10
(D)	Radio Trans-mitters BC-640-A,-B,-D	M.O.	5555.5–8666.6	F_1				100	0.05	50	25		
(E)	Radio Trans-mitter T-171B/FR	M.O.	125–525	F_1			Holder FT-249	5000	1	100	100		

Circuit Data for Figure 1-151. F in kc. R in kilohms. C in $\mu\mu$f. L in μh.

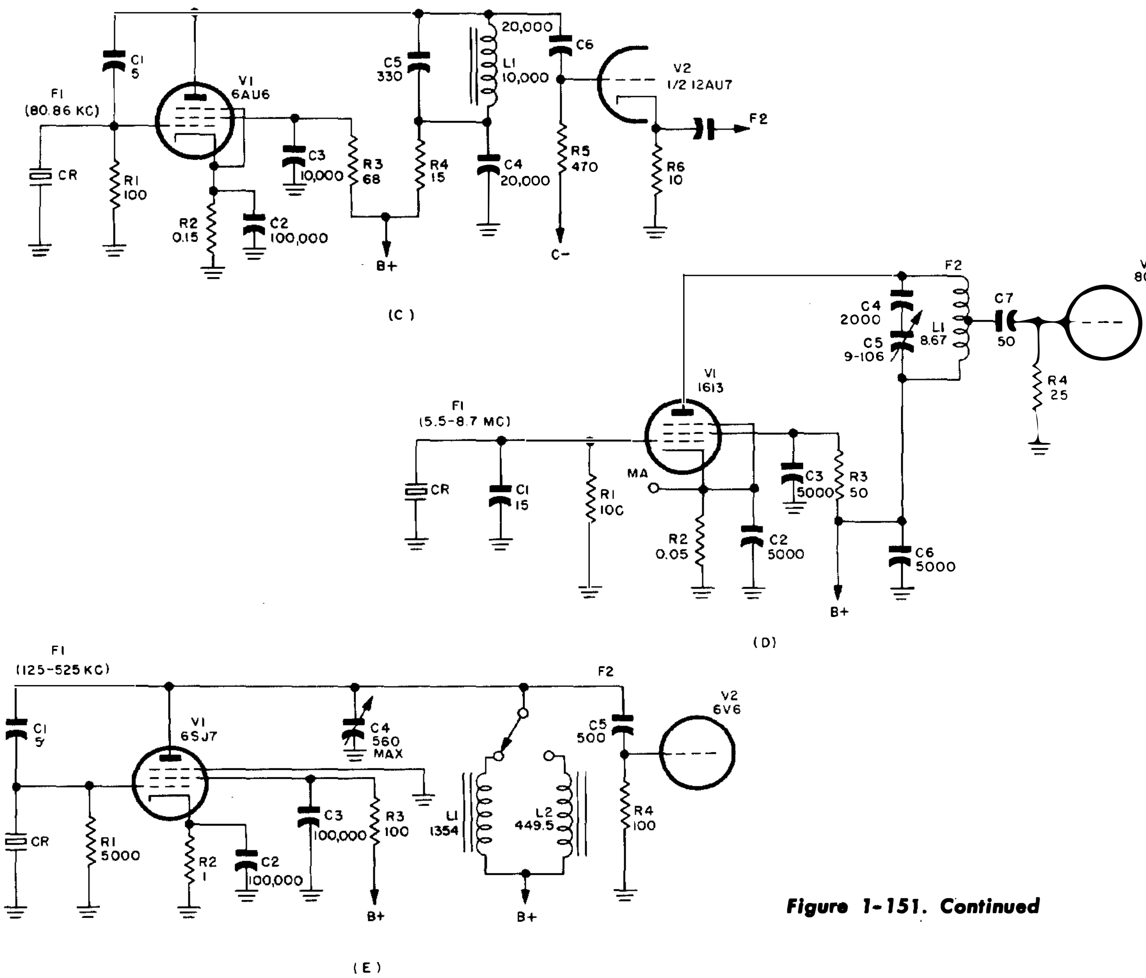
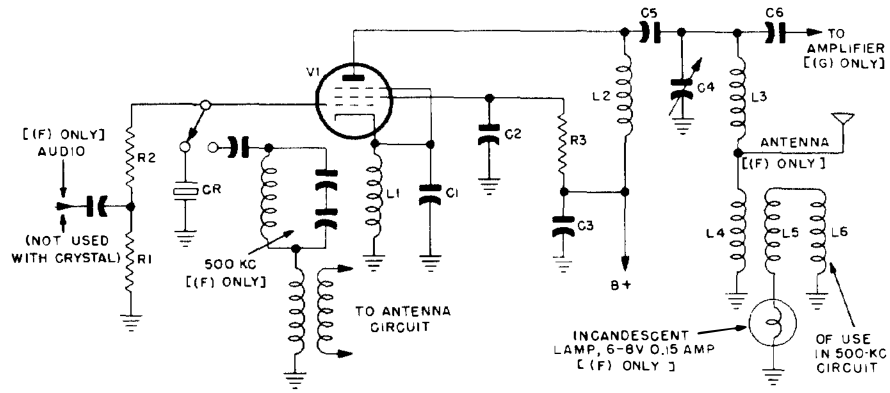
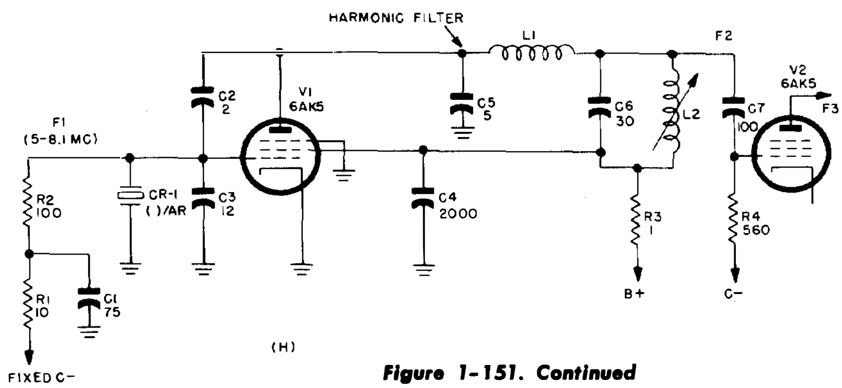


Figure 1-151. Continued

C_1	C_2	C_3	C_4	C_5	C_6	C_7	C_8	L_1	L_2	L_3	L_4	L_5	V_1	V_2
5-50	2	500,000	2000	500,000	10,000								6AC7	6AC7
5000	10,000	20,000	10,000	250	25	365	105	2500	2500				6F6	837
5	100,000	10,000	20,000	330	20,000			10,000					6AU6	1/2 12AU7
15	5000	5000	2000	9-106	5000	50		8.67					1613	807
5	100,000	100,000	560	500				1354 (7Ω)	449.5 (3.75Ω)				6SJ7	6V6



(F) (G)

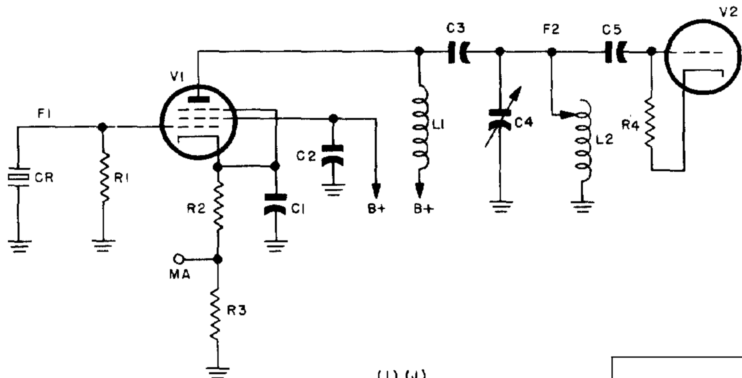


(H)

Figure 1-151. Continued

Fig.	Equipment	Purpose	F_1	F_2	F_3	F_4	CR	R_1	R_2	R_3	R_4	R_5	R_6
(F)	Radio Set AN/CRT-3	"Gibson Girl" power osc	8280	F_1			CR-1A/ AR	33	27	33			
(G)	Radio Trans- mitter T-91/ VRC-4	M.O.	1700– 8000	F_1			Sig Stock #2Z3531B	50	0	50			
(H)	Radio Set AN/ARC- 1A	Guard- channel hetero- dyne freq gen- erator	5020– 8120	F_1	$18F_1$		CR-1( )/ AR	10	100	1	560		
(I)	Radio Trans- mitter T-4/FRC	M.O.	2000– 6000	F_1			Holder FT-249	100	0.5	0.0238	100		
(J)	Radio Trans- mitter T-5/FRC	M.O.	150– 550	F_1			Sig Stock #2X41	100	0.5	0.0238	100		
(K)	Radio Trans- mitter T-10/ CRN-5	M.O.	5555.5– 8666.6	F_1	$3F_1$	$6F_1$	CR-1A/ A R or DC-11-( )	0.1	100	0.5	0.0008	20	100

Circuit Data for Figure 1-151. F in kc. R in kilohms. C in $\mu\mu$f. L in μh.



(I) (J)

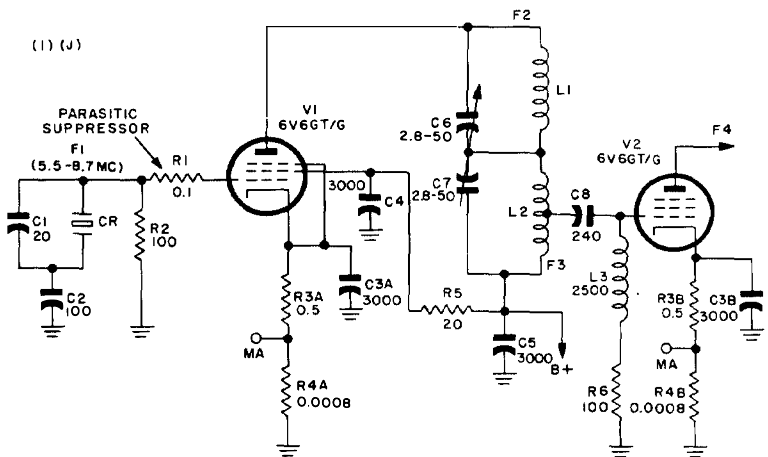


Figure 1-151. Continued

(K)

C_1	C_2	C_3	C_4	C_5	C_6	C_7	C_8	L_1	L_2	L_3	L_4	L_5	V_1	V_2
300	1000	1000	4.5–43.7	1000	0			2350	2500	3-1/2 or 12 turns 1-1/2 in.	3-1/2 or 12 turns 1-1/2 in.	1-1/4 turn, 1-1/2 in.	12A6	
250	10,000		7.2–143 7	100	100			2500	13, 25, 33, 63 turns	13, 25, 33, 63 turns	0	0	6V6 GT	
75	2	12	2000	5	30	100							6AK5	6AK5
10,000	10,000	500	9–150	500				2500					6V6G	807
10,000	10,000	10,000	11–250	10,000				10,000	Pie wound				6V6G	807
20	100	3000	3000	3000	2.8–50	2.8–50	240	37 turns	14 turns	2500			6V6GT/G	6V6GT/G

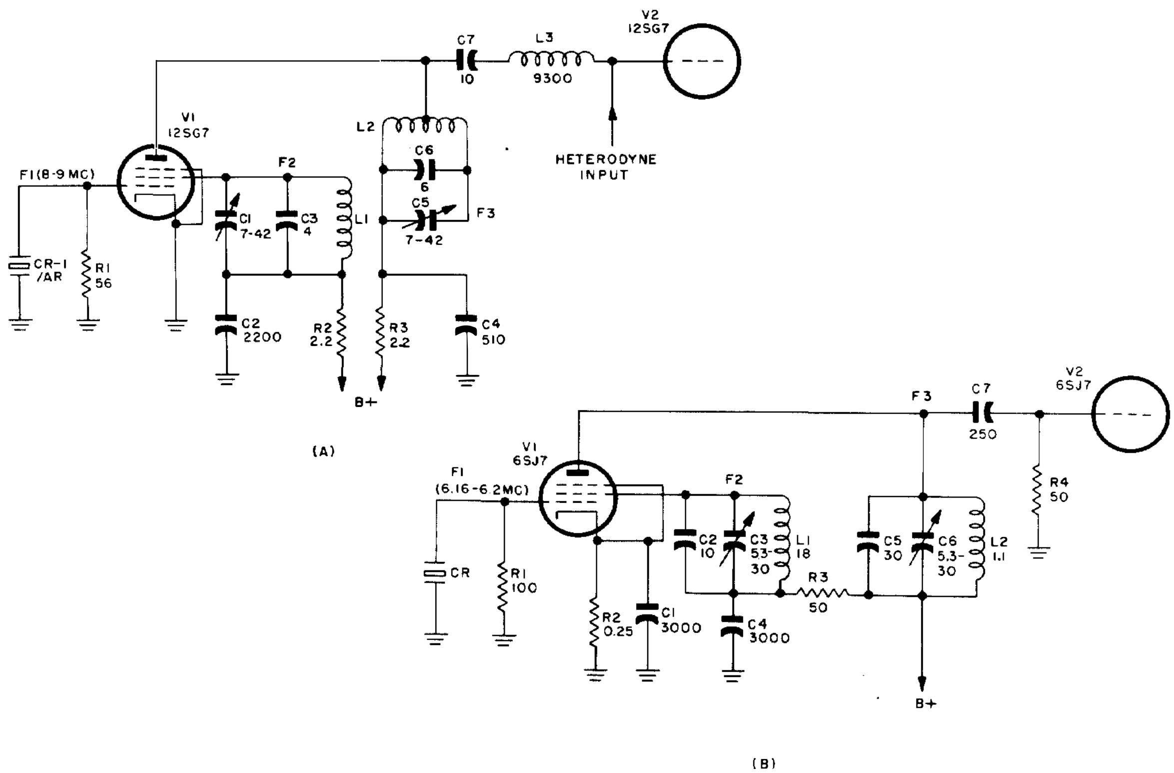


Figure 1-152. Electron-coupled Miller oscillators. All circuits except (A), (B), and (C) are tri-tet modifications

Fig.	Equipment	Purpose	F_1	F_2	F_3	CR	R_1	R_2	R_3	R_4	R_5
(A)	Radio Receiver and Selector BC-617-AZ	Heterodyne oscillator	8000–9000	F_1	$4F_1$	CR-1( )/AR	56	2.2	2.2		
(B)	Radio Transmitters T-3/CRN-2 and T-3A/CRN-2	M.O. and multiplier	6203.7–6159.26	F_1	$3F_1$	CR-1A/AR or DC-17-B	100	0.25	50	50	
(C)	Radio Transmitter BC-329-J	M.O.	200–410	F_1		Holder FT-164	1000	1.5	0.001	5	17.5
(D)	Radio Receiver R-146A/ARW-35	1st heterodyne oscillator	7833–10,000	$6F_1$	$6F_1$ + 5000	Holder FT-243	100	100	4700	0	
(E)	Radio Sets AN/TRC-1, -1A, -1B, -1C	Heterodyne osc in Radio Receiver R-19( )/TRC-1	7300–8750	$5F_1$ and $6F_1$	NA	CR-6/U	100	100	250	0	
(F)	Radio Sets AN/TRC-1D, -1E	Heterodyne osc in Radio Receiver R-19( )/TRC-1	7300–8750	$5F_1$ and $6F_1$	NA	CR-6/U	100	100	250	50	
(G)	Radio Receiver R-19H/TRC-1	Heterodyne oscillator	7300–8750	$5F_1$ and $6F_1$	NA	CR-6/U	100	100	240	51	

Circuit Data for Figure 1-152. F in kc. R in kilohms. C in $\mu\mu$f. L in μh. NA: Not Applicable.

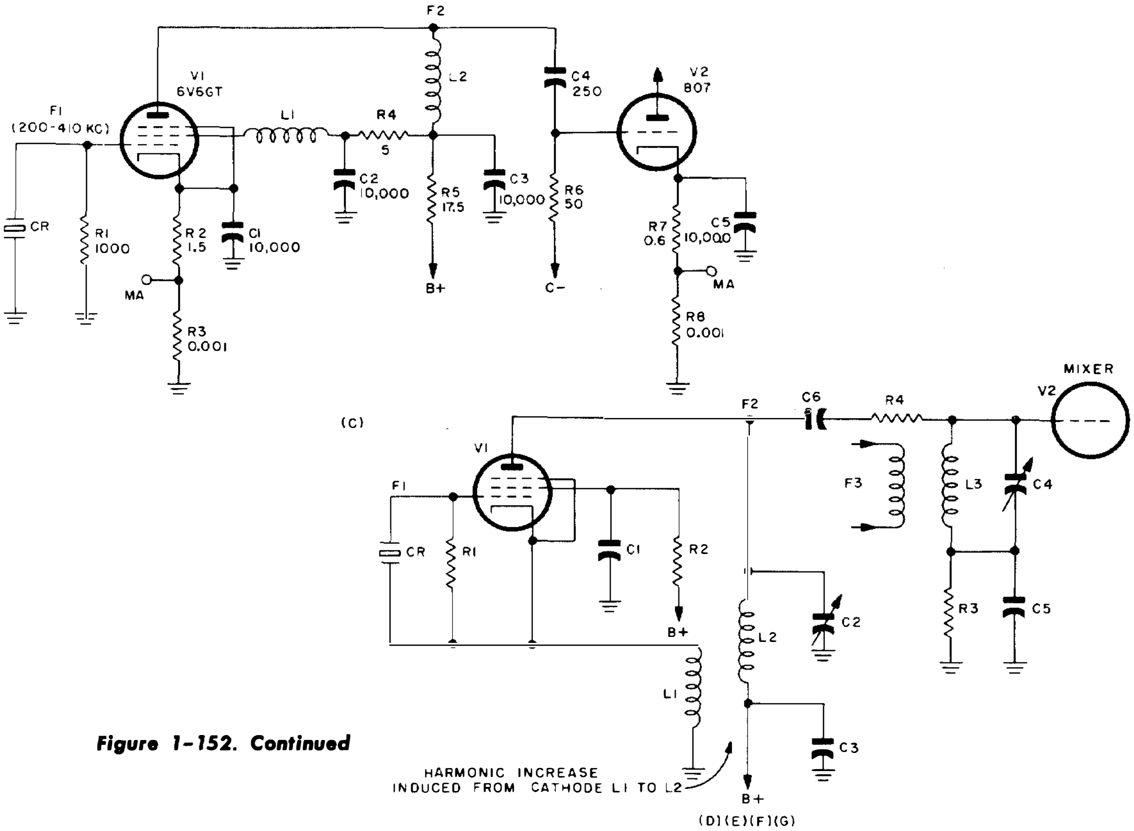
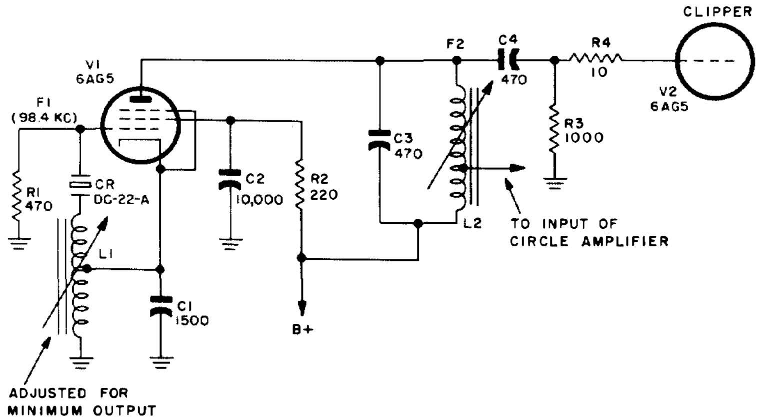


Figure 1-152. Continued

R_6	R_7	R_8	C_1	C_2	C_3	C_4	C_5	C_6	C_7	L_1	L_2	L_3	V_1	V_2
			7-42	2200	4	510	7-42	6	10			9300	12SG7	12SG7
			3000	10	5.3-30	3000	30	5.3-30	250	18	1.1		6SJ7	6SJ7
50	0.6	0.001	10,000	10,000	10,000	250	10,000			2500	2500		6V6GT	807
			2000	3-54	2000	5-30	250	4					12SH7	6AC7
			1500	3-54	1500	0	0	100				0	6SH7	6SH7
			1500	3-54	1500	0	0	100				0	6SH7	6SH7
			1500	3-54	1500	0	0	100				0	6SH7	6SH7



(H)

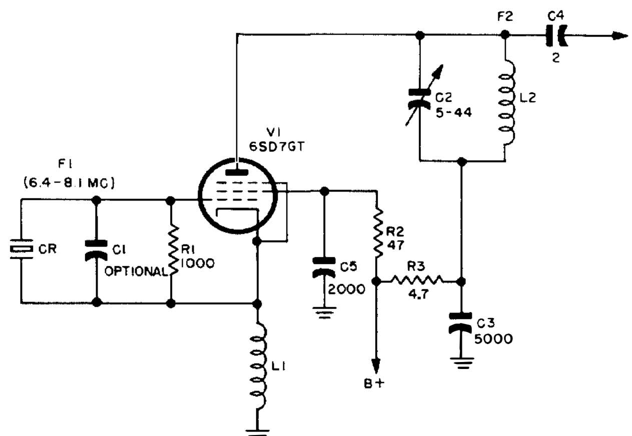
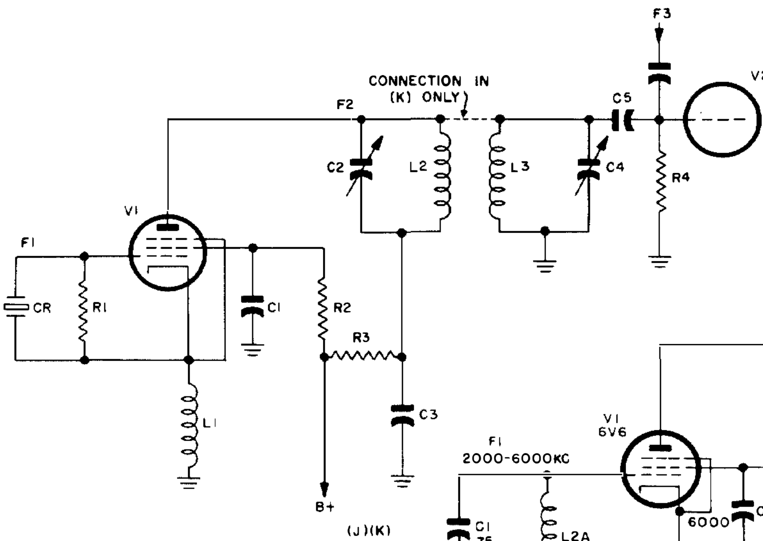


Figure 1-152. Continued

(I)

Fig.	Equipment	Purpose	F_1	F_2	F_3	CR	R_1	R_2	R_3	R_4	R_5
(H)	Radio Receiver and Transmitter BC-788-A	Timing osc and 5000-ft range control	98.356	F_1		DC-22-A	470	220	1000	10	
(I)	Galvin Radio Receivers PA-8098 and PA-8245	1st l— cal osc in Communication Equipment AN-CRC-3	6425-8120	$4F_1$ and $5F_1$		Holder FT-243	1000	47	4.7		
(J)	Radio Receiver R-368/FRC-6A	Heterodyne oscillator	6425-8425	$4F_1$ and $5F_1$	F_2+ 4.3		1000	47	4.7	47,000	
(K)	Radio Set AN/VRC-2	1st heterodyne osc	6425-8425	$4F_1$ and $5F_1$	F_2+ 4.3		47	47	4.7	47,000	
(L)	Radio Transmitter BC-610-E	M.O.	2000-6000	F_1		Holder FT-171-B	33	100	15	5.6	

Circuit Data for Figure 1-152. F in kc. R in kilohms. C in $\mu\mu$f. L in μh. NA: Not Applicable.



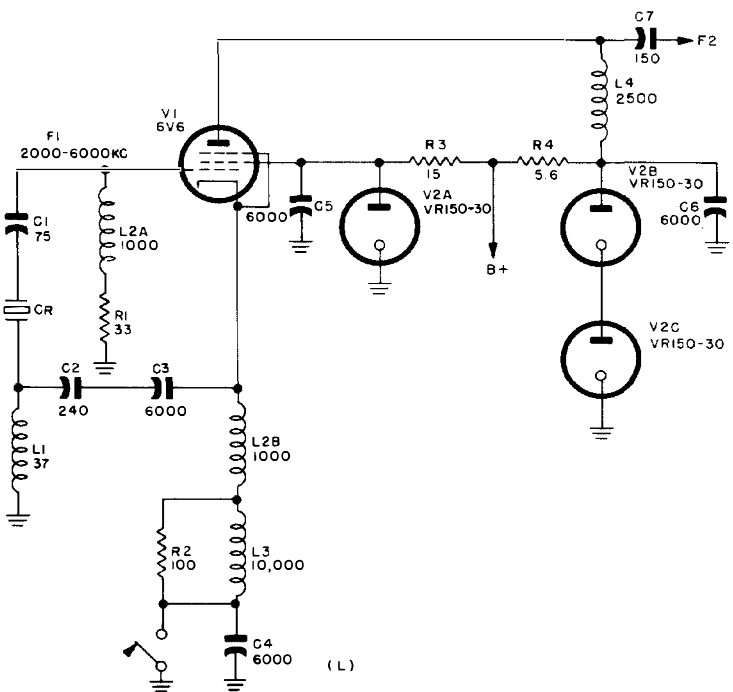


Figure 1-152. Continued

R_6	R_7	R_8	C_1	C_2	C_3	C_4	C_5	C_6	C_7	L_1	L_2	L_3	V_1	V_2
			1500	10,000	470	470				150 and 400 turns	Four 190-turn sections		6AG5	6AG5
			Optional	5-44	5000	2				(0.3Ω)			6SD7 GT	
			5600	6-45	5600	6-45	2						6SG7GT	6SG7GT
			2000	5-44	5000	0	2					∞	6SD7GT	6SD7GT
			75	240	6000	6000	6000	6000	150	37	1000	10,000	6V6	VR150-30

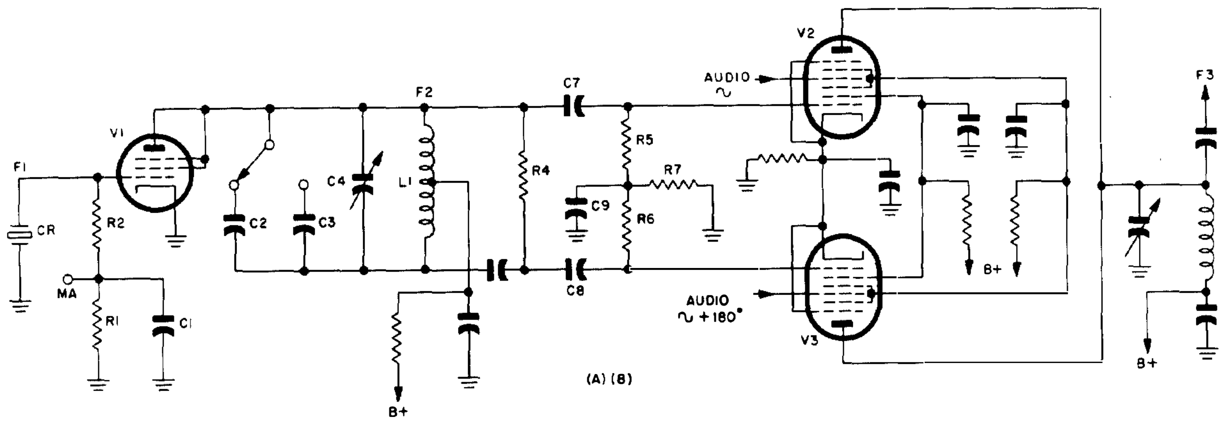
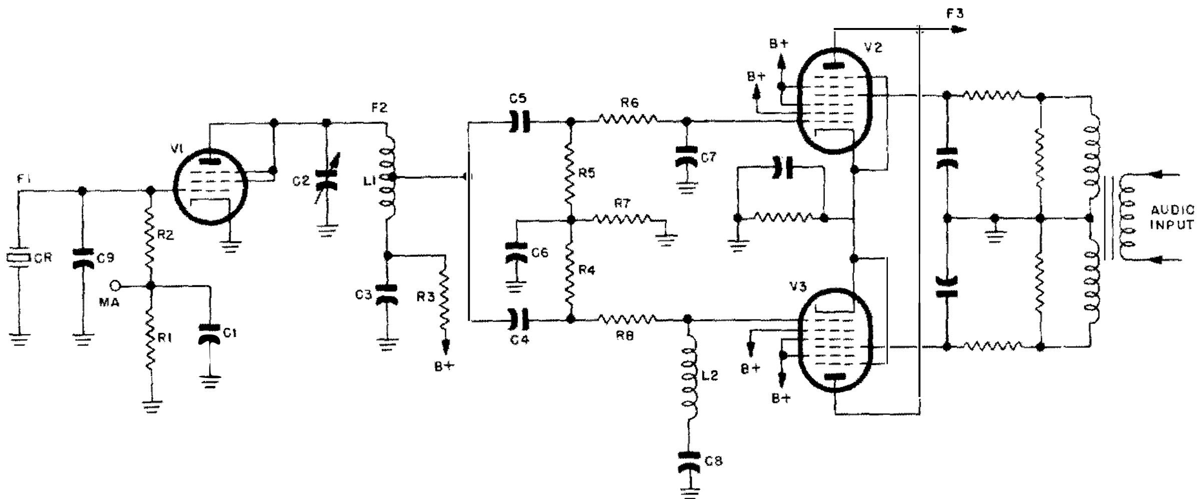


Figure 1-153 Miscellaneous Miller-oscillator modifications

Fig.	Equipment	Purpose	F_1	F_2	F_3	CR	R_1	R_2	R_3	R_4	R_5	R_6	R_7	R_8
(A)	Galvin Radio Transmitter PA-8218 P/O AN/CRC-3	M.O. for phase modulator circuit	937 5 1250	F_1	F_1 (mod.)	Motorola FMT	1	470	4 7	10	47	47	0 1	
(B)	Radio Transmitter T-264/FRC-6A	M.O. for phase modulator circuit	937.5 1250	F_1	F_1 (mod.)		1	470	4.7	10	47	47	0 1	
(C)	Galvin Radio Transmitter PA-8244 P/O AN/CRC-3	M.O. for phase modulator circuit	3750–5000	F_1	F_1 (mod.)	Holder FT-243	1	470	4 7	47	47	4.7	0 1	4 7
(D)	Galvin Radio Transmitter PA-8026 P/O AN/VRC-2	M.O. for phase modulator circuit	3750–5000	F_1	F_1 (mod.)	Holder FT-243	1	470	4.7	47	47	4.7	0 1	4.7

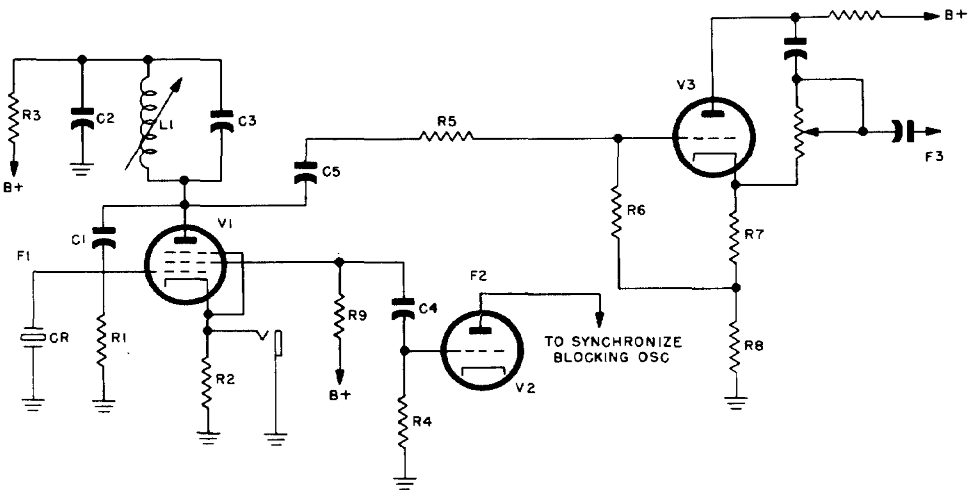
Circuit Data for Figure 1-153. F in kc. R in kilohms. C in $\mu\mu$f. L in μh, unless otherwise noted.



(C) (D)

Figure 1-153. Continued

R_9	C_1	C_2	C_3	C_4	C_5	C_6	C_7	C_8	C_9	L_1	L_2	L_3	V_1	V_2 V_3
	5000	25	50	5-44	5000	10	100	100	2000				7C7	7A8
	5600	24	0	6-45	5600	24	100	100	2200				7C7	7A8
	5000	5-44	5000	100	100	2000	5	2000	5		R-F choke		7C7	7A8
	5000	5-44	5000	100	100	2000	5	2000	5		R-F choke		7C7	7A8

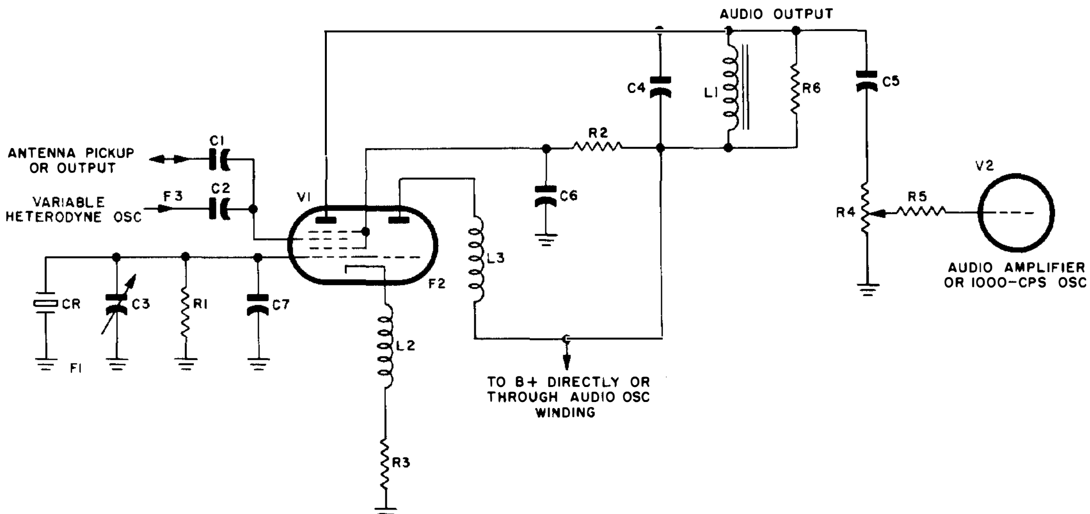


(E) (F) (G)

Figure 1-153. Continued

Fig.	Equipment	Purpose	F_1	F_2	F_3	CR	R_1	R_2	R_3	R_4	R_5	R_6	R_7	R_8
(E)	Crystal Calibrator TS-177/CPS-1	Calibration of mile range markers	93.12	F_1	F_1	GE# 32C401G43	1000	0.012	18	1000	2000	1000	2.2	10
(F)	Crystal Calibrator TS-241/CPS-5	Calibration of mile range markers	93.12	F_1	F_1	GE# 32C401G43	1000	0.012	18	1000	2000	1000	2.2	10
(G)	Calibrator-Generator Group OA-96/CPS-6B	Calibration of nautical mile range markers	80.867	F_1	F_1		1000	0	15	1000	2200	1000	2.2	10
(H)	Frequency Meter TS-174/U	Crystal calibrator and heterodyne circuit	1000	nF_1	20,000–250,000	DC-9	1000	8.75	0.075	1000	150	∞		
(I)	Frequency Meter TS-175/U	Crystal calibrator and heterodyne circuit	5000	nF_1	85,000–1,000,000	CR-1A/AR	1000		0.75	1000	0	100		
(J)	Frequency Meters BC-221-AG,-AK	Crystal calibrator and heterodyne circuit	1000	nF_1	1000–20,000	DC-9-AD or DC-9-P	1000	10	0.15	500	0	∞		

Circuit Data for Figure 1-153. F in kc. R in kilohms. C in $\mu\mu f$. L in μh, unless otherwise noted.



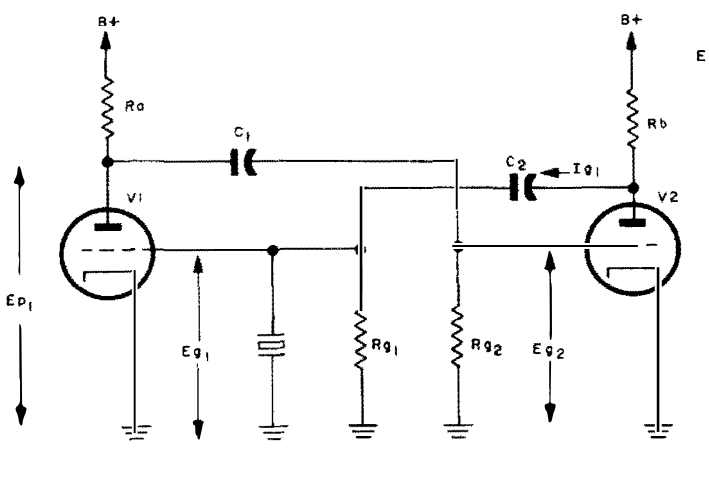
(H)(I)(J)

Figure 1-153. Continued

R_9	C_1	C_2	C_3	C_4	C_5	C_6	C_7	C_8	C_9	L_1	L_2	L_3	V_1	V_2 V_3
100	3.8	500,000	330	51	10,000					8500			6SJ7	1/2 6SN7GT
100	3.8	500,000	330	51	10,000					8500			6SJ7	1/2 6SN7GT
100	5	500,000	470	47	10,000					8500			6SJ7	1/2 6SN7W
	12	12	3-25	1000	50,000	2000	0			80h	0	850	6K8	6SJ7
	2.5	12	3-25	0	6000	2000	0			∞	175	70	6K8	6C8G
	25	50	12	1000	20,000	250,000	6			150h	0	1000	6K8	6SJ7

Two-Tube Parallel-Resonant Crystal Oscillators

1-345. Most of the two-tube oscillator circuits are designed so that the crystal is operated at or near its series-resonance frequency. An exception is the multivibrator type of circuit in which the crystal unit is connected between the grid and cathode of one of the tubes. The basic circuit is shown in figure 1-154(A). If there were no capacitive effects to consider, E_{g2} would be 180 degrees out of phase with E_{g1} and the 180-degree feed-back inversion would be accomplished entirely by V_2. In this case, the input impedance of V_1 would have to be purely resistive—that is, the crystal unit would operate at parallel resonance with the input capacitance. The proper phase of E_{g1} could also be obtained with the crystal unit operating near series resonance, and unless the V_1 input impedance were so reduced that oscillations could not be sustained under such condition there would always be the risk that the oscillator would jump from one equilibrium state to the other. In an actual circuit, the circuit capacitances will prevent V_1 from operating into a purely resistive load, so that E_{p1} slightly lags the equivalent generator voltage $-\mu E_{g1}$. The larger the values of R_a, R_{p1}, R_{g2}, C_1, and C_{g2}, the nearer will the lag in E_{p1} approach the 90-degree limit. If it is assumed that R_{g2} is very large compared with the reactance of the V_2 input capacitance, C_{g2}, the phase of E_{g2} is approximately the same as the phase of E_{p1}. Also, if the ratio, C_1/C_{g2}, is very large, the magnitudes of E_{g2} and E_{p1} are very nearly equal. As can be seen in figure 1-154 (B), the lag in E_{p1} causes the equivalent generator voltage of V_2, equal to $-\mu E_{g2}$, to lag E_{g1}. The circuit will oscillate at that frequency at which the crystal impedance creates the necessary phase difference between E_{g1} and $-E_{g2}$. Note, that except for the possible coupling between the output circuits of V_1 and V_2 because of the grid-to-plate capacitance of V_2 (which can be made negligible by the use of screen-grid tubes), the phase of E_{g2}, and hence of $-\mu E_{g2}$, is entirely independent of impedance changes in the V_2 plate circuit. Thus, to predetermine the angle θ in figure 1-154(B), it is only necessary to consider the V_1 stage as a conventional vacuum-tube amplifier circuit. In turn, the V_2 stage can be treated separately as an equivalent circuit driven by a generator of voltage $-\mu E_{g2}$. The design must be such that when the crystal unit is operating at its rated X_o, the voltage across the crystal unit differs from the equivalent generator voltage, $-\mu E_{g2}$, by the desired angle θ. If E_{g1} leads E_{p2}, as indicated in figure 1-154(B), the feed-back current through C_2 will be very nearly in phase with E_{p2}. The input impedance of V_1 will appear somewhat inductive, and close to series resonance with C_2. The smaller that C_2 is made, the higher will be the frequency. If the plate circuit of V_2 is made inductive, E_{p2} (not by changing θ) can be shifted to be more nearly in phase with E_{g1}; however, such operation would tend to become unduly critical. For example, assume that E_{p2} were rotated to where it was in phase with E_{g1}. This would mean that the phase of the feed-back current, I_{g1}, with respect to E_{p2} was equal to its phase with respect to E_{g1}, and this in turn would require that the over-all Q of Z_{g1} and C_2 in series be the same as the Q of Z_{g1} alone—an impossibility unless C_2 is infinite. But C_2 cannot be made large without the risk that the circuit will operate as an RC controlled multivibrator. Thus, in the circuit of figure 1-154, E_{p2} must lag E_{g1} if oscillations are to be maintained. The smaller the phase difference between E_{p2} and E_{g1}, the less will be the leading component of I_{g1} with respect to E_{p2}, and



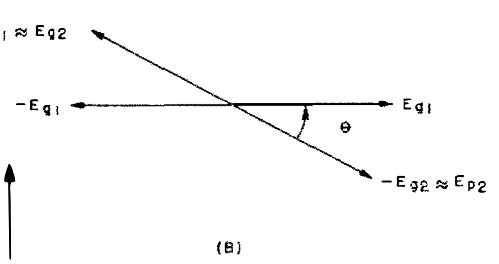


Figure 1-154. (A) Basic circuit of the multivibrator type of parallel-resonant crystal oscillator. (B) Phasor diagram of r-f voltages of circuit shown in (A). $-E_{g1}$ and $-E_{g2}$ represent the phases of the equivalent V_1 and V_2 generator voltages

the less will be the leading (inductive) component of E_{g1} with respect to I_{g1}. Hence, the smaller the angle θ, the more nearly must the crystal approach parallel resonance with the input capacitance of V_1. Any change in the V_1 plate circuit that tends to decrease θ therefore tends to raise the frequency. C_{g1} can be increased by the insertion of a fixed capacitance to make the total approach the rated load capacitance of the crystal unit.

1-346. Since the crystal unit is effectively a capacitance at all frequencies except those near its points of mechanical resonance, some precaution must be taken in the circuit design to ensure that the crystal maintains control over the frequency and that no danger exists that the two-stage circuit can perform as a free-running multivibrator with the frequency controlled by the RC constants. When C_2 in figure 1-154 is small and C_{g1} is large by comparison, and when the plate impedances R_a and R_b are small by comparison with the R_p of the tube, the feed-back voltage at the low frequencies corresponding to the RC time constants can quite easily be kept below the requirements for sustained oscillations.

1-347. No data is available concerning the relative frequency stability of the parallel-resonant multivibrator type oscillator, but, from qualitative considerations only, it would seem that a performance equal to, and very possibly superior to, that of the average Pierce circuit could be expected, although the operation of the circuit would certainly be much more critical. The V_1 amplifier stage can be designed to operate into a practically purely resistive load, so that fluctuations in the V_1 plate resistance will have little or no effect upon the frequency. On the other hand, under these conditions θ will be slightly negative and $-\mu E_{g2}$ will lead E_{g1} unless a large reactance, or, preferably, a resistance, is connected in series with C_1. If θ is not critically small, the external circuit of V_2 can also be designed to appear as a pure resistance, and any variations in the plate resistance of the tube will have a negligible effect on frequency. The annulling of the stray-capacitance effects in the output circuits of the two tubes will require the use of coils, which may not be desirable if a wide frequency range is intended. The phase-shifting Q_f of the feed-back circuit is computed in the same manner as in the Pierce and Miller circuits except that the required phase shift is much smaller.

1-348. In the design of such a circuit, the value of θ can be arbitrarily predetermined. Assume that when R_e of the crystal unit is a maximum, the feed-back current, I_{g1}, is to be in phase with $-\mu E_{g2}$. With R_{g1} assumed to be large compared with the average PI of the crystal unit, estimate the capacitance C_{g1} required across the crystal unit for the Q of Z_{g1} to equal the tangent of θ when the X_e of the crystal unit corresponds to the reactance of the rated load capacitance. Make C_2 such that its reactance at the operating frequency is equal to the reactive component of Z_{g1} computed above. Under these conditions I_{g1} will lag $-\mu E_{g2}$ for all values of R_e less than maximum, since the Q of Z_{g1} would be greater than tan θ if the frequency did not increase to bring Z_{g1} nearer its antiresonant value. But the increase in the frequency as a result of a decrease in R_e is less than that which would occur if the circuit were designed so that I_{g1} would lead $-\mu E_{g2}$ for most values of R_e. Also, if the feed-back circuit is designed for series resonance when R_e is a maximum, the values of C_2 should prove more practical, the operation will be less critical, and there is the assurance that all values of R_e will permit oscillation. The plate-circuit impedances of the two tubes are next determined so that the feed-back voltage is sufficient to maintain oscillations when R_e of the crystal unit is a maximum. Generally, neither tube should operate into a load exceeding 5000 ohms. The plate voltages are chosen so that the crystal unit cannot be driven beyond the rated drive level. With the phase characteristics of the circuit determined, more or less by design, to ensure a proper load capacitance for the crystal, it may be that the optimum operating voltage will be more readily determined through experiment. At equilibrium, the total voltage gain of the loop, from E_{g1} to E_{p1} to E_{g2} to E_{p2} and back to E_{g1}, must be equal to unity. Thus,

$$G_1\,G_2\,G_3\,G_4 = \frac{E_{p2}}{E_{g1}} \cdot \frac{E_{g2}}{E_{p1}} \cdot \frac{E_{p2}}{E_{g2}} \cdot \frac{E_{g1}}{E_{p2}} = 1 \qquad 1\text{—}348\ (1)$$

where, referring to the circuit in figure 1—154,

$$G_1 = E_{p1}/E_{g1} = \frac{\mu_1\,Z_{p1}}{R_{p1} + Z_{p1}} \approx g_{m1}\,R_a \qquad 1\text{—}348\ (2)$$

$$G_2 = E_{g2}/E_{p1} \approx 1 \qquad 1\text{—}348\ (3)$$

$$G_3 = E_{p2}/E_{g2} = \frac{\mu_2\,Z_{p2}}{R_{p2} + Z_{p2}} \approx g_{m2}\,R_b \qquad 1\text{—}348\ (4)$$

$$G_4 = E_{g1}/E_{p2} \approx 1 \qquad 1\text{—}348\ (5)$$

So

$$G_1\,G_2\,G_3\,G_4 \approx g_{m1}\,g_{m2}\,R_a\,R_b \approx 1 \qquad 1\text{—}348\ (6)$$

Equation (6) is only a first-order approximation in which it is assumed the plate resistances of the tubes are very large compared with the external plate impedances, Z_{p1} and Z_{p2}, which, in turn, are approximately equal to R_a and R_b, respectively. Also, it is assumed that X_{c1} is small compared with Z_{g2}, and that either X_{c2} is small compared with Z_{g1} or the tendency towards a series-resonant rise in voltage across Z_{g1} is sufficient to make equation (5) approximately correct.

1-349. The output is most often taken from across a 500- to 1000-ohm resistance between the cathode of V_2 and ground. In this event, R_{g2} connects directly to the cathode of V_2—not to ground. The cathode output is quite useful for matching to low-impedance inputs, such as would occur, for example, when feeding a coaxial line. Regardless of where the output is obtained, it can be seen that its amplitude cannot be expected to greatly exceed that of the Pierce circuit. Since two amplifier stages are required and no additional gain is produced, there can be little advantage in using the multivibrator circuit unless thermostatic control of the temperature is employed and the design is such as to ensure greater frequency stability than can be achieved in the Pierce circuit. The use of a single dual-type tube offers greatest economy. The principal advantages of the circuit are its relative independence of fluctuations in the tube voltage, and its adaptability for impedance-matching to low-impedance output circuits.

1-350. Figure 1-155 illustrates three multivibrator-type crystal oscillators that were designed for use in Diversity Receiving Equipment AN/FRR-3 ( ). Circuit (A) is a later-model replacement of circuit (B). Very possibly the preference for (A) is at least partly due to a desire to eliminate the variable effects of the inductors in (B) with changes in frequency. From the data available it cannot be said that the crystal in (B) is not actually operating at or very near its series-resonance frequency. The state of operation of the

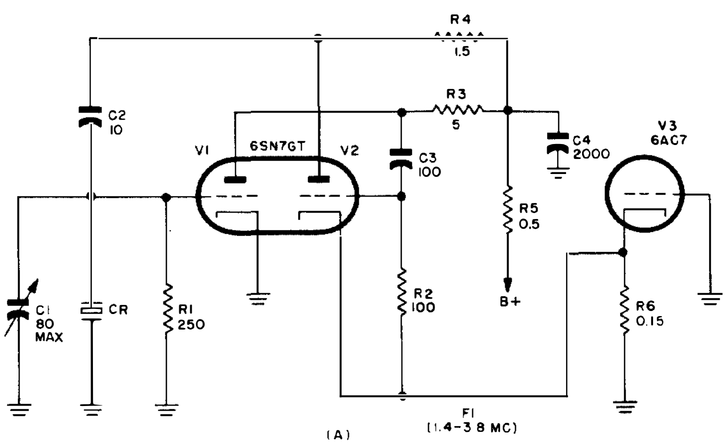
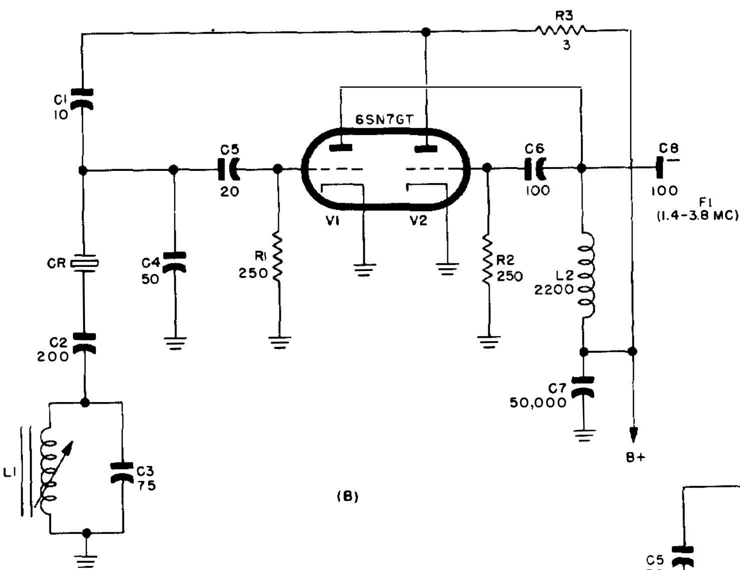


Figure 1-155. Two-stage parallel-resonant oscillators of the multivibrator type.

Fig.	Equipment	Purpose	F_1	CR	R_1	R_2	R_3	R_4	R_5	R_6	R_7	R_8
(A)	Diversity Receiving Equipment AN/FRR-3A	Local oscillator	1400–3800	Holder FT-249	250	100	5	1.5	0.5	0.15		
(B)	Diversity Receiving Equipment AN/FRR-3	Local osc.	1400–3800	Holder FT-249 (Entire circuit in 55° C oven)	250	250	3					
(C)	Diversity Receiving Equipment AN/FRR-3A	BFO with AFC reactance tube, V_3	462.45		50	0.6	10	100	250	250	5	5

Circuit Data for Figure 1-155. F in kc. R in kilohms. C in $\mu\mu$f. L in μh.



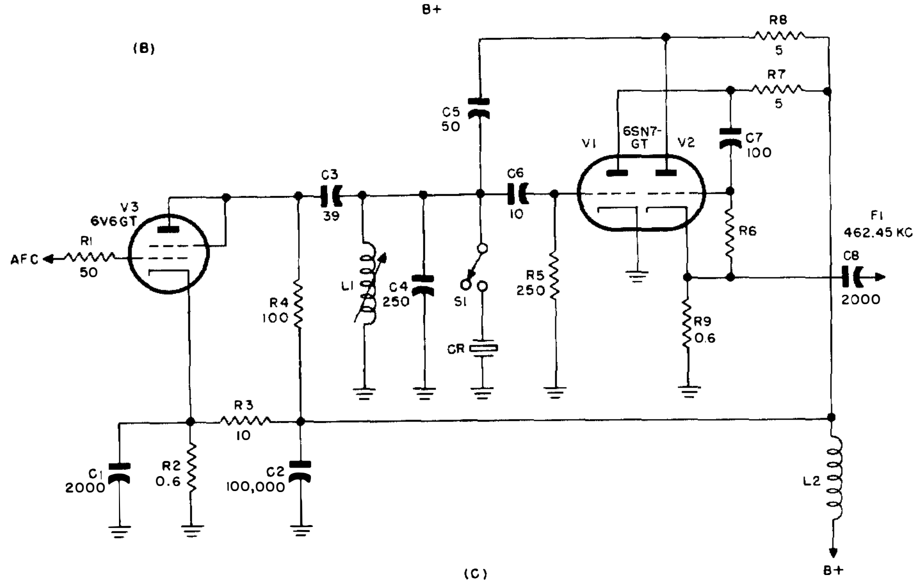


Figure 1- 155. Continued

R_9	C_1	C_2	C_3	C_4	C_5	C_6	C_7	C_8	L_1	L_2	V_1-V_2	V_3
	80	10	100	2000							6SN7GT	6AC7
	10	200	75	50	20	100	50,000	100		2200	6SN7GT	
0.6	2000	100,000	39	250	50	10	100	2000		2200	6SN7GT	6V6GT

crystal will vary quite widely with changes in the tuning of the L_1C_3 tank. Circuit (C) is a beat-frequency oscillator which is crystal-stabilized when switch S_1 is in the crystal position, as shown. The triode-connected beam power tube serves as a reactance tube, which effectively shunts the crystal with a capacitance that varies with the bias supplied to the control grid of the tube. The bias, in turn, is controlled by the a-f-c discriminator circuit in a teletype terminal. The purpose of the circuit is to ensure that the beat frequency remains constant even though the frequency of the incoming signal should vary slightly. If the beat frequency tends to drift, the sign and magnitude of the discriminator output causes the bias of the reactance tube to effectively change the load capacitance of the crystal unit in such a direction that the frequency of the oscillator rises or falls by approximately the same number of cycles per second as does the incoming signal.

Oscillators with Crystals Having Two Sets of Electrodes

1-351. The original crystal oscillator devised by Dr. Nicolson, as well as a number of the earlier crystal oscillators tested by Dr. Cady, employed crystals with, effectively, two pairs of electrodes. The basic circuit is shown in figure 1-156. The required phase inversion of the amplifier output voltage is provided by the crystal unit operating at a mode for which the polarities of the plate and grid terminals with respect to ground are 180 degrees out of phase. The circuit shown operates the crystal unit very near its series-resonance frequency. In practice, a capacitor is normally connected between crystal and ground, so that the circuit is more commonly employed for parallel-mode tested crystal units. Still, it is not without some license that we classify this type of oscillator as a parallel-mode type. The crystals most applicable for this class of circuit are the very-low-frequency elements of the X group, which vibrate in lengthwise extensional or flexural modes. The electrode connections that permit the desired phase inversion depend upon the particular crystal element. Assume that electrodes numbers 1 and 3 are on one side of the crystal, and that 2 and 4 are on the opposite side, as indicated in figure 1-156(A). For a flexure element, such as element N, where electrodes 1 and 3 parallel each other down the length of the crystal, as shown in figure 1-156(B), the flexure mode is excited when the potential across 1 and 2 is oppositely polarized to that across 3 and 4. If the same electrode arrangement is to be used to excite an extentional mode (or the flexural mode of the duplex element J) the polarities of the two sets of electrodes must be in phase. In this case, the connections of one set of electrodes should be reversed in the circuit shown in figure 1-156(A). For example, plates 2 and 3 should be connected to ground and plate 4 should be connected to the grid, if the proper phase inversion is to be obtained. A crystal having the two sets of electrodes at opposite ends of the crystal, as shown in figure 1-156(C), would be driven at the second harmonic of the length extensional mode (or of the flexural mode of a duplex crystal), if connected as shown in figure 1-156(A). Greater stability and a smaller crystal are possible for a given frequency by operating at the fundamental mode. To permit this, if the crystal unit is plated as shown in figure 1-156(C), the connections of one pair of electrodes should be the reverse of those shown in figure 1-156(A). If it can be assumed that the current in the grid circuit is negligible compared with the crystal current between terminals 1 and 2, and if the stray capacitance between the two sets of electrodes is ignored, the equivalent circuit between terminals 1 and 2 will appear approximately as shown in figure 1-156(D). L, C, and C_0 represent the parameters of a fully plated crystal. A more exact analysis of this type of crystal unit can be found in the book "Electromechanical Transducers and Wave Filters" by W. P. Mason, D. Van Nostrand Co.

1-352. Figure 1-157 shows a practical oscillator design employing crystal units having two sets of electrodes. Although the electrode connections shown for CR would indicate that the plate and

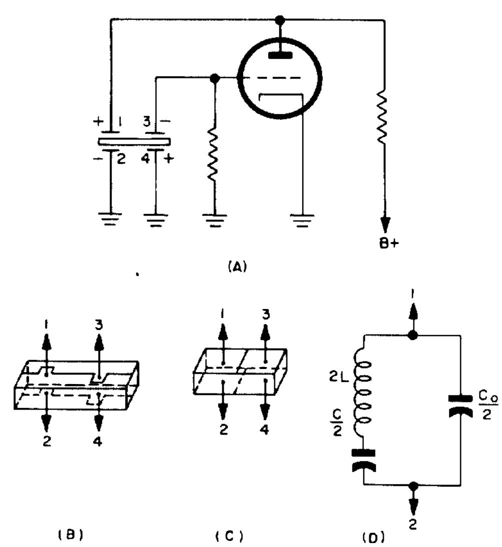


Figure 1-156. Basic circuit of oscillator using crystal with two pairs of electrodes

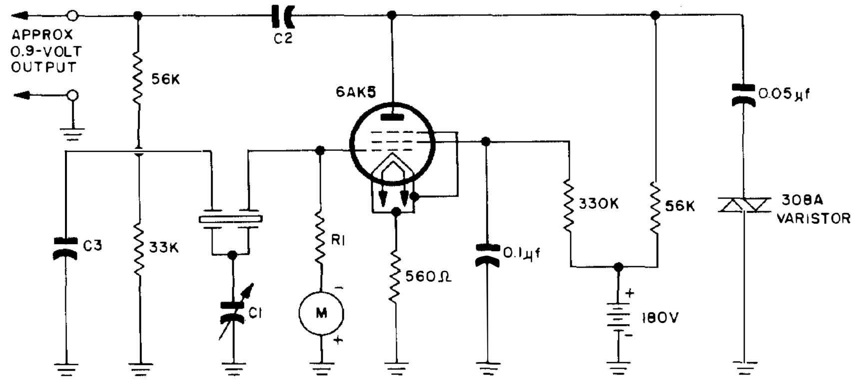


Figure 1-157. Practical crystal-oscillator design employing very-low-frequency crystal unit with two sets of electrodes

grid terminals always connect to the same side of the crystal unit, the actual connections will depend upon the particular element used. The variable capacitor C_1 permits a frequency adjustment of approximately 60 parts per million. To ensure that the crystal unit is operating into its rated load capacitance, the exact frequency of a test crystal should be known when it is at series resonance with its rated capacitance. To a first approximation, the terminal that is to be connected to the grid can be assumed to be open-circuited, so that the resonance to be tested is that between the rated C_x and one half of the crystal. With the test crystal connected in the oscillator circuit, C_1 can be adjusted to provide an output at the previously measured "rated" frequency of the test crystal. This adjustment therefore will provide the rated load capacitance for all crystal units of the same type. The varistor is inserted to protect the crystal unit from overdrive, and to ensure a stable output voltage. As recommended by Bell Telephone Laboratory engineers, the nominal values of R_1, C_1, C_2, and C_3 that provide satisfactory operation in the 1.2- to 10-kc frequency range are given in the following table. The values shown will provide a direct current in M of approximately 12 microamperes.

Frequency Range (kc)	R_1 (kil-ohms)	C_1 ($\mu\mu$f)	C_2 ($\mu\mu$f)	C_3 ($\mu\mu$f)
1.2 - 1.5	100	180	4000	500
1.5 - 2.0	100	180	3000	500
2.0 - 2.5	100	150	2000	500
2.5 - 3.2	100	150	1500	500
3.2 — 4.5	100	120	1000	500
4.5 - 6.7	100	120	700	250
6.7 - 8.0	100	90	500	250
8.0 —10.0	51	90	1000	0

1-353. Figure 1-158 shows the crystal oscillator in Test Set TS-251/UP, which employs a duplex crystal element. The crystal circuit is used to synchronize the blocking oscillator at a frequency of 1818.18 cps. The output of the blocking oscillator is for counting down to 303.03 pulses per second, which, in turn, are used in checking Loran pulse-repetition rates. A CR-11/U crystal unit is used which has a resonant frequency of 1817.44 $\pm$0.3 cps at 75° Fahrenheit. The rated maximum effective resistance of the crystal is 30,000 ohms, and its rated maximum permissible current is 0.03 milliampere. The fixed capacitance paralleling the variable capacitance is used only if necessary. The varistor is rated at 1 ma/14V for temperatures between 75 and 86 degrees Fahrenheit.

Crystal and Magic-Eye Resonance Indicator

1-354. An interesting application of a parallel-resonant crystal circuit is the tuning indicator shown in figure 1-159. When the tuned frequency, F_1, of a variable oscillator is equal to the antiresonant frequency of the crystal unit in parallel with the input capacitance of V_1, a magic-eye tube, the excitation of V_1 is a maximum, as is the current through R_4, and, hence, also the shadow angle of the indicator. The circuit thus provides a constant visual crystal check on the tuned oscillator frequency. Different crystal units can be switched in for different channels.

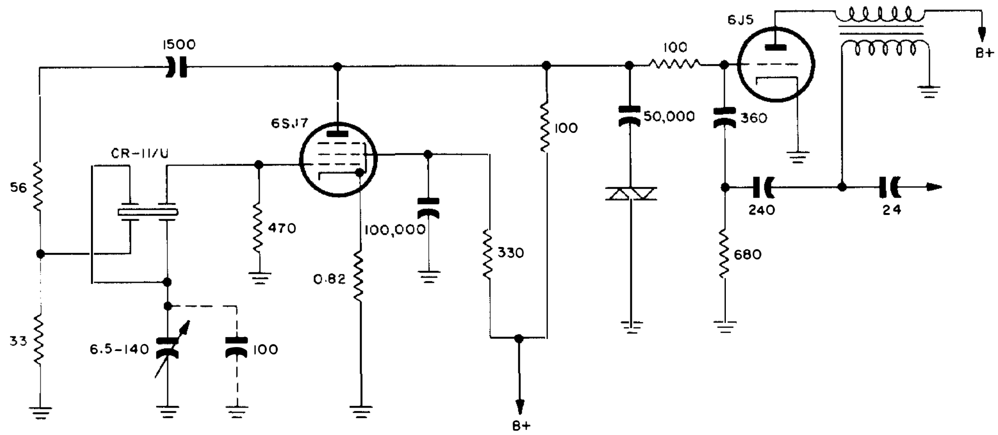


Figure 1-158. Duplex-electrode crystal circuit in Test Set TS-251/UP for synchronizing blocking oscillator at 1818.18 cps. Resistance is given in kilohms, capacitance in micromicrofarads

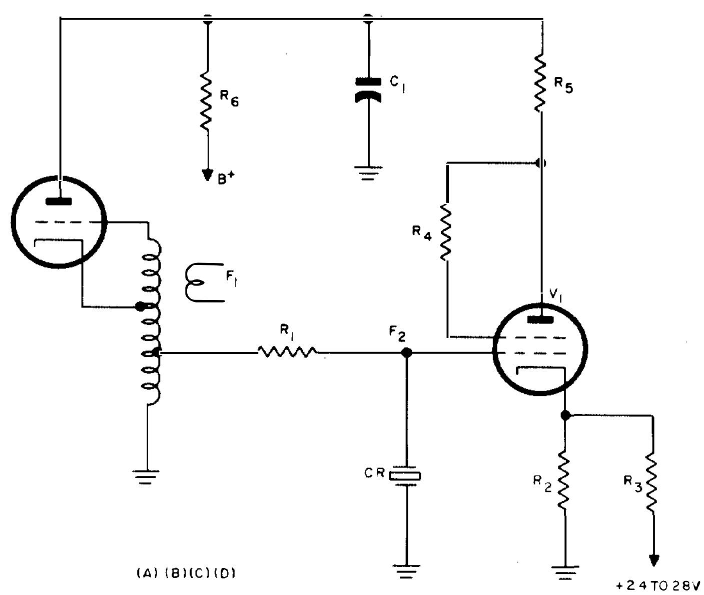


Figure 1-159. Crystal and magic-eye resonance indicator

Fig.	Equipment	F_1	CR	R_1	R_2
(A)	Radio Transmitter BC-696-A	3000- 4000	DC-8-C,- D,-K	5.1	0.39
(B)	Radio Transmitter BC-457-A	4000- 5300	DC-8-C,- D,-K	10	0.39
(C)	Radio Transmitter BC-458-A	5300- 7000	DC-8-C, D,-K	15	0.39
(D)	Radio Transmitter BC-459-A	7000- 9100	DC-8-C, D,-K	5.1	0.39

Circuit Data for Figure 1-159. F in kc. R in kilohms. C in $\mu\mu$f.

SERIES-RESONANT CRYSTAL OSCILLATORS

1-355. For maximum frequency stability it is generally preferable to operate a crystal unit at its series-resonance frequency, but series-mode circuits are most widely used for overtone operation. At series resonance the crystal element appears as a resistance, so that in the normal circuit it can be short-circuited or replaced by a comparable resistance without stopping oscillations. Series-resonant oscillators generally have smaller outputs than do oscillators of the parallel-resonant type. Also, series-resonant oscillators usually require more circuit components, and hence are not often used except in the very-high-frequency range. In general, the operation of the series-mode circuits is less complicated than that of the parallel-mode oscillators. Nevertheless, the circuit design becomes increasingly critical at the higher frequencies and higher overtones. The stray capacitances must be kept to a minimum, and all leads must be as short as possible. It may be necessary to nullify the crystal shunt capacitance, C_o, by connecting across the crystal unit an inductor that is antiresonant with C_o at the operating frequency. It may also be desirable to connect a capacitor in series with the crystal unit, to tune out the stray inductance of the crystal leads. Tuned circuits must be provided if a crystal unit is to be driven at a particular overtone mode. Quite often, satisfactory operation is obtained simply by designing a conventional variable-tuned oscillator to operate at the desired frequency, and then inserting the crystal unit in a plate tank or feed-back circuit. There will be a range of tuning adjustments in which the crystal can assume control and hold the frequency very nearly constant. As the tuning adjustments are varied beyond this range, the control becomes quite unstable or ceases altogether. Usually, the region of stable control becomes smaller as the overtone order is increased. If broad-band operation is desired with no tuning adjustment other than the selector switch for changing the crystal, additional precautions must be taken to ensure that oscillations cannot be maintained except when the crystal impedance is small —that is, the crystal unit is operating near series resonance. For maximum frequency stability, the effective resistance of the circuit facing the crystal unit should be as small as possible. At the higher frequencies, the stray capacitances limit the impedances obtainable from the tuned circuits, thereby making them more selective and hence more effective in influencing the frequency and in increasing the instability.

1-356. The series-mode oscillators most widely recommended are listed in the following table, and rated according to their relative design and performance characteristics. A rating of 1 represents the top relative superiority in the corresponding characteristic. It should be understood that the ratings are based upon average qualitative results which might well be contradicted by the data of individual investigators. Any one of the series-mode circuits expertly designed could surpass the performance of a poorly designed circuit rated higher in a particular characteristic. The frequency-stability rating assumes average oven and voltage regulation.

R_3	R_4	R_5	R_6	C_1	V_1
1.5	1000	51	0.02	50,000	1629
1.0	1000	51	0 02	50,000	1629
1.0	1000	51	0.02	50,000	1629
1 5	1000	51	0 02	50,000	1629

Symbol of Oscillator	A	B	C	D	E	F	G	H	I	J	K	L
Frequency Stability (%)	0.0001	0.002	0.0005	0.0004	0.0004	0.0004	0.0004	0.0002	0.0015	0.002	0.002	0.0015
Power Output	6	5	3	1	3	5	4	4	4	2	1	3
Versatility	4	4	2	1	2	3	2	3	3	3	2	3
Upper Frequency Level	5	1	3	2	2	2	2	3	3	3	4	3
High-Resistance Crystals	2	5	3	4	4	4	2	4	4	4	1	3
Ease of Adjustment	6	7	1	3	2	2	2	5	5	5	4	5
Untuned Bandwidth	5	5	2	2	1	2	1	6	6	6	4	6
Frequency Multiplication	5	5	2	4	4	1	5	5	3	3	5	6
Low Harmonic Output	1	2	2	5	5	5	5	4	4	4	3	2
Circuit Simplicity	4	5	3	2	1	2	2	3	3	3	4	4
Isolation from Load	1	2	3	4	4	2	4	4	4	4	4	4
Low-Frequency Operation	1	6	3	5	5	5	2	4	4	4	1	2

The oscillator symbols in the foregoing table correspond to the respective index letters of the oscillators listed below.

Names of Series-Mode Oscillators

A. Meacham Bridge
B. Capacitance Bridge
C. Butler, or Cathode-Coupled
D. Grounded-Cathode, Transformer-Coupled Type
E. Grounded-Grid, Transformer-Coupled Type
F. Grounded-Plate, Transformer-Coupled Type
G. Transitron
H. Impedance-Inverting Transitron
I. Impedance-Inverting Pierce
J. Impedance-Inverting Miller
K. Grounded-Cathode Two-Stage Feedback
L. Modified Colpitts, C.I. Meter Type

Meacham Bridge Oscillator

1-357. The Meacham bridge oscillator, illustrated in figure 1-160, provides the greatest frequency stability of any vacuum-tube oscillator yet devised, but the region of maximum frequency stability is limited to the lower frequencies because of the increased effect of the stray circuit capacitances when the frequency becomes greater than a few hundred kilocycles per second. The oscillator is of the crystal-stabilized type employing tuned circuits. At frequencies above 1000 kc the effect of the stray capacitance is sufficient to reduce the stability to a point where little is to be gained by the use of the Meacham circuit. The oscillator is principally employed with GT-cut crystals in frequency standards, to generate frequencies of 100 kc. In figure 1-160, it can be seen that if the bridge were perfectly balanced there would be no excitation voltage. At the start of oscillations, the ratio of R_1 (practically equal to the series-arm R of the crystal unit) to R_2 is smaller than the ratio of the R_3 to R_4. But R_4 is a thermistor — it is the resistance of a tungsten lamp which sharply increases in value as the temperature rises. (A semiconductor such as carbon, silicon, or germanium can be used, in which case the resistance will decrease with temperature. The negative temperature coefficient of the semiconductor is generally larger than the positive coefficient of tungsten, but the semiconductor thermistor is more expensive and is much more difficult to duplicate because of its great sensitivity to impurities.) As oscillations build up, the current through R_4 increases to a point where the heat generated from the power losses raises the the thermistor temperature, and hence the resistance, to a point where the bridge is almost balanced. Equilibrium is reached when the imbalance of the bridge is just sufficient to supply heat to the thermistor at the same rate at which it escapes. For maximum amplitude stability, the ambient temperature of R_4 should not be permitted to vary over a wide range. Normally, the tungsten lamp will heat to a dull red of approximately 600 degrees centigrade. A variation of over 100 degrees in the ambient temperature could have a significant effect on the equilibrium power losses in the thermistor, if extreme precision were desired. The operating temperature of the lamp is very low compared with the rated temperature, and consequently the lamp can be expected to last indefinitely. The oscillator should be designed and adjusted so that the phase shift occurs entirely in the bridge. That is, the tube should operate into

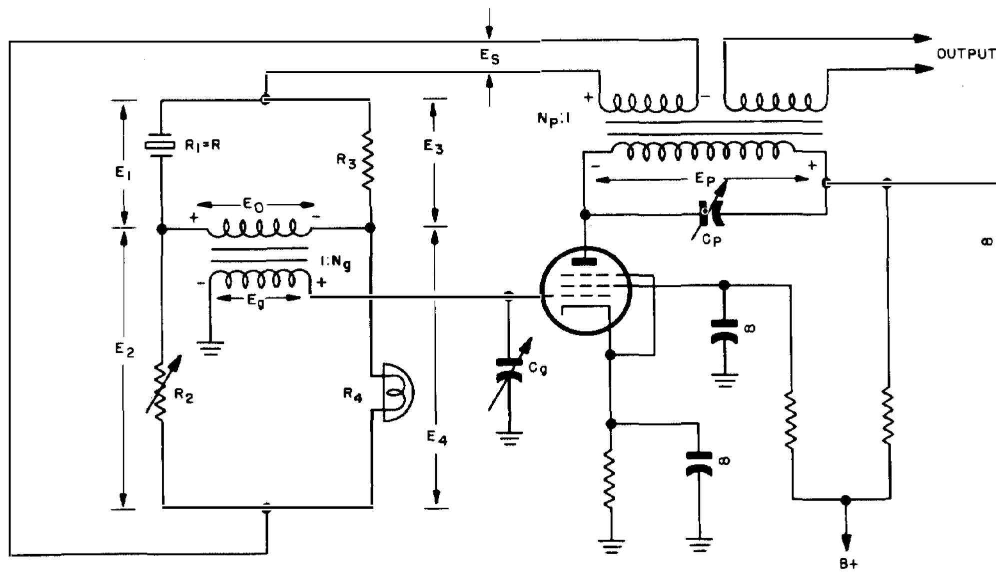


Figure 1-160. Basic circuit of Meacham bridge-stabilized oscillator

a pure resistance, so that the instant the plate current is maximum the peak transformer voltages should occur with polarities as indicated in figure 1-160. Transformers having powdered-iron, toroidal cores can provide a coefficient of coupling very close to unity in the low-frequency range. The following analysis of the frequency stability and the activity stability of the Meacham oscillator, except for minor deviations and extensions, has been guided by the postulates and basic considerations as presented by W. A. Edson.*

FREQUENCY STABILITY OF MEACHAM BRIDGE OSCILLATOR

1-358. First, we shall assume that the vacuum tube in figure 1-160 operates into a purely resistive load, and that the entire phase reversal takes place in the bridge transformer. The phasor diagram in figure 1-161 (A) shows the relation of the voltage E_o to the other voltages of the bridge network. (Refer to figure 1-160 for voltage symbols.) Next, assume that some change in the capacitance of the circuit requires that E_o be shifted in phase by a very small angle equal to ϕ, but that the change is so small that the magnitude of all the bridge voltages can be assumed to remain constant. In order to produce the phase shift ϕ, it can be seen that E_2, and hence the current through R_2, must be rotated by an angle θ. In a triangle with angles A and B opposite to sides a and b, respectively, $\frac{a}{\sin A} = \frac{b}{\sin B}$ (Law of Sines). Likewise, in the triangle $E_oE_2E_4$,

$$E_o/\sin\theta = E_4/\sin\phi$$

or

$$\sin\theta = \frac{E_o}{E_4}\sin\phi \qquad 1\text{—}358\ (1)$$

Since we are assuming that both θ and ϕ are very small, equation (1) can be written, approximately,

$$\theta = \frac{E_o}{E_4}\phi \qquad 1\text{—}358\ (2)$$

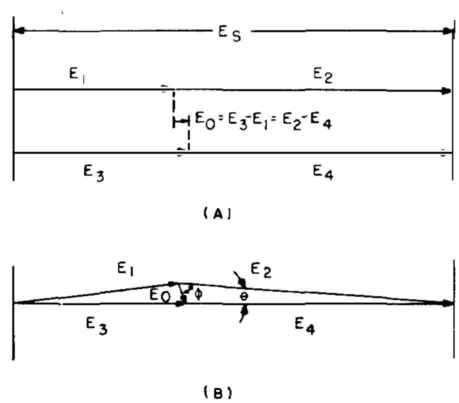


Figure 1-161. Phasor diagrams of bridge voltages in Meacham oscillator. Angles ϕ and θ (greatly magnified) represent small shifts in phase when crystal unit is operating slightly off series resonance

* *Vacuum Tube Oscillators*, John Wiley and Sons, 1953.

The current through the input transformer can be assumed to be negligible compared with the total current through the crystal, so the current through R_2 is essentially the same as that through the crystal. Under these conditions, as is shown in paragraph 1-241,

$$\theta \approx \tan\theta = \frac{X_e}{R_1 + R_2} = \frac{2L\Delta\omega}{R + R_2} \qquad 1\text{—}358\ (3)$$

where L and R are series-arm parameters of the crystal unit. By rearranging equation (3) and dividing by ω, the fractional change in frequency required to produce an angle θ is found to be

$$\frac{\Delta\omega}{\omega} = \frac{\theta\,(R + R_2)}{2\omega L} = \frac{\theta\, R_c\sqrt{C}}{2\sqrt{L}} \qquad 1\text{—}358\ (4)$$

where $R_c = R + R_2$ is the total resistance of the crystal side of the bridge, and C is the series-arm capacitance. On substitution of equation (2) in equation (4), it is found that

$$\frac{\Delta\omega}{\omega} = \frac{\phi\, E_o\, R_c}{2E_4} \cdot \sqrt{\frac{C}{L}} \qquad 1\text{—}358\ (5)$$

Equation (5) indicates that the more nearly balanced the bridge (the smaller the E_o/E_4 ratio), the greater will be the frequency stability.

Now,

$$E_4 - \frac{R_4\, E_s}{R_3 + R_4} \qquad 1\text{—}358\ (6)$$

Letting $(R_3 + R_4)/R_4 \approx (R_1 + R_2)/R_2 = m$, we see that $E_o/E_4 = \dfrac{m\, E_o}{E_s}$.

Also, since

$$E_o = E_s\left(\frac{R_3}{R_3 + R_4} - \frac{R_1}{R_1 + R_2}\right)$$

$$= E_s\left(\frac{R_3}{R_T} - \frac{R_1}{R_c}\right) \qquad 1\text{—}358\ (7)$$

where $R_T = R_3 + R_4$, then

$$E_o/E_4 = m\left(\frac{R_3}{R_T} - \frac{R_1}{R_c}\right) = R_3/R_4 - R_1/R_2$$

On substitution in equation (5), we have

$$\frac{\Delta\omega}{\omega} = \left(\frac{R_2\, R_3 - R_1\, R_4}{R_2\, R_4}\right)\frac{R_c\,\phi}{2\sqrt{L/C}} \qquad 1\text{—}358\ (8)$$

It now remains to determine the magnitude of ϕ for a small change in the capacitance of the circuit. The most likely changes in capacitance take place in the grid circuit, the average ΔC_g usually being on the order of 10 times the average ΔC_p in the plate circuit. Looking away from the grid, it can be seen that when the bridge is very nearly balanced, the grid faces a resistive impedance

$$R_g = \frac{N_g^2\,(R_1 + R_3)\,(R_2 + R_4)}{R_1 + R_2 + R_3 + R_4} \qquad 1\text{—}358\ (9)$$

The capacitance C_g will have been adjusted to be effectively antiresonant with the leakage inductance of the transformer, which inductance can be imagined to be in parallel with R_g. Still looking away from the grid, we can imagine a generator connected between grid and cathode. If the capacitance should change by a small amount ΔC_g, the ratio of the excess reactive component of current to the resistive component becomes $R_g\omega\Delta C_g$. This will equal the tangent of the phase shift, ϕ, which is sufficiently small for tan ϕ to be assumed to equal ϕ. Thus, with $\phi = R_g\,\Delta C_g$, on substituting equation (9) for R_g, equation (8) becomes

$$\frac{\Delta\omega}{\omega} = \left(\frac{R_2\, R_3 - R_1\, R_4}{R_2\, R_4}\right) \cdot \frac{(R_1 + R_3)\,(R_2 + R_4)}{(R_1 + R_2 + R_3 + R_4)}$$

$$\cdot\ \frac{R_c\, N_g^2\, \Delta C_g}{2\,L} \qquad 1\text{—}358\ (10)$$

Now, let us assume that R_1 and R_3 remain fixed. As R_2 is varied, R_4 must vary in direct proportion to keep the bridge balanced. If $R_3 = kR_1$, R_4 will always approximately equal kR_2. Substituting these values of R_3 and R_4 in all terms where the error introduced can be considered negligible, equation (10) becomes

$$\frac{\Delta\omega}{\omega} = \frac{R_1^2\,(kR_2 - R_4)\,(1 + k)}{kR_2} \cdot \frac{N_g^2\,\Delta C_g}{2L}$$

$$1\text{—}358\ (11)$$

If R_4 is expressed as being equal to R_2 (k − i), equation (11) becomes

$$\frac{\Delta\omega}{\omega} = \frac{(1 + k)\, i\, R_1^2\, N_g^2\, \Delta C_g}{2kL} \qquad 1\text{—}358\ (12)$$

1-359. Equation 1—358 (12) does not quite indicate the relations among all the circuit parameters that are effective in providing an optimum frequency stability. It is first necessary to determine how i, which is a measure of the imbalance in the circuit, is dependent upon the other parameters. For this purpose, it is necessary to find that value of i which must exist in order for the feed-back voltage to be at equilibrium. It will be assumed that the r-f plate current, I_p, is equal to $g_m E_g$. If this assumption is not warranted, g_m in any of the following equations can be replaced by $\mu/(R_p+Z_p)$. To a first approximation,

$$Z_p = N_p^2 R_c R_T/(R_c + R_T) \qquad 1\text{—}359\ (1)$$

and

$$E_s = E_p/N_p = I_p Z_p/N_p = g_m E_g N_p R_c R_T/(R_c + R_T) \qquad 1\text{—}359\ (2)$$

also, $E_g = N_g E_o$, and E_o is given by equation 1—358(7). On substitution in equation (2)

$$1 = g_m N_p N_g (R_c R_3 - R_1 R_T)/(R_c + R_T) \qquad 1\text{—}359\ (3)$$

Equation (3) is the equilibrium feed-back equation for the Meacham oscillator. On expressing R_3 and R_4 as functions of R_1 and R_2, it is found that at equilibrium

$$i = \frac{(1 + k) R_c}{g_m N_p N_g R_1 R_2} \qquad 1\text{—}359\ (4)$$

Substituting (4) in equation 1—358 (12)

$$\frac{\Delta\omega}{\omega} = \frac{(1 + k)^2 R_1 R_c N_g \Delta C_g}{2 k g_m R_2 N_p L} \qquad 1\text{—}359\ (5)$$

In a similar manner, in equating ϕ to $Z_p \omega \Delta C_p$, it can be shown that for small changes in the plate capacitance

$$\frac{\Delta\omega}{\omega} = \frac{R_c^3 N_p \Delta C_p}{2 (1 + k) g_m R_2^2 N_g L} \qquad 1\text{—}359\ (6)$$

It can be seen that greater stability is to be had when g_m is a maximum and when the ratio $m = R_c/R_2$ is small. For changes in C_g, the optimum value of k is 1 (when $(1 + k)^2/k$ passes through a minimum). For changes in C_p, it would be desirable to have k as large as practicable. A further consideration is to so proportion the parameters that the expected variations in C_g and C_p will have the maximum opportunity to cancel in their effects. Also, caution must be taken that in improving the stability in one respect, it is not impaired to a greater extent in another. Since the expected ΔC_g is approximately 10 times the expected ΔC_p, the ratio of the right-hand sides of equations (5) and (6) can be equated to 1, with 10 ΔC_p substituted for ΔC_g. On thus dividing (5) by (6)

$$1 = \frac{10 (1 + k)^3 R_1 R_2 N_g^2}{k R_c^2 N_p^2}$$

or

$$N_p^2/N_g^2 = \frac{10 (1 + k)^3 R_1}{k m R_c} = \frac{10 (1 + k)^3 (m - 1)}{k m^2} \qquad 1\text{—}359\ (7)$$

With the oscillator designed according to equation (7), average capacitance variations in the plate and grid circuits will have approximately equal effects upon the frequency. When the square root of equation (7) is combined with equations (5) and (6), and R_c is expressed as $mR_1/(m - 1)$,

$$\frac{\Delta\omega}{\omega} = \frac{\Delta C_g m^2 R_1}{2 L g_m}\sqrt{\frac{1 + k}{10 k (m - 1)}} \qquad 1\text{—}359\ (8)$$

and

$$\frac{\Delta\omega}{\omega} = \frac{\Delta C_p m^2 R_1}{2 L g_m}\sqrt{\frac{10 (1 + k)}{k (m - 1)}} \qquad 1\text{—}359\ (9)$$

If the oscillator is to be designed on the basis of equation (7), it can be seen that k should be made as large as is practicable, and m should be such that the factor $m^2/\sqrt{m - 1}$ is a minimum. It can be shown that this occurs when

$$m = 4/3 \qquad 1\text{—}359\ (10)$$

Frequency stability, of course, is not the only consideration; there are also the vacuum-tube and thermistor characteristics and the power rating of the crystal unit that must be taken into account in deciding upon the optimum parameter relations. Remember, that in equations (4), (5), and (6), g_m can be replaced by the more exact term $\mu/(R_p + Z_p)$. Certainly, a high-mu tube is to be preferred, and when it is operated class A the second-harmonic output can be expected to be at least 65 db below the fundamental. The screen voltage should be fairly high, in order to increase g_m. Normal operating voltages can be employed, but E_g should not be allowed to drive the grid positive.

ACTIVITY STABILITY OF MEACHAM BRIDGE OSCILLATOR

1-360. Starting with E_o, the input to the grid transformer, we see that at equilibrium the product of the gains of all the stages, from E_o back to E_o, must be equal to 1. Thus,

$$\text{Gain:}\quad G_1\,G_2\,G_3\,G_4 = N_g \times \frac{\mu Z_p}{R_p + Z_p} \times 1/N_p \times \left(\frac{R_3}{R_T} - \frac{R_1}{R_c}\right) = 1$$

$$\text{Voltage:}\quad E_o \xrightarrow{G_1 = E_g/E_o} E_g \xrightarrow{G_2 = E_p/E_g} E_p \xrightarrow{G_3 = E_s/E_p} E_s \xrightarrow{G_4 = E_o/E_s} E_o$$

From equation 1—358 (12) it can be seen that in the interest of frequency stability, i, and hence the imbalance of the bridge should be as small as possible. Fortunately, this condition also agrees with the requirements of high activity stability, for the smaller the *difference* of the actual thermistor resistance from a value equal to kR_2, the larger will be the percentage change in that *difference* for a small change in the thermistor voltage. Equation 1—358 (7) can be written

$$E_o = G_4\,E_s = i\,R_1\,R_2\,E_s/k\,R_c^2 \qquad 1\text{—}360\ (1)$$

where $i = (kR_2 - R_4)/R_2$ and G_4 is the gain of the stage. In the over-all gain equation, above, G_1 and G_3 can be considered constant, so that G_4 primarily has the function of compensating any changes in G_2 of the vacuum tube. From equation (1), it can be seen that iE_s/E_o can be considered a constant. Or, in the over-all gain equation we see that

$$G_1\,G_2\,G_3\,i\,R_1\,R_2/k\,R_c^2 = 1 \qquad 1\text{—}360\ (2)$$

or that

$$G_2\,i = k\,R_c^2\,N_p/R_1\,R_2\,N_g = \text{constant}$$

On differentiating,

$$i\,dG_2 + G_2\,di = 0$$

or

$$\frac{di}{i} = -\frac{dG_2}{G_2} \qquad 1\text{—}360\ (3)$$

Equation (3) shows that for a given percentage change in the gain of the tube, the smaller the value of i, the smaller need be the change in R_4 to restore equilibrium. Since $G_2 \approx g_m Z_p$, we can write

$$di/i = -\,dg_m/g_m \qquad 1\text{—}360\ (4)$$

On differentiating $iE_s/E_o = C$, where C is a constant, we find that

$$i\,dE_s + E_s\,di - C\,dE_o = 0 \qquad 1\text{—}360\ (5)$$

If the effects on E_o due to the changes in E_s and i exactly cancel so that $dE_o = 0$, then, by equations (3) and (5)

$$-\,dG_2/G_2 = di/i = -\,dE_s/E_s$$

Under these circumstances it can be seen that the percentage change in the activity is exactly equal to the change in the gain of the tube. If the thermistor is to be effective in preventing the amplitude of the output from changing significantly with changes in G_2, clearly an increase in E_s must produce a *decrease* in E_o. We can define the activity sensitivity of the bridge to be

$$s = -\,E_s\,di/i\,dE_s \qquad 1\text{—}360\ (6)$$

The sensitivity is thus defined as the percentage variation in i per percentage variation in the voltage across the bridge. The problem now is to convert equation (6) into a function (equation 14) of the circuit constants so that s can be predetermined by the design engineer. From equations (5) and (6) we find that the percentage change in E_o per percentage change in E_s is

$$-\frac{E_s\,dE_o}{E_o\,dE_s} = s - 1 \qquad 1\text{—}360\ (7)$$

In practice, E_o/E_s (= G_4) can be on the order of 0.003 or smaller; so, if the change in E_o is comparable to that of E_s in magnitude, excellent amplitude stability will be achieved. The stability depends first upon the magnitude of i, and secondly, upon the sensitivity of the thermistor. The latter is defined as

$$S = \frac{E_4\,dR_4}{R_4\,dE_4} = \frac{d(\log R_4)}{d(\log E_4)} \qquad 1\text{—}360\ (8)$$

Figure 1-162 shows the resistance-voltage characteristics of a number of tungsten lamps for ambient temperatures at room values. For lower ambient temperatures, the curves would be shifted to the right somewhat, and for higher temperatures, to the left. Since the curves are plotted on log paper, according to equation (8) it can be seen

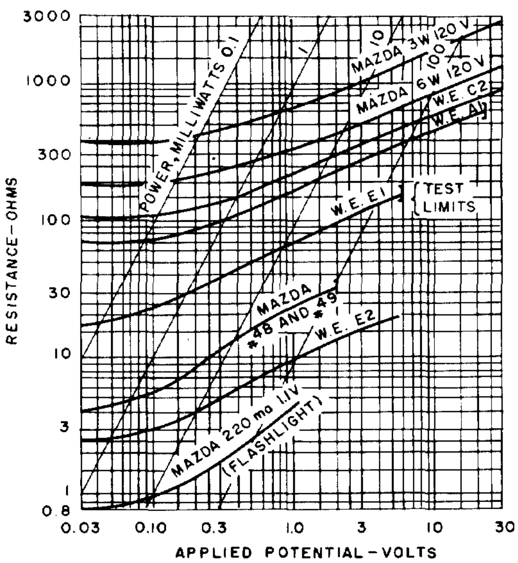


Figure 1-162. Resistance of typical tungsten lamps versus applied voltage and power dissipation when the ambient temperature is 300° Kelvin scale (approximately 27° C)

that the thermistor sensitivity S at a given value of E_4 is the actual slope of a curve at that point. It is important, of course, to operate the thermistor at a voltage where the slope approaches a maximum. It is convenient to express the bridge s as a function of the thermistor S.

Since $i = \frac{k\,R_2 - R_4}{R_2}$, then

$$di = -\frac{dR_4}{R_2} \approx -\frac{k\,dR_4}{R_4} \qquad 1\text{—}360\ (9)$$

or

$$s = -\frac{E_s di}{i\,dE_s} = \frac{k\,E_s\,dR_4}{i\,R_4\,dE_s} = \frac{kS}{i}\left(\frac{E_s\,dE_4}{E_4\,dE_s}\right) \qquad 1\text{—}360\ (10)$$

Now, $E_4 = E_2 - E_o \approx E_2$, but $dE_4\ (= dE_2 - dE_o)$ is not approximately equal to dE_2. So $dE_4/E_4 = (dE_2 - dE_o)/E_2 = (dE_s - m dE_o)/E_s$, where $m = \frac{R_c}{R_2} \approx \frac{E_s}{E_2}$. On substitution for dE_4/E_4 in equation (10), we find

$$s = \frac{kS}{i}\left(1 - \frac{m dE_o}{dE_s}\right) \qquad 1\text{—}360\ (11)$$

By equation (7) we see that $dE_o/dE_s = E_o(1 - s)/E_s$. On substitution in (11) and after rearranging, we have

$$s = \frac{kSE_s - mkSE_o}{iE_s - mkSE_o} \qquad 1\text{—}360\ (12)$$

The term $mkSE_o$ in the numerator can be considered negligible, and dropped. After expressing E_o in terms of equation (1) and rearranging, it is found that

$$s = kSR_c/i(R_c - SR_1) \qquad 1\text{—}360\ (13)$$

Finally, on substituting for i its value given by equation 1—359(4), we are able to express the activity sensitivity entirely in terms of the known circuit parameters. Thus,

$$s = \frac{kSg_mN_pN_gR_1R_2}{(1 + k)\ (R_c - SR_1)}$$

or

$$s = kSg_mN_pN_gR_1/(1 + k)\ [m - S(m - 1)] \qquad 1\text{—}360\ (14)$$

The reciprocal of s can be considered the percentage gain in the output voltage (or in E_p or E_s) for a unit percentage change in the gain of the tube, since dG_2/G_2 is equal to $-di/i$. In the equation for s, note that if the thermistor sensitivity were equal to R_c/R_1, the stability mathematically would be infinite. Since R_c/R_1 is greater than 1, a single tungsten lamp could not provide the thermistor sensitivity for the above condition to hold unless special measures were available to reduce the heat leakage from the filament. The effective sensitivity could be increased if R_1 were replaced by another tungsten lamp, and the crystal unit were inserted in the place of R_2. Theoretically, the sensitivity can be made much larger than unity simply by varying the ambient temperature together with the operating temperature of the filament; for instance, by constructing a thermistor with the filament mounted inside a heater sleeve and controlling the heater current by feedback from a later amplifier stage. If equation (14) is taken apart, it will be found that the denominator term, $(R_c - SR_1)$, originates from that component of dE_4 that is equal to $-dE_o$. When there is an increase in E_s, the voltage E_4 changes in two ways: one is due to the change in the current through R_4, and the other is due to the increase in the resistance, itself. It is the latter component that is approximately equal to $-dE_o$. Mathematically, the change in E_4 is expressed by the differential equation

$$dE_4 = d(I_4\,R_4) = R_4\,dI_4 + I_4\,dR_4 \qquad 1\text{—}360\ (15)$$

Since dR_4/R_4 is equal to SdE_4/E_4, on substitution in equation (15) it can be shown that

$$dR_4 = S\left(\frac{R_4\,dI_4}{I_4} + dR_4\right) \qquad 1\text{—}360\ (16)$$

If S is greater than 1, an increase in voltage across R_4 must result in a decrease in current. (Incidentally, since the change in R_4 is actually due to a change in the temperature brought about by an increase in power, a value of S greater than unity implies that the percentage increase in resistance is at least equal to twice the percentage decrease in current.) Now, assuming that the current through the input transformer is negligible, $E_s = I_4R_T$, and $dE_s = I_4dR_4 + R_TdI_4$, where $dR_4 = dR_T$. If E_s is to remain constant, that is, if dE_s is to equal zero for a small change in the gain of the tube, dI_4/I_4 must equal $-\frac{dR_4}{R_T}$. If the latter value is substituted in equation (16), it will be found that for conditions of $s = \infty$:

$$S = \frac{R_T}{R_T - R_4} = \frac{R_T}{R_3} = \frac{R_c}{R_1} \qquad 1\text{—}360\ (17)$$

This is the explanation of the term $(R_c - SR_1)$ in the denominator of equation (14). Other than the assumption that the changes in the current I_o through the grid transformer can be considered negligible in their effect upon dE_4, the term is entirely a function of the R_3 and R_4 arms of the bridge, and is not related to the gain characteristics of the rest of the circuit. No experimental data is available concerning the operation of the Meacham bridge oscillator with values of S greater than unity, when R_4 behaves as a negative resistance (an increase in E_4 is accompanied by a decrease in I_4). Theoretically, if S were greater than R_c/R_1, an increase in the g_m of the tube would ultimately result in a decrease in the output voltage and in the voltage applied across the bridge. In an actual circuit, whether stable values of R_4 would be maintained under such conditions is open to question. Perhaps the thermal lag of the filament and the extreme sensitivity of E_o would so influence the operation that R_4 would periodically overshoot its mark and prevent an unmodulated equilibrium from being reached. In practice, the values of S will be on the order 0.5, so such considerations do not arise. For s to be as large as possible, referring to equation (14), it can be seen that $[k/(1 + k)]$ should be as large as practicable. This agrees with the equations for frequency stability if the circuit is to be designed according to equation 1—359(7). The term $[k/(1 + k)]$ has no maximum, but approaches unity as k increases indefinitely. Assume that s is equal to 50. This means that a change in the gain of the tube of 1 per cent will cause a change of only one-fiftieth of 1 per cent in the output voltage. Or in terms of db, since

$$s = -\frac{E_s}{dE_s}\cdot\frac{di}{i} \approx -\frac{\Delta(\log i)}{\Delta(\log E_s)}$$

$$= \frac{\Delta(\log G_2)}{\Delta(\log E_p)} = \frac{\Delta\text{db in tube gain}}{\Delta\text{db in output}} \qquad 1\text{—}360\ (18)$$

an increase of 0.5 db in the gain of the tube will cause only a 0.01-db increase in the output.

CRYSTAL DRIVE LEVEL CONSIDERATIONS IN MEACHAM BRIDGE OSCILLATOR

1-361. A starting consideration in the design of a Meacham bridge oscillator is that the crystal unit is not to be overdriven. If P_1 is the crystal power, $I_c = \sqrt{P_1/R_1}$ is the crystal current, and

$$E_s = I_c\,R_c = \frac{mR_1}{m-1}\sqrt{\frac{P_1}{R_1}} = \frac{m\sqrt{P_1\,R_1}}{m-1} \qquad 1\text{—}361\ (1)$$

With R_1 determined by the crystal unit, it is desirable, from the point of view of frequency stability, for m, and hence R_2 to have small optimum values. If S approaches unity, the small R_2 will also be an important consideration in activity stability, but for normal values of S the activity stability is improved slightly if R_2 is large. The term $\left(\frac{R_2}{R_c - SR_1}\right)$ in equation 1—360 (14) has no maximum, but approaches unity as R_2 is increased indefinitely and R_1 and S are held constant. Usually, the requirements of frequency stability are the more important, and R_2 should be kept as small as practical thermistor resistances and values of k permit. At low frequencies, values of R_1 may be in the neighborhood of 1000 ohms or more. The voltage across the thermistor will be

$$E_4 = \frac{E_s}{m} = \frac{I_c\,R_1}{m-1} = \frac{E_1}{m-1} = \frac{\sqrt{P_1\,R_1}}{m-1} \qquad 1\text{—}361\ (2)$$

where E_1 is the voltage across the crystal unit. For convenience, we repeat equation 1—360 (1), but expressed as a function of m and k:

$$E_o = (m-1)\,i\,E_s/km^2 \qquad 1\text{—}361\ (3)$$

The power dissipation in R_4 is

$$P_4 = E_4 I_4 = E_1 I_c/(m - 1)\,k = P_1/(m - 1)\,k \qquad 1\text{—}361\ (4)$$

The impedance of the bridge in terms of R_1 is

$$Z_s = R_c R_T/(R_c + R_T) = kmR_1/(1 + k)(m - 1) \qquad 1\text{—}361\ (5)$$

The plate impedance of the tube is

$$Z_p = N_p^2 Z_s = \frac{N_p^2\, kmR_1}{(1 + k)(m - 1)} \qquad 1\text{—}361\ (6)$$

The plate voltage is

$$E_p = I_p Z_p = \frac{g_m E_g N_p^2\, k\, m\, R_1}{(1 + k)(m - 1)} \qquad 1\text{—}361\ (7)$$

Also,

$$E_p = N_p E_s = mN_p\sqrt{P_1 R_1}\,/(m - 1) \qquad 1\text{—}361\ (8)$$

and

$$E_g = N_g E_o \qquad 1\text{—}361\ (9)$$

Finally, we repeat equation 1—359 (3), the over-all equation for feed-back equilibrium, but expressed in terms of R_1, m, and k:

$$G_1 G_2 G_3 G_4 = \frac{g_m N_p N_g\, i\, R_1}{m(1 + k)} \qquad 1\text{—}361\ (10)$$

R_2 can be adjusted to provide the same value of m for each different crystal unit. Under these circumstances, E_s and E_p will be the same in each oscillator, even though R_1 varies. Two fundamental problems are that the design must ensure that the crystal current does not overdrive the crystal unit when R_1 is small, and that the thermistor current is sufficient for S to be a maximum.

DESIGN PROCEDURE FOR MEACHAM BRIDGE OSCILLATOR

1-362. *The fixed point of reference for estimating the current and voltage at any point in the Meacham circuit is the thermistor voltage E_4. This is the voltage that is required to make $R_4 = R_3/(m - 1)$. If R_3 and m are held constant, E_4 as well as E_s $(= mE_4)$ and E_p $(= N_pE_s)$ will also be constant.* If P_{cm} is the rated crystal power, R_m is the maximum series resistance of the crystal unit, and $\frac{R_m}{N}$ is the minimum expected resistance of the crystal unit, then, by equation 1—361 (2), E_4 must not be greater than the value

$$(\max)\ E_4 = \frac{\sqrt{P_{cm} R_m}}{(m - 1)\sqrt{N}} \qquad 1\text{—}362\ (1)$$

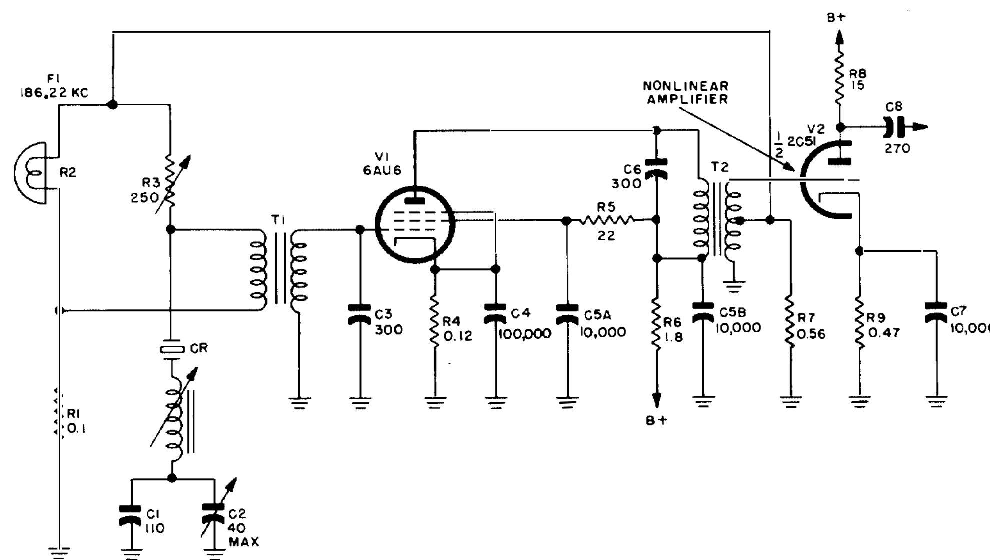
Since the Meacham oscillator is most applicable for use in the low-frequency range where crystal units having resistances in the neighborhood one or more thousand ohms are not uncommon, the risk is greatly increased that an exceptionally well-mounted crystal will have a resistance of as little, as, perhaps, $R_m/25$. Also, since the Meacham circuit is primarily useful as a precision oscillator, an additional safety factor should be allowed to prevent the crystal unit from being driven beyond its test specifications. For these reasons, it is suggested that in the absence of prior experience or manufacturer's recommendations for a given type of crystal unit, the Meacham design for frequencies below 200 kc assume a minimum R of $R_m/25$, rather than $R_m/9$ as was assumed in the case of the parallel-resonant oscillator design. However, it can still be assumed that the most probable crystal unit will have an $R = R_m/3$. If the crystal unit to be used is a precision GT cut, a safety factor as large as $N = 25$ need not be made. In any event, crystal units having resistances less than $R_m/9$ can be expected to be extremely rare, and if $N = 9$ is considered a sufficient safety factor, an output voltage two-thirds greater can be realized than if N is assumed to be 25. A crystal unit having a resistance less than $R_m/9$ would be driven beyond its test level, but far below a level that could damage the crystal. Since the resistance is already low, an increase in resistance with overdrive would do more good than harm. The only concern is that the frequency of a borderline crystal may deviate beyond the tolerance limits. Such a risk could be checked during a production test, but would subtract from the reliability of crystal replacements in the field. In equation 1—361 (10), it can be seen that when k is a minimum (when $R_1 = R_m$), the imbalance, as measured by i, is a minimum. When k is large, the percentage changes in $(k + 1)$ and in R_1 are very nearly equal, so that the imbalance tends to vary as the square of R_1. k should be chosen for maximum frequency stability under variations of C_g, assuming that the crystal unit resistance is its most probable value (approximately $R_m/3$). According to equation 1—359 (5), with all else fixed, the percentage change in frequency is a minimum when $k = 1$. The most probable optimum value of k, therefore, fixes R_3 as equal to $R_m/3$. Thus, for any random value of R_1,

$$k = R_m/3R_1 \qquad 1\text{—}362\ (2)$$

Next, a value of m equal to 4/3 (see equation 1—359 (10)) should be chosen, to provide maximum frequency stability on the assumption that equations 1—359 (8) and (9) are to apply when R_1 is its most probable value. After this is done, a safety factor of N should be selected, and the maximum value of E_4 should be determined by equation (1), such that it will not require a bridge voltage sufficient to overdrive the crystal unit when R_1 is equal to R_m/N. Next, a thermistor is chosen that will provide a value of $R_4 = \frac{R_3}{m - 1}$ when E_1 is equal to, or less than, the maximum value determined above. Next, the ratio N_p/N_g can be determined, using equation 1—359 (7) with the assumption that k = 1 and m = 4/3. This gives a ratio $\frac{N_p}{N_g} = \sqrt{15} \approx 4$, which value thus provides the greatest probability that random changes in C_g and C_p can cancel when R_1 is its most probable value. The next step is to select a tube with high class-A g_m and R_p. A 6AC7 would be very satisfactory. Using equation 1—361 (6), determine N_p on the assumption that $Z_p = \frac{R_p}{10}$ when $R_1 = \frac{R_m}{3}$. Now, N_g can be made equal to $N_p/4$. R_2, of course, must be variable over a percentage range comparable to that to be expected from the crystal unit. Normal tube voltages are used. The other circuit components can be determined according to the tube specifications for class-A operation and the special output requirements of the oscillator. E_b, E_g, E_o, I_b, I_c, etc can be determined from the equations in paragraph 1-361, the frequency stability from equations 1—359 (5) and (6), and the activity stability from equation 1—360 (14) for maximum, most probable, and minimum values of R_1.

MODIFICATIONS OF MEACHAM BRIDGE DESIGN

1-363. Two designs of the Meacham bridge stabilized oscillator are shown in figure 1-163. In each



(A)

Figure 1-163. Meacham bridge-stabilized oscillators

Fig.	Equipment	Purpose	F_1	CR	R_1	R_2	R_3	R_4
(A)	Control-Monitor IP-68/CPN-2A	Timing osc. Controls indicator sweep freq and prr of shoran station	186.22	Oven controlled	0 1	Thermistor	250	0 12
(B)	Radio Set AN/FRC-10	Carrier osc and phase-shift circuit	100	WECo D-163897 or D-169649	0 1	0.1	0.5	2

Circuit Data for Figure 1-163. F in kc. R in kilohms. C in $\mu\mu$f. L in μh.

of these circuits inductor-capacitor combination has been connected in series with the crystal unit. Obviously the combination is intended to be resonant at the crystal frequency. The variable arrangement shown in figure 1-163 (A) permits the frequency to be pulled to a more exact value if desired, the crystal unit (if necessary) operating with a reactive component in its impedance. Or, in case the tube operates into a partly reactive load, the tuning elements in the bridge could permit the crystal, itself, to operate at exactly series resonance. The series inductor and capacitor are effective in aiding the initial build up of oscillations and in ensuring that the crystal assumes control at the frequency of the desired mode. It can also be presumed that the LC combination in the bridge improves the waveform somewhat and reduces the small distortion introduced by the tungsten lamp. This distortion is due to the fact that the filament cools at least to some extent during the time that the current alternates from its effective value in one direction to its effective value in the opposite direction. At frequencies above 100 cycles per second this distortion in the waveform is not serious. At radio frequencies it is normally small compared with the distortion introduced by the tube. The resistance R_7 in figure 1-163 (A) appears to be inserted in order to maintain a constant tube load by minimizing the variations in the bridge impedance due to adjustments and to crystal units having difference resistances. In figure 1-163 (B) note that the crystal unit is grounded. This is the usual arrangement. The parallel primary windings of the grid transformer in the same figure suggest that the arrangement is an adaption of a readily available transformer, very probably of the same construction as the one in the plate circuit. The parallel primary connection is in the direction of phase addition. Because the near-unity coupling between the coils effectively doubles the

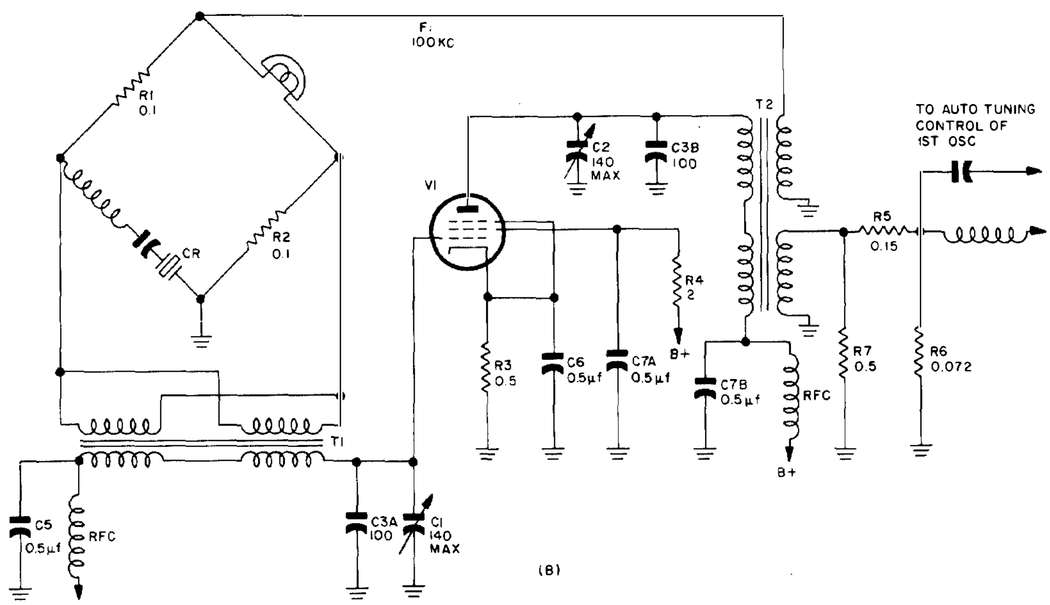


Figure 1-163. Continued

R_5	R_6	R_7	C_1	C_2	C_3	C_4	C_5	C_6	C_7	V_1	V_2
22	1.8	0.56	110	40	300	100,000	10,000	300	10,000	6AU6	½ 2C51
0.15	0.072	0.5	140	140	100	100	500,000	500,000	500,000	WECo 337A	

inductance of each, the parallel connection provides the same step-up arrangement and primary impedance that would be provided by only one of the coils if used alone, but with a reduction in the leakage inductance. A Meacham bridge-stabilized oscillator can be designed employing two or more tubes. On the average, slightly better frequency stability can be achieved with a two-tube circuit, but only in rare instances are the additional cost, space, and weight requirements worth the small improvement in performance. Perhaps, at frequencies in the neighborhood 1000 kc or higher the two-tube arrangement could be more profitable than the one-tube stage. The design of a multistage bridge oscillator can be practically the same as that of the one-stage circuit except that the tube gain, G_2, is replaced by $G_{21} G_{22} \ldots G_{2n}$, where G_{2k} is the voltage gain of a transformation stage between the output and input of the bridge, and where n is the total number of such transformations. By increasing the number of positive-db stages, the bridge i can be made as small as desired, and the frequency and activity stability will be increased in proportion to the gain. It is because the possible gain is unlimited for all practical purposes that the Meacham oscillator represents the ultimate in precision control of the frequency. In the final analysis the limiting condition is the degree to which the crystal parameters, themselves, can be kept constant. Figure 1-164 shows the basic circuit of a two-tube Meacham oscillator that employs no transformers and offers the advantage of only a single tuned stage. Although the design equations are somewhat different from those of the conventional one-tube stage, the same basic approach is to be employed, and the problems to be encountered can be solved similarly to those of the transformer-coupled circuits.

Capacitance-Bridge Oscillators

1-364. Capacitance-bridge oscillators may possibly prove suitable for use in the v-h-f range. Their advantage lies in the fact that a properly balanced capacitance bridge cannot provide sufficient feedback of the proper phase to sustain oscillations at any frequency other than the tuned frequency of the circuit, *provided a crystal unit is connected in the circuit that has a resonant mode of vibration at the tuned frequency.* A properly balanced capacitance-bridge oscillator is thus crystal-controlled, rather than crystal-stabilized. On the other hand, if the bridge is not balanced, the circuit can operate as a free-running oscillator, which may or may not be crystal-stabilized. For the purpose of ensuring operation of crystal units at designated very high harmonic modes, the capacitance bridge, if not the most dependable, is at least as dependable as any other so far tested. The principal disadvantage of this type of circuit is that rather critical tuning adjustments must be made, and one crystal unit cannot replace another unless these adjustments are repeated. Largely on this account the circuit is not to be preferred for frequencies below 50 mc, and perhaps not below 75 mc. Nevertheless, once the bridge is properly adjusted, the operation with a crystal unit free of spurious modes is dependable under any extremes in temperature that can be reasonably expected.

BASIC CIRCUIT OF CAPACITANCE-BRIDGE OSCILLATOR

*1-365. Figure 1-165 illustrates the basic circuit

* The discussion in paragraph 1-365 is based upon the analysis of the basic circuit appearing in the report, *H.F. Harmonic Crystal Investigation,* by S. A. Robinson and F. N. Barry of Philco Corporation, on Army Contract #W33-038 ac-14172, 1947.

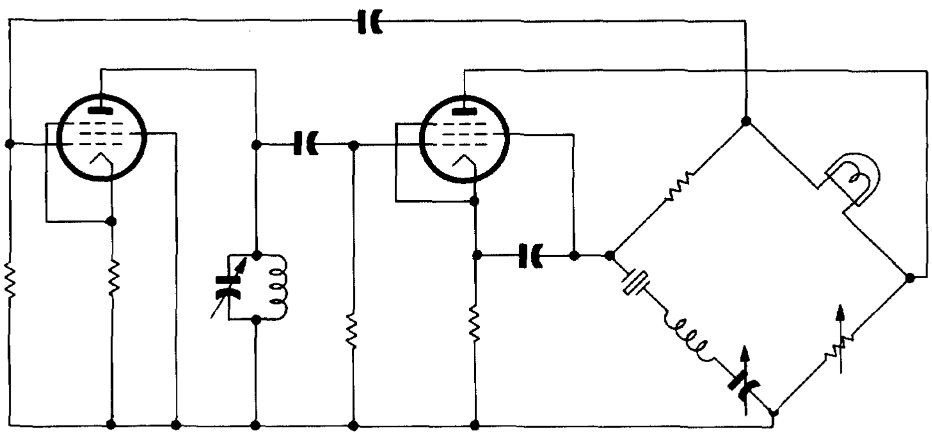


Figure 1-164. Two-stage Meacham bridge-stabilized oscillator

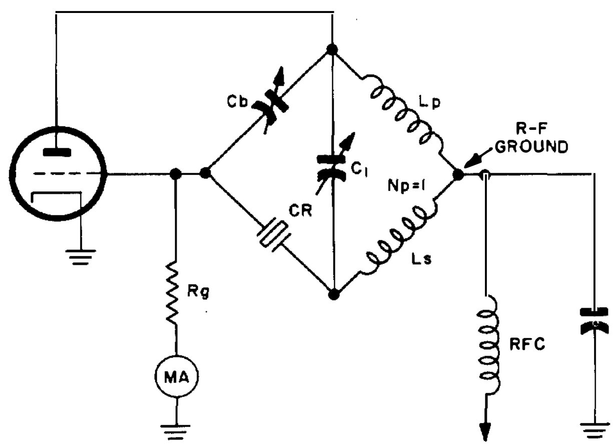
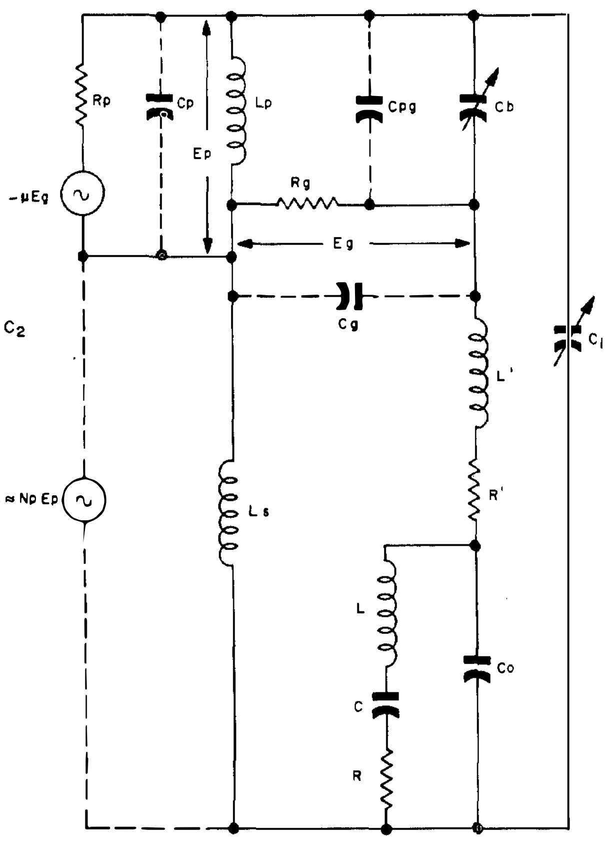


Figure 1-165. (A) Basic circuit of capacitance-bridge oscillator. (B) Equivalent circuit of capacitance-bridge oscillator

of a capacitance-bridge oscillator. The first design consideration in this oscillator, as in all others that are to operate in the v-h-f range, is to keep all the leads as short as possible. Those circuit arrangements and circuit components that provide a minimum of stray inductance and capacitance are to be preferred. In the circuit of figure 1-165 (A), L_p and L_s are actually a single, tapped inductor with the two sides wound on the same form and tightly coupled together. The induced-voltage effect is equivalent to that of a single generator connected across both coils and driving the bridge with an emf $(E_p + E_s) \approx E_p(N_p + 1)$. N_p, the *turns* ratio of L_p to L_s, is usually, and most conveniently, equal to 1. In case the shunt capacitance of the crystal unit is greater than 10 $\mu\mu$f, it would be desirable to make N_p slightly greater than 1. An N_p greater than 1 but less than 2 can be expected to provide a higher output, but the operation will tend to be more critical and the frequency less stable. Before the circuit is placed in operation, the bridge must be balanced at an off-resonance frequency, so that no voltage can appear across the grid circuit. At an off-resonance frequency the crystal unit appears as a capacitance C_o, so that under the conditions of balance

$$\frac{X_{Cb}}{X_{Co}} = \frac{X_{Lp}}{X_{Ls}} = \frac{C_o}{C_b} = \frac{L_p}{L_s} = N_p^2 \qquad 1\text{—}365\ (1)$$

With $N_p = 1$, C_b is adjusted to equal C_o. (C_p is here assumed to include C_{pg}, and L_p to account for C_p. See figure 1-165 (B).) The total capacitance



in the circuit is thus,

$$C_T = C_1 + \frac{C_o}{2} \qquad 1\text{—}365\ (2)$$

Since the distributed inductance of the crystal leads, L′, tends to increase the effective value of C_o, the frequency at which the bridge is balanced should not be greatly different from the intended operating frequency. If L′ is unduly large, a series capacitance should be connected in the crystal arm of the bridge sufficient to annul the stray inductance in the vicinity of the operating frequency. Once the bridge is balanced, C_b should not be adjusted again. The initial adjustment is rather critical, requiring an accuracy of a few tenths of a micromicrofarad. C_o and C_b in the v-h-f range should be as small as possible. The coaxially-mounted crystals, such as those contained in the HC-10/U holder, are to be preferred on this account. Values of C_o in the neighborhood of 4 or 5 $\mu\mu$f are quite feasible. C_o can be further reduced by connecting an inductor across the crystal unit to annul part, but not all, of the shunt capacitance; however, this should be avoided, because the presence of the inductor would narrow the frequency range over which the bridge can be considered balanced. C_b must be adjustable over the expected capacitance range of the particular type of crystal unit to be used. C_1 is the tuning capacitance. For crystal-controlled operation, C_1 is adjusted so that the total circuit capacitance C_T is approximately resonant with the total inductance at the operating frequency. To balance the circuit, C_1 is set to a position that tunes the bridge to a frequency well off the resonance frequency of the crystal unit. Referring to figure 1-165 (C), it can be seen that E_g equals $(E_s + E_{CR})$. At the tuned frequency, R can be neglected and the crystal unit considered as a capacitance, C_o. Approximately, E_s and E_{CR} are 180 degrees out of phase, and therefore tend to annul each other. Now, assume that C_b is made to approach zero. I_4 and E_{CR} therefore become negligible, and the circuit behaves as if the crystal side of the bridge were open-circuited at C_b. The remaining circuit would be simply a Hartley oscillator with the crystal unit serving as a blocking capacitor between the inductor and the grid. If C_b is gradually increased, E_{CR} builds up until a point is reached where E_s effectively is canceled and E_g is insufficient to sustain oscillations. C_b should then be increased one more increment beyond the oscillation cutoff. At this setting of C_b, the bridge can be considered properly balanced, but a check should first be made that oscillations do not occur at other settings of C_1 well removed from its value for crystal control. If such oscillations do occur, the adjustment of C_b should be repeated. The free-running oscillations can be distinguished from the crystal-controlled oscillations by the continuous nature of their activity curves as measured by grid current and output meters when C_1 is varied above and below a discontinuous region. A discontinuous point indicates an abrupt change to crystal control, where the frequency begins to change at a much slower rate with variations in circuit capacitance. However, once the bridge is balanced, no oscillations occur except near the crystal resonance frequency, in which region the bridge balance is upset.

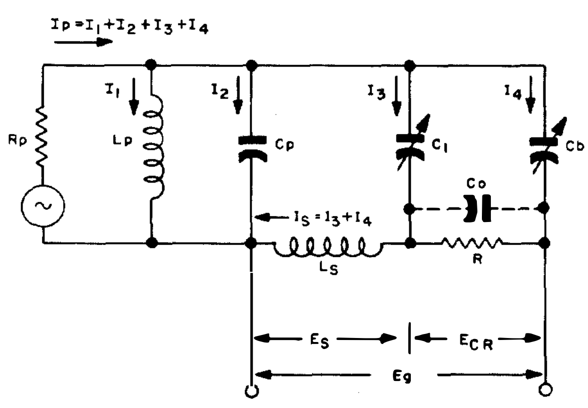


Figure 1-165. (C) Simplified equivalent circuit

1-366. Referring again to figure 1-165 (C), with the circuit balanced, suppose that C_1 is gradually increased from its minimum value. At some point oscillations suddenly start; as C_1 is further increased, the activity builds up to a maximum and then sharply declines, as is illustrated in figure 1-166. Note also the sharp decrease in frequency when maximum amplitude is approached. Apparently, when oscillations first begin, the crystal appears inductive. E_{CR} therefore has a large component in phase with E_s, and the circuit is essentially a modified Miller oscillator. Also, the ratio of I_4 to I_3 is a maximum, since E_{CR} tends to cancel the voltage across C_b. As C_1 is slowly increased, the frequency and the inductive reactance of the crystal drop. This means that the effective Q of the grid circuit also decreases. Although the presence of the capacitance C_1 modifies the phase relations, the circuit performs fundamentally as a Miller oscillator. L_s can be interpreted as something of a booster inductor to increase the effective inductance of the crystal unit, and I_3 can similarly be viewed as a booster current to boost the voltage across the inductive component in the grid circuit without, at the same time, increasing the voltage across the crystal R_e. That the capacitance-bridge circuit actually has the same characteristics as does the Miller circuit is well illustrated by the similarities between the curves of figure 1-166 and the equivalent curves for the Miller oscillator shown in figure 1-144. Note that for both oscillators, the circuit capacitance for maximum excitation does not coincide with, but is smaller than, the value for maximum output. One significant difference between the two circuits is the fact that the Miller circuit cannot maintain the proper feedback phase if the crystal is operated at series resonance, whereas the capacitance-bridge circuit can, because of the presence of L_s. Where oscillation cutoff for the Miller circuit is above the series-resonance frequency of the crystal, it is below the series-resonance frequency in the capacitance-bridge circuit.

1-367. If the crystal control is to be fully effective, the series-arm resistance must be small compared with the shunt reactance, X_{Co}. Although this requirement becomes increasingly difficult at the higher harmonics, it can be achieved, even at frequencies well above 100 mc. Assuming that

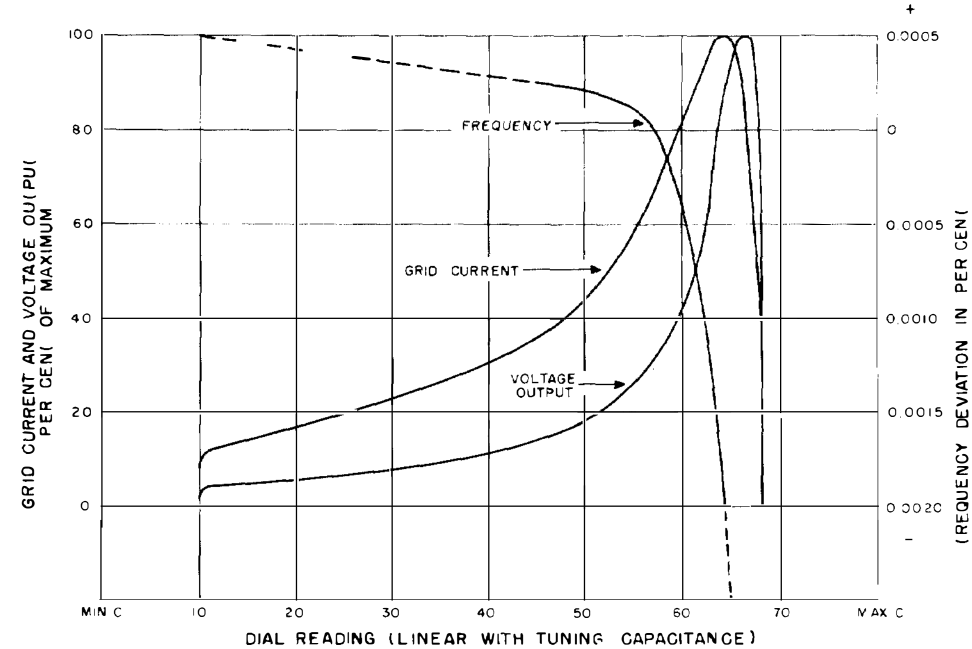


Figure 1-166. Typical performance curves of capacitance-bridge oscillator, showing effects of change in bridge tuning capacitance on voltage output, activity (d-c grid current), and frequency

the series-arm R is not more than one-tenth the magnitude of X_{Co}, then the approximate equation for the effective crystal reactance, $X_e = X_s X_{Co}/(X_{Co} + X_s)$, where $X_s = 4\pi L\Delta f$ series-arm reactance, is sufficiently close for an interpretation of the capacitance-bridge performance. Now, oscillations cannot start unless $|X_{Cb}| > X_e + X_{Ls}$. X_{Cb} we shall assume is equal to X_{Co} under the conditions of balance. X_e is equal to $|X_{Co}|$, and hence to $|X_{Cb}|$, when $X_s = -X_{Co}/2$, that is, when the crystal unit is halfway between series resonance and antiresonance. Thus, when oscillations start, the crystal frequency must be much nearer to the resonant than to the antiresonant state. Also, the plate circuit must appear inductive to the vacuum tube to a degree dependent upon the effective Q of the grid circuit. This means that I_1 must be slightly greater than $(I_2 + I_3 + I_4)$. In figure 1-165 (A), it can be seen that the crystal unit operates into a load reactance approximately equal to the parallel combination of C_1 and the inductor $(L_p + L_s)$ in series with C_b. As the reactance of C_1 approaches that of the inductor, the reactance of the parallel combination rises very sharply, and a small change in C_1 can make a large change in the load reactance across the crystal unit. More than any other factor, this is the reason for the sharp dip in the frequency curve as C_1 approaches a maximum.

1-368. It is not possible to tell at which point in the curve the crystal passes through series resonance. Since at series resonance the reactance of C_1 in parallel with L_p and L_s is equal, approximately, to $-X_{Cb}$, the resonance frequency may well be below the knee of the curve for a crystal having a very small C_o (conditions for large X_{Cb} and near-parallel resonance of C_1 with the inductor) and above the knee for crystals of larger C_o. At series resonance, if R is small I_4 approximately equals I_3C_b/C_1. Assuming that $E_s (= I_s X_{Ls})$ leads $E_{CR} (= I_4 R)$ by 90 degrees, the effective Q of the grid circuit at series resonance is equal to E_s/E_{CR}. When E_s and E_{CR} are expressed as functions of I_3, C_b, C_1, and X_{Ls}, it can be shown that (series resonance) $Q_g = \dfrac{E_s}{E_{CR}} = \dfrac{X_{Ls}(C_1 + C_b)}{RC_b}$. The above equation is only a broad approximation in the v-h-f range, since all the distributed parameters have been ignored, particularly the grid capacitance and the resistance of the inductor. However, it does indicate that the larger the ratio of C_1 to C_b, or, equivalently, X_{Cb}/X_{C1}, the smaller the inductive phase shift that will be required in E_p, and the more nearly will the bridge tank approach parallel resonance. If R, or rather the total grid losses, should increase or decrease, the frequency will decrease or increase, respectively. It seems safe to assume that crystal units having the larger values of RC_o products will operate fairly near their series-resonant state. This is due partly to the fact that the smaller the X_{C_0}/R ratio, the smaller the frequency range between resonance and antiresonance. Crystal units having the smaller values of RC_0 will perform with greater amplitude and frequency stability if operated above series resonance. Unfortunately, crystal units in the v-h-f range are tested only for series resonance. The greater likelihood of the occurrence of unwanted modes increases the importance of having the circuit designed so that the operation of the crystal unit lies within its tested specifications. While the capacitance bridge is excellent for preventing all modes of oscillation except the one desired, it is not a true series-mode oscillator, although it is so classified because its v-h-f application requires the use of crystals that are only series-tested. Rather, the oscillator is something of a hybrid between a Miller and a stabilized Hartley circuit. In the interest of frequency and amplitude stability, the circuit should be adjusted to operate above the knee of the frequency curve. A setting of the tuning capacitance corresponding to a grid current of 50 per cent of the maximum possible provides, approximately, the optimum output voltage and operating state nearest series resonance that are consistent with the operating region of better stability. The peak of the voltage-output curve in figure 1-166 corresponds closely to the adjustment for maximum tank impedance, which certainly occurs below series resonance where the crystal unit appears as a capacitance. The larger the capacitive reactance that the crystal unit can have and still permit oscillations, the more nearly can series-resonance oscillations fall within the higher stability region. For this purpose, the ratio of L_s to L_p and of C_b to C_0 should be as large as unity, or greater, when the capacitance-bridge oscillator is to be used with series-tested crystal units.

DESIGN MODELS FOR CAPACITANCE-BRIDGE OSCILLATORS

1-369. The circuits shown in figures 1-167 through 1-171 represent five different modifications of the capacitance-bridge oscillator. These circuits were designed and tested by the research team of S. A. Robinson and F. N. Barry of Philco Corporation. No single type of circuit was found to be superior for operation over the entire tested frequency range of 50 to 200 mc, but each circuit has advantages for certain applications. The inductive arms of the bridge can be a single, self-supporting tapped inductor having an inside diameter of one-quarter inch or greater. Silver-plated AWG No. 16 wire can be used. The tuning and balancing capacitances are small air capacitors. The fixed capacitances are, for the most part, the button-mica Erie type. Composition resistances are used, having nominal values of ±10 per cent. Successful operation of any of the circuits depends largely upon arranging the circuit components to permit the shortest possible leads, and all components should be of small physical size. Silver-plating of the components is desirable, and careful shielding and the use of low-loss insulating materials is necessary. Without good shielding and well-insulated capacitor shafts, it may be impossible to adjust the bridge properly because of the effects of hand capacitance. Transmit-time effects become quite significant as the frequency is increased beyond 50 mc. The lag in the response of the plate current with rapid changes in the grid voltages is equivalent to the circuit behavior that would result if an inductance were connected in series with R_p of the tube. The lower the plate voltage, the larger is the apparent inductance and its accompanying tendency to lower the frequency. Usually, this effect makes it easier to operate the crystal unit at series resonance, but the need for careful B^+ regulation becomes all the more important. For normal voltages, transit lag is approximately 0.2 degree per megacycle in v-h-f tubes such as the 6AK5.

SINGLE-TUBE, 50- TO 90-MC CAPACITANCE-BRIDGE OSCILLATOR

1-370. The circuit shown in figure 1-167 has been operated at frequencies as high as 135 mc, but its particular merit lies in its performance at frequencies between 50 and 90 mc. When operated in the high-stability region, output up to 10 volts can be obtained, although care should be taken that the rated drive level of the crystal is not exceeded. Outputs of 2 milliwatts into an inductively coupled 100-ohm resistor can be obtained in the same operating region.

COMPACT, MINIATURE, 50- TO 120-MC CAPACITANCE-BRIDGE OSCILLATOR

1-371. The circuit shown in figure 1-168 is particularly suited for construction as a small, packaged, plug-in oscillator. If desired, several such oscillators of different frequencies can be designed as interchangeable units of the associated equipment. The entire shielded unit need not occupy a space greater than 6 cubic inches. The maximum frequency at which this circuit was found to oscillate was 156 mc, but the activity at that frequency was less than one-tenth that at 50 mc. At 120 mc the activity is approximately one-fourth of that at

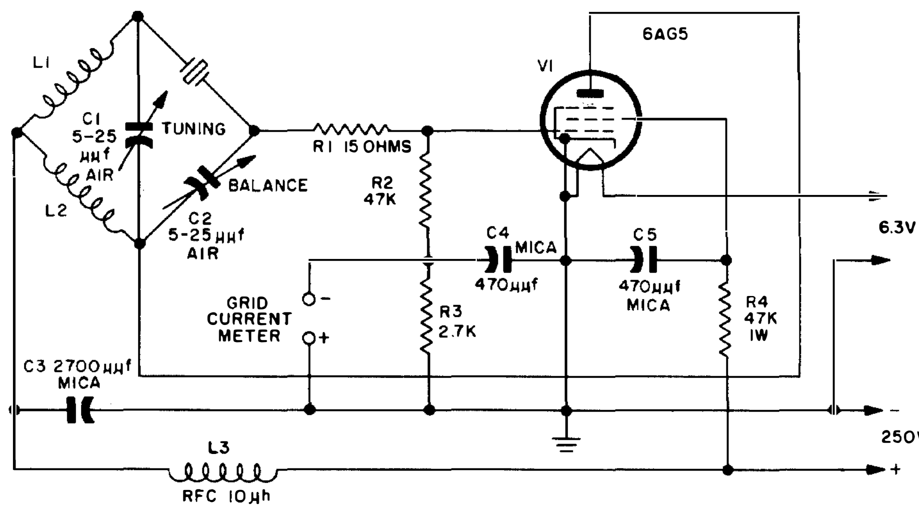


Figure 1-167. A single-tube capacitance-bridge oscillator which is practical for operation in the 50–90-mc frequency range. Resistors not otherwise specified are ½ w. L-1 and L-2 are a single center-tapped coil of suitable inductance

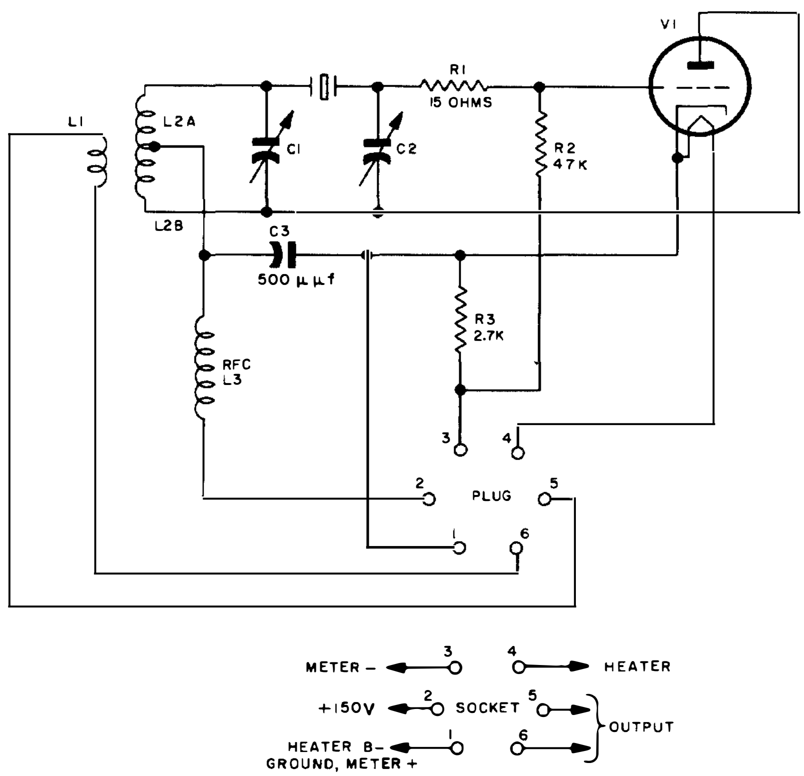


Figure 1-168. A plug-in capacitance-bridge oscillator which is practical for 50–150-mc frequency range

50 mc, so 120 mc appears to be the most practical upper frequency limit. A subminiature tube having high transconductance is used. Greater output is to be achieved with a triode, but greater frequency stability is to be had with a pentode. With a triode, the comparatively large plate-to-grid capacitance which shunts the balancing condenser may make it difficult to achieve a balancing capacitance as small as that of the crystal unit. This condition requires that the L-2A section of the bridge inductor be somewhat larger than the L-2B section. The possible output is reduced thereby, but the crystal unit will be operated nearer its series-resonance frequency. The output secondary, L_1, can be a single turn coupled to the plate end of L_2.

CAPACITANCE-BRIDGE OSCILLATOR FOR GREATER POWER OUTPUT IN THE 50- TO 80-MC RANGE

1-372. The circuit shown in figure 1-169 was designed for the purpose of achieving a maximum power output without regard to the rated drive level of the crystal unit. However, none of the crystals used were fractured during the experiments. The higher-power circuit is essentially the same as that of figure 1-167 except that the N_p ratio of the bridge inductor is greater, higher voltages are used, and a 6AG7 replaces the 6AG5 tube. Although the 6AG7 has a higher transconductance and power rating than the 6AG5, the interelectrode capacitances are greater, the internal leads are longer, and the base is constructed of higher-loss material. The circuit operated at frequencies as high as 102 mc, but above 80 mc the disadvantages introduced by the vacuum-tube construction make the circuit impractical. Better performance might be expected with a 6AH6. With the tube operated near its maximum rated dissipation, a one-watt inductively coupled output was obtained at 54 mc, and one-third watt at 80 mc. These outputs are representative of the peak obtainable. Much less power is to be had if the oscillator is adjusted for operation in the higher-stability region.

TWO-TUBE, 50- TO 100-MC CAPACITANCE-BRIDGE OSCILLATOR

1-373. The circuit shown in figure 1-170 is similar in operation to the one-tube circuit except that the feedback has an additional amplifier stage to boost the gain. There is a significant difference in that the crystal unit is connected to the plate side of the bridge. Under this arrangement, the excitation voltage of V_1 lags the r-f plate voltage of V_2, which means that if the plate load is resistive the r-f plate voltage of V_1 would tend to lag the required excitation voltage of V_2. For oscillations to occur, the plate tuning tank of V_1 must appear inductive in order to shift the input of V_2 to the proper phase. After equilibrium is reached, a slight increase in the value of C_1 causes the plate impedance of V_1 to become more nearly resistive, and therefore the input of V_2 becomes more nearly 180 degrees out of phase with the input of V_1. This requires that the frequency drop to a point where the voltage across C_3 is more nearly in phase with the r-f plate voltage for V_2. For both tubes to operate into resistive loads, the crystal unit must appear as a capacitance. For the crystal unit to operate near series resonance and at the same time maintain the oscillations in the higher-stability region, it would seem that R_1, the parasitic damping resistor in the input circuit of V_2 can be replaced, if necessary, by a resistance comparable in value to the V_2 input reactance. The effect will be to shift the V_2 input phase in a lagging direction, which would require the V_1 tank to be more detuned, and hence less critically adjusted. This, in turn, will require a comparable shift in the phase of the input to V_1, which is to be had by a decrease in frequency, thereby permitting the bridge to be less critically tuned in the vicinity of the crystal resonance point. The circuit in figure 1-170 was found quite practical for use as a test oscillator for measuring the relative performance characteristics of harmonic-mode crystal units. During the temperature runs, even though frost had collected on various components, the operation of the circuit was little affected. For duplicate units of this circuit to provide essentially the same meter readings for tests of the same crystal unit, it is necessary that the vacuum tubes used in the twin circuits show the same plate characteristics within ± 5 per cent. A breadboard model of the oscillator having different values of tuning inductances was able to operate at 140 mc. L_1, in figure 1-170, is a 5-turn coil, approximately one-quarter inch in diameter; L_2 and L_3 are the two halves of a 4-turn, center-tapped coil, approximately one-half inch in diameter.

MULTITUBE CAPACITANCE-BRIDGE OSCILLATOR OPERABLE AT FREQUENCIES UP TO AND ABOVE 200 MC

1-374. The circuit shown in figure 1-171 has been used to generate crystal-controlled frequencies as high as 219 mc, the seventy-third harmonic of a 3-mc crystal. This frequency approaches the ultimate directly obtainable with quartz crystals at the present state of the art. A large part of the

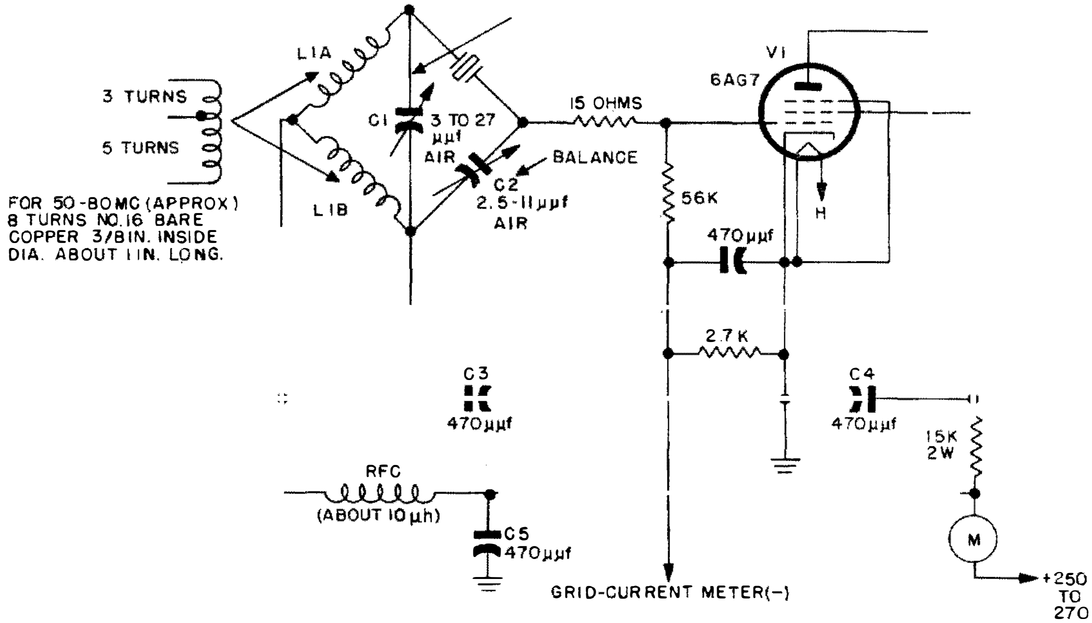


Figure 1-169. A capacitance-bridge oscillator for higher power output which is practical for operation in the 50–80-mc frequency range. Resistors not otherwise specified are ½ w. All fixed capacitors have mica dielectrics

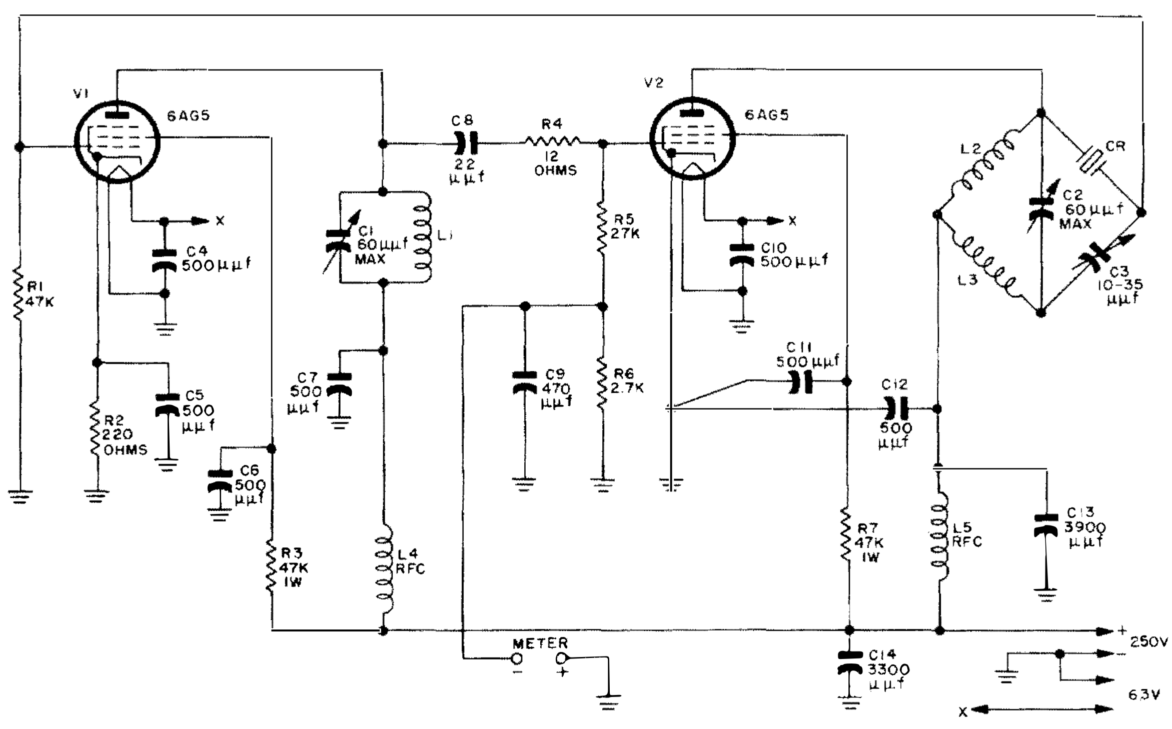


Figure 1-170. A two-tube capacitance-bridge oscillator which is practical for operation in the 50–100-mc frequency range

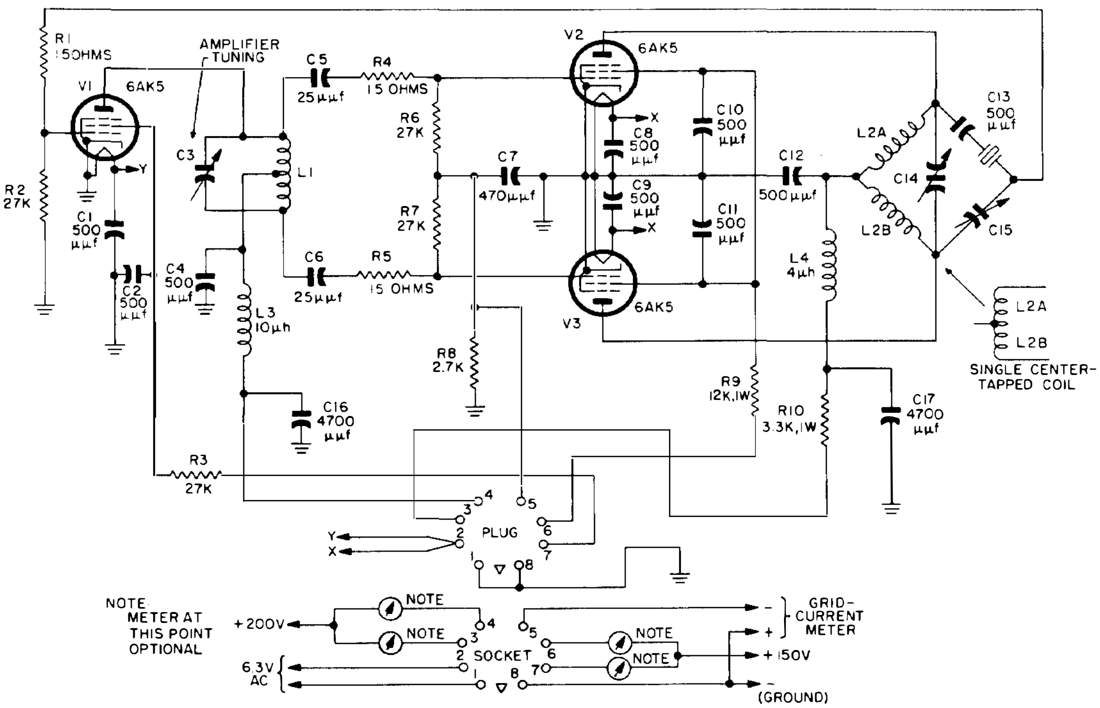


Figure 1-171. A multitube capacitance-bridge oscillator which is practical for operation in the 50–200-mc range. Resistors not otherwise specified are rated at ½ w

success of the oscillator in figure 1-171 is due to the balanced electrical and mechanical nature of the push-pull capacitance-bridge circuit. The operation is very much the same as that of the circuit in figure 1-170 except that the bridge stage is operated in push-pull. With different values of inductance, the circuit provides reliable frequency control anywhere in the v-h-f range, from 200 mc on down. Probably its most practical application is as a harmonic test oscillator. The upper frequency obtainable is not limited by the circuit itself, but by the resistances and shunt capacitances of the crystal units.

OTHER MODIFICATIONS OF THE CAPACITANCE-BRIDGE OSCILLATOR

1-375. A number of capacitance-bridge modifications have been successfully attempted, four of which are illustrated in figure 1-172. The circuits are largely self-explanatory, and will not be discussed here. Probably of most importance is the electron-coupled circuit, since it permits frequency multiplication in the plate circuit. The triode connection of the crystal circuit probably prevents the crystal, itself, from being operated at frequencies above 75 mc.

The Butler Oscillator

1-376. At the present time, probably the most widely used of the series-mode oscillators is the Butler, cathode-coupled, two-stage oscillator. The basic design and equivalent circuits are shown in figure 1-173. Although the single-tube, transformer-coupled type of oscillator will probably outrank the two-tube circuit eventually, the Butler oscillator is the more popular at present because of its simplicity, versatility, frequency stability, and, of most importance, its comparative reliability. With the older types of crystal units, it was generally found that the Butler circuit was the least critical to design and to adjust for operation of the crystal at a given harmonic. The balanced arrangement of the circuit and the fact that twin triodes can be obtained in a single envelope contribute a saving in space and cost, and permit the use of short leads. For greater frequency stability than is normally to be had from parallel-mode oscillations, the cathode-coupled circuit can be used quite satisfactorily at any of the lower frequencies provided the resistance of the crystal unit is not greater than a few hundred ohms. However, the power output is small by comparison

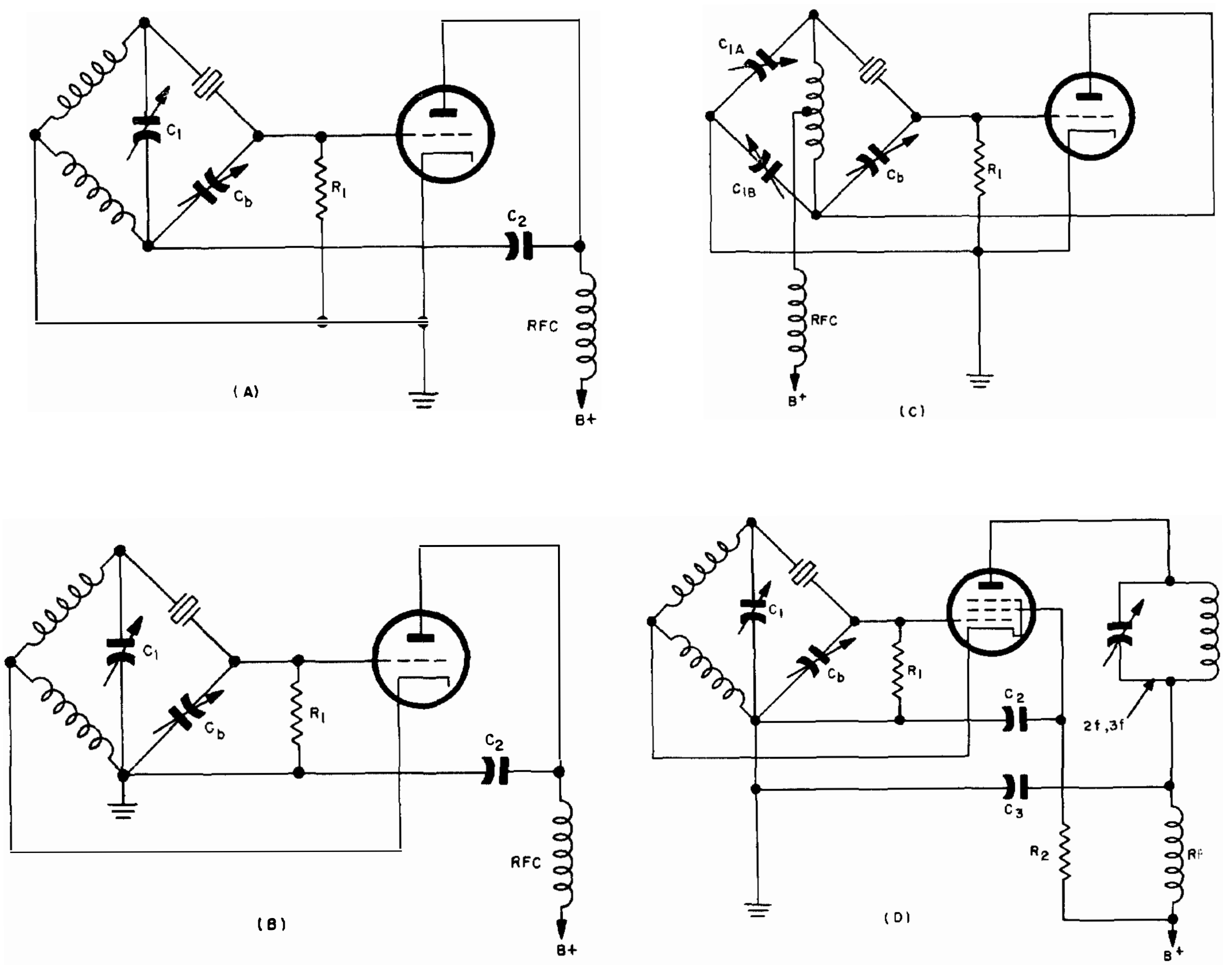
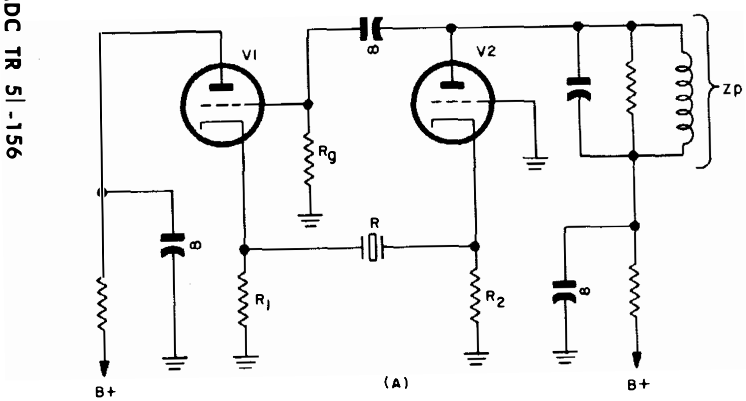


Figure 1-172. Miscellaneous capacitance-bridge oscillators

with that of the Miller circuit for the same crystal power, and the broad bandwidth without plate tuning of the Pierce circuit is not matched. The Butler circuit is usually designed for class-A operation, but class C is possible if greater output and plate efficiency are desired. The output may be taken from almost any part of the circuit—the plate or cathode of either tube. Quite often the cathode follower, V_1, in figure 1-173, is a pentode, with the screen, control grid, and cathode forming a triode section electron-coupled to a plate circuit that usually is tuned for frequency multiplication, although the electron coupling can be employed simply to obtain greater output amplitude and to isolate the load from the oscillator circuit. At very high frequencies, where the shunt reactance of the crystal unit approaches the magnitude of the series-arm R, the operation is generally improved by shunting the crystal unit with an inductor that is antiresonant with the shunt capacitance of the crystal at the operating frequency. When properly designed and adjusted, the two tubes operate 180 degrees out of phase into resistive loads, and the crystal unit acts as a pure resistance.

1-377. In figure 1-173, V_1 is connected as a grounded-plate cathode follower. The V_1 output current, I_c, enters the feed-back path through the crystal unit, which is operating at series resonance. The impedance of the crystal unit is thus approximately equal to the equivalent series-arm resistance, V_2, a grounded-grid amplifier connected in the feed-back circuit, is excited by I_2, the component of I_c that passes through R_2. I_{p2}, the remaining component of the feed-back current passes through V_2. It can be seen that the input voltage of V_1, E_{g1}, is equal to the output voltage of V_2 if we assume that the coupling capacitance is infinite. The plate circuit of V_2 is broadly tuned to the de-



NOTE:

(1) $Z_L = \frac{Z_p R_g}{Z_p + R_g}$

(2) $\rho = -\frac{\mu_2 E_{g2}}{I_{p2}} = -\frac{\mu_2 (R_{p2} + Z_L)}{\mu_2 + 1}$

(3) $Z_1 = \frac{R_{p1}}{\mu_1 + 1}$

(4) $Z_2 = \frac{R_{p2} + Z_L}{\mu_2 + 1}$

(5) $E_{g1} = E_p - E_1$

(6) $Z_{g2} = \frac{R_2 Z_2}{R_2 + Z_2}$

(7) $Z_f = R + Z_{g2}$

(8) $Z_k = \frac{R_1 Z_f}{R_1 + Z_f}$

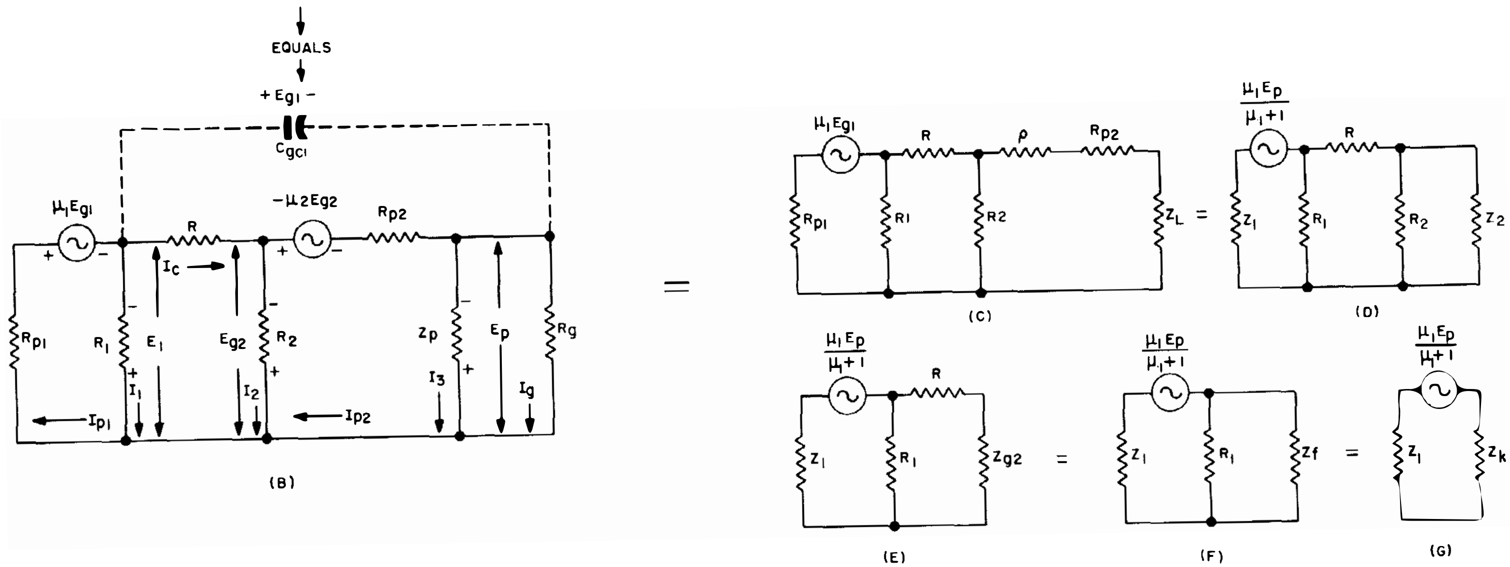


Figure 1-173. (A) Basic diagram of Butler two-stage cathode-coupled oscillator. (B) Equivalent r-f circuit of Butler oscillator. Current arrows indicate instantaneous electron flow when r-f voltages have polarities shown. (C) Simplified equivalent circuit where generator of grounded-grid amplifier is replaced by a negative resistance. (D), (E), (F), and (G) Progressive simplifications of equivalent Butler circuit

sired frequency. If Z_p were simply a resistance, the circuit could still oscillate at the first crystal harmonic. If the crystal unit were shorted out, the circuit could also oscillate, but with the frequency controlled by the tuned plate circuit. R_1 and R_2 are usually equal, having values between 50 and 200 ohms. V_1 and V_2 are also usually of the same tube type. As will be seen, frequency stability is improved with large values of transconductance. Note in figure 1-173(B) that the r-f plate current in V_1 is greater than that in V_2. As in all other vacuum-tube oscillators, there are two fundamental equilibrium conditions to consider: the over-all gain must equal unity, and the over-all phase shift must equal zero. We shall first consider the factors affecting loop-gain.

LOOP GAIN OF BUTLER CIRCUIT

1-378. At equilibrium we can say that

$$1 = \frac{E_1}{E_p} \cdot \frac{E_{g2}}{E_1} \cdot \frac{E_p}{E_{g2}} = G_1\, G_2\, G_3 \text{ (respectively)} \qquad 1\text{—}378\ (1)$$

The immediate problem is to find the values of G_k in terms of the circuit parameters. First, referring to figure 1-173 (ignore the capacitance C_{gc1} in circuit (B)), assume that the voltage across R_g is approximately equal to E_p, then

$$E_{g1} = E_p - E_1 \qquad 1\text{—}378\ (2)$$

$$E_1 = I_{p1} Z_k \qquad 1\text{—}378\ (3)$$

and

$$I_{p1} = \frac{\mu_1 E_{g1}}{R_{p1} + Z_k} = \frac{\mu_1 E_p - \mu_1 I_{p1} Z_k}{R_{p1} + Z_k} = \frac{\mu_1 E_p}{R_{p1} + Z_k(\mu_1 + 1)} \qquad 1\text{—}378\ (4)$$

On combining equations (3) and (4)

$$E_1 = \frac{\mu_1 E_p Z_k}{R_{p1} + Z_k(\mu_1 + 1)} \qquad 1\text{—}378\ (5)$$

and

$$G_1 = E_1/E_p = \frac{\mu_1 Z_k}{R_{p1} + Z_k(\mu_1 + 1)} \approx \frac{g_{m1} Z_k}{1 + g_{m1} Z_k} \qquad 1\text{—}378\ (6)$$

The approximation in equation (6) is made on the assumption that $(\mu_1 + 1) \approx \mu_1$. If the numerator and denominator of equation (4) are divided by $(\mu_1 + 1)$, we have

$$I_{p1} = \frac{\dfrac{\mu_1}{\mu_1 + 1} E_p}{\dfrac{R_{p1}}{\mu_1 + 1} + Z_k}$$

Note that with Z_k fixed by the external circuit, the plate current is related to the excitation voltage, E_p, in such a way that the tube behaves as if it had an effective amplification factor of $\frac{\mu_1}{\mu_1 + 1}$ and an effective plate resistance equal to $\frac{R_{p1}}{\mu_1 + 1}$. This resistance is given the symbol Z_1 in figure 1-173. If an additional resistance, R_L, were connected between the plate of V_1 and r-f ground, Z_1 would equal $\frac{R_L + R_{p1}}{\mu_1 + 1}$. Now, to find the value of $G_2 = E_{g2}/E_1$, we start with

$$E_{g2} = I_2 R_2 \qquad 1\text{—}378\ (7)$$

$$E_1 = I_c R - \mu_2 E_{g2} + I_{p2} R_{p2} + I_{p2} Z_L \qquad 1\text{—}378\ (8)$$

$$I_c = I_2 + I_{p2} \qquad 1\text{—}378\ (9)$$

and

$$I_{p2} = \frac{E_{g2} + \mu_2 E_{g2}}{R_{p2} + Z_L} = \frac{(\mu_2 + 1) E_{g2}}{R_{p2} + Z_L} \qquad 1\text{—}378\ (10)$$

On rearranging equation (10) to express E_{g2} as a function of I_{p2}, and substituting this function for E_{g2} in equation (8), we have

$$E_1 = I_c R + I_{p2}\left[-\frac{\mu_2 (R_{p2} + Z_L)}{\mu_2 + 1} + R_{p2} + Z_L\right] \qquad 1\text{—}378\ (11)$$

From equation (11) we find that the equivalent generator of V_2 can be represented by an equivalent negative resistance

$$\rho = \frac{-\mu_2 (R_{p2} + Z_L)}{\mu_2 + 1} \qquad 1\text{—}378\ (12)$$

It can be seen that ρ is smaller in magnitude than $(R_{p2} + Z_L)$, so that the total V_2 branch resistance is positive. Defining the V_2 branch impedance to be Z_2, we have

$$Z_2 = \rho + R_{p2} + Z_L = \frac{R_{p2} + Z_L}{\mu_2 + 1} \qquad 1\text{—}378\ (13)$$

On substituting equations (7) and (13) into equation (10), we have

$$I_{p2} = \frac{R_2 I_2}{Z_2} \qquad 1\text{—}378\ (14)$$

so that equation (9) may be written

$$I_c = \left(\frac{Z_2 + R_2}{Z_2}\right) I_2 \qquad 1\text{—}378\ (15)$$

On substituting equations (13), (14), and (15) into (11),

$$E_1 = \frac{I_2 R (Z_2 + R_2)}{Z_2} + \frac{I_2 R_2 Z_2}{Z_2}$$

$$= I_2 \left[\frac{R (Z_2 + R_2) + R_2 Z_2}{Z_2}\right] \qquad 1\text{—}378\ (16)$$

Thus

$$G_2 = \frac{E_{g2}}{E_1} = \frac{R_2 Z_2}{R (Z_2 + R_2) + R_2 Z_2} \qquad 1\text{—}378\ (17)$$

To find G_3 ($= E_p/E_{g2}$), we see that

$$E_p = I_{p2} Z_L \qquad 1\text{—}378\ (18)$$

which, on substituting the value of I_{p2} given in equation (14), becomes

$$E_p = \frac{I_2 R_2 Z_L}{Z_2} \qquad 1\text{—}378\ (19)$$

Dividing by E_{g2} ($= I_2 R_2$), we have

$$G_3 = \frac{E_p}{E_{g2}} = \frac{Z_L}{Z_2} \qquad 1\text{—}378\ (20)$$

The conditions for equilibrium as expressed by equation (1) are thus found to be

$$G_1 G_2 G_3 =$$

$$\frac{\mu_1 Z_k R_2 Z_L}{[R_{p1} + Z_k (\mu_1 + 1)] [(Z_2 + R_2) R + R_2 Z_2]} = 1 \qquad 1\text{—}378\ (21)$$

By a slightly different approach, in which the equilibrium is expressed as

$$G_1' G_1'' G_2 G_3 = \frac{E_{g1}}{E_p} \cdot \frac{E_1}{E_{g1}} \cdot \frac{E_{g2}}{E_1} \cdot \frac{E_p}{E_{g2}} = 1$$

where

$$G_1' = \frac{E_{g1}}{E_p} = \frac{R_2 Z_L - (Z_2 + R_2) R - R_2 Z_2}{R_2 Z_L} \qquad 1\text{—}378\ (22)$$

and

$$G_1'' = \frac{E_1}{E_{g1}} = \frac{\mu_1 Z_k}{R_{p1} + Z_k} \qquad 1\text{—}378\ (23)$$

it will be found that

$$G_1' G_1'' G_2 G_3 = G_1 G_2 G_3 =$$

$$\left[\frac{R_2 Z_L}{(Z_2 + R_2) R + R_2 Z_2} - 1\right] \frac{\mu_1 Z_k}{R_{p1} + Z_k} = 1 \qquad 1\text{—}378\ (24)$$

Since R_{p1} and R_2Z_L are very large compared with Z_K and $[(Z_2 + R_2) R + R_2 Z_2]$, respectively, we can simplify equations (21) and (24) by writing

$$\frac{R_2 Z_L Z_k g_{m1}}{(Z_2 + R_2) R + R_2 Z_2} \approx 1 \qquad 1\text{—}378\ (25)$$

On dividing both numerator and denominator by $(Z_2 + R_2)$ and substituting for the values of Z_{g2} and Z_f as defined in equations (6) and (7) of figure 1-173, equation (25) can be simplified somewhat. Thus

$$\frac{Z_{g2} Z_L Z_k g_{m1}}{Z_2 Z_f} = 1 \qquad 1\text{—}378\ (26)$$

1-379. The design of a Butler oscillator must be such that under no-signal conditions the left side of equation 1—378(26) is greater than unity. As oscillations build up, the principal effects will be a decrease in the effective g_{m1} and g_{m2} as the signal swings farther into the lower bend of the E_cI_b curve. How large the equilibrium amplitude will be depends upon how much greater than unity the left side of equation 1—378(26) is at the start. The larger the left-side magnitude, the greater must be the decrease in g_{m1}, and hence the greater the equilibrium activity must be. If the oscillator is to operate class A, as is usual, the gain equilibrium should very nearly hold for no-signal conditions, with due allowance made for a maximum Z_f ($= R + Z_{g2}$) when $R = R_m$, the maximum series resistance permissible for the particular type of crystal unit chosen. With all else constant, maximum activity is to be obtained when g_{m1} and g_{m2} are maximum under no-signal conditions. Assurance that the crystal unit will not be driven beyond its rated power can be approximately predicted from the plate characteristics of the tube to be used. If no grid current is drawn, the bias on V_1 will be

$$E_{c1} = - I_{b1} R_1 \qquad 1\text{—}379\ (1)$$

where I_{b1} is the average d-c plate current of

V_1. Grid current can be drawn if [(max)E_p - (max)E_1] is greater than $I_{b1}R_1$, in which case the bias will be

$$E_{c1} = (\max)\ E_1 - (\max)\ E_p \qquad 1\text{—}379\ (2)$$

Greater amplitude stability is achieved if R_1 is sufficiently small for equation (2) to apply. Using the appropriate equations in paragraph 1-378, a maximum value of E_1 can be determined that will not allow the crystal current, I_c, to become greater than $\sqrt{P_{cm}/R}$, where P_{cm} is the maximum recommended power level of the crystal, and R is any crystal resistance between R_m and $R_m/9$. Very possibly, a twin triode may be preferred, or perhaps the choice of tubes will be dictated by the h-f type of tube most readily available. R_1 and R_2 are to be kept as small as possible in the interest of frequency stability. In the final analysis, the plate voltage permitting an optimum output for the average crystal unit, without risking an overdrive for any expected value of crystal R, is most easily checked by experiment.

DESIGN CONSIDERATIONS TO MAXIMIZE THE FREQUENCY STABILITY OF THE BUTLER OSCILLATOR

1-380. If the cathode follower operates into a purely resistive network, as is indicated in the equivalent circuit of figure 1-173(B), maximum stability in the phase characteristics is obtained. As nearly a resistive circuit as possible is desirable, for under these conditions the frequency is independent of the plate resistance of the tubes, and a small increment of reactance requires the least adjustment of the crystal to restore a phase equilibrium. E_1, the output voltage of the cathode follower, is in phase with the excitation voltage, E_p. In a resistive circuit, E_p will be 180 degrees out of phase with E_{g2}. Thus, E_{g2} must be 180 degrees out of phase with E_1. Since E_{g2} is the voltage of ground with reference to the cathode of V_2, and E_1 is the voltage of the cathode of V_1 with reference to ground, I_1 and I_2 must be in phase. For example, imagine that R is zero, then R_1 and R_2 could be assumed to be two halves of a single resistance. In this case E_{g2} would equal $-E_1$, and the proper phase relation would exist.

1-381. To maintain as nearly as possible a resistive circuit, Z_p must tune as broadly as is practicable; the tendency of the input capacitances of V_1 and V_2 to shift the phase must be compensated; the transit time of the vacuum tubes must be minimized; and, if an inductor is connected across the crystal unit to antiresonate with the shunt capacitance, C_o, the resulting parallel-resonant circuit must also tune very broadly. The tuned plate circuit, Z_p, must be sufficiently selective to ensure that the circuit can oscillate only at the desired harmonic of the crystal frequency, but, beyond this, any increase in the tank selectivity only results in a greater phase shift, and consequently a greater frequency shift, for a given percentage change in the plate capacitance. The use of a damping resistance as indicated in figure 1-173(A), or a low-Q coil, will broaden the tuning of the tank. The stray capacitance from the plate of V_2 to ground should be kept to a minimum.

1-382. To annul the input capacitance of V_2, which is equal to the total capacitance between the cathode of V_2 and ground, we can connect an inductor in series with R_2, or replace R_2 with a low-Q inductor and employ gridleak bias for V_2 (while keeping the grid at r-f ground by the use of an r-f bypass capacitor), or shunt R_2 with an inductor in series with an r-f bypass capacitor. In any event, the inductor is to be antiresonant with the cathode-to-ground capacitance at the operating frequency. With R_2 acting as a damping resistance, a broadband response is ensured for the antiresonant combination.

1-383. The grid-to-cathode capacitance and the cathode-to-ground capacitance of V_1 can also be annulled by the use of antiresonant inductors. However, a more effective and economical method is to design the circuit so that the two cathode capacitances of V_1 neutralize each other regardless of the particular frequency. The grid-to-cathode capacitance, C_{gc1}, is illustrated by the dotted-line circuit in figure 1-173(B). The voltage across C_{gc1} is E_{g1}, so that the leading component of current through the grid circuit is

$$I_{gx} = \frac{E_{g1}}{X_{cg1}} \qquad 1\text{—}383\ (1)$$

For convenience, let it be imagined that all of I_{gx} flows through R and R_{p2} in completing its circuit. If it is not to upset the phases of the voltages across these resistances, I_{gx} must be annulled by an equal lagging current through the R-R_{p2}-Z_p circuit. Thus, assuming the transit-time effect is negligible, the plate tank must be slightly inductive if V_2 is to operate into a purely resistive load. This much can be controlled by the adjustment of the V_2 plate circuit. With the circuit properly adjusted, it can now be imagined that I_{gx} is no longer a part of I_c, but circulates directly through Z_L and C_{gc1} in series. The design problem is to ensure that no part of I_{gx} flows through R_1 or V_1, but returns to Z_L by flowing entirely through the V_1 cathode-to-ground capacitance, C_1—not shown in figure 1-173. Furthermore, the design should be such that I_{gx} is all the current that flows through C_1; otherwise, there will be a net unneutralized leading component upsetting the voltage phases in the rest of the circuit. With proper neutralization, the only reactive current will be confined to a series circuit comprised of C_{gc1}, C_1, ground, and an effective inductance shunting Z_L. The voltage across each of the reactive impedances due to I_{gx} will be equal in magnitude and phase to the voltages caused by the in-phase currents flowing through the corresponding resistive impedances. Thus, to neutralize the circuit, the leading current, I_{gx}, flowing through C_1 must of itself produce the voltage E_1. This occurs when

$$E_1 = I_{gx} X_{C1} \qquad 1\text{—}383\ (2)$$

or, using equation (1) to replace I_{gx},

$$E_1 - \frac{E_{g1} X_{C1}}{X_{cg1}} \qquad 1\text{—}383\ (3)$$

Now,

$$E_1 = \frac{\mu_1 E_{g1} Z_k}{R_{p1} + Z_k} = g_{m1} E_{g1} Z_k \qquad 1\text{—}383\ (4)$$

Using equation (4) to eliminate E_1 in equation (3), we have

$$g_{m1} E_{g1} Z_k = \frac{E_{g1} X_{C1}}{X_{cg1}}$$

or

$$g_{m1} Z_k = \frac{C_{gc1}}{C_1} \qquad 1\text{—}383\ (5)$$

Equation (5) defines the ratio for the V_1 input to output capacitance that should exist for maximum frequency stability.

1-384. To minimize the tendency of the transit time to cause the respective plate currents of V_1 and V_2 to lag the equivalent generator voltages, small-dimensioned h-f tubes should be used, and the plate voltages should be as high as practicable. If the additional expenditure in the design and production of the circuit are warranted, suitable networks can be devised to neutralize the transit effects.

1-385. At the higher frequencies the series resistance of the crystal unit tends to increase, since the lagging component of current through the series arm must increase in order to annul the increased leading component through the shunt capacitance, C_o. For this to occur, the frequency may need to be increased considerably above the natural resonance of the motional arm, so that the effective series resistance approaches the value of a parallel-resonant impedance. To reduce the resistance to the series-arm value, C_o should be annulled by an antiresonant inductor having an inductance

$$L_o = \frac{1}{\omega^2 C_o} \qquad 1\text{—}385\ (1)$$

To prevent the crystal shunt reactances from being more frequency sensitive than the plate circuit of V_2, a shunt resistance should also be connected across the crystal unit to dampen the L_oC_o tank. A suitable resistance, R_o, that can interfere very little with the crystal stabilizing effect is

$$R_o = 5\, R_m \qquad 1\text{—}385\ (2)$$

where R_m is the rated maximum permissible crystal series resistance.

STABILIZING EFFECT OF CRYSTAL IN BUTLER CIRCUIT

1-386. From equation 1—241(2) we found that for a crystal operating at series resonance the fractional change in frequency required to produce a small change in phase, $d\theta$, is expressed by

$$\frac{d\omega}{\omega\, d\theta} \quad \frac{R_c}{2\sqrt{L/C}}$$

where R_c is the total resistance the crystal faces, including the crystal's own resistance, and L and C are the series-arm parameters of the crystal unit. In the Butler circuit (refer to figure 1-173) the crystal operates into a resistance

$$R_c = Z_f + \frac{R_1 Z_1}{R_1 + Z_1} \qquad 1\text{—}386\ (1)$$

where Z_f is the resistance of the feed-back circuit, and $\frac{R_1 Z_1}{R_1 + Z_1}$ is the output resistance of V_1. On substituting the values for Z_1 and Z_f, we have

$$R_c = R + \frac{R_2 (R_{p2} + Z_L)}{R_2 (\mu_2 + 1) + R_{p2} + Z_L} + \frac{R_1 R_{p1}}{R_1 (\mu_1 + 1) + R_{p1}} \qquad 1\text{—}386\ (2)$$

If we assume that $R_{p1} \approx R_{p2} = R_p >> Z_L$, and that

$\mu_1 \approx \mu_2 = \mu >> 1$, equation (2) becomes

$$R_c = R + \frac{R_2 R_p}{R_2\mu + R_p} + \frac{R_1 R_p}{R_1\mu + R_p}$$

or

$$R_c = R + \frac{R_2}{R_2 g_m + 1} + \frac{R_1}{R_1 g_m + 1}$$

1—386 (3)

From equation (3) it is seen that the Q of the crystal circuit, and hence the frequency stability, is to be improved if R_2 and R_1 are kept as small as possible and the transconductance of each tube is high. If R_1 and R_2 are of such values that the denominators in equation (3) are large compared with 1, a limiting value is approached, where

$$(\text{max})\ R_c = R + \frac{2}{g_m} \qquad 1\text{—}386\ (4)$$

DESIGN PROCEDURE FOR BUTLER OSCILLATOR

1-387. In considering the use of a Butler oscillator for controlling frequencies below 20 mc, the principal factor to consider is whether the frequency stability required is greater than that which is normally obtained with a Pierce circuit. If not, there is little to gain by using two tubes and a frequency-sensitive tuned circuit, unless it is very important that the waveform in the output be more nearly sinusoidal and less influenced by the variations in the crystal resistance. The frequency stability of an average Butler circuit can be expected to be approximately 0.0005 per cent as compared with a stability of approximately 0.001 to 0.0015 per cent for an average Pierce circuit. Above 20 mc, the principal competitor of the Butler is the transformer-coupled type of oscillator. The chief advantage of the Butler is its relative ease of adjustment and dependability. A borderline replacement crystal unit or an aging crystal unit, as a general rule, is more likely to be operative in the two-stage, cathode-coupled circuit than in any of the other types of v-h-f oscillators. Once the Butler circuit has been selected as the most appropriate to use, a crystal unit that has been series-tested at the intended frequency should be selected. The required minimum frequency tolerance and the operating conditions to be expected determine whether the crystal unit, or perhaps the entire oscillator, is to be oven-controlled. For the next step, it is probably best to select the types of vacuum tubes to use. Insofar as space, weight, and cost are concerned, a single tube envelope for both amplifier stages is desirable. On the other hand, it may be found that a more balanced arrangement and more direct circuit connections can be had with separate tubes, particularly if the crystal unit is to be oven-mounted. The transconductance and plate resistance of the tubes should both be high, for maximum frequency stability. For the same amplification factor, the tube with the larger g_m is usually to be preferred. For h-f and v-h-f operation miniature tubes are preferable, in order to reduce the transit time and the electrode-to-ground capacitances. For class-A operation, both tubes can be of the same type. For class-C operation, the power rating of the cathode follower should be greater than that of the grounded-grid amplifier. For maximum stability it may be desirable to isolate the load from the rest of the circuit, or to tune the load circuit for frequency multiplication. In this case a pentode can be used for either the cathode follower (usually) or for the grounded-grid amplifier, with the load taken from the electron-coupled plate circuit, and with the screen grid serving as the oscillator plate. Either pentodes or triodes can be used in the basic Butler circuit, as desired. In the v-h-f range, triodes have the advantage of smaller transit-time effects. Assume that it is intended to operate the tubes class A. To reduce the transit time, to increase g_m, and to permit a minimum value of Z_L (low-Q tank), the plate voltages should be as high as practicable. Determine the values of R_1 and R_2 that will provide a normal cathode bias for class-A operation. With all else equal, the feedback transmission losses are a minimum if $R_1 = R_2$. For class-C operation, R_2 should equal approximately $4R_1$. Assume that R of the crystal unit is the maximum permissible value, and that the effective g_m's of the vacuum tubes are 25 per cent less than their rated values for class-A operation at the selected plate and grid voltages. With these assumptions, determine the value of Z_L that is required to make the gain equation, 1—378 (26), hold. R_g should be large compared with Z_p, so that $Z_L \approx Z_p$. The plate tank represented by Z_p can be designed as a high-Q circuit, antiresonant at the operating frequency, and shunted by a simulated load resistance, R_L, much smaller than the antiresonant impedance of the tank, itself. In this case, $Z_L \approx Z_p \approx R_L$. The approximations above will be sufficient to build an experimental circuit that should oscillate in a free-running state. A variable resistance can be connected to simulate a crystal unit at series resonance. By varying the simulated crystal resistance over the range possible for a random selection of crystal units, R_L in the plate circuit can be adjusted, if necessary, to ensure that oscillations occur at all possible values of crystal resistance without driving the crystal at a higher than recommended level. The empirical optimum value of R_L can be accepted as the value of Z_L to achieve in the design of the output circuit of the grounded-grid amplifier. The actual design of the output stage depends, of course, upon the type of load into which the oscillator is to operate. The important consideration is that an effective resistance having the value of the experimental R_L is to be introduced in one way or another across the plate tank. The final problem is to neutralize the various circuit capacitances. In neutralizing the V_1 cathode capacitances, the adjustment which permits the feed-back circuit to be purely resistive can be expected to coincide with the conditions for maximum output amplitude and maximum crystal current. The design procedure discussed above should be accepted simply as a suggestion. Individual engineers may well prefer that primary attention be given to fitting the design to meet special requirements.

MODIFICATIONS OF THE BUTLER OSCILLATOR

1-388. As in the case of other conventional oscillator designs, the number of modifications of the Butler circuit appear to be unlimited. In figure 1-174, the basic electron-coupled circuit is shown in (A), and a circuit employing a common ground return for the two tubes is shown in (B). In the electron-coupled circuit, the load, represented by R_L', is effectively isolated from the oscillator circuit, in which the screen of V_1 serves as the cathode-follower anode. The plate circuit of V_1 can be tuned to the second or third harmonic of the oscillator frequency, if desired, in which case V_1 should be operated class C. For maximum output voltage, the plate impedance of V_1 should be high. In figure 1-174 (B), the low-Q inductor, L_1, is

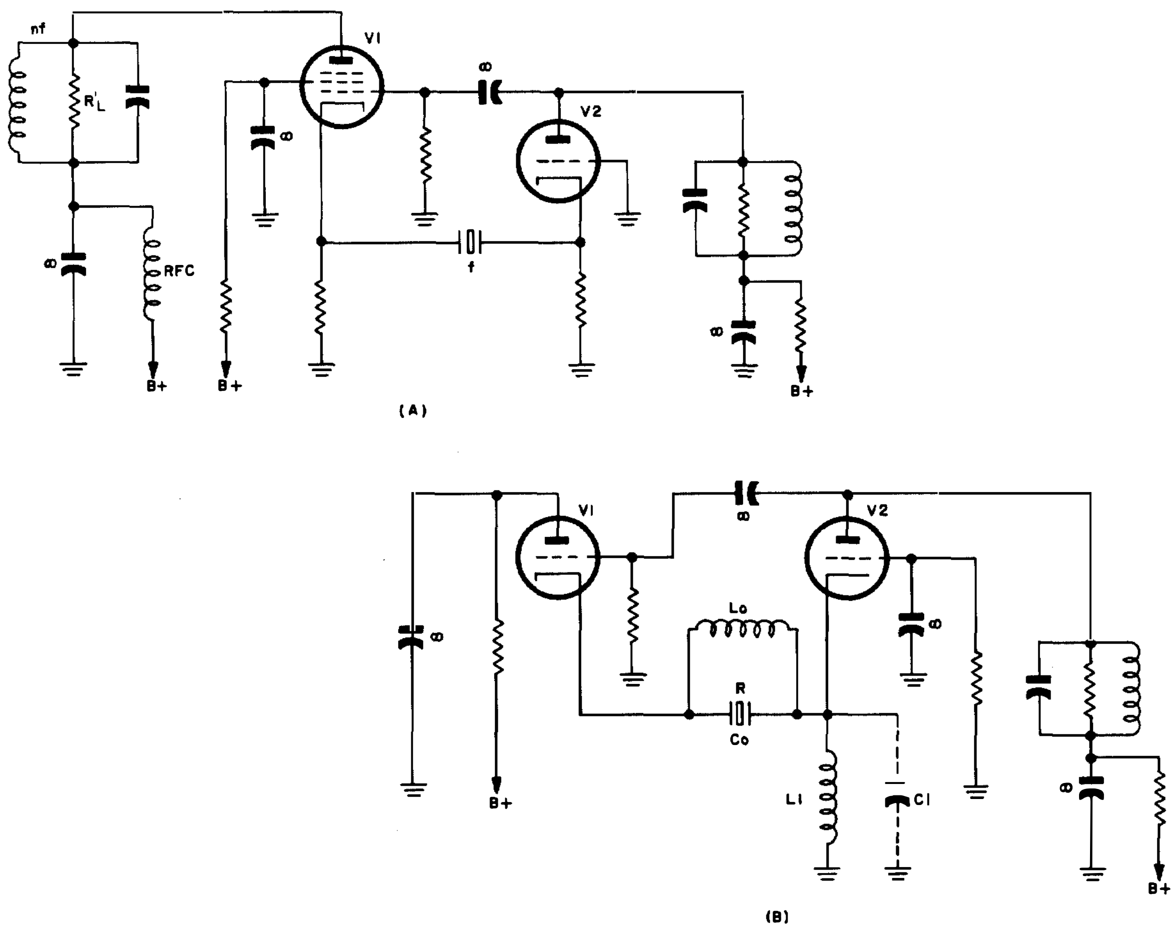


Figure 1-174. (A) Electron-coupled Butler circuit. (B) Butler circuit having common cathode ground return

antiresonant with the distributed capacitance, C_1, as is L_o with C_o. Gridleak bias is employed with both tubes. However, the grid of V_2 is still kept at r-f ground through the bypass capacitance. It can be seen that equation 1—378 (26),

$$\frac{Z_{g2}\, Z_L\, Z_k\, g_{m1}}{Z_2\, Z_f} = 1$$

when applied to the common-ground return circuit, becomes

$$\frac{Z_{g2}\, Z_L\, g_{m1}}{Z_2} = 1 \qquad \text{1—388 (1)}$$

since Z_k is equal to Z_f. Also, since $Z_2 = \dfrac{R_{p2} + Z_L}{\mu_2 + 1} \approx \dfrac{1}{g_{m2}}$, since $Z_{g2} = \dfrac{R_2 Z_2}{R_2 + Z_2}$, where R_2 now represents the antiresonant impedance of the L_1C_1 combination, it can be shown that equation (1) can be expressed as

$$Z_L\, g_{m1} - \frac{1}{R_2\, g_{m2}} = 1 \qquad \text{1—388 (2)}$$

If the product $R_2 g_{m2}$ is large compared with 1, at equilibrium g_{m1} must approximately equal $1/Z_L$.

1-389. The schematics of a number of Butler circuits employed by the military services are shown in figure 1-175. Circuit (A) is a receiver heterodyne oscillator that can be switched from crystal to manual operation simply by shorting out the crystal. C_1 is an r-f bypass capacitor that prevents the inductor L_1 from shorting out the cathode bias developed across R_1. L_1 is designed to be antiresonant with the grounded-grid amplifier cathode-to-ground capacitance. C_2 is inserted to resonate with the distributed inductance of the crystal leads and feed-back circuit. C_3 is a split-stator capacitor which permits the use of a grounded rotor, thereby reducing intersectional capacitances. Since the r-f current through C_3 is small, wiping contacts can be used for grounding without introducing noise. C_4 is simply a trimmer which permits adjustment of the effective inductance of L_2. The plate side of C_3 effectively has a relatively large fixed component due to the stray plate-to-ground capacitance. To keep the plate tank balanced, an equal amount of fixed capacitance, C_5, is added to the other side of C_3. R_2 is added to suppress parasitic oscillations. As can be seen, the output is obtained by split-load operation of the cathode follower. Since the output is delivered to a mixer circuit, where the effective load might be expected to undergo slight changes, it is probable that the loaded cathode-follower plate circuit is less frequency sensitive than the finely balanced tank in the V_2 plate circuit. Since the balanced tank must be sufficiently selective to stabilize the frequency during manual operation, it cannot be loaded as would normally be done. Although the crystal is not active in the circuit during manual operation, it cannot be removed without resulting in an increase in frequency. This is due to the decrease in the cathode-to-ground capacitance that results when the crystal unit is removed.

1-390. Figure 1-175 (B) (C) (D) (E) (F) is a composite arrangement of five different oscillators, none of which have all the components shown. For example, the B^+ return of V_1 is through R_9 in (B), (E), and possibly (F); it is through R_6 in (C), and also in (D), except that R_2 and R_6 are one and the same in the latter circuit, although the actual circuit is not indicated in the schematic shown. In circuit (E) the F_4 output is cathode-coupled to a mixer tube (6AK5W). An r-f choke is connected between the cathode and ground, not the resistance R_{10}. In circuit (B), C_5 is actually composed of two 1.5-$\mu\mu$f capacitors in series. The F_3 output is developed across a tuned tank identical with and also inductively coupled to, the plate tank in the V_2 plate circuit. L_5 in circuit (B) thus serves as a transformer primary. The F_3 output is fed to the grid of one and to the cathode of a second 6AG5 mixer stage. The heterodyned output of the first is 20 to 30 mc, and that of the second is 4.8 to 5.7 mc.

1-391. The circuit shown in figure 1-175 (G) is a carefully designed experimental model that was built and tested during an investigation of h-f and v-h-f oscillators by a research team headed by W. A. Edson at the Georgia Institute of Technology. A breadboard model of this circuit was operative at frequencies as high as 150 mc, with crystal resistances as high as 500 ohms. Circuit (G) was found to have a frequency stability of 0.22 parts per million per volt change in the high-voltage supply. The frequency was controlled at 126 mc, the ninth harmonic of a 14-mc fundamental. The shunt capacitance of the test crystal was 12 $\mu\mu$f, and the series resistance after tuning out the capacitance was 300 ohms. Oscillations could not be sustained at plate voltages below 50 volts, and the frequency instability increased greatly at voltages above 85 volts. Note that circuit (K) in figure 1-175, which has been designed to operate with low- and medium-frequency, fundamental-mode crystal units, is not a true Butler circuit in that neither tube is operated as a cathode follower.

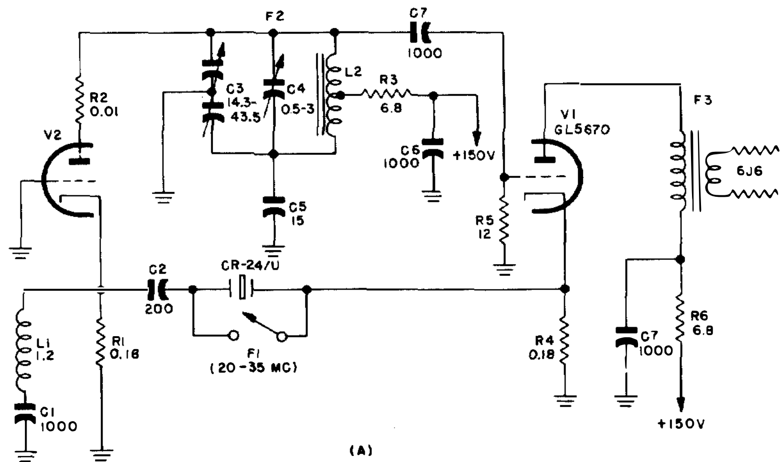
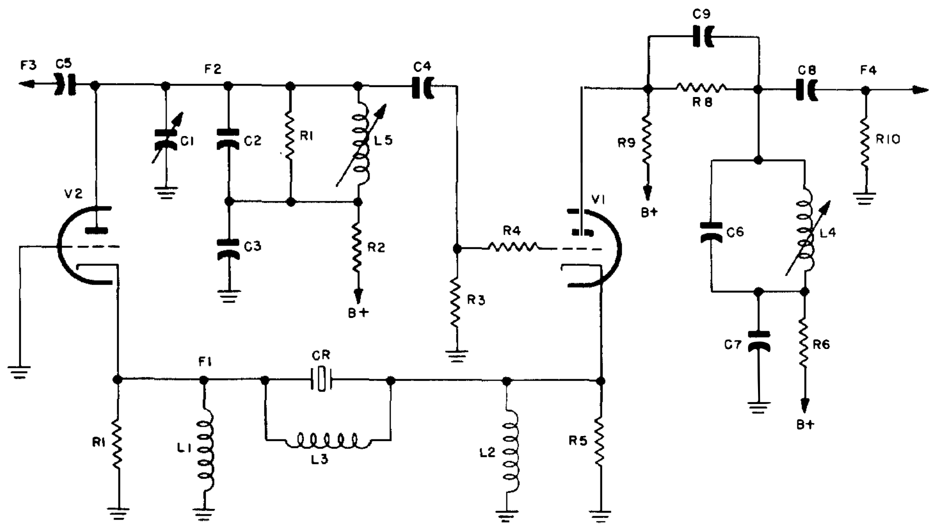


Figure 1-175. Modifications of Butler oscillator. Dotted lines in circuit (G) indicate stray capacitances

Fig.	Equipment	Purpose	F_1	F_2	F_3	F_4	CR	R_1	R_2	R_3	R_4	R_5	R_6	R_7	R_8
(A)	Radio Receiver R-266/ URR-13	Hetero-dyne oscillator	20.3–34.9	$2F_1$			CR-24/U (5th mode)	0.18	0.01	6.8	0.18	12	6.8		
(B)	Receiver Trans-mitter RT-178/ ARC-27	3rd trans-mitter osc or 2nd re-ceiver osc for hetero-dyning	25.7–34.7	F_1	?	NA	CR-23/U	0.1	?	100	0	0.15	NA	12	NA
(C)	Receiver-Trans-mitter RT-178/ ARC-27	1st guard receiver local osc	37.266	F_1	NA	$2F_1$	CR-23/U	0.39	0.12	47	0	0.22	0.12	6.8	0
(D)	Receiver-Trans-mitter RT-173/ ARC-33	1st mon-itor osc of trans-mitter M.O.	0.8333	F_1	$4F_1$–$13F_1$	F_1	CR-28/U	3.3	33	100	0	1	Same re-sistor as R_2	∞	0.22
(E)	Receiver-Trans-mitter RT-173/ ARC-33	Guard-channel hetero-dyne oscill-ator-doubler	55.668–58.169	F_1	NA	$2F_1$ and $4F_1$	CR-32/U	0.12	8.2	33	0	0.33	NA	27	∞
(F)	Radio Receiver R-252A/ ARN-14	1st heter-odyne oscillator	44.275–57.275	F_1	NA	$2F_1$	CR-23/U		8.2	4.7	0.1		?	∞	?

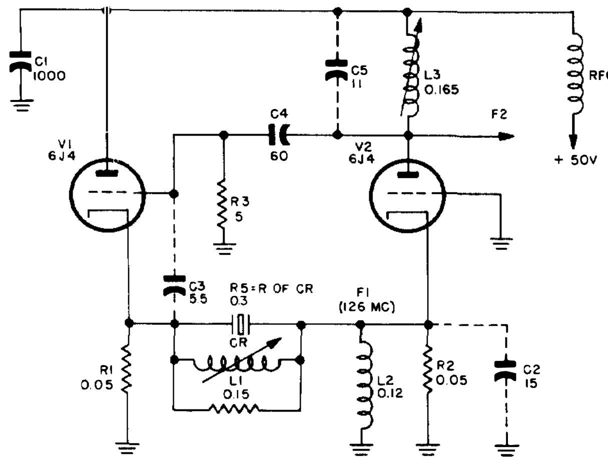
Circuit Data for Figure 1–175. F in mc. R in kilohms. C in $\mu\mu f$. L in μh. NA (not applicable) means that no connections of any kind exist between points indicated. Question mark (?) indicates that schematic of the associated part of the circuit is not available.



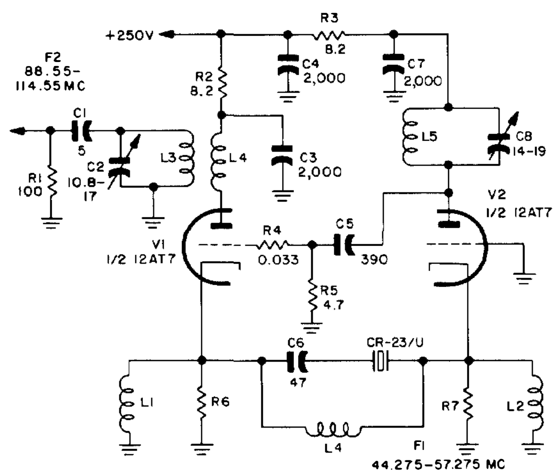
(B) (C) (D) (E) (F)

Figure 1-175. Continued

R_9	R_{10}	C_1	C_2	C_3	C_4	C_5	C_6	C_7	C_8	C_9	L_1	L_2	L_3	L_4	L_5	V_1 V_2
		1000	200	14.3-43.5	0.5-3.0	15	1000	1000			1.2					GL5670
1	NA	1-8	7	1500	47	0 75	NA	NA	NA	NA	∞	∞	∞	NA	2-4	12AT7
NA	27	0	20	3000	100	NA	24	500	20	0	∞	∞	∞			12AT7
NA	470	0	104		470	470	100	25,000	470	0	∞	∞	∞			5670
8 2	Cathode of mixer	0	5		10	NA	1.6-5	∞	10	10	∞	∞				5670
?	?	2-7	12	2000	100	NA	?	?	?	?				?		12AT7



(G)

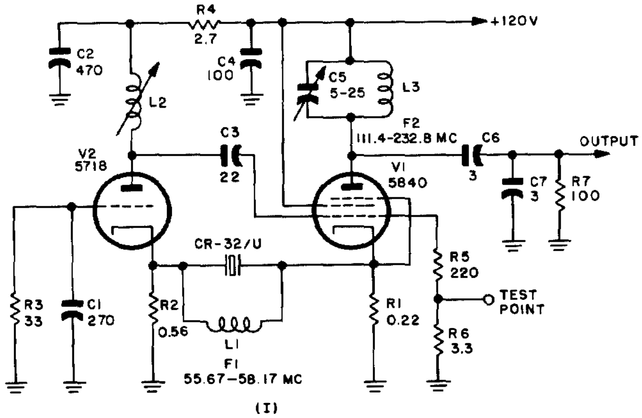


(H)

Figure 1-175. Continued

Fig.	Equipment	Purpose	F_1	F_2	F_3	F_4	CR	R_1	R_2	R_3	R_4	R_5	R_6	R_7	R_8
(G)	Experi-mental osc		126	F_1			9th har-monic; series resist-ance equal to R_5	0.05	0.05	5	1.2	0.3			
(H)	Radio Receiver R-540/ARN-14C	H-F in-jection osc	44.275-57.275 (14 crystals)	$2F_1$			CR-23/U	100	8.2	8.2	0.033	4.7			
(I)	Radio Set AN/ARC-34 (XA-1)	Guard channel injection osc and multiplier	55.67-58.17	$2F_1$ and $4F_1$			CR-32/U	0.22	0.56	33	2.7	220	3.3	100	
(J)	Radio Set AN/ARN-21(XN-2)			F_1			CR-23/U	0.22	0.22	10	1				

Circuit Data for Figure 1-175. F in mc. R in kilohms. C in $\mu\mu$f. L in μh. NA (not applicable) means that no connections of any kind exist between points indicated. Question mark (?) indicates that schematic of the associated part of the circuit is not available.



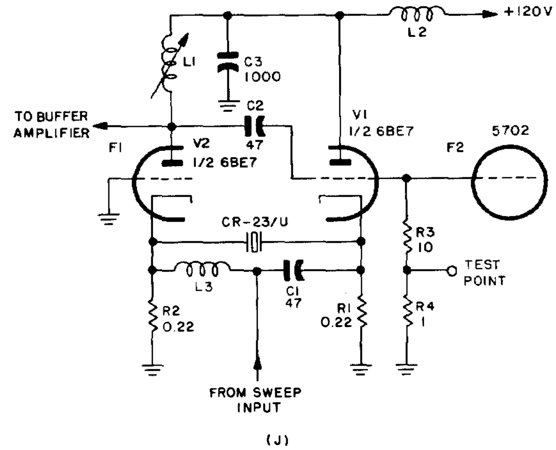
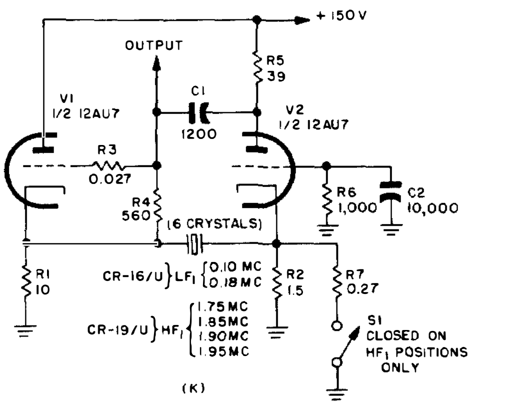


Figure 1-175. Continued

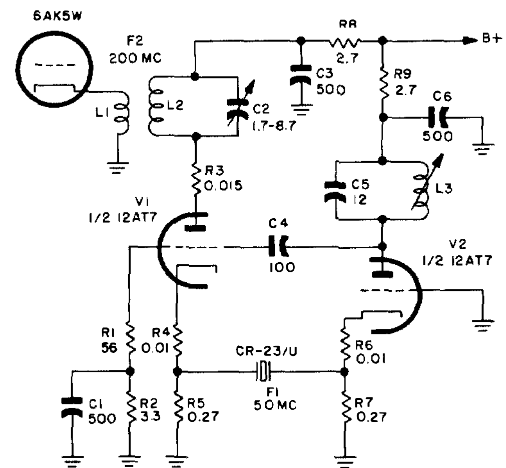
R_9	R_{10}	C_1	C_2	C_3	C_4	C_5	C_6	C_7	C_8	C_9	L_1	L_2	L_3	L_4	L_5	V_1 V_2
		1000	15	5 5	60	11					0 15	0.12	0.165			6J4 each
		5	10.8-17	2000	2000	390	47	2000	14-19							12AT7
		270	470	22	100	5-25	3	3								5840 (V_1) 5718 (V_2)
		47	47	1000												6BE7

Fig.	Equipment	Purpose	F_1	F_2	F_3	F_4	CR	R_1	R_2	R_3	R_4	R_5	R_6	R_7	R_8
(K)	Signal Generator SG-34(XA)/UP	L-F and m-f osc	0.10 0.18 1.75 1.85 1.90 1.95				CR-16/U (LF_1) CR-19/U (HF_1)	10	1.5	0.027	560	39	1000	0.27	
(L)	Signal Generator SG-13/ARN	200-mc generator for mixing with lower freq. signals	50	$4F_1$			CR-23/U	56	3.3	0.015	0.01	0.27	0.01	0.27	2.7

Circuit Data for Figure 1—175. F in mc. R in kilohms. C in $\mu\mu f$. L in μh. NA (not applicable) means that no connections of any kind exist between points indicated. Question mark (?) indicates that schematic of the associated part of the circuit is not available.



(K)



(L)

Figure 1-175. Continued

Transformer-Coupled Oscillator

1-392. At the present time, the transformer-coupled crystal oscillator (see figure 1-176) is not being widely used. It was during the v-h-f oscillator investigation at the Georgia Institute of Technology for the Signal Corps in 1950, mentioned in the last paragraph, that the transformer-coupled oscillator appeared to be the most promising for all-around versatility and general-purpose use. First, there is the advantage of a single-tube oscillator. Secondly, for low-power output (about four times the crystal power), the frequency stability has been found to be slightly superior to that of the average Butler circuit. Thirdly, with properly designed phase-compensating networks, an untuned pass band of 10 mc is possible. Finally, with a relatively small sacrifice in frequency stability, the circuit design can be such that the power output is increased several fold without exceeding the recommended maximum drive level of the crystal unit. Although the transformer-coupled oscillator can perform satisfactorily at lower frequencies, its chief application is for control and generation of harmonic-mode frequencies above 20 mc. The oscillator is generally designed for class-C operation. A significant disadvantage is that the circuit design for optimum performance characteristics —characteristics that can be approximately duplicated from one oscillator to another of similar design—is generally more difficult to achieve than in other oscillator circuits. This is due chiefly to the difficulty in predicting the effective input impedance of tubes operated class C at frequencies where transit-time and stray-capacitance effects become appreciable. As a result, the theoretical and actual equilibrium conditions frequently are found to differ to a greater degree than in the average series-mode oscillator. More cut-and-try experimentation may prove necessary than would otherwise be the case. The operating principle of the grounded-cathode, transformer-coupled oscillator is closely allied to that of the grounded-grid and grounded-plate versions. The discussion and equations for the transformer-coupled oscillators

R_9	R_{10}	C_1	C_2	C_3	C_4	C_5	C_6	C_7	C_8	C_9	L_1	L_2	L_3	L_4	L_5	V_1 V_2
		1200	10,000													12AU7
2.7		500	1 7-8.7	500	100	12	500									12AT7

are, for the most part, based upon the analysis by W. A. Edson.

PHASE CONSIDERATIONS OF TRANSFORMER-COUPLED OSCILLATOR

1-393. Referring to figure 1-176, the useful load, represented by R_L, is connected across the secondary of the plate transformer. The chief function of R_1 is to improve the frequency stability by lowering the resistance of the crystal circuit, and to improve the amplitude stability by reducing the effect of variations in the input resistance of the tube. The parameter a is simply the constant of proportionality relating R_L to R_1. C_1 and C_2 are capacitors for tuning out the leakage inductance of the plate and grid transformers, respectively. The leakage inductance is equal to the high-side inductance multiplied by $(1—k^2)$, where k is the coefficient of coupling. It can be directly measured at the low side of the transformer when the high side is shorted. Both transformers can be, simply, tapped coils. The crystal impedance is assumed to be the series-resonance impedance, R. The effective turns ratios, N_p and N_g, can be so chosen that C_{pg} and C_o annul each other's effects. With the circuit properly designed, the tube operates into a resistive load, C_p being antiresonant with the dampened coil L_p. I_s, I_L, I_c, E_L, and E_o (the voltage across the crystal) are in phase with E_p. The grid transformer thus provides the required 180-degree phase shift between E_p and E_g.

1-394. Where the resistance of the crystal series arm is not small compared with the shunt reactance, X_{Co}, the effects of C_o can be annulled by the conventional method of connecting an inductor across the crystal unit, by means of mutual inductance between the plate and grid transformers (to be discussed in connection with the grounded-grid oscillator), or by balancing the effects of C_o against those of C_{pg}. When the circuit is properly balanced, the crystal unit operates at the resonant frequency of the series arm. When C_o is balanced against C_{pg}, the leading component, I_o, of the current through the crystal—that part through C_o—passes in its entirety, through C_2 and the primary of the grid transformer. Similarly, the current through C_{pg}, I_{pg}, passes in its entirety, through the secondary of the grid transformer. For this to occur, the voltages induced by the two currents in each section of the grid transformer must exactly

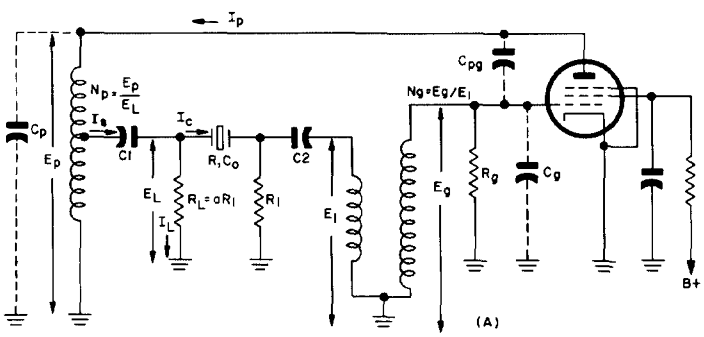


Figure 1-176. (A) Basic circuit of transformer-coupled oscillator

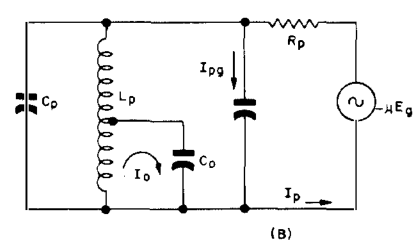


Figure 1-176. (B) Equivalent parallel-resonant plate circuit when crystal capacitance, C_o, is balanced by plate-to-grid capacitance, C_{pg}

annul each other. The transformer then appears as a short circuit to both currents. As indicated in figure 1-176 (B), C_o becomes equivalent to a capacitor shunting the ground-connected half of L_p, thereby effectively increasing L_p, and C_{pg} becomes equivalent to an increase in C_p equal to C_{pg}. The balanced state is reached when

$$I_o = N_g I_{pg} \qquad 1\text{—}394\ (1)$$

or

$$\omega C_o E_L = N_g \omega C_{pg} E_p \qquad 1\text{—}394\ (2)$$

or

$$C_{pg} = \frac{C_o}{N_g} \cdot \frac{E_L}{E_p} = \frac{C_o}{N_g N_p} \qquad 1\text{—}394\ (3)$$

Equation (3) can generally be realized with practical values of N_g and N_p if C_{pg} is on the order of 0.05 to 0.1 $\mu\mu$f. Such values of C_{pg} can be obtained with screen-grid tubes but not with triodes unless a d-c blocked inductive arm is connected between plate and grid to annul most of the capacitance.

GAIN REQUIREMENTS OF TRANSFORMER-COUPLED OSCILLATOR

1-395. Referring to figure 1-176 (A), it can be seen that at equilibrium,

$$G_1 G_2 G_3 G_4 = \frac{E_p}{E_g} \cdot \frac{E_L}{E_p} \cdot \frac{E_1}{E_L} \cdot \frac{E_g}{E_1} = 1 \qquad 1\text{—}395\ (1)$$

where

$$G_1 = \frac{E_p}{E_g} = \frac{I_p Z_p}{E_g} = \frac{\mu Z_p}{R_p + Z_p} = g_m Z_p \qquad 1\text{—}395\ (2)$$

$$G_2 = \frac{E_L}{E_p} = \frac{1}{N_p} \qquad 1\text{—}395\ (3)$$

$$G_3 = \frac{E_1}{E_L} = \left(\frac{R_1 R_g'}{R_1 + R_g'}\right) \cdot \left(\frac{1}{R + \dfrac{R_1 R_g'}{R_1 + R_g'}}\right) \approx \frac{R_1}{R + R_1} \qquad 1\text{—}395\ (4)$$

$$G_4 = N_g \qquad 1\text{—}395\ (5)$$

R_g' in equation (4) is the effective resistance of the grid transformer as it appears in parallel with R_1. It is equal to E_1^2 divided by the grid losses. Thus,

$$R_g' = \frac{E_1^2}{P_g} = \frac{(\text{eff}) R_g E_1^2}{E_g^2} = \frac{(\text{eff}) R_g}{N_g^2} \qquad 1\text{—}395\ (6)$$

where (eff) R_g is the effective grid resistance which takes into account the gridleak losses and the transit-time loading. For small values of N_g, the R_g' losses can be considered negligible. The plate power can be assumed to be,

$$P_p = I_s E_L \approx \frac{E_L^2 (R_L + R + R_1)}{R_L (R + R_1)} = \frac{E_L^2 (R + R_1 + a R_1)}{a R_1 (R + R_1)} \qquad 1\text{—}395\ (7)$$

and

$$Z_p = \frac{E_p^2}{P_p} = \frac{a N_p^2 R_1 (R + R_1)}{R + R_1 + a R_1} \qquad 1\text{—}395\ (8)$$

on combining equations (2), (3), (4), (5), and (8), we find that at equilibrium

$$G_1 G_2 G_3 G_4 = \frac{g_m a N_p N_g R_1^2}{R + R_1 (1 + a)} = 1 \qquad 1\text{—}395\ (9)$$

Q DEGRADATION IN TRANSFORMER-COUPLED OSCILLATOR

1-396. In Edson's analyses of series-mode oscillators, he employs a useful term which he calls the Q degradation of the crystal unit. It is defined

$$D = \frac{R_c}{R} \qquad 1\text{—}396\ (1)$$

where R_c is the total resistance which the crystal must operate into. In the transformer-coupled oscillator, assuming that the transformer impedances that the crystal faces are large compared with R_L and R_1 of figure 1-176,

$$D = \frac{R + R_1 (1 + a)}{R} \qquad 1\text{—}396\ (2)$$

As discussed in paragraph 1-241, the frequency stability of the series-resonant crystal unit is directly proportional to the Q of the crystal circuit. It is therefore directly proportional to $\frac{Q}{D}$, where Q is the Q of the crystal, itself. Thus, with a given crystal unit, the frequency stability varies inversely with D. If the minimum Q of a crystal unit is estimated from the maximum permissible C_o, from the frequency, the harmonic, the particular crystal element, and the maximum permissible series resistance, and if the required frequency stability is known, then the maximum permissible D can be determined from the random phase shifts to be expected during operation. In the average circuit, it is sufficient, simply, to keep D as low as possible consistent with the output desired. When R_1 is expressed as a function of a, R, and D, the loop gain as defined by equation 1—395 (9) becomes

$$G_1\,G_2\,G_3\,G_4 = \frac{g_m\,a\,N_p\,N_g\,R\,(D-1)^2}{D\,(a+1)^2} \qquad 1\text{—}396\ (3)$$

LOAD-TO-CRYSTAL POWER RATIO OF TRANSFORMER-COUPLED OSCILLATOR

1-397. The ratio of the load to the crystal power is

$$P_L/P_c = \frac{I_L^2\,R_L}{I_c^2\,R} = \frac{(D+a)^2}{a\,(a+1)\,(D-1)} \qquad 1\text{—}397\ (1)$$

It is generally desired to have the power ratio as high as is consistent with satisfactory frequency stability. With a given value of D, the ratio becomes large as a is made small. If the design is based on obtaining a given minimum output with a given value of D when the crystal-unit resistance is a maximum (minimum D), the value of a can be determined by equation (1), and equation 1—396 (3) can be used to determine the value of N_pN_g most likely to produce the required g_m for the crystal to be driven at the desired level.

BROAD-BAND CONSIDERATIONS IN THE TRANSFORMER-COUPLED OSCILLATOR

1-398. For broad-band untuned operation it is important to have Z_p and Z_g (the impedance *faced* by the grid) and the plate and grid capacitances as small as possible. Assuming that R_g is large compared with the resistance appearing across the secondary of the grid transformer.

$$Z_g = \frac{N_g^2\,R_1\,(R + a\,R)}{a\,R_1 + R_1 + R} \qquad 1\text{—}398\ (1)$$

or

$$Z_g = \frac{N_g\,(D + 1/a)}{g_m\,N_p\,(D-1)} \qquad 1\text{—}398\ (2)$$

Also, Z_p can be expressed as

$$Z_p = \frac{N_p\,(D+a)}{g_m\,N_g\,(D-1)} \qquad 1\text{—}398\ (3)$$

To keep Z_p and Z_g low, it is desirable that N_p and N_g be as small as possible. For a given value of g_m, the product N_pN_g can be a minimum when $\frac{a}{(a+1)^2}$ (see equation 1—396 (3)) is a maximum. The maximum occurs when a = 1. With a, D, and g_m decided upon, equations (2) and (3) give mutually minimum values when $N_p = N_g$. Thus, for broad-band operation, let

$$Z_p = Z_g = \frac{D+1}{g_m\,(D-1)} \qquad 1\text{—}398\ (4)$$

D should be as large as possible consistent with the required frequency stability.

FREQUENCY STABILITY OF THE TRANSFORMER-COUPLED OSCILLATOR

1-399. The frequency-stability equations for the transformer-coupled oscillator are

$$\frac{d\omega}{\omega} = -\frac{\omega Z_p D}{2Q}\,dC_p \qquad 1\text{—}399\ (1)$$

and

$$\frac{d\omega}{\omega} = -\frac{\omega Z_g D}{2Q}\,dC_g \qquad 1\text{—}399\ (2)$$

where Q is the Q of the crystal. When equations 1—398 (2) and (3) are multiplied by D, it can be shown the DZ_p is a minimum when

$$D = 1 + \sqrt{1+a} \qquad 1\text{—}399\ (3)$$

and DZ_g is a minimum when

$$D = 1 + \sqrt{1 + 1/a} \qquad 1\text{—}399\ (4)$$

For both to be a minimum simultaneously, *a* must be equal to 1, which means that

$$D = 1 + \sqrt{2} = 2.414 \qquad 1\text{—}399\ (5)$$

Under these conditions, the power ratio, as given by equation 1—397 (1), becomes

$$\frac{P_L}{P_c} = 4.12 \qquad 1\text{—}399\ (6)$$

For larger power outputs, the value of *a* must be decreased and N_p increased. Since the expected variations in the grid capacitance are generally larger than those in the plate capacitance, it is usually desirable to favor the grid circuit insofar as the frequency stability is concerned.

1-400. The greatest probability that the effect of a random variation in the grid capacitance will be canceled by a random variation in the plate capacitance occurs when the fractional change in frequency due to the average ΔC_p is equal to that due to the average ΔC_g. If the average ΔC_g is on the order of 10 times the average ΔC_p, then Z_p should be equal to $10Z_g$. Equations 1—399 (1) and (2) will then represent equal average variations in frequency.

DESIGN PROCEDURE FOR TRANSFORMER-COUPLED OSCILLATOR

1-401. The procedure to follow in designing a transformer-coupled oscillator depends upon the principal objectives to be sought in the design. That is, some fixed requirement serves as a starting point, and the design proceeds from there. One limitation that will be common to all the circuits is that the crystal power rating not be exceeded. This requirement, then, in the general case, can be the initial design consideration. The crystal power is

$$P_c = I_c^2 R = \frac{E_g^2 R}{N_g^2 R_1^2} \qquad 1\text{—}401\ (1)$$

Also

$$P_c = I_c^2 R = \frac{E_L^2 R}{(R_1 + R)^2} = \frac{E_p^2 R}{N_p^2 (R_1 + R)^2}$$

$$= \frac{g_m^2 E_g^2 Z_p^2 R}{N_p^2 (R + R_1)^2}$$

or

$$P_c = \frac{g_m^2 E_g^2 a^2 R_1^2 N_p^2}{R D^2} \qquad 1\text{—}401\ (2)$$

Multiplying equation (1) by equation (2) and taking the square root, we have

$$P_c = \frac{g_m E_g^2 a N_p}{D N_g} \qquad 1\text{—}401\ (3)$$

Now, $(g_m E_g^2)$ is assumed equal to $(I_p E_g)$, which, in turn, is principally a function of the excitation voltage and the plate characteristics of the tube to be used. By equation 1—395 (9) (also by equating equation (1) to equation (2))

$$g_m = \frac{R_c}{a R_1^2 N_p N_g} \qquad 1\text{—}401\ (4)$$

R_c, remember, is equal to RD, the total resistance that the crystal unit operates into. From equation (4) it can be seen that if $R_c (= R + R_1 (1 + a))$ is large compared with the maximum $R (= R_m)$, the equilibrium transconductance will be approximately the same for all values of R. On the other hand, if the minimum D is small, the value of g_m at minimum R may be as much as one-half its value for $R = R_m$. Note that some change must occur in g_m, and hence in E_g, if R varies and the rest of the parameters remain constant. To ensure class-C operation for all values of R, let the class-A value for g_m equal twice the equilibrium g_m according to equation (4), with R_c assumed to be a maximum. With the circuit so designed, the amplitude of the oscillations will build up until the tube is cut off a fair proportion of each cycle. Even if class-A operation is desired, a reasonable difference should be allowed between the rated transconductance of the tube and the estimated equilibrium value when R_c is maximum. This should be sufficient to allow for all expected tolerances in the plate characteristics and in the tuning of the oscillator circuit. The percentage variations in R from one crystal unit to the next is not quite as great in the v-h-f crystals as in the lower-frequency elements, since it is more important that the maximum permissible resistance be kept as small as practicable.

1-402. The ideal design would permit the percentage variations in E_g^2 to exactly equal in magnitude the percentage variations in R. Under these conditions, the crystal power, as indicated in equation 1—401 (1) would be the same for each crystal unit. To approach such a design, g_m, as a function of E_g, would have to be known for the particular tube. Such an analysis is beyond the present discussion, but the method to be used would be quite similar to that employed in the analysis of the effects of different values of crystal resistance for the Pierce circuit. As a rule of thumb, the average crystal R in the v-h-f range can be assumed to equal $\frac{R_m}{2}$. If the transformer-coupled oscillator is designed to drive the crystal unit at 50 per cent of its maximum rated power for the average R, there is little danger that crystal units having other values of resistance will be overdriven. For broadband operation the crystal power will be approximately directly porportional to R. When D is small the crystal power increases as R decreases as long as the percentage increase in E_g^2 is greater than the percentage decrease in R. With $R = \frac{R_m}{2}$ and $P_c = \frac{P_{cm}}{2}$, equation 1—401 (1) gives an assumed value for E_g^2, thus

$$(\text{ave})\ E_g^2 = \frac{P_{cm} R_1^2 N_g^2}{R_m}$$

or

$$(\text{ave})\ E_g = N_g R_1 \sqrt{P_{cm}/R_m} \qquad 1\text{—}402\ (1)$$

Note that equation (1) also gives the value of E_g which would exist if a crystal of maximum resistance were driven at its rated level. However, equation (1) is less important for determining E_g than it is for determining N_g. It is assumed that a class-C value of g_m has been agreed upon. An approximate value of E_g corresponding to the chosen g_m is thus already determined. After the value of R_1 is decided upon, equation (1) can serve to determine N_g.

1-403. The design procedure followed so far can generally be applied to any transformer-coupled oscillator. A v-h-f pentode and a harmonic series-mode crystal unit are selected. Regulated screen and plate voltages for the tube are decided upon. An approximate class-C value of g_m and the corresponding E_g are estimated, and average values of P_c and R are assumed. Once that D and a have been selected according to the particular requirements of the oscillator, the $\frac{N_p}{N_g}$ ratio can be determined from equation 1—401 (3), and N_g from equation 1—402 (1). Or, in case a particular $\frac{N_p}{N_g}$ ratio is to be preferred, a and D can be determined with the aid of equation 1—401 (3). An alternative approach, which is the one to be followed when using the table in paragraph 1-404, is to first determine optimum values for D, a, and N_p/N_g, assume an average crystal $R = \frac{R_m}{2}$ and a crystal power equal to $\frac{P_{cm}}{2}$, and use equation 1—401 (3) to determine the value of $g_m E_g^2$ ($= I_p E_g$). The next problem is to determine what value of E_g will produce the required value of $I_p E_g$ when the tube is operating into a plate impedance that is small relative to the tube R_p. The equations relating I_p to E_g in the analysis of the Pierce circuit, or rather the basic methods used to derive the equations, are applicable here if modified properly. Since some trial-and-error will be required regardless, it may well be preferable to determine the correct E_g empirically. The selected vacuum tube can be driven by an external generator having a variable output at a frequency near the actual frequency for which the oscillator is to be designed. The tube should operate into a small resistive impedance and the gridleak resistance should be the same as that to be used in the final design. E_g should be varied until the measured I_p is such that the product $I_p E_g$ agrees with the value computed from equation 1—401 (3). With E_g approximately known, N_g can be determined by means of equation 1—401 (1).

1-404. The dimensionless equations listed below relate the various parameters. These equations, some of which have already been given, are also useful in determining the various circuit voltages and currents when E_g is known, or the requirements thereof, and in comparing the characteristics of oscillators of different design.

$$g_m Z_p = \frac{N_p (D + a)}{N_g (D - 1)} \qquad 1\text{—}404\ (1)$$

$$g_m Z_g = \frac{N_g (D + 1/a)}{N_p (D - 1)} \qquad 1\text{—}404\ (2)$$

$$N_p^2 g_m R = \frac{N_p D (a + 1)^2}{N_g a (D - 1)^2} \qquad 1\text{—}404\ (3)$$

$$N_g^2 g_m R = \frac{N_g D (a + 1)^2}{N_p a (D - 1)^2} \qquad 1\text{—}404\ (4)$$

$$R_l/R = \frac{D - 1}{a + 1} \qquad 1\text{—}404\ (5)$$

$$P_L/g_m E_g^2 = \frac{N_p (D + a)^2}{N_g D (D - 1) (a + 1)} \qquad 1\text{—}404\ (6)$$

$$P_L/P_c = \frac{(D + a)^2}{a (a + 1) (D - 1)} \qquad 1\text{—}404\ (7)$$

The following table, prepared by the v-h-f oscillator research team at the Georgia Institute of Technology, lists the quantitative relations that hold for five typical designs of the transformer-coupled oscillator. D, a, and the N_p/N_g ratio are predetermined to provide optimum or practical operating characteristics according to five different objectives.

A. Symmetrical circuit design that yields minimum values of $Z_p D$ and $Z_g D$. By equations 1—399 (1) and (2), this design permits maximum frequency stability if the average variations in C_p and C_g are approximately equal. Low power output. Narrow bandwidth.

B. Nonsymmetrical grid and plate impedances. Designed for optimum frequency stability when (ave) ΔC_g = 10 (ave) ΔC_p. Low power output. Narrow bandwidth.

C. Nonsymmetrical grid and plate impedances. Maximum frequency stability when (ave) ΔC_g = 10 (ave) ΔC_p and the power output is 10 times that in design B, but the stability is less than that in designs A and B.

D. Symmetrical circuit. Broad-band untuned operation. Provides small values for $Z_p C_p$ and $Z_g C_g$. Tubes require large transconductance and small input and output capacities. Z_p and Z_g are one-half those in design A. Average ΔC_p assumed equal to average ΔC_g. Frequency stability below average. Power output low, but greater than that in designs A and B.

E. Nonsymmetrical design except that $Z_p = Z_g$. High power output—same as that in design C. Broad-band untuned operation. Frequency stability below average.

TRANSFORMER-COUPLED OSCILLATOR DESIGNS

Parameter	Value of Parameter				
	Design A	Design B	Design C	Design D	Design E
Z_p/Z_g	1	10	10	1	1
a	1	1	0.0985	1	0.24
D	2.414	2.414	2.414	10.65	10.65
N_p/N_g	1	$\sqrt{10}$	7.07	1	1.165
g_mZ_p	2.414	7.63	12.55	1.207	1.315
g_mZ_g	2.414	0.763	1.255	1.207	1.315
g_mZ_pD	5.83	18.4	30.3	12.85	14.10
g_mZ_gD	5.83	18.4	30.3	12.85	14.0
$N_p^2g_mR$	4.828	15.25	104.0	0.457	0.854
$N_g^2g_mR$	4.828	1.525	2.10	0.457	0.63
P_L/P_C	4.12	4.12	41.2	7.05	41.2
R_1/R	0.707	0.707	1.285	4.825	7.78
$\frac{P_L}{g_mE_g^2}$	1.705	5.4	11.9	0.662	1.08

MODIFICATIONS OF TRANSFORMER-COUPLED OSCILLATOR

1-405. Four modifications of the basic transformer-coupled oscillator are shown in figure 1-177. Circuits (A), (B), and (C) are experimental models that were designed at the Georgia Institute of Technology, but not in accordance with the designs given in paragraph 1-404. Circuit (D) is an oscillator in actual use that has been designed specifically to operate with Crystal Unit CR-24/U without driving the crystal beyond its Military-Standard level. Figure 1-177(A) is a low-power circuit intended to be operated within a band of ±2 per cent of 55 mc. When tested with a number of crystal units having overtone frequencies between 50 and 60 mc, the circuit, with L_p and L_g adjusted to be antiresonant with C_p and C_g, respectively, at 55 mc, showed the following operating characteristics. Figure 1-177(D) shows a slug-tuned transformer-coupled oscillator that employs a battery-operated subminiature tube. This oscillator is used in Radio Receiver-Transmitter RT-159A/URC-4.

f (mc)	Harmonic	C_o ($\mu\mu$f)	R (ohms)	Stability (ppm/volt)	Δf (kc)	P_L (mw)
50	5	5	48	0.6	1.5	100
54	9	5	60	0.18	0.5	80
54.8	7	6	80	0.11	0.0	50
58.3	7	8	82	0.625	1.4	40
60	3	13	25	0.66	1.7	40

Δf is the difference between the operating frequency and the series-resonance frequency of the crystal unit as measured with Crystal Impedance Meter TS-683/TSM. Figure 1-177(B) is a circuit intended for broad-band untuned operation with a center frequency at 63 mc. Satisfactory operation was obtained on a crystal plug-in basis from 53 to 73 mc. The designers recommend that the phase-compensating network, which was selected with the help of chart I, page 445, "Network Analysis and Feedback Amplifier Design," Bode, be shifted from the grid to the plate circuit. Figure 1-177(C) is a circuit designed to provide an output approaching 1 watt at 50 mc. Due largely to difficulties in predicting the input resistance of the vacuum tube, the differences between the initial

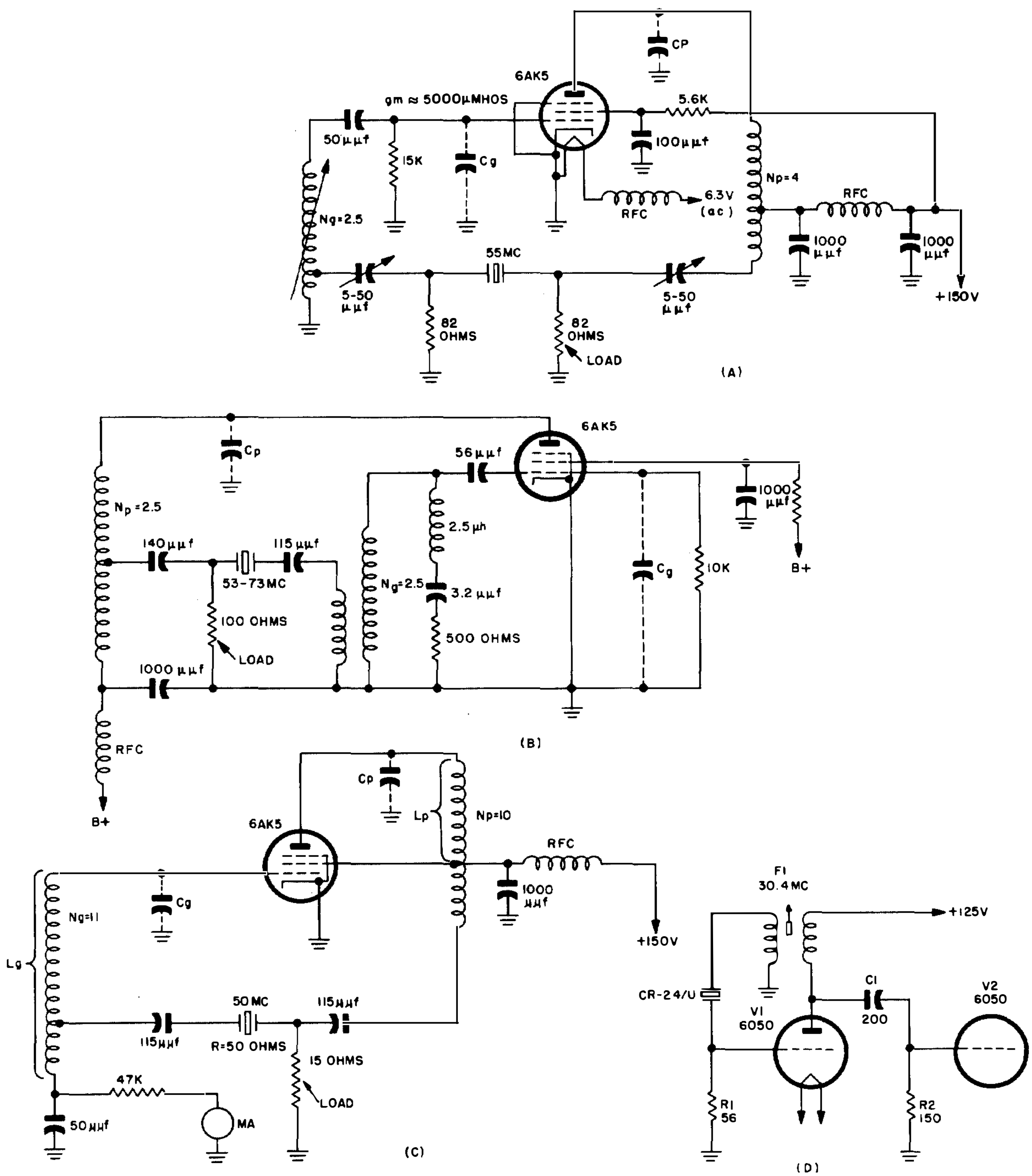


Figure 1-177. Modifications of transformer-coupled oscillator for: (A) High stability. (B) Untuned broad band. (C) Large output. (D) Slug-tuned, narrow band

design and the final adjustments (the latter shown in figure 1-177(C)) were quite large. The operating characteristics of the circuit are given below:

Frequency stability = 0.3 ppm/volt
D-C power input = 1.8 watts
R-F power output = 0.6 watt
Crystal power = 0.02 watt
Efficiency = 33 per cent
$P_L/P_c = 30$
Grid bias = —19 volts

Grounded-Grid Oscillator

1-406. The grounded-grid oscillator, or single-tube Butler oscillator (see figure 1-178) is very similar in principle to the transformer-coupled oscillator. In fact, it can be described as a transformer-coupled oscillator in which the feedback is cathode-coupled to the input. The cathode coupling eliminates the need for phase reversal in either the plate or input transformer. As indicated in figure 1-178, the input transformer can be reduced to an r-f choke. For the plate circuit, an autotransformer permits a maximum coefficient of coupling. Either a triode or pentode can be used. Although operable at lower frequencies, the oscillator is usually employed in the v-h-f range. The low input impedance reduces the effect of variations in the cathode capacitances. Unless the tube is operated class C, the crystal operates into a relatively low-resistance circuit. The power output is lower than that obtainable with a transformer-coupled oscillator, but for small outputs the frequency stability is quite high and the load is more readily shielded from the input. This latter feature reduces the possibility that the oscillator will operate at frequencies other than that of the desired mode of the crystal. The shunt capacitance of the crystal unit should be compensated by a broad-band anti-resonant inductor, or by other means. The most dependable method for use over a wide range of frequencies is to introduce mutual inductance between the input transformer or choke and the plate transformer. This can be attained by winding the cathode inductor on the same form as the plate transformer is wound. The correct coefficient of coupling between the input and plate inductors is the one that permits the crystal to vibrate at its true motional-arm resonance. For a theoretical discussion of this mode of capacitance compensation, see Edson et al. The grounded-grid oscillator is most advantageous to use when maximum compactness and simplicity are desired in a low-power, broad-band, untuned v-h-f oscillator. There is the very important additional advantage that the oscil-

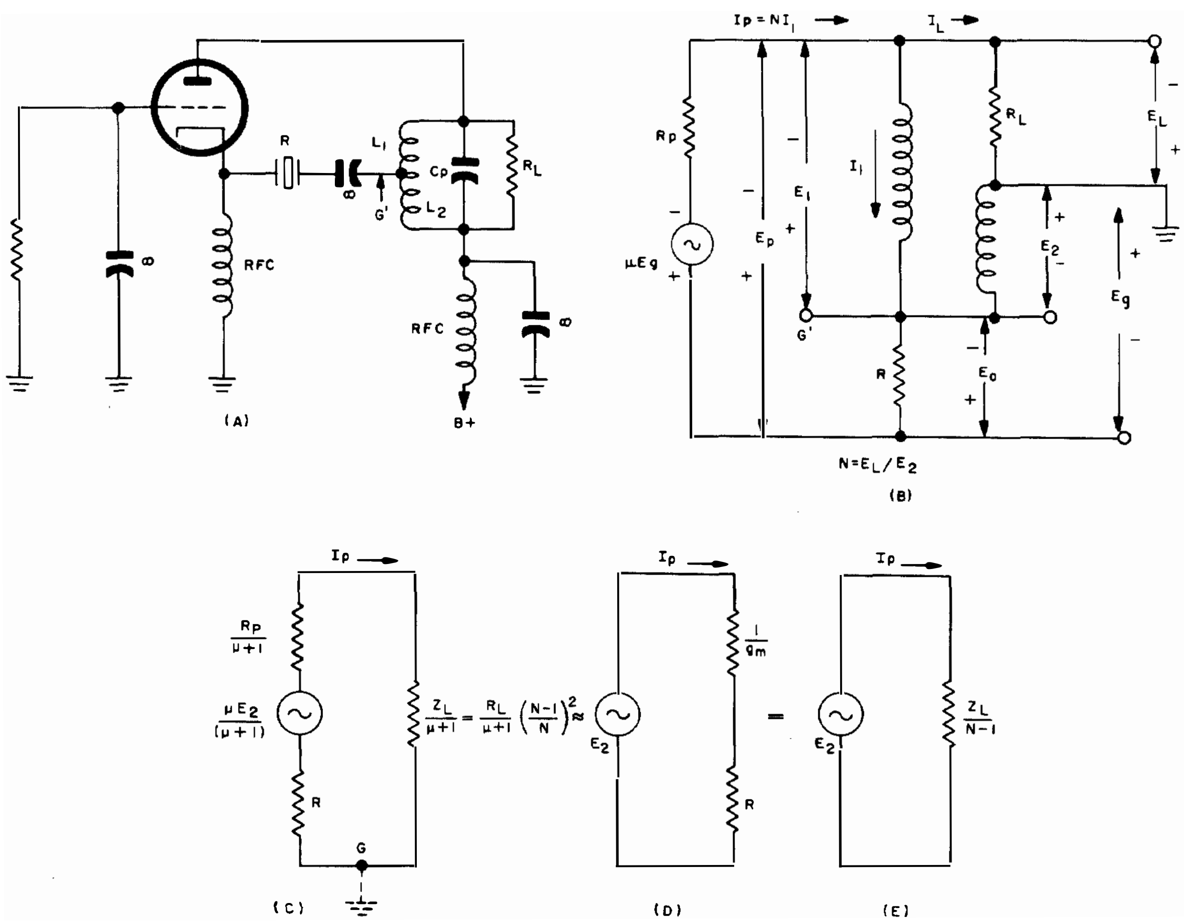


Figure 1-178. Basic grounded-grid oscillator and equivalent circuit progressively simplified

lator can be readily designed so that both the output amplitude and the frequency stability are virtually independent of the crystal resistance.

ANALYSIS OF GROUNDED-GRID OSCILLATOR

1-407. Referring to figure 1-178, it will be assumed that the plate circuit is tuned to the desired harmonic, series-resonance frequency of the crystal, that the resistance in the plate tank is negligible, that the grid current is negligible, that the autotransformer coefficient of coupling is unity, that the cathode-to-ground capacitance is antiresonant with the cathode choke, and that the r-f current through the choke is negligible. Under these conditions the equivalent r-f circuit of the oscillator is that shown in figure 1-178(B). All voltage symbols are treated as unsigned. The polarities shown correspond to the instantaneous polarities that hold during the positive alternations of the r-f grid voltage, E_g. The current arrows point in the instantaneous direction of the in-phase electron flow. The reactive component of the plate tank current is not represented, although in reality it is primarily the "flywheel" current that produces the voltages E_1 and E_2. This reactive current in the tank is that which flows through C_p and is equal to $\omega C_p E_L$. If R_L were reduced to zero, there would be no reactive current and the transformer would effectively short-circuit the crystal to the plate. Of course, oscillations could not exist under these conditions, if for no other reason than the fact that E_2 would be reduced to zero and E_g, being simply the voltage across the crystal unit, would be displaced 180 degrees from the phase required for oscillations to be maintained. E_2 must be greater than E_o. The important feature to remember is that insofar as the resistive component of the current is concerned the action of the autotransformer is the same as that of a conventional transformer when the primary and secondary circuits are connected in parallel as shown in figure 1-178(B). Note, however, that the turns ratio, N, is defined as the ratio of the *total* turns to the turns comprising L_2.

1-408. The power fed to the transformer is simply

$$P_L = I_L E_L = I_p E_1 \qquad 1\text{—}408\ (1)$$

Now,

$$E_L = E_1 + E_2 = E_1 + \frac{E_L}{N}$$

or

$$E_L = \frac{NE_1}{N-1} \qquad 1\text{—}408\ (2)$$

So

$$I_L = \frac{I_p E_1}{E_L} = \frac{(N-1)\, I_p}{N} \qquad 1\text{—}408\ (3)$$

And since

$$I_p = I_L + I_1 \qquad 1\text{—}408\ (4)$$

We have

$$I_p = \frac{(N-1)\, I_p}{N} + I_1 \qquad 1\text{—}408\ (5)$$

or

$$I_p = NI_1 \qquad 1\text{—}408\ (6)$$

and

$$I_L = I_1\,(N-1) \qquad 1\text{—}408\ (7)$$

The ratio of the output to the crystal power is

$$P_L/P_c = I_L^2\, R_L/I_p^2\, R = \frac{R_L}{R}\left(\frac{N-1}{N}\right)^2 = Z_L/R \qquad 1\text{—}408\ (8)$$

Note that for a given turns ratio, the power ratio is directly proportional to the resistance ratio. The term $R_L\left(\frac{N-1}{N}\right)^2$ is simply the equivalent load resistance, Z_L, that the transformer presents to I_p. The total impedance across the vacuum tube is thus

$$Z_p = R_L\left(\frac{N-1}{N}\right)^2 + R = Z_L + R \qquad 1\text{—}408\ (9)$$

1-409. It is convenient to imagine that the ground connection is at the point G′ in figure 1-178. That is, let G′ be our point of reference. Insofar as the r-f circuit is concerned such a supposition requires no alteration in the currents and voltages involved, but it does simplify the visualization of the circuit charactistics. The supposition does not mean that there is no difference in the r-f potential between G′ and the actual ground. With G′ in figure 1-178 (B) assumed to be the ground connection, it can be seen that the tube is effectively connected as a cathode follower except that no load is taken from the cathode circuit. The crystal R is the cathode resistance, and E_2 is the voltage input to the grid circuit. As discussed in the analysis of the two-tube Butler circuit and as illustrated in figure 1-173, the cathode-follower type of circuit can be represented by an equivalent circuit in which the plate-circuit resistance, exclusive of the cathode resistance, is equal to $\frac{1}{\mu + 1}$ times the actual resistance. Also, the equivalent generator voltage is $\frac{1}{\mu + 1}$ times the conventional value that would be assumed if the cathode resistance were not present. As applied to the grounded-grid oscillator, the equivalent circuit is that shown in figure 1-178 (C). Care must be taken not to interpret the power supplied to the plate-circuit resistances as being reduced by a factor of $\frac{1}{\mu + 1}$. With μ and R_p assumed large compared with unity and Z_p, respectively, circuit (C) reduces to circuit (D).

LOOP GAIN IN GROUNDED-GRID OSCILLATOR

1-410. At equilibrium

$$G_1 G_2 G_3 G_4 = \frac{E_o}{E_2} \cdot \frac{E_1}{E_o} \cdot \frac{E_L}{E_1} \cdot \frac{E_2}{E_L} = 1 \qquad 1\text{—}410\ (1)$$

where

$$G_1 = \frac{E_o}{E_2} = \frac{\mu R}{R_p + Z_L + R\,(\mu + 1)} \approx \frac{R}{\frac{1}{g_m} + R} \qquad 1\text{—}410\ (2)$$

$$G_2 = E_1/E_o = Z_L/R = \frac{R_L\,(N-1)^2}{RN^2} \qquad 1\text{—}410\ (3)$$

$$G_3 = E_L/E_1 = N/(N-1) \qquad 1\text{—}410\ (4)$$

$$G_4 = E_2/E_L = 1/N \qquad 1\text{—}410\ (5)$$

Equation (2) simply represents the gain of a cathode follower. It can be derived from the equivalent circuits in figure 1-178 in a manner similar to the cathode follower gain derivation in the analysis of the two-tube Butler circuit. In combining equations (2), (3), (4), and (5), we have

$$G_1 G_2 G_3 G_4 = \frac{N^2\,(R_p + R + \mu R)}{R_L\,(N-1)\,(\mu + 1 - N)} = 1 \qquad 1\text{—}410\ (6)$$

If the approximate value for G_1 can be assumed, the gain equation can be expressed as

$$\frac{1}{g_m} = \frac{R_L\,(N-1)}{N^2} - R = \frac{Z_L}{N-1} - R \qquad 1\text{—}410\ (7)$$

ACTIVITY CONSIDERATIONS IN GROUNDED-GRID OSCILLATOR

1-411. From equation 1—410 (7) we can predict the relative activity to be expected with different load and crystal resistances. When oscillations first start, the term $1/g_m$ is a minimum. The amplitude increases until g_m is reduced by the increase in grid bias to the value that makes equation 1—410 (7) hold. The greater the difference between the initial, zero-bias value of g_m and the equilibrium value, the greater will be the final amplitude. The initial g_m of the tube should therefore be as large as possible if a maximum output is desired. For a given crystal R, the output is increased by increasing R_L and making the ratio $\frac{N-1}{N^2}$ a maximum, which occurs when $N = 2$. Assuming that $N = 2$, i.e., that the transformer coil is center-tapped at the connection to the crystal, equation 1—410 (7) becomes

$$1/g_m = R_L/4 - R = Z_L - R \qquad 1\text{—}411\ (1)$$

In equation 1—408 (8), it was found that the power ratio, to which we shall assign the symbol $r\left(= \frac{P_L}{P_c}\right)$, is equal to Z_L/R. On substituting the value rR for Z_L in equation (1) we find that

$$r = 1 + \frac{1}{g_m R} \qquad 1\text{—}411\ (2)$$

By equations (1) and (2), which hold only when $N = 2$, we see that increasing the amplitude by increasing R_L results in simultaneously increasing the power ratio. Also, note that the minimum power ratio can be predetermined as a function of the class-A value of g_m and the maximum permissible R of the crystal unit. If variations in R from crystal unit to crystal unit are not to have a large effect upon the output amplitude, $Z_L\left(= \frac{R_L}{4}\right)$ must be large compared with the maximum crystal R. In other words, the minimum power ratio should be as large as possible consistent with the requirements of frequency stability.

CRYSTAL DRIVE-LEVEL CONSIDERATIONS IN GROUNDED-GRID OSCILLATOR

1-412. If P_{cm} is the rated maximum power of the crystal unit, the maximum permissible I_p for a given R is

$$(\max)\ I_p = (\max)\ E_g\, g_m = \sqrt{P_{cm}/R} \qquad 1\text{—}412\ (1)$$

Consequently,

$$(\max)\ E_1 = (\max)\ I_p\, Z_L = Z_L \sqrt{P_{cm}/R} \qquad 1\text{—}412\ (2)$$

By equations 1—410 (4) and (5)

$$(\max)\ E_L = \frac{(\max)\ E_1\ N}{N-1} = \frac{Z_L\ N\sqrt{P_{cm}/R}}{N-1} \qquad 1\text{—}412\ (3)$$

and

$$(\max)\ E_2 = (\max)\ E_L/N = \frac{Z_L\sqrt{P_{cm}/R}}{N-1} \qquad 1\text{—}412\ (4)$$

Assuming that N = 2, we have

$$\frac{(\max)\ E_L}{2} = (\max)\ E_1 = (\max)\ E_2 = \frac{R_L}{4}\sqrt{\frac{P_{cm}}{R}} \qquad 1\text{—}412\ (5)$$

The maximum permissible excitation voltage of the tube with a center-tapped transformer is

$$(\max)\ E_g = \frac{(\max)\ I_p}{g_m} = (\max)\ E_2 - (\max)\ E_o = \left(\frac{R_L}{4} - R\right)\sqrt{\frac{P_{cm}}{R}} \qquad 1\text{—}412\ (6)$$

Equation (6) is to be interpreted as giving the maximum *permissible* E_g for a *given* value of R. The smaller the value of R, the larger will be the permissible value of E_g. For v-h-f crystal units in which the shunt capacitance of the crystal is not compensated the minimum R encountered may be on the order of $R_m/5$, where R_m is the rated maximum. However, with capacitance compensation all values of R will be less than R_m, and the minimum may well be on the order of $R_m/9$ or less. The plate characteristics of the vacuum tube and the electrode voltages must be such that an excitation voltage equal to the maximum permissible E_g, as defined by equation (6), does not cause an effective I_p greater than $\sqrt{P_{cm}/R}$. With a sharp-cutoff, grid-leak-biased tube operating into a plate impedance that is small compared with R_p, the effective I_p remains essentially constant as the peak-to-peak amplitude of E_g is increased from a value equal to $|E_{co}|$ to a value equal to $2|E_{co}|$, where E_{co} is the cutoff bias. As discussed in paragraph 1-312 in connection with the Pierce circuit, the crest amplitude of I_p between the crest values of E_g equal to $\frac{|E_{co}|}{2}$ and $|E_{co}|$ remains approximately equal to $\frac{I_{bm}}{2}$, where I_{bm} is the zero-bias plate current. There is a small maximum approximately equal to 0.54 I_{bm} (see equation 1—312 (21)), but for all practical purposes it can be assumed that I_p is constant for all values of g_m between class-A and class-B operation. Since the effective g_m is equal to I_p/E_g, the doubling of E_g without changing I_p is equivalent to halving g_m. If oscillations can be maintained at all, the slightest tolerance allowed in g_m ensures that the amplitude will build up until the excitation voltage overlaps the lower bend in the E_cI_b curve. Thus, for a sharp-cutoff tube the minimum equilibrium E_{gm} $(= \sqrt{2}E_g)$ will very nearly equal $\frac{|E_{co}|}{2}$. Since $1/g_m = Z_L - R$ when $N = 2$, any value of Z_L greater than $2R_m$ can ensure that the r-f plate current, and hence the output, will be the same for all values of R falling within the crystal specifications. All that need be done is to design the circuit for class-A operation on the assumption that $R = R_m$. This requires the use of a sharp-cutoff vacuum tube with tube voltages such that

$$I_{bm} < 2\sqrt{\frac{2\ P_{cm}}{R_m}} \qquad 1\text{—}412\ (7)$$

and the design of the load and transformer network such that

$$\left|\frac{E_{co}}{I_{bm}}\right| = \frac{Z_L}{N-1} - R_m \qquad 1\text{—}412\ (8)$$

Under the conditions defined by equations (7) and (8), a crystal having a maximum R will be driven at or under the rated maximum drive, depending upon whether I_{bm} is equal to, or less than, the value specified in equation (7). If the crystal unit is replaced by another of lower resistance, the crystal current and the output will remain essentially the same. Since E_2 will also be unchanged, the increase in the excitation voltage will be entirely that due to the decrease in the voltage across the crystal. The driving power of the crystal will be directly proportional to the crystal R.

FREQUENCY STABILITY OF GROUNDED-GRID OSCILLATOR

1-413. In the equivalent circuit shown in figure 1-178(D), it can be seen that the effective resistance, R_c, of the crystal circuit is $\left(\frac{1}{g_m} + R\right)$. But, $1/g_m = \frac{Z_L}{N-1} - R$, thus

$$R_c = \frac{Z_L}{N-1} - R + R = \frac{Z_L}{N-1} \qquad 1\text{—}413\ (1)$$

If N = 2,

$$R_c = Z_L = R_L/4 \qquad 1\text{—}413\ (2)$$

Equations (1) and (2) assume that the mu of the tube is large compared with unity, and that $R_p >> Z_L$. From equation 1—241 (2), the fractional change in frequency required to compensate a small change, $d\theta$, in the feed-back phase is

$$\frac{d\omega}{\omega} = \frac{R_c\, d\theta}{2\sqrt{L/C}} = \frac{Z_L\, d\theta}{2\,(N-1)\sqrt{L/C}} = \frac{R_L\, d\theta}{8\sqrt{L/C}} \qquad 1\text{—}413\ (3)$$

where L and C are series-arm parameters of the crystal, and equation (2) is assumed to hold. Of significance is the fact that for a given Z_L, the frequency stability is independent of the resistance of the crystal. However, $\left(\frac{Z_L}{N-1}\right)$ must always be greater than R, else the conditions for oscillation as defined by equation 1—410 (7) cannot hold. Thus, although the frequency stability can be considered independent of R for a given Z_L, the effective Q of the crystal circuit must always be less than the lowest actual $Q\left(=\frac{2\omega L}{R_m}\right)$ of the crystal. The larger the g_m of the tube, the more nearly can this limiting value for the effective Q be reached, since the more nearly $\left(\frac{Z_L}{N-1}\right)$ can be made to approach R_m in magnitude. But a large g_m must be accompanied by a low cutoff voltage for the tube, else equations 1—417 (7) and (8) cannot be made to hold and the crystal will be overdriven. Unfortunately — yet not unexpectedly — the requirements for maximum output are the reverse of those for maximum frequency stability. If the only frequency-stability problem were to maintain the circuit Q as high as possible, the output could be increased without decreasing the stability, by making both R_L and N large. This could permit an increase in Z_L without affecting the value of $\left(\frac{Z_L}{N-1}\right)$. On the other hand, if $d\omega/\omega$ in equation (3) is to be kept small, not only $\left(\frac{Z_L}{N-1}\right)$ but also $d\theta$ must be kept to a minimum. Because the input impedance of the tube is very low, changes in the cathode capacitance have a negligible effect on the feed-back phase. The principal variations in the phase are due to changes in the plate and load capacitance. To reduce these effects to a minimum, Z_L must be as small as possible. Its smallest permissible value will occur when $\left(\frac{N-1}{N^2}\right)$ is a maximum; that is, when $N = 2$. Letting $N = 2$ and $R_L = 8R_m$, Z_L will equal $R_L/4 = 2R_m$. The crystal-circuit Q for all values of R will then be one-half the minimum crystal Q to be expected for the particular type of crystal unit. The output will approximately equal $2P_{cm}$ for all values of crystal R. Much larger outputs can be obtained without greatly reducing the frequency stability, by the use of remote-cutoff tubes. With these tubes the r-f plate current can be made to vary inversely with the square root of the crystal resistance. Under these conditions, it would be the crystal power that remains constant and the output power that varies with R. If $P_L = 2P_{cm}$ when $R = R_m$, $P_L = 18P_{cm}$ when $R = R_m/9$.

DESIGN PROCEDURE FOR GROUNDED-GRID OSCILLATOR

1-414. The design procedure depends considerably upon the special requirements to be met by the circuit. As a concrete example assume that a low-power, 50-mc oscillator requiring a minimum of circuit components and an output amplitude that will not be greatly affected by a replacement of the crystal unit with another of the same frequency is desired. The grounded-grid oscillator is probably the best suited for such a purpose. Assume further that a frequency tolerance of ±0.01 per cent is required without temperature control for all temperatures between —40 and +90 degrees centigrade. Crystal Unit CR-24/U with a frequency tolerance of ±0.005 per cent between —55 and +90 degrees centigrade should be able to provide the required stability. So also will Crystal Unit CR-23/U, but the former unit is mounted in the coaxial holder, the HC-10/U, which is generally to be preferred because its lower inherent shunt capacitance should permit a higher average Q. There is no guarantee of this, since the maximum C_o is 7 $\mu\mu$f in each case; however, the CR-24/U employs the 5th harmonic and the CR-23/U the 3rd harmonic (thinner crystal) for the 50-mc frequency. The greater CR-24/U L/C ratio should more than offset its slightly higher R_m. Nevertheless, a check should be made to see if crystal units of either type having the desired frequency are currently being manufactured or have been manufactured in the past. If not, serious consideration should be given to the possibility of employing a different frequency. The cost of the crystal unit will be less if it is already in production, and the risk that an undue amount of experimentation will be required to produce a crystal unit that meets the military standards at an unexplored frequency can be avoided. If the crystal unit is expected to withstand considerable mechanical shock, the CR-24/U must be used, regardless.

1-415. Assume that a 50-mc CR-24/U crystal unit has been selected. According to Military Standard MS91380, $R_m = 75$ ohms and $P_{cm} = 2$ mw. By equation 1—412(1)

$$\text{(max) } I_p = 10^3\sqrt{\frac{0.002}{75}} \approx 5.2 \text{ ma}$$

According to equation 1—412(7)

$$I_{bm} \leqslant 2\sqrt{2} \times 5.2 = 14.7 \text{ ma}$$

Assume that the most available tube is the 6AU6, sharp-cutoff, miniature pentode. Operated at 250 plate volts, a screen voltage of 140 volts provides a zero-bias plate current of approximately 15 ma. The cutoff bias will be approximately —5 volts. Assuming a value of $N = 2$, by equation 1—412(8)

$$Z_L = 75 + \frac{5}{0.015} = 410 \text{ ohms}$$

and

$$R_L = 4\ Z_L = 1640 \text{ ohms}$$

The power ratio when $R = R_m$ will be

$$\text{(min) } r = \frac{Z_L}{R_m} \approx 5.5$$

The power output for all values of R will be

$$P_L = \text{(min) } r\ P_{cm} = 11 \text{ mw}$$

The crystal unit will operate into an effective resistance equal to 410 ohms. The effective g_m of the tube will vary from approximately 3000 μmhos, when R is a maximum, to approximately 2500 μmhos, when R is a minimum. If greater frequency stability is required, R_L can be decreased by approximately three-fourths, so that $Z_L = 300$ ohms. With this value of Z_L, when R is maximum g_m will be 4450 μmhos. By increasing the screen voltage to 150 volts or slightly greater, an r-f plate current very nearly equal to the maximum permissible for R_m can be attained. As R is decreased I_p and P_L increase somewhat, but the crystal unit will not be overdriven.

MODIFICATIONS OF THE GROUNDED-GRID OSCILLATOR

1-416. Figure 1-179 shows four different designs of the grounded-grid oscillator which were built and successfully tested at the Georgia Institute of Technology. Because of the high initial transconductances, 0.011 μmho for the 6J4 and 0.009 for the 6AH6, the oscillation amplitude of these circuits would drive the average Military Standard crystal unit beyond the recommended maximum level. This does not mean that a standard crystal unit will necessarily be in danger of being shattered by the circuits shown, but that the frequency, resistance, and freedom from spurious modes could not be guaranteed by the test standards. To employ the circuits illustrated in figure 1-179, different tubes, or plate-supply voltages may need to be used.

1-417. Figure 1-179(A) is a narrow-band oscillator with the load connected across the secondary of the plate transformer. Except for the fact that the input "transformer" (L_k, having a 1:1 voltage ratio) provides no phase reversal, the circuit is very similar to that of the basic transformer-coupled oscillator. R_L, connected as shown, is equivalent to a load resistance of N^2R_L connected across L_p. The variable capacitance is for tuning out the transformer leakage inductance. L_k is antiresonant with the cathode capacitance, and the

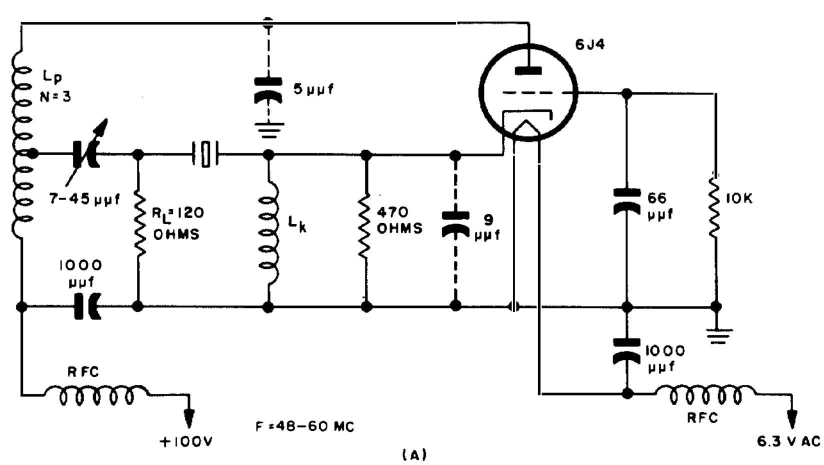
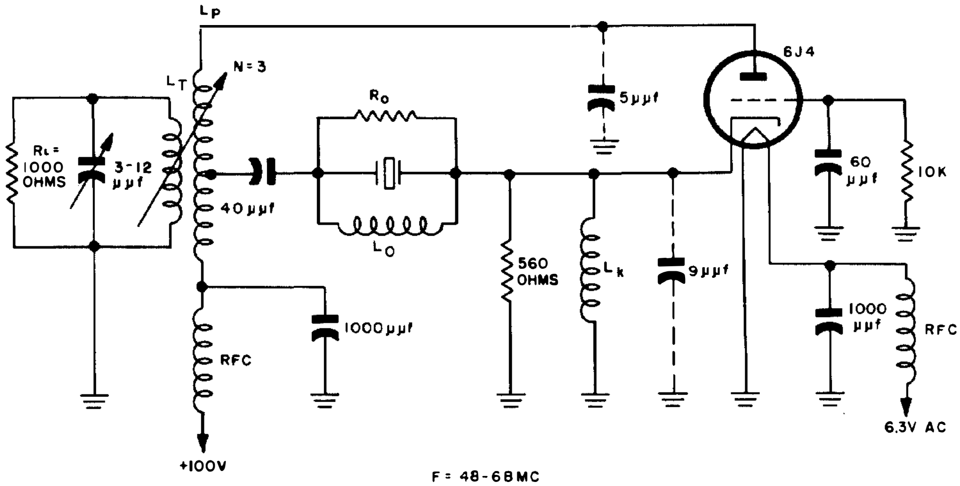
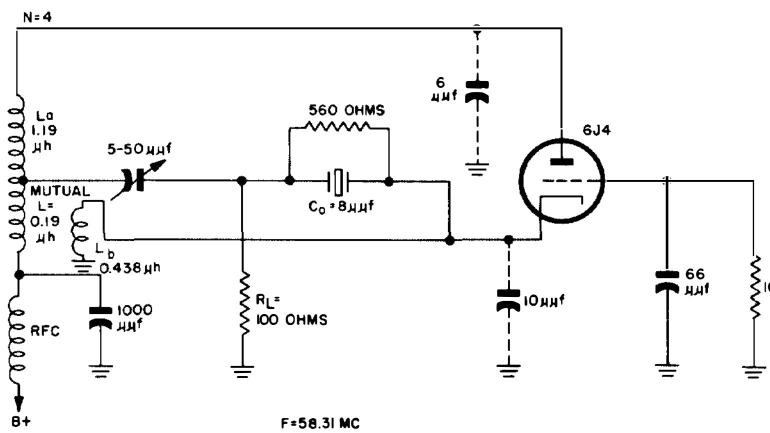


Figure 1-179. Modifications of grounded-grid oscillator. (A) Narrow-band circuit



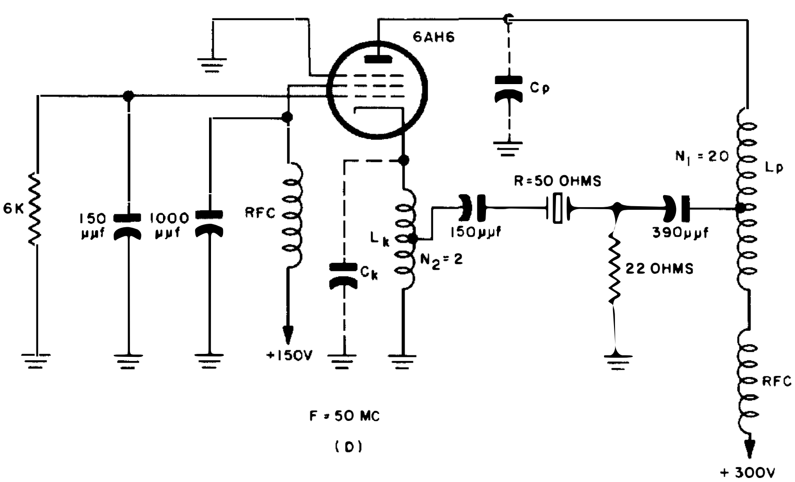
F = 48-68MC

(B)



F=58.31 MC

(C)



F = 50 MC

(D)

Figure 1-179. Modifications of grounded-grid oscillator. (B) Broad-band circuit. (C) Circuit for compensating crystal capacitance by mutual inductance. (D) High-efficiency class-C circuit cathode resistance broadens the tuning. The circuit was tuned for operation at a center frequency of 55 mc. The plate transformer consists of 12 turns of AWG No. 26 PE wire wound on a Miller type 69048 slug-tuned coil form and tapped at 4 turns. The circuit operates class B. The operating data for several different crystal units is given below.

f (mc)	Harmonic	C_o ($\mu\mu$f)	R (ohms)	Stability (ppm/volt)	Δf (kc)	P_L (mw)
48	3	14	45	0.42	3	55
50	3	12.5	35	0.20	1.5	103
54	3	13.5	45	0.11	9	78
58	7	9	32	0.14	0.6	28
60	3	13	25	0.24	1.6	40

The frequency stability is measured in average parts per million per volt when the voltage is changed by 50 volts. Δf gives the deviation observed between the series-resonance frequency, when measured with CI Meter TS-683/TSM, and the actual oscillator frequency. It would seem that the 54-mc crystal, which should show the smallest value of Δf, was influenced by a spurious mode.

1-418. The circuit in figure 1-179 (B) is designed for broad-band untuned operation. L_p consists of 10 turns of No. 30 PE wire, tapped at 3.3 turns and wound on a 0.4-inch-diameter form. The tertiary winding, L_T, consists of 17 turns on a 0.24-inch form that can be slipped inside the L_p core by a screw adjustment. The circuit is first tuned with a 1000-ohm load connected directly across L_p with L_T open. Next, with the circuit connected as shown, the coupling between L_T and L_p is adjusted until the same grid current as before is obtained. All tuning adjustments were made at 57.5 mc. The performance data of this circuit for several different crystals is given below.

f (mc)	Harmonic	C_o ($\mu\mu$f)	R (ohms)	Grid I_C (μamp)	Stability (ppm/volt)	Δf (kc)	P_L (mw)
48	3	14.5	25	58	0.21	5	41
50	3	11	28	85	0.25	—0.5	45
58.31	7	8	80	60	0.01	0	52
65.31	7	10	80	65	0.15	0.2	46
66.65	5	4	65	90	0.21	2.0	50
67.2	7	14	80	62	0.07	4.0	47

1-419. The circuit in figure 1-179 (C) is designed to compensate the capacitance of the crystal unit by mutual inductance between the plate and cathode inductors instead of by a shunt inductor as in circuit (B). L_a consists of 9 close-wound turns of AWG No. 30 PE wire, tapped at 2.5 turns; L_b consists of 4.5 turns of AWG No. 30 PE wire on a thin spacer. The proper coupling adjustment is obtained by substituting a capacitance equal to C_o in place of the crystal and adjusting the circuit to oscillate at the true series-resonance frequency of the motional arm, but only after L_a and L_b have separately been adjusted to resonate with C_p and C_k, respectively, at their computed resonant frequencies ($\omega_a^2 = \frac{1}{L_a C_p}$, $\omega_b^2 = \frac{1}{L_b C_k}$). It can be shown that

$$L_a = \frac{L_p N_1^2 L_o}{(1 - M^2)(L_p + N_1^2 L_o)} \qquad 1\text{—}419\ (1)$$

$$L_b = \frac{L_k N_2^2 L_o}{(1 - M^2)(L_k + N_2^2 L_o)} \qquad 1\text{—}419\ (2)$$

and

$$M^2 = \frac{L_p L_k/N_2^2}{(L_p + N_1^2 L_o)(L_o + L_k/N_2^2)} \qquad 1\text{—}419\ (3)$$

where L_p and L_k are the values of the plate-to-ground and cathode-to-ground inductances, respectively, that would occur if there were no coupling between them, $L_{\bullet}$ is the imaginary shunt inductance that would be required to antiresonate C_o, M is the coefficient of coupling between the plate and cathode inductors, and N_1 and N_2 are the plate and cathode turns ratios, respectively. N_2 is simply equal to unity in the circuit shown. The performance data for the circuit is as follows:

- f = 58.31 mc
- stability = 0.28 ppm/volt
- P_L = 90 mw
- Δf = operating freq minus tested series-resonance freq = —100 cycles
- Δf when C_k increased from 10 to 13 $\mu\mu f$ = —60 cycles
- Δf when C_o increased from 8 to 11 $\mu\mu f$ = —45 cycles
- Δf when C_p increased from 6 to 6.5 $\mu\mu f$ = —90 cycles

1-420. The circuit shown in figure 1-179 (D) is designed for high-efficiency operation as a small class-C power oscillator. L_p consists of 20 turns of AWG No. 28 PE wire wound on a 0.25-inch coil form and tapped at 1 turn. L_k is a 10-turn, 0.25-inch-diameter coil of AWG No. 28 PE wire, tapped at 5 turns. The observed performance data for this circuit is as follows:

- f = 50 mc
- load voltage = 9 volts
- grid bias = —10 volts
- load power = 1.9 watts
- crystal power = 0.08 watts
- frequency stability = 0.6 ppm/volt
- plate dissipation = 3 watts
- efficiency = 63 per cent

The Grounded-Plate Oscillator

1-421. The vacuum-tube circuit of the grounded-plate oscillator shown in figure 1-180 is essentially the same as the two-tube Butler oscillator except that a step-up transformer replaces the grounded-grid amplifier of the Butler circuit. The gain of the Butler grounded-grid tube is thus replaced by the gain, N, of the transformer in figure 1-180 (A). The grounded-plate oscillator is most advantageous when used in the electron-coupled form, as shown in figure 1-180 (C), where the plate circuit can be tuned to provide frequency multiplication. Otherwise, the larger output of the basic transformer-coupled circuit or the greater simplicity of the grounded-grid circuit make these oscillators preferable to the grounded-plate design insofar as obtaining the same order of frequency stability is concerned. The grounded-plate oscillator can be designed for larger outputs by providing a step-up transformer in the cathode circuit and removing the r-f voltage from the gridleak resistor, as is shown in figure 1-180 (B). This permits the cathode-follower to operate into the same output impedance but with a greatly reduced load resistance across the crystal circuit. The output per milliwatt of crystal power is thereby increased. Increasing the power output in this manner makes the oscillator more critical to design and adjust so as to prevent free-running oscillations, particularly if the tube is to be operated class C, where the effective input impedance becomes more or less unpredictable at very high frequencies.

1-422. The over-all gain equation of the oscillator in figure 1-180 (A) is

$$\frac{\mu N R_2 Z_k}{(R + R_2)(R_p + Z_k + \mu Z_k)} = 1 \qquad 1\text{—}422\ (1)$$

where Z_k is the total effective resistance between the cathode and ground. Assuming that the resistance presented by the transformer is equal to R_g/N^2 and is much greater than R_2, we have

$$Z_k \quad \frac{R_1 (R + R_2)}{R_1 + R + R_2} \qquad 1\text{—}422\ (2)$$

The effective resistance into which the crystal operates is

$$R_c = Z_k' + R + R_2 \qquad 1\text{—}422\ (3)$$

where Z_k' is the output impedance of the cathode follower as faced by the crystal. If μ is very large compared with unity,

$$Z_k' = \frac{R_1}{1 + g_m R_1} \qquad 1\text{—}422\ (4)$$

For crystal resistances on the order of 75 ohms or smaller, R_1 and R_2 can also be approximately 75 ohms each. Values of R_1 = 68 ohms, R_2 = 100 ohms, R_g = 200K, and N = 9 have been recommended for use with a 6J4 triode. The shunt capacitance of the crystal unit, as well as that of the grid and cathode, can be compensated if need be by conventional antiresonant inductors. To be preferred is the method described in the discussion of the two-tube Butler circuit—designing the circuit so that

$$g_m Z_k = \frac{C_{gc}}{C_k} \qquad 1\text{—}422\ (5)$$

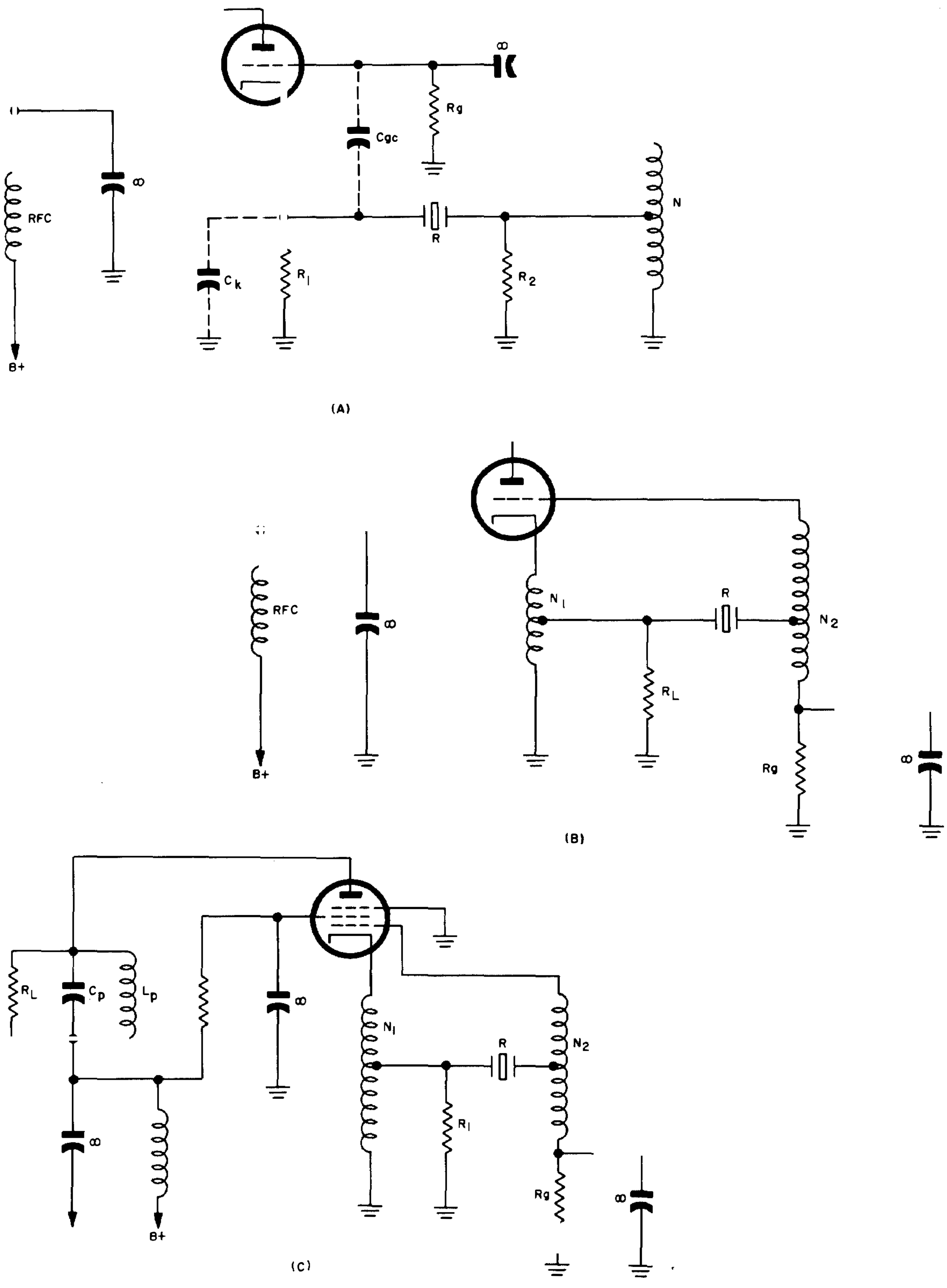


Figure 1-180. Grounded-plate oscillators. (A) Basic circuit. (B) Circuit for increased power output. (C) Electron-coupled circuit

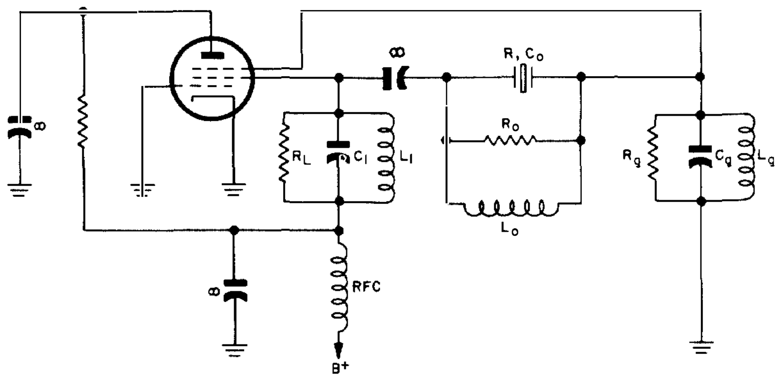


Figure 1-181. Transitron crystal oscillator

Transitron Crystal Oscillator

1-423. The transitron oscillator (see figure 1-181) operates by virtue of the negative transconductance between the suppressor and screen grids of a pentode. The total cathode current of the pentode is little affected by variations in the suppressor voltage, being primarily a function of the potential between the screen and cathode. However, as the suppressor voltage is made more negative, the fraction of the total space current diverted to the screen circuit is increased. The screen voltage therefore tends to follow the suppressor voltage. By connecting a resonant feed-back network between the screen and suppressor, oscillations can be maintained and no phase reversal is necessary. The principal advantage of this circuit is its simplicity and its ability to oscillate with series-mode crystals having comparatively high series resistances. It can be employed in the v-h-f range, but unless the crystal resistances are expected to be abnormally high, the relatively large electrode capacitances and the small transconductance make the performance inferior to that of the transformer-coupled oscillator.

1-424. When the circuit is used with high-resistance crystals it is very important that the crystal shunt capacitance be properly compensated, in order to eliminate the possibility of free-running oscillations. As has been demonstrated by W. A. Edson with the aid of Nyquist diagrams (graphical representations of the over-all loop gain and phase rotation as the frequency is varied from 0 to ∞), the circuit can be designed to permit only one mode of oscillation if, treating g_m as unsigned,

$$\frac{g_m R_L R_g}{R_o + R_L + R_g} < 1 \qquad 1\text{—}424\ (1)$$

and

$$\frac{R_L R_g}{R_o}\left(\frac{C_1 + C_g}{R_o} + \frac{C_1}{R_g} + \frac{C_g}{R_1}\right) \leqslant C_o \qquad 1\text{—}424\ (2)$$

The condition implied by equation (1) when g_m is its maximum possible value means that the loop gain is insufficient to start or maintain oscillations at any frequency unless R_o is effectively decreased (such as being bypassed by the series-resonance R of the crystal) so that the left side of the equation is greater than or equal to unity. Equation (2), when satisfied, means that a zero phase shift in the feedback can occur at only one frequency. Thus, if the circuit is tuned for operation at the desired series-mode frequency of the crystal and equations (1) and (2) are satisfied, spurious oscillations will not be possible.

1-425. Note that C_o is effectively increased by the suppressor-to-screen capacitance, so that L_o must be smaller than would otherwise be the case. C_1 and C_g are simply distributed capacitances to ground. Each of the three parallel combinations are antiresonant at the crystal frequency, so $L_1C_1 = L_oC_o = L_gC_g = \frac{1}{\omega^2}$. Assuming that the antiresonant circuits have impedances R_L, R and R_g, respectively, then if I_{g2} is the r-f screen current, the voltage across the load is

$$E_{g2} = \frac{I_{g2} R_L (R + R_g)}{R_L + R + R_g} \qquad 1\text{—}425\ (1)$$

The r-f suppressor voltage is

$$E_{g3} = \frac{E_{g2} R_g}{R + R_g} = \frac{I_{g2} R_L R_g}{R_L + R + R_g} \qquad 1\text{—}425\ (2)$$

If we assume that E_{g2} is small compared with the d-c screen voltage, and define the suppressor-to-screen transconductance as the change in screen current per change in suppressor-to-cathode voltage—not per change in the suppressor-to-screen voltage—the gain conditions for equilibrium are, by equation (2),

$$\frac{E_{g3}}{I_{g2}} = \frac{1}{g_m} = \frac{R_L R_g}{R_L + R + R_g} \qquad 1\text{—}425\ (3)$$

Of the vacuum tubes available, the 6AS6, which has a suppressor-to-screen transconductance of 1600 μmhos, is probably to be preferred. With this tube, oscillations can be maintained with crystal units having series resistances of well over 1000 ohms. Although oscillations can also be maintained with large values of R_L and R_g, these resistances should be kept as small as practicable so as not to unnecessarily degrade the crystal Q and reduce the frequency stability. The transitron oscillator is also quite useful at low frequencies, particularly with high-resistance crystal units. When a fundamental-mode crystal element is employed, the tuned circuits may not be necessary; but to avoid the possibility of free-running oscillations or unwanted crystal modes, at least the screen circuit should be broadly tuned. (See paragraph 1-590 for discussion of negative-resistance limiting of transitron circuit.)

Impedance-Inverting Crystal Oscillators

1-426. Impedance-inverting oscillators employ a network similar to that shown in figure 1-182(A), to permit conventional lower-frequency oscillators to be operated with crystal control in the v-h-f range. A number of these oscillators were designed and tested at the Georgia Institute of Technology under the direction of Mr. W. A. Edson. The discussion to follow is based on the final report of this research. The impedance-inverting network is designed to behave as a quarter-wave line having a characteristic impedance, $Z_o = \omega L_1 = \frac{1}{\omega C_n} = \frac{1}{\omega C_o}$. With this design, the network always appears as an inverted Z_s equal to $Z_n = \frac{Z_o^2}{Z_s}$, where Z_s is the series-arm impedance of the crystal. If $Z_s = 0$, C_o is shorted out and Z_n is infinite. (L_1 is assumed to have a zero loss.) If Z_s is infinite, L_1 is series-resonant with C_o, and $Z_n = 0$. If $Z_s = Z_o$, the network appears as an infinite line with $Z_n = Z_o$. When Z_s is a small inductive reactance, Z_n is a large capacitive reactance, and vice versa. With $Z_o >> R$ of the crystal, the network

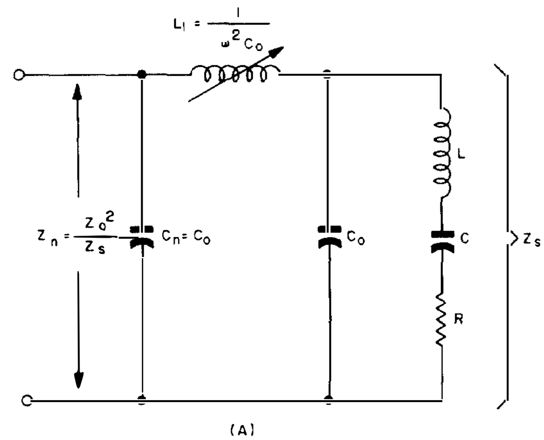


Figure 1-182. Impedance-inverting oscillator circuits. (A) Basic impedance-inverting network

serves to invert the crystal resistance to a high impedance equal to Z_o^2/R. For a given C_n, maximum frequency stability is to be had under the quarter-wave line conditions ($C_o = C_n$), but, if desired, higher impedances can be had by making C_n less than C_o, or by reducing the effective values of C_n and C_o with the use of shunt inductors. The shunt inductances, however, should be considerably larger than the values required for antiresonance at the crystal frequency. With ω equal to the series-arm resonance frequency, and $L_1 = \frac{1}{\omega^2 C_o} = \frac{1}{\omega^2 C_n}$, Z_n appears as an antiresonant resistance when the series arm of the crystal is resonant and $Z_s = R$. At frequencies well removed from crystal resonance, the crystal behaves simply as a capacitance, C_o, so that the network has a second antiresonant frequency, the square of which is $\omega_2^2 = \frac{C_n + C_o}{C_n C_o L_1}$. To ensure that this second frequency is damped out, a resistance equal to Z_o can be connected across the crystal unit.

1-427. Even though the *equivalent* impedance-inverting network is designed to be antiresonant at approximately the crystal frequency, the operating frequency may well require that the crystal network facing the *actual* terminal connections be reactive if the necessary phase reversal is to be accomplished. For example, it is necessary that the actual plate-to-grid network appear inductive when used in the Pierce circuit. In the Pierce circuit the fundamental modification introduced by the impedance-inverting circuit is simply the addition of an inductor having a reactance $\omega L_1 = \frac{1}{\omega C_o}$ in series with the crystal. It can be imagined that the reactance of the inductor replaces the X_e of a parallel-resonant crystal unit, and the low series-resonant R of the crystal approximately replaces its parallel-resonant value, R_e. C_n of the network is C_{pg} of the vacuum tube, if we view the circuit literally. By this interpretation, C_n is not antiresonant with the inductive branch, but must offer a higher impedance than does the inductor at the operating frequency. On the other hand, if the entire external circuit is viewed in toto by the negative-resistance method, which is the impedance-inverting interpretation, C_n appears as an equivalent capacitance equal to C_{pg} plus the additional amount required to make the network antiresonant. Since this latter interpretation can be employed to illustrate any oscillator circuit that contains a crystal connected in series with an inductance $L_1 = 1/\omega^2 C_o$, the presence of the series inductance alone could be sufficient to define an impedance-inverting oscillator. The inverted impedance, Z_n, is related to the impedance of the inductor and crystal branch as the PI of a crystal is related to the equivalent impedance of the crystal unit. Although these questions are somewhat academic, for some readers it may be more helpful to interpret the network in figure 1-182(A) as an impedance-*converting* circuit rather than as an *inverting* circuit. In the transitron circuit, the network is directly used to invert the crystal R to a higher effective resistance, but in other applications the designer may prefer to treat the actual network as an equivalent X_e and R_e of a parallel-mode crystal unit, transferring the equivalent impedances directly to the equations of the basic parallel-resonant oscillators.

1-428. There are two significant advantages to the impedance-inverting type of design. One is that the conventional parallel-resonant circuits can be operated with excellent frequency stability in the v-h-f range. Another is that by using series-mode crystals at the fundamental frequencies, the design restrictions regarding the parallel-resonant type of crystal unit can be avoided. No data is available, but experimentation may show that even in the fundamental-frequency range larger outputs can be obtained with an inductor and a series-mode crystal without degrading the over-all frequency stability. The chief disadvantage of the impedance-inverting network is that it cannot be used for broad-band untuned operation.

IMPEDANCE-INVERTING TRANSITRON OSCILLATOR

1-429. An experimental 50-mc impedance-inverting transitron oscillator is shown in figure 1-182 (B). In this circuit, the network, consisting primarily of L_1, C_n, and the crystal, is adjusted to present a resistive impedance between the screen and ground. Since R_L is very large compared with the crystal R, it can be assumed that R_L is effectively connected in parallel with the antiresonant network. C_n includes a 3—12 $\mu\mu$f padding capacitor adjusted at 5 $\mu\mu$f, the screen-to-ground capacitance, and the suppressor-to-plate capacitance (the latter is added because the screen is practically bypassed to the suppressor and the plate is at r-f ground). E_{g3} in this circuit can be considered equal to E_{g2}. Thus, the condition required for oscillations to build up is simply that $1/g_m$ be smaller than the actual plate-to-ground resistance. At the plate voltage used, the initial g_m is approximately 1500 μmhos, so $1/g_m = 667$ ohms. Ignoring the suppressor-to-ground resistance, the screen operates into an impedance of $R_L Z_n/(R_L + Z_n) =$ 1350 ohms, where $Z_n = X_{C_n}{}^2/R$. The margin of gain is therefore on the order of two to one. The power delivered to R_L was observed to be 15 mw. The frequency deviation was measured at 0.1 ppm/volt. When R_1 was replaced by a 40,000-ohm resistor, the frequency deviation was found to be only 0.004 ppm/volt. Although the power output is low, the extraordinary independence of the frequency under variations in the supply voltage marks the impedance-inverting transitron oscillator as the most stable to use in the v-h-f range. One of the chief reasons for this stability is very probably the fact that the r-f screen current need contain no reactive component. The impedance-inverting network, as faced by the screen, can appear as a pure resistance.

IMPEDANCE-INVERTING PIERCE OSCILLATOR

1-430. A 50-mc impedance-inverting Pierce oscillator is shown in figure 1-182(C). This circuit supplied 70 mw to the 1500-ohm load, and had a frequency deviation of 0.6 ppm/volt. Note that the total C_g is equal to the total C_p. The antiresonant C_n for the inductive branch of the impedance-inverting network is thus very nearly $(C_{pg} + C_p/2)$, which in turn is equal to C_o. This value of C_n neglects the equilavent negative capacitance due to the reactive component of the r-f plate current. Viewed only as an impedance-*converting* network connected between the plate and grid, $C_n = C_{pg}$, and the network appears as an inductive reactance numerically equal to $2/\omega C_p$ or $2/\omega C_g$. The upper useful limit of this type of circuit is approximately 100 mc.

1-431. Figure 1-182(D) shows an electron-coupled modification of the impedance-inverting Pierce

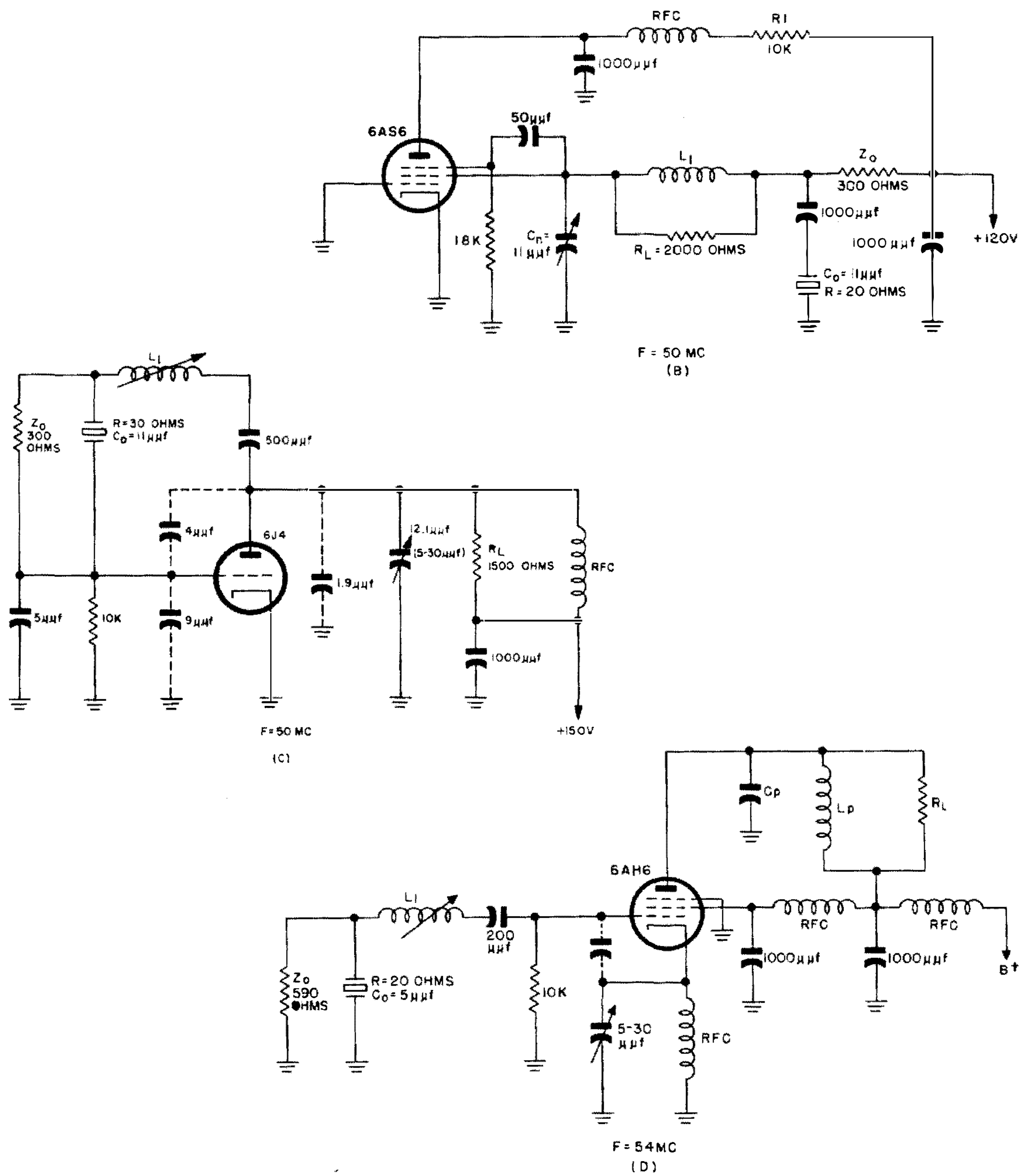
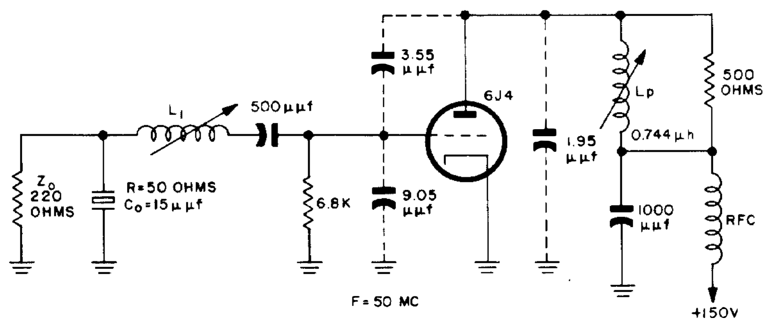


Figure 1-182. Impedance-inverting oscillator circuits. (B) Impedance-inverting transitron oscillator. (C) Impedance-inverting Pierce oscillator. (D) Impedance-inverting electron-coupled Pierce oscillator

oscillator. The frequency of the oscillator circuit is virtually independent of the tuning adjustments in the plate circuit. With the plate circuit tuned to the 1st, 2nd, 3rd, and 4th harmonics successively, the power supplied R_L was found to be 400 mw at 54 mc, 225 mw at 108 mc, 50 mw at 162 mc, and 10 mw at 218 mc. The frequency stability is approximately the same as that of the triode circuit in figure 1-182(C). The upper frequency limit of the grounded-screen circuit in (D) was found to be 70 mc.



(E)

Figure 1-182. Impedance-inverting oscillator circuits. (E) Impedance-inverting Miller oscillator

IMPEDANCE-INVERTING MILLER OSCILLATOR

1-432. Figure 1-182(E) is an experimental design of a 50-mc, impedance-inverting, Miller oscillator. The oscillator is designed so that the C_n of the equivalent negative resistance circuit is equal to C_o. Assuming that the reactive component of the plate current is negligible, $C_n = C_o = C_g + C_1$, where C_1 is the equivalent capacitance of C_{pg} in series with the parallel combination of C_p and L_p. C_1 is thus given by the equation

$$C_1 = \frac{C_{pg}\,(1 - \omega^2 L_p C_p)}{1 - \omega^2 L_p (C_p + C_{pg})} \qquad 1\text{—}432\ (1)$$

C_{pg} and C_p are fixed by the tube capacitances, and C_1 is equal to $C_o - C_g$, so the solution of equation (1) requires a definite value of L_p, which in circuit (D) was found to be 0.744 µh. The circuit supplies R_L with a power output of 0.5 watt for a crystal drive of 0.07 watt. The frequency deviation was found to be 0.6 ppm/volt.

Grounded-Cathode Two-Stage Feed-Back Oscillator

1-433. The two-stage feed-back oscillator (see figure 1-183) is used primarily for high-resistance, series-mode crystals operating at fundamental frequencies not higher than 500 kc and usually below 300 kc. The design is rather straightforward. V_1 and V_2 are tubes of the same type and can be contained in the same envelope. Although pentodes should permit slightly greater frequency stability, triodes are quite satisfactory for most purposes. Since V_2 alone can provide the necessary phase reversal, both tubes can operate into resistive loads. The V_1 plate circuit is thus tuned to the crystal resonance frequency. The proper adjustment of C_p is indicated by a maximum reading on the meter, M. R_2 is connected across the L_p-C_p tank, to broaden the tuning and reduce the frequency effects of variations in the V_1 plate capacitance. The resistance, R_e, of the crystal circuit is approximately $R + 2R_1$. On the assumption that $R = R_m$ (the maximum permissible crystal resistance), R_1 should be made as small as possible consistent with stable oscillations. This is desirable in order for the effective Q of the crystal circuit, and hence the frequency stability, to be maximum.

1-434. The loop-gain requirement for equilibrium is

$$G_1\,G_2\,G_3 = \frac{E_{p1}}{E_{g1}} \cdot \frac{E_{p2}}{E_{p1}} \cdot \frac{E_{g1}}{E_{p2}} = 1 \qquad 1\text{—}434\ (1)$$

where

$$G_1 = \frac{E_{p1}}{E_{g1}} = g_{m1}\,R_2 \qquad 1\text{—}434\ (2)$$

$$G_2 = \frac{E_{p2}}{E_{p1}} = \frac{g_{m2}\,R_1\,(R + R_1)}{R + 2R_1} \qquad 1\text{—}434\ (3)$$

and

$$G_3 = \frac{E_{g1}}{E_{p2}} = \frac{R_1}{R + R_1} \qquad 1\text{—}434\ (4)$$

Equations (2), (3), and (4) assume that V_1 and V_2 operate into plate impedances approximately equal to R_2 and $\frac{R_1(R + R_1)}{R + 2R_1}$, respectively, and that these impedances are very small compared with the R_p of the tubes. Combining equations (2), (3), and (4), we find that at equilibrium, the tube transconductances are such that

$$G_1\,G_2\,G_3 = \frac{g_{m1}\,g_{m2}\,R_1^{\,2}\,R_2}{R + 2R_1} \qquad 1\text{—}434\ (5)$$

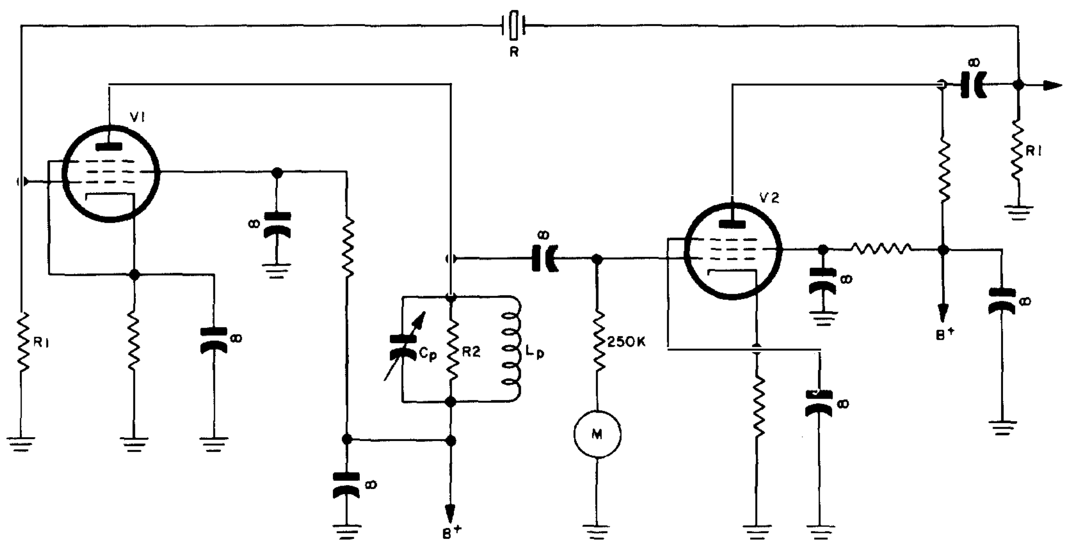


Figure 1-183. Grounded-cathode two-stage feed-back oscillator

Assuming that $g_{m1}\, g_{m2} = g_m^2$, where g_m is the nominal class-A transconductance of the V_1 and V_2 type of tube, and that $R = R_m$, we can select values of R_1 and R_2 so that equation (5) will equal 1.5. This provides a 3-to-2 margin of gain, which should be sufficient to ensure operation with all but completely defective tubes. The cathode resistors can be selected so that the amplitude of oscillations does not overdrive a crystal of maximum R. An alternative, and possibly a simpler approach, is first to select a cathode resistor for V_1, with the intention of operating that tube at a fixed class-A bias. The gain of the V_1 stage can then be treated as a predetermined constant and the V_2 stage designed to provide the necessary limiting by gridleak bias. The class-A gain of the V_2 stage must be sufficient to permit oscillations when the crystal unit has a maximum resistance, and the excitation current must not be sufficient to overdrive the crystal unit when the crystal resistance is a minimum. If desired, a parallel-mode crystal unit connected in series with its rated load capacitance can be substituted for a series mode crystal unit. Such operation increases the average effective feedback resistance, but the presence of the capacitor can reduce the tendency of the circuit to oscillate at unwanted frequencies.

MODIFIED TWO-STAGE FEED-BACK OSCILLATOR

1-435. A modification of the two-stage, feed-back oscillator to reduce the higher harmonics and thereby improve the quality of the sine-wave output for sync control is shown in figure 1-184. It can be seen that the tuned tank, undamped, is connected in the plate circuit of V_2 instead of that of V_1 as is conventionally done. The output, E_o, is taken from a different part of the tank in each of the three circuits represented. The non-bypassed cathode resistors are inserted for their degenerative effect on the higher harmonics and parasitic frequencies. They also reduce the effective input capacitance of the tubes. It would seem that the degradation of the crystal Q is somewhat large. The Miller effect in V_1 is probably significant in determining the impedance that the crystal faces —certainly so in circuit (C), where R_3 is one megohm and C_9 is inserted to increase the plate-to-grid capacitance by 25 $\mu\mu$f. However, the chief purpose of C_9 is to serve as a neutralizing capacitance for all free-running oscillations where the crystal unit would behave as a capacitance, C_o.

Colpitts Oscillators Modified for Crystal Control

1-436. Figure 1-185 illustrates a number of special-purpose circuits which are basically Colpitts oscillators modified for crystal control. Circuits (A), (B), and (C) are conventional CI-meter oscillators (see paragraph 1-220). The tank inductance, equal to $2L_1$, is split into two equal inductances, L_{1A} and L_{1B}. Each of the variable inductors in circuit (A) actually represent seven fixed inductors which can be connected into the circuit by a range switch. The capacitors C_1 (A and B) are continuously variable, and are so ganged that C_{1A} is always equal to C_{1B}. Circuits

Fig.	Equipment	Purpose	F_1	CR	R_1	R_2	R_3	R_4	R_5
(A)	Range Calibrators TS-102/AP and TS-102A/AP	Crystal control of 500-yd marker and sync pulses	327.8	Sig C Stock No. 2X62-327.8; WECo No. D-168342	1.8	10	18	1000	1 8
(B)	Calibrator TS-19/APQ-5	Crystal control of 1000-ft marker and sync pulses	491.04	WECo No. D-164868	1.8	10	18	1000	1 8
(C)	Range Calibrator TS-293/CPA-5	Osc for radar IFF. P/O Radar Sets AN/CPX-1 and AN/CPX-2	186.3	Belmont Drawing No. A-8K-3577	1.8	10	1000	1000	1.8

Circuit Data for Figure 1-184. F in kc. R in kilohms. C in $\mu\mu$f. L in μh.

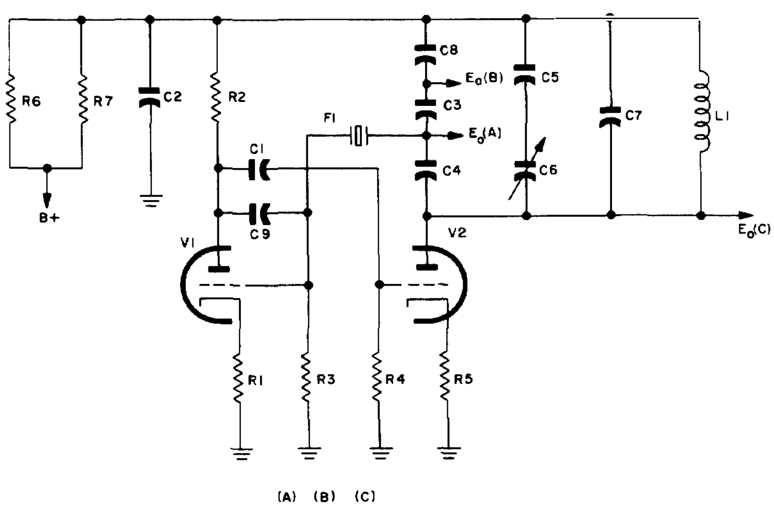
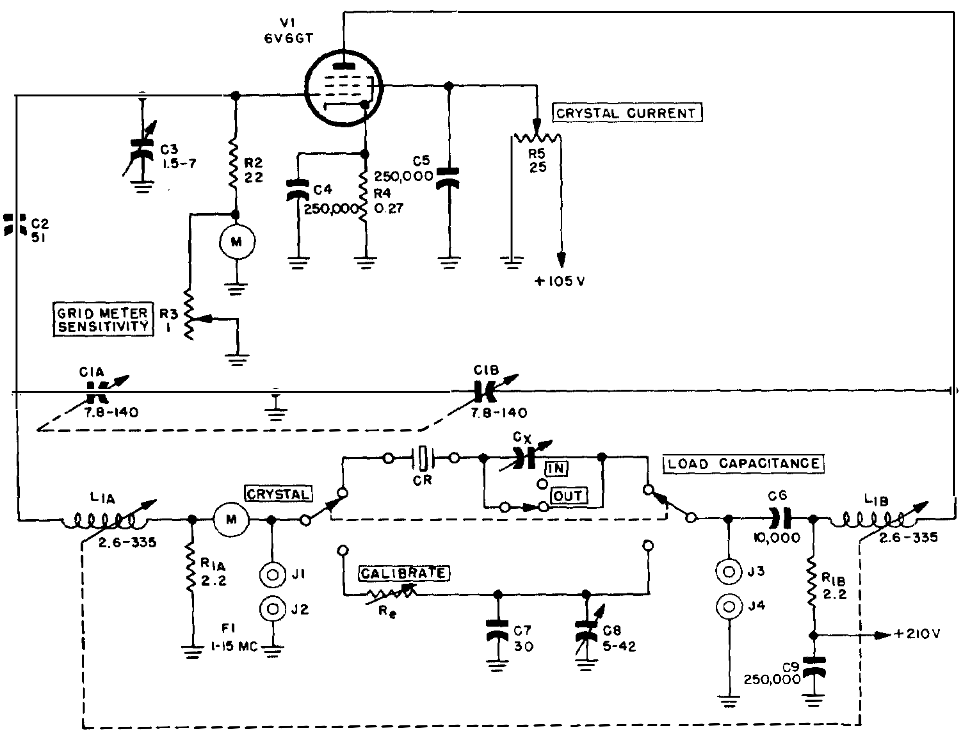


Figure 1-184. Modifications of two-stage feed-back oscillator to improve sine-wave output

Fig.	Equipment	Purpose	F_1	CR	R_1	R_2	R_3	R_4	R_5	R_6	R_7	R_8	R_9	R_{10}
(A)	Crystal Impedance Meter TS-330/TSM	Substitution circuit for measuring parameters of crystal unit	1-15	Military Standard quartz crystal units	2.2 each	22	1	0.27	25					

Circuit Data for Figure 1-185. F in mc. R in kilohms. C in $\mu\mu$f except where otherwise noted. L in μh.

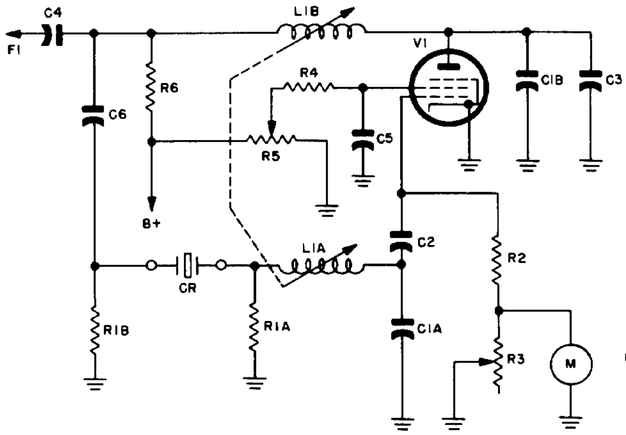
R_6	R_7	C_1	C_2	C_3	C_4	C_5	C_6	C_7	C_8	C_9	L_1	V_1 V_2
2.2	2.2	3300	100,000	3900	390	10,000	75	20	∞	0	500	6SN7GT
2.2	2.2	3300	100,000	3900	200	10,000	75	10	3900	0	380	6SN7GT
1	∞	6000	50,000	350	75	0	0	25	∞	25	Sig C Stock No. 2C-638-1C1	6SN7GT



(A)

Figure 1-185. Colpitts circuits modified for series-mode crystal control

C_1	C_2	C_3	C_4	C_5	C_6	C_7	C_8	C_9	L_1	L_2	L_3	L_4	V_1	V_2
7.8-140 each	51	1.5-7	250,000	250,000	10,000	30	5-42	250,000	2.6-335 each				6V6GT	



(B)(C)

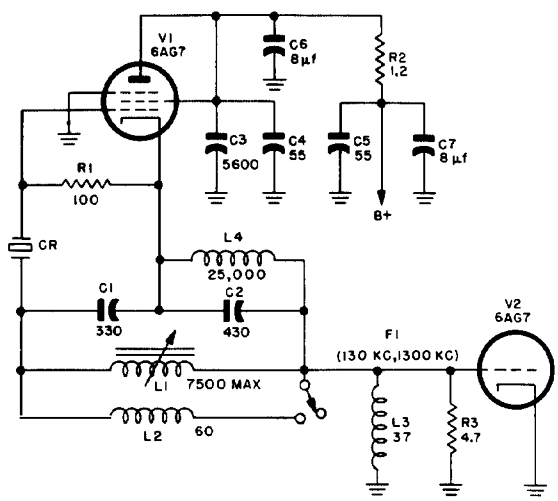


Figure 1-185. Continued

(D)

Fig.	Equipment	Purpose	F_1	CR	R_1	R_2	R_3	R_4	R_5	R_6	R_7	R_8	R_9	R_{10}
(B)	Crystal Impedance Meter TS-683/TSM	Substitution circuit for measuring parameters of crystal unit	10-60	Military Standard quartz crystal units	0.1 each	15	1	15	25	4.7				
(C)	Crystal Impedance Meter TS-683/TSM	Substitution circuit for measuring parameters of crystal unit	55-140	Military Standard quartz crystal units	0.056 each	15	1	15	25	2.2				
(D)	Test Set TS-250/APN	Range osc. Output consists of positive range pips	0.13 and 1.3	Bliley No. 122-5006 (octal base)	100	1.2	4.7							
(E)	Diversity Receiving Equipment AN/FRR-3	BFO with afc; manual or crystal operation	0.46245		100	5	1	10	20	2	0.5	0.035	500	1.5

Circuit Data for Figure 1-185. F in kc. R in kilohms. C in $\mu\mu f$ except where otherwise noted. L in μh.

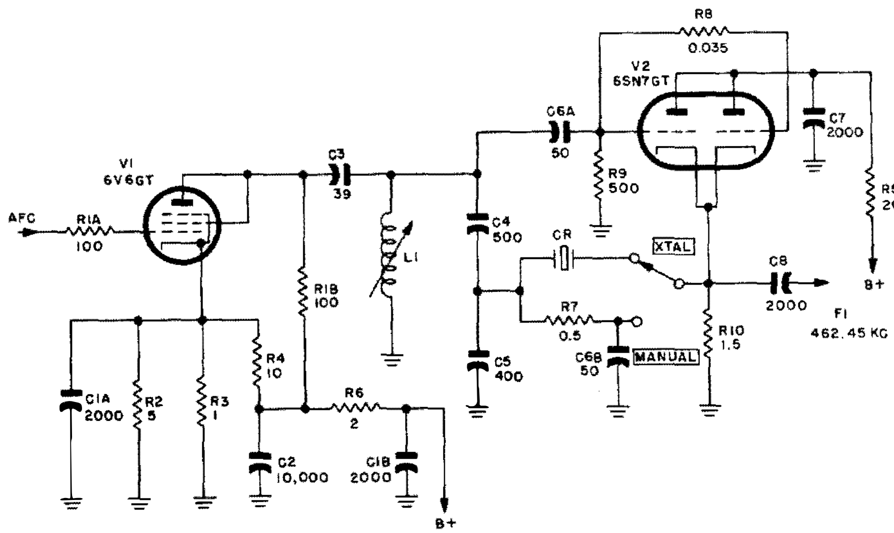


Figure 1-185. Continued

(E)

C_1	C_2	C_3	C_4	C_5	C_6	C_7	C_8	C_9	L_1	L_2	L_3	L_4	V_1	V_2
33 each	22	1	5	470	470				0–7.35 each				5654	
5 each	12	0	1	470	470				0–0.987				5654	
330	430	5600	55	55	8μf	8μf			7500 max	60	37	25,000	6AG7	6AG7
2000	10,000	39	500	400	50	2000	2000						6V6GT	6SN7GT

(B) and (C) are substantially of the same basic design as circuit (A). Except for resistors R_4 and capacitances C_3 and C_4, the parameters of circuits (B) and (C) have the same numbers as their functional analogues in circuit (A). Circuits (B) and (C) are not designed for parallel-resonance measurements. For crystal resistance measurements, the calibrating resistor must be substituted externally for the crystal unit. The capacitors C_1 are fixed and the inductors L_1 are continuously variable and are so ganged as always to be equal. In each of the CI-meter circuits shown, it can be seen that if the resistance of the tank, including the crystal, were zero, and if the tank were perfectly balanced, no voltage would exist between the crystal and ground. The voltage across L_{1A} plus that across C_{1A} would equal zero, and no current would flow through the resistors R_1, which effectively form a bridge between the inductance arm to the grounded connection of the capacitance arm. In practice, a net voltage does exist across L_{1A} and C_{1A} in series, and this voltage appears across R_{1A}, being measurable at the jacks J1 and J2 in circuit (A). The r-f voltage across R_{1B} is approximately that across J3 and J4, which in turn is equal to the R_{1A} voltage plus that across the crystal resistance. The R_1 resistors are not essential insofar as maintaining oscillations is concerned, but they load the circuit, thereby reducing the effect of the variations in crystal resistance upon the oscillator activity, and they serve to protect the crystal, to balance the circuit to ground, and to facilitate measurements of the crystal voltage ($E_{Re} = E_{J3} - E_{J1}$) without unduly interfering with the effective circuit parameters. The CI-meter oscillator can be analyzed as a particular type of transformer-coupled oscillator, as an impedance-inverting oscillator, or as an equivalent Pierce oscillator having a crystal $X_e = \omega(L_{1A} + L_{1B})$ and an effective crystal resistance accounting for the losses in the resistances R_1 as well as in the R_e of the actual crystal.

1-437. Figures 1-185(D) and (E) are examples of grounded-plate Colpitts circuits which have been modified for series-mode crystal control. Circuit (D) is designed to provide positive range pips to the grid of V_2. The circuit operates class C at either one of two frequencies, the appropriate crystal being connected between the cathode tank and the grid of V_1. Circuit (E) is designed for either manual or crystal control. During manual control the resistor R_7 replaces the crystal unit. V_1 is operated as a reactance tube. The a-f-c bias varies in such a way that the b-f-o frequency tends to follow any changes in the frequency of the teletype signal being received.

CRYSTAL CALIBRATION

1-438. The design of a crystal oscillator to be used for calibrating the frequency of other oscillators generally is directed toward obtaining outputs rich in harmonics. Where tuned-plate circuits are required the L/C ratios should be high, so that high impedances are also presented to the overtone frequencies. The oscillator should be operated class C, and often the gridleak resistance is a megohm, or higher. If the crystal calibrator is to serve as a frequency standard of greater-than-average precision, this precision becomes the principal design problem insofar as the oscillator is concerned; if need be, the required harmonics can be developed in nonlinear amplifier stages that follow the oscillator stage. The higher the overtone, the weaker will be its effective output power, but with proper design useful outputs up to and above the 100th harmonic can be obtained. With the addition of frequency multiplier and/or divider circuits a single crystal can provide a useful calibrator frequency range as broad as desired. For maximum precision, a G element, usually cut for 100 kc, should be used.

1-439. Figure 1-186 illustrates a simply designed crystal calibrator employing an electron-coupled Miller oscillator operating into a resistive plate load. Such a circuit will ensure sufficient frequency stability for most purposes. Harmonic outputs in steps of 100 kc are provided up to frequencies of 10,000 kc. For higher frequencies, the 1000-kc crystal can be used to provide calibration points in multiples of 1000 kc. The variable grid capacitor is employed to ensure that the crystal operates into the correct load capacitance.

Crystal Calibrator Employing Regenerative Frequency Divider

1-440. Figure 1-187 shows the regenerative frequency-divider circuit of the crystal frequency indicator (CFI) used in Radio Transmitting Set AN/ART-13A. This circuit employs a 200-kc crystal to control a rich mixture of harmonics, providing useful check points spaced as close as 25 kc apart. The crystal oscillator, utilizing the triode section, V_1, seems best described as a modification of an impedance-inverting Pierce circuit. When oscillations first start, the output of the oscillator is fed to grid No. 1 of the pentagrid mixer, V_3. The 50-kc and 150-kc components of the noise voltages that are mixed with the 200-kc signal are amplified by V_3 and fed to the input of the V_2 triode section. The V_2 plate circuit, which is tuned

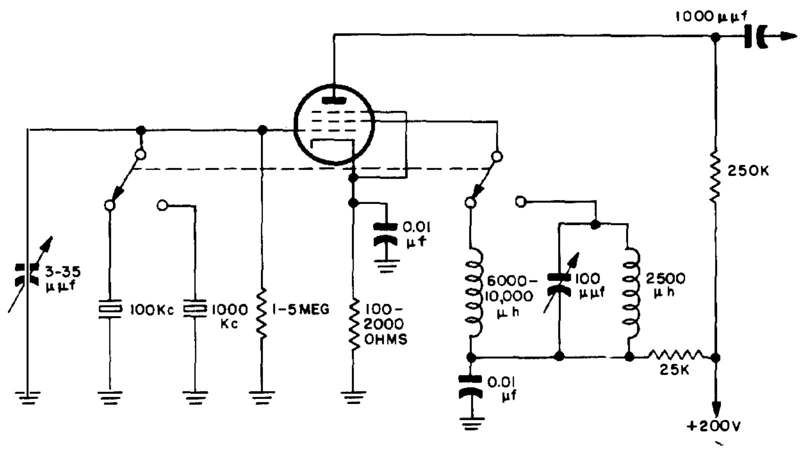


Figure 1-186. Typical design of single-tube general-purpose crystal calibrator circuit

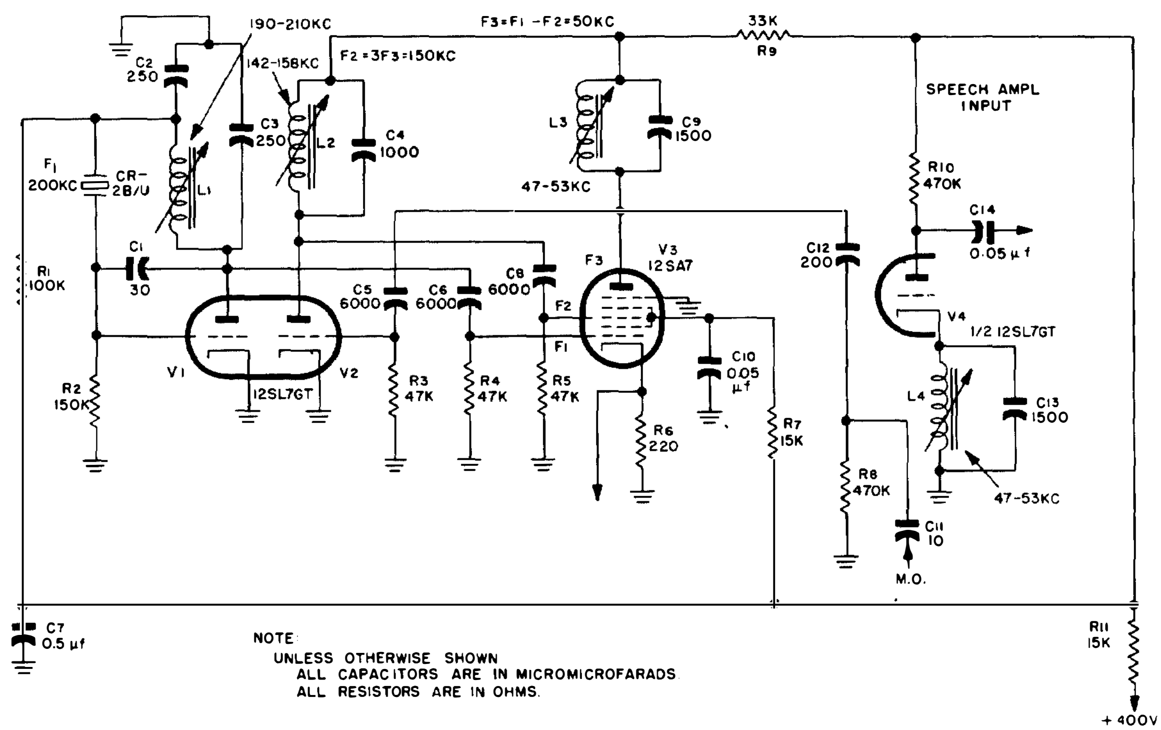


Figure 1-187. CFI regenerative frequency divider in Radio Transmitter T-47A/ART-13 (P/O Radio Transmitting Set AN/ART-13A)

to 150 kc, amplifies the 150-kc noise input and triples the 50-kc input. The 150-kc output of V_2 is then fed back to the pentagrid mixer at grid No. 3. It is again amplified and fed back to V_2. However, the direct amplification and regeneration of the 150-kc signal alone is not sufficient nor properly phased to maintain oscillations at this frequency. The 150-kc oscillations are sustained principally by tripling the 50-kc feedback, which builds up as the amplified difference frequency of the 200-kc and 150-kc inputs to V_3. The output of V_3 is effectively a 50-kc fundamental frequency standard of large harmonic content that is fed to the grid of triode section V_4, where it is mixed with signals from the variable oscillators of the transmitter. The output of V_4 is fed to the input of an audio amplifier, which amplifies the beat note whenever the variable oscillator approaches the frequency of one of the CFI harmonics. In practice, the recommended check-point harmonics are spaced 25 kc apart in the 200—600-kc frequency range, 100 kc apart from 2000 to 3000 kc, 150 kc (3000—4000 kc), 200 kc (4000—6000 kc), 300 kc (6000—9000 kc), 450 kc (9000—12,000 kc), and 600 kc (12,000 to 18,100 kc). The presence of the harmonics of an apparent 25-kc fundamental, which is used in the low-frequency calibrations, is not readily explained on the basis of the foregoing discussion of the circuit. A complete analysis of the nonlinear characteristics of the circuit is not available, but it appears possible that if a 25-kc signal appears at the plate of V_3, it can conceivably be sustained by being fed to V_2, mixed with f_2 to form a sum frequency of 175 kc, fed back to V_3 and mixed with the 200-kc injector signal to regenerate a difference frequency of 25 kc. It should be understood that $f_2 = 3f_3 = \frac{3}{4}f_1$ is a necessary relation, and that f_2 and f_3 are synchronized and controlled by the crystal oscillator. The phase and frequencies of the regenerative circuits automatically follow the phase and frequency of the V_1 output. For a more analytical study of regenerative frequency dividers, see discussions by R. L. Miller, R. L. Fortescue, and W. A. Edson.

SYNTHESIZING CIRCUITS

1-441. Of great promise, particularly for use in airborne radio equipment in the v-h-f range where crystal control is necessary to maintain the required frequency stability, has been the development of synthesizing circuits, in which a very few crystals are able to control a large number of channels. In the discussions to follow we shall use the term *frequency synthesis* very loosely to apply to any type of frequency-control circuit or system in which a few fixed-frequency oscillators are used to control or to stabilize a large number of radio frequencies. If the term were used rigorously, it would apply only to those cases where an output frequency is produced entirely from heterodyned combinations of internally generated frequencies. Examples of this type are provided by the Plessey frequency generator and by the Collins transmitter frequency-control system employed in Radio Set AN/ARC-27. For our purposes we shall extend the term to cover such systems as the Bendix frequency-control circuit in Radio Set AN/ARC-33, where the output frequency is not actually synthesized but is obtained from a variable-tuned master oscillator that is crystal-stabilized at many frequencies. Also implied by the term will be such systems as the Collins crystal-controlled multichannel receiver circuits. In these latter circuits only one end-product frequency is desired — a fixed superheterodyne intermediate frequency. But the system design is such that with the use of a very few crystals the desired intermediate frequency can be synthesized under crystal control from received signals on any one of hundreds of possible radio channels.†

The Plessey Synthesizing System*

1-442. The first crystal-controlled frequency synthesizer in commercial usage appears to have been

† Not all types of synthesizers in current use are covered in the above discussion. Other recently developed and equally important circuits include:

The General Radio Company synthesizer, developed under Signal Corps Contract No. DA-36-039-sc-15542. The GR synthesizer operates on a principle fundamentally different from those described in this report. In the GR system an oscillator is phase- and frequency-locked through a variable scale-of-N divider. The pulse output of the divider is compared by coincidence methods with a pulse derived from a crystal oscillator. The frequency range of this synthesizer is 0.1 mc to 10 mc.

The Matawan Synthesizer ME-447, of the Lavoie Laboratories Instrument Company. This system generates any multiple of 1 kc within the range of 1.0 mc to 2.0 mc.

The Rohde and Schwarz decade synthesizer and exciter system (Federal Telephone and Radio Company Types HS-431, HS-441, and HS-471), which covers a range of 50 kc to 30 mc.

The Telefunken Precision Frequency Meter. This meter is used in the measurement of frequencies between 1 kc and 300 mc. The circuitry contains a frequency synthesizer capable of generating sine-wave outputs between 1 kc and 30 mc. It is claimed that harmonics and sidebands of the output frequency are at least 80 db below the selected signal, and that the synthesizer accuracy is ± 0.2 cps for frequencies between 1 kc and 3 mc, and ± 2.0 cps between 3 mc and 30 mc.

* Note: The discussion of the Plessey synthesizing system is based primarily upon the report, "The Frequency Synthesizer", by Mr. H. J. Finden of the Plessey Company, Ltd., England, published in the *Journal of the Institution of Electrical Engineers*, Vol. 90, Part III, 1943.

that developed by the Plessey Company, Ltd. of England. This synthesizer has been designed as a frequency generator to be used in making precise radio-frequency measurements. The synthesizing system employed is nevertheless quite applicable for other uses, such as providing multichannel excitation voltages for radio communication equipment. As designed by the Plessey engineers, the synthesizer generates a sequence of harmonic signals of much greater precision and purity than is obtainable with conventional types of frequency generators. The original model permits a direct-reading dial selection of any of the first 10,000 harmonics of 1 kilocycle per second; a later and larger model extends the range to the first 100,000 harmonics, i.e., any harmonic of 1 kc up to 100 mc. All these frequencies are made available singly as pure sine waves (unmixed with other harmonics or frequency products) by a decade system of frequency dividers and multipliers, mixing stages, and filters where all the generated frequencies are under the control of a single 1000-kc precision crystal standard. Theoretically the system could be extended to cover a broader or a different frequency range; or could be changed to permit steps between adjacent frequencies that are smaller or larger than 1 kc. If required, it would be quite practicable for the Plessey generator, itself, to be expanded to cover also the 100-to-1000-mc range in 10-mc steps. In 1955 the Schomandt Company of Munich, Germany placed on the market a similar type of synthesizer frequency generator covering the 0—30-mc range in 1-kc steps. The output of the Schomandt synthesizer is equivalent to that of the Plessey synthesizer in quality, having at least a 60-db attenuation of all unwanted frequencies. It was the demand for such narrowly spaced pure output frequencies for use in making frequency measurements that originally led to the development of the synthesizer circuits.

SYNTHESIZER ADVANTAGES IN RADIO-FREQUENCY MEASUREMENTS

1-443. Prior to the development of the frequency synthesizer there were two conventional methods for measuring radio frequencies—the "interpolation" method and the "successive heterodyning" method.* Briefly, the interpolation method, which is satisfactory where extreme accuracy is not required, consists of mixing the unknown frequency with the two nearest harmonics of a frequency standard, and zero-beating the difference frequencies obtained against the output of a linearly tuned variable oscillator. It is then possible to interpolate the unknown frequency by determining its relative position between the known harmonics of the standard. In the successive heterodyning method, the unknown frequency is mixed with a known harmonic of a frequency standard; the difference frequency is then heterodyned with a second standard harmonic to obtain a second and lower difference frequency; and the process is repeated, if necessary, until a difference frequency is obtained that lies within an accurately measureable audio range. Although the successive heterodyning method can be quite accurate, occasions arise where the operator cannot be certain without undue checking that the difference frequencies being measured are not the products of unwanted harmonics contained in the heterodyned signals. The use of a frequency synthesizer that permits individual pure sine-wave outputs of a sequence of narrowly spaced frequencies, instead of a simultaneous mixture of many harmonics, can be said to offer a third and greatly superior means of measuring radio frequencies.

1-444. With the use of decade dial control, greater operating simplicity is possible than with the interpolation method; and when the pure sine-wave frequencies are spaced only 1 kc apart, the interpolation accuracy of the successive heterodyning method is maintained, but with the elimination of those chance difference products that can result from harmonic mixtures of multiple stages of heterodyning.

FUNCTIONAL OPERATION OF PLESSEY SYNTHESIZER

1-445. The circuit system by which the Plessey synthesizer produces thousands of frequencies, all controlled by a single 1000-kc crystal standard, is illustrated in figure 1-188. The block diagram shown is that of the original, single-cabinet model that permits the operator a choice of any one of the first 10,000 harmonics of 1 kc. It can be seen that there are three successive stages in which the input frequency is divided by 10, so that the last divider represents an over-all division of the original standard (1000 kc) by 1000. The dividers and the 1000-kc harmonic generator are of the synchronized, free-running, multivibrator type whose outputs are rich in harmonics. Each of these multivibrator circuits forms the first stage of a sequence which can be tuned to pass any one of the first 10 harmonics of its respective multivibrator fundamental. These sequences are labeled A, B, C, and D in figure 1-188. In the synthesis of a frequency, we can say generally that sequence A

* Note: See paragraph 2-66 to 2-151 for detailed descriptions of frequency-measuring systems in current use.

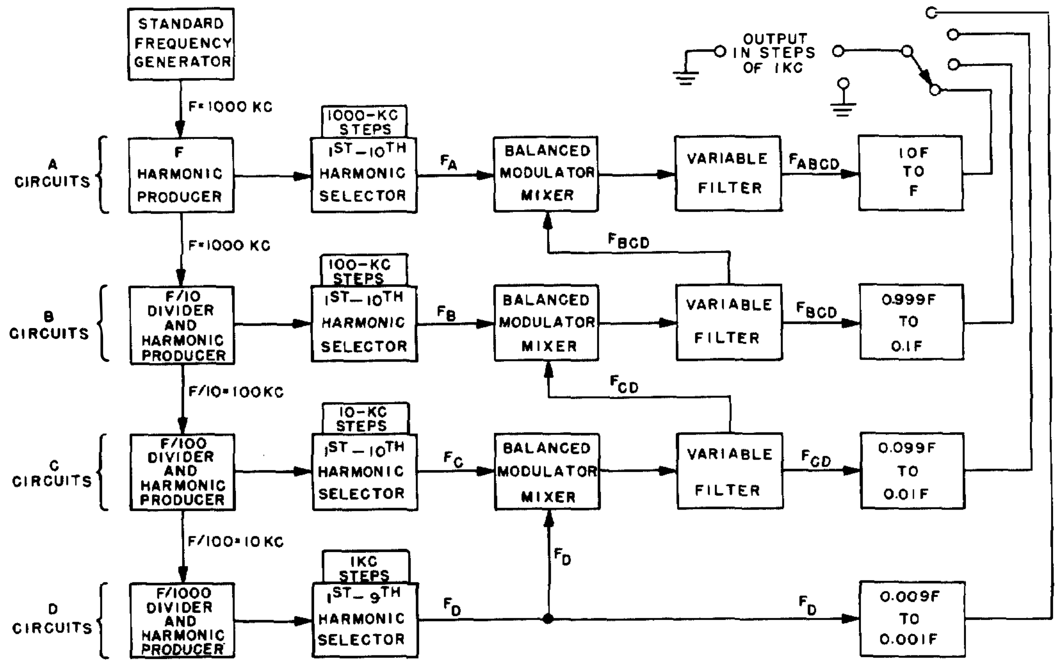


Figure 1-188. Block diagram of a Plessey synthesizer designed to cover the 0—10-mc spectrum in 1-kc steps

supplies that part of the final frequency which is a multiple of 1000 kc, B that part which is a multiple of 100 kc, C that part which is a multiple of 10 kc, and D that part which is a multiple of 1 kc.

1-446. For example, assume that an output frequency of 6789 kc is desired. The A, B, C, and D harmonic selectors, respectively, will be decade-set to pass the 6th, 7th, 8th, and 9th harmonics of their respective input signals from the preceding multivibrator stages. In balanced modulator C, the output of selector D, 9 kc, is mixed with the 80-kc output of selector C. (The signals are heterodyned in a balanced modulator circuit rather than in a more efficient type of mixer in order to eliminate the two input frequencies from the modulator output. In this manner the sum and difference products become the dominant frequencies in the modulator output.) Filter C is dial-set to pass the desired frequency product, 89 kc, which it feeds to balanced modulator B. In modulator B, the 89-kc signal is heterodyned with the 700-kc output of the decade-set harmonic selector B. Filter B is dial-set to pass the sum product, 789 kc, from the B modulator output to the A modulator input, where it is mixed with the 6000-kc output of harmonic selector A. Filter A is dial-set to pass the sum product, 6789 kc, of the mixed signals, which product is then amplified and fed through a phase inverter to the synthesizer output jack.

1-447. The foregoing example of the operation of the Plessey synthesizer suggests that the sum rather than the difference products of the mixed signals are always selected. In practice this is not the case, even though the decade dialing system is so designed that the operator is always provided a direct reading of the output frequency as if he were only adding the decade units together. In order to sufficiently filter out the unwanted product, it is important that the signals to be mixed are so selected that there is at least a 10 per cent difference in frequency between the sum and difference products. Since the filters must be capable of suppressing all adjacent harmonics of the mixed signals, it can be assumed that they are also capable of suppressing the unwanted heterodyne product if it differs from the desired product by as much as the fundamental harmonic of the modulator input from the harmonic selector. For example, in modulator C, the space between the sum ($f_C + f_D$) and the difference ($f_C - f_D$) frequencies should not be less than 10 kc, the fundamental of the harmonic from selector C. Since

$$(f_C + f_D) - (f_C - f_D) = 2f_D > 10 \text{ kc}$$

then f_D must never be less than 5 kc if it is to be mixed with f_C. Similarly, f_{CD} must not be less than 50 kc if it is to be mixed with f_B, and f_{BCD} must not be less than 500 kc if it is to be mixed with f_A.

1-448. To illustrate, let us suppose that a frequency of 91 kc is desired. It would not do for f_C and f_D to be 90 kc and 1 kc, respectively, for then the sum product, 91 kc, would be separated from the difference product, 89 kc, by only 2 kc. Rather, 100 kc should be selected as f_C and 9 kc as f_D. The variable filter C would be set to pass the difference product, 91 kc; which product differs from the sum product, 109 kc, by 18 kc, well beyond the minimum permissible limit of 10 kc.

1-449. As a more involved example we shall determine the heterodyne frequencies that would be used in the synthesis of an 8136-kc output. For a mental calculation of the correct frequency combinations the easiest method is to start with the output frequency, f_{ABCD}, and from this determine f_A, f_{BCD}, f_B, f_{CD}, f_C, and f_D in that order, working from the larger units to the smaller. Each of the above six frequencies is determined by remembering that none of the input frequencies to the A, B, and C modulators can be less than 500, 50, and 5 kc, respectively. Thus, we see at once that 8136 kc is not to be the sum product of 8000 kc and 136 kc in the A modulator, since 136 kc is less than 500 kc. So f_A must be 9000 kc and f_{BCD} must be 1000 minus 136 kc, that is, 864 kc; which means that filter A will be adjusted to pass the difference product (9000 kc minus 864 kc). Since 64 kc is greater than 50 kc, the required 864-kc output of modulator B can be obtained as the sum product of 800 kc and 64 kc, f_B and f_{CD}, respectively. Since 4 kc is less than 5 kc, the required 64-kc output of modulator C must be obtained as the difference product of 70 kc and 6 kc, f_C and f_D, respectively. We see that in order to select an output of 8136 kc, the decade dials of the A, B, C, and D harmonic selectors must be set to pass, respectively, the 9th, 8th, 7th, and 6th harmonics. In other words, the output frequency would be a synthetic product of the four frequencies, 9000 kc, 800 kc, 70 kc, and 6 kc. So also would be an output frequency of 9876 kc. Since the decade dials that control the harmonic selectors may be set at the same positions for two or more frequencies, some arrangement must be made so that the decade reading presented to the operator identifies correctly the particular frequency being synthesized. This convenience is accomplished in the Plessey synthesizer by manually operated range adjustments that alter the correspondence of the dial readings with the dial positions. Thus, in the example above, with the proper range settings, decade dial A in position 9 would give a reading of 8, decade dial B in position 8 would give a reading of 1, decade dial C in position 7 would give a reading of 3, and decade dial D in position 6 would give a reading of 6. The mechanics of exactly how this feature is incorporated in the Plessey synthesizer, although relatively simple in principle, is somewhat beyond the subject matter of our assignment here.

CIRCUIT DESIGN OF PLESSEY SYNTHESIZER

1-450. The general circuit design employed in a Plessey synthesizer is shown in the schematic diagram of figure 1-190. The circuit shown, when synchronized by a 1000-kc standard (whose circuit is not shown), is capable of covering the 0-10-mc range in 1-kc steps. Note that each of the four decade harmonic sequences begins with a multivibrator-type of harmonic generator. Rheostats are furnished for adjusting the natural oscillation period of each multivibrator, in order to allow for aging effects and the like. More elaborate or reliable harmonic-generator circuits are not required since the failure of any of the multivibrators would be immediately apparent by the reading in the output meter. Figure 1-189 shows in detail the circuit parameters of the 100-kc multivibrator, which also acts as the 1st divider. Note that the 1st divider is synchronized by the output of the 1000-kc amplifier and not directly by the frequency standard. The output of the 1st divider in turn is used to synchronize the 2nd divider, and that of the 2nd to synchronize the 3rd.

1-451. In figure 1-190 it can be seen that harmonic selection is achieved by switching to the correct tuning capacitor from a bank of 10. The same inductance is used for each of the harmonics. Since the percentage difference between adjacent harmonics is less as the order of the harmonic becomes higher, it is more difficult to eliminate the 9th and 11th harmonics when selecting the 10th, than it is to eliminate the 1st and 3rd when selecting the 2nd. For this reason, the value of each of the fixed tuning inductors is chosen to provide an optimum Q at the 10th harmonic. This permits a relative magnification of the 10th harmonic over its adjacent harmonics of approximately 200, which is equivalent to a 32-db attenuation of the 9th and 11th harmonics and more than that for all others. The attenuation of adjacent harmonics becomes greater as the selected harmonic becomes lower, so that in any event it is never less than 32 db at each tuned circuit. Two tuned circuits in series provide more than a 60-db attenuation,

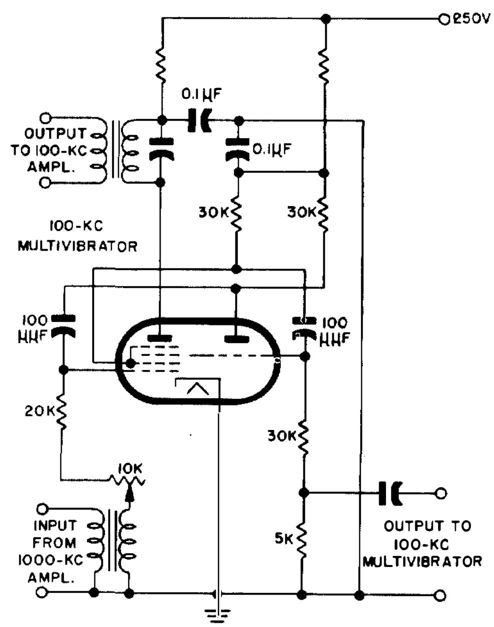


Figure 1-189. Schematic diagram of a Plessey synthesizer designed to cover the 0—10-mc spectrum in 1-kc steps. The circuit of the crystal oscillator standard is not shown

which for all practical purposes is sufficient to consider the selected harmonic a pure sine wave.

1-452. With the use of a balanced modulator it is not necessary to use as many tuned circuits as would otherwise be necessary to eliminate all unwanted harmonics and frequency products. Note in figure 1-190 that the balanced modulator design is such that two matched amplifiers have a common output circuit, but that they are excited by equal signals 180 degrees out of phase, so that the amplified signals cancel each other in the load. Thus, even though the heterodyne efficiency of the balanced modulator is less than that of other types of mixers, the balanced circuit is greatly advantageous in helping to eliminate all the unwanted frequencies, particularly the unwanted harmonics, that originate in the circuits preceding a mixer stage. In the Plessey synthesizer it can be seen that the modulators are provided with a switching arrangement by which one of the tubes of each modulator can be cut out of the circuit by opening its cathode return. One of these switches is opened whenever a modulator stage must pass an unmixed signal. With one tube removed, the balanced arrangement is destroyed, and since only one input signal is being handled, the vacuum tube still connected in the circuit will be operated as a conventional amplifier. If, for example, the desired output were a 2000-kc signal, none of the modulators would be in operation except modulator A, which would be unbalanced and operated simply as an amplifier of the 2nd harmonic from the 1000-kc harmonic generator.

1-453. Variable-tuned circuits are provided as bandpass filters. These must be adjusted manually in selecting the proper heterodyne product to be passed. The selectivity is sufficient to provide at least a 30-db attenuation of any unwanted signal that differs as much as 10 per cent from the desired signal.

1-454. The phase inverters are inserted for proper impedance matching. They permit an output at any frequency within the operating range of 100 millivolts across a 75-ohm load. The system as a whole insures at least a 60-db attenuation of all unwanted frequencies.

The Bendix Synthesizing System

1-455. In America, much of the pioneering in the field of frequency synthesis has been done by the research staff of the Bendix Corporation.

The following discussion is based upon the synthesizing circuit originally described by W. R. Hedeman of Bendix in the magazine *Electronics.* Figure 1-191 shows a block diagram of the synthesizer circuit developed at Bendix for use in controlling the frequency of a continuously variable v-h-f receiver heterodyne oscillator. In this circuit, the first crystal oscillator employs but one crystal. The harmonic generator that follows this oscillator produces a rich output of harmonics, the first of which is f_c, the fundamental of the first crystal oscillator. The harmonics selector is composed of a number of band-pass circuits, each circuit designed to pass a particular harmonic of the crystal frequency. The number of frequencies controlled by the synthesizer is directly proportional to the number of harmonic channels in the selector. Let f_h equal the harmonic selected and f_o equal the frequency of the variable oscillator. The value of f_o is always higher than that of f_h. These two frequencies are mixed in the first frequency converter to form the sum-and-difference frequencies, which, in turn, are fed to the input of the first band-pass amplifier. The first band-pass amplifier amplifies and passes only the difference frequency, $f_o - f_h$. This difference frequency is fed to the second frequency converter, where it is mixed with the output, f_x, of the second crystal oscillator. The second crystal oscillator is generally provided with more than one crystal unit, but only the fundamental frequency of the oscillator is used when a

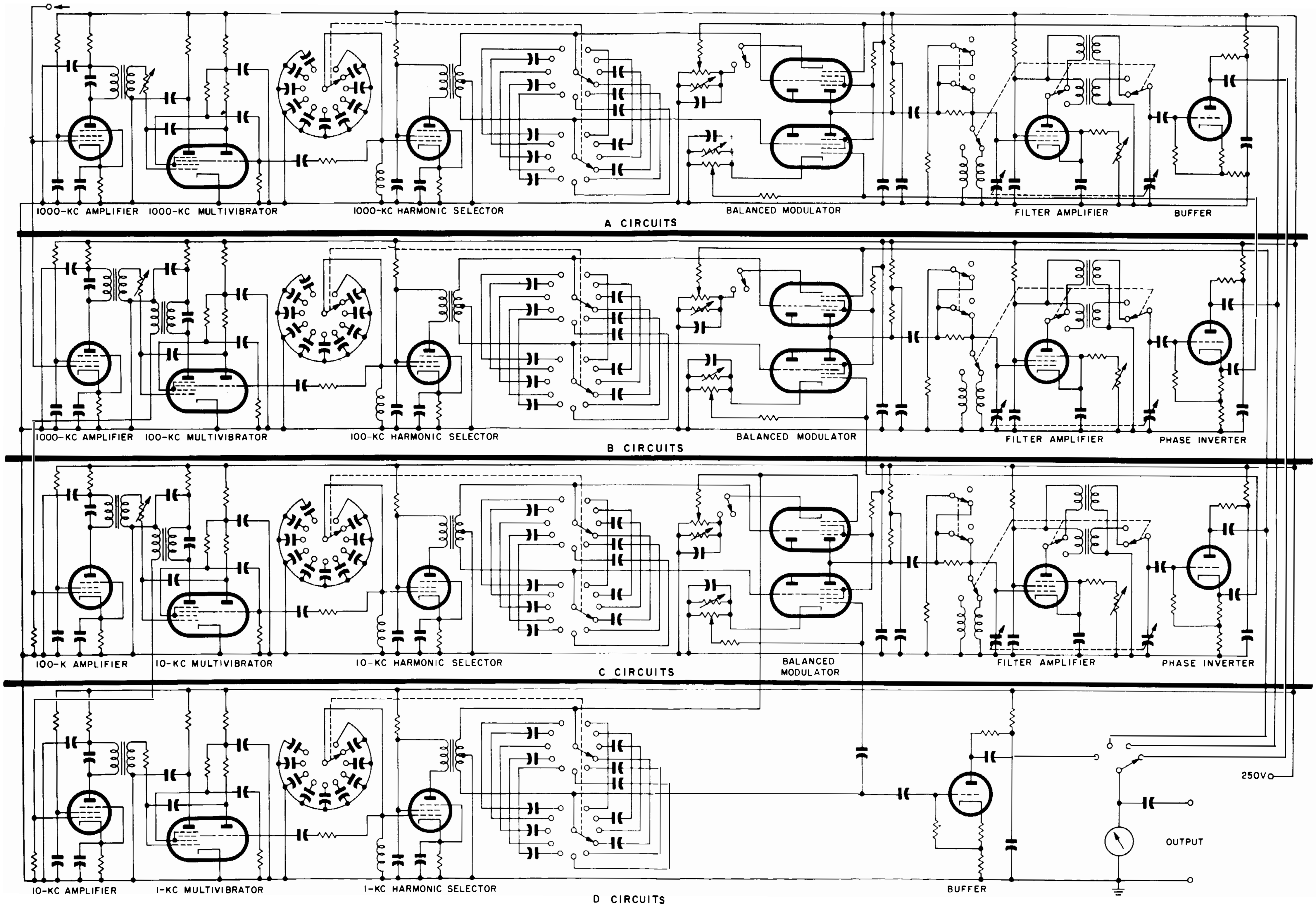


Figure 1-190. Schematic diagram of the 100-kc synchronized multivibrator used in the Plessey synthesizer as a decade divider of a 1000-kc standard and as a 100-kc harmonic generator

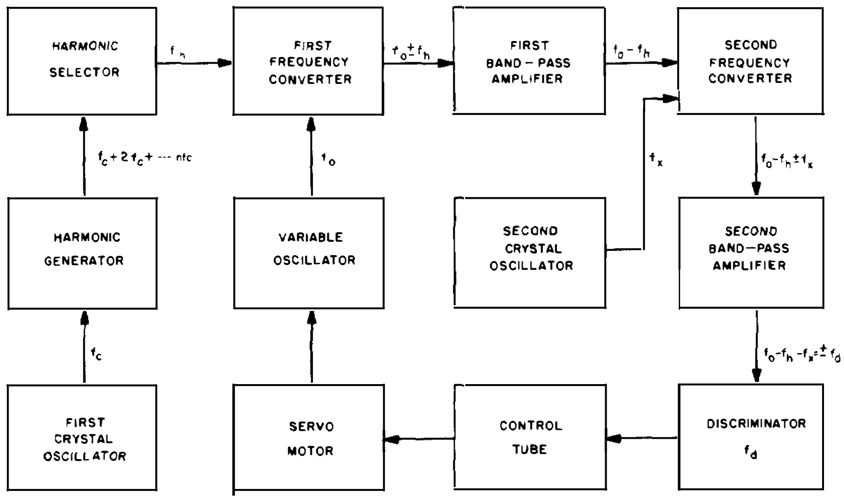


Figure 1-191. Block diagram of frequency-synthesizer circuit

particular crystal is selected. The number of controlled channels is directly proportional to the number of second-oscillator crystals. The sum and difference frequencies of the frequency converter are fed to the second band-pass amplifier, which amplifies and passes only the difference frequency, $(f_o - f_h) - f_x$. This difference frequency is fed to a discriminator. The number of channels controlled is directly proportional to the number of discriminators used. The d-c a-f-c output of the discriminator is used to control the bias of the control tube. The plate current of the control tube determines the rotor position of a servo motor, and the rotor is mechanically coupled to control the tuning elements of the variable oscillator. The servo motor continues to turn and thereby continues to change the frequency, f_o, until the output of the discriminator is zero. This occurs when the output of the second band-pass amplifier is equal to f_d, the frequency of the discriminator circuit. By reversing the polarity of the discriminator output leads, f_o can be made to vary in the opposite direction in order to reach equilibrium. Thus, for each value of f_h, f_x, and f_d, there are two equilibrium values of f_o. These are given by the equation

$$f_o = f_h + f_x \pm f_d \qquad 1\text{—}455\ (1)$$

1-456. Let

N = Total number of channels, f_o.
H = Number of harmonics used (1st crystal oscillator).
X = Number of crystals (2nd crystal oscillator).
D = Number of discriminators.

It can be seen from equation 1—455 (1) that

$$N = 2HXD \qquad 1\text{—}456\ (1)$$

The factor 2 is introduced by the fact that for each discriminator there are two values of f_o for each combination of f_h and f_x. One value is $f_o = f_h + f_x + f_d$, and the other is $f_o = f_h + f_x - f_d$.

1-457. As a concrete example, let us imagine that it is desired to cover the frequency range between 100 and 156 mc with the channels spaced 200 kc apart. The lowest value of f_o is to be 100.2 mc, and the highest is to be 156 mc. Thus,

$$N = \frac{(\text{max})\ f_o - (\text{min})\ f_o}{\Delta f_o} + 1 = \frac{156 - 100}{0.2} = 280 \qquad 1\text{—}457\ (1)$$

By equation 1—456 (1),

$$HXD = \frac{N}{2} = 140 \qquad 1\text{—}457\ (2)$$

The smallest total number of elements occurs when H, X, and D can be made as nearly equal to each other as possible, but in an actual design problem, this may not be the most practical solution. In our particular example let us assume that the first seven harmonics of f_c are to be used.
With H = 7, then by equation (2)

$$XD = \frac{140}{7} = 20 \qquad 1\text{-}457\ (3)$$

The combinations (X,D) possible are (20,1), (10,2), (5,4), (4,5), (2,10), and (1,20). For our problem we shall suppose that the combination ($X = 10$, $D = 2$) proves the most practical. Thus, with the use of 11 crystals in all and 2 discriminators, 280 crystal-controlled channels are to be obtained.

1-458. Let $f_1, f_2, f_3 \ldots . f_{280}$ designate the values of f_o from the lowest to the highest, in that order. Let $f_{x1}, f_{x2} \ldots . f_{x10}$ designate the values f_x from the lowest to the highest, in that order. The values of f_h in ascending order are $f_c, 2f_c \ldots . 7f_c$. Finally, let f_{d1} and f_{d2} designate the lower and the higher discriminator frequencies, respectively. To avoid the possibility of spurious conversion frequencies, the highest value of f_h should be less than the minimum associated 1st band-pass amplifier frequency, $f_o - f_h$. Also, the lowest value of f_x should be higher than the highest 2nd band-pass amplifier frequency, which, of course, will equal the highest discriminator frequency, f_d. The order of the variable-oscillator frequencies and the sequence of circuit connections required to provide each frequency is indicated by the following sequence of equations.

$$f_1 = f_c + f_{x1} - f_{d2} = 100.2 \text{ mc}$$
$$f_2 = f_c + f_{x1} - f_{d1} = 100.4 \text{ mc}$$
$$f_3 = f_c + f_{x2} - f_{d2} = 100.6 \text{ mc}$$
$$f_4 = f_c + f_{x2} - f_{d1} = 100.8 \text{ mc}$$

etc

$$f_{20} = f_c + f_{x10} - f_{d1} = 104 \text{ mc}$$
$$f_{21} = f_c + f_{x1} + f_{d1} = 104.2 \text{ mc}$$
$$f_{22} = f_c + f_{x1} + f_{d2} = 104.4 \text{ mc}$$
$$f_{23} = f_c + f_{x2} + f_{d1} = 104.6 \text{ mc}$$
$$f_{24} = f_c + f_{x2} + f_{d2} = 104.8 \text{ mc}$$

etc

$$f_{40} = f_c + f_{x10} + f_{d2} = 108 \text{ mc}$$
$$f_{41} = 2 f_c + f_{x1} - f_{d2} = 108.2 \text{ mc}$$
$$f_{42} = 2 f_c + f_{x1} - f_{d1} = 108.4 \text{ mc}$$
$$f_{43} = 2 f_c + f_{x2} - f_{d2} = 108.6 \text{ mc}$$

etc

$$f_{280} = 7 f_c + f_{x10} + f_{d2} = 156 \text{ mc}$$

1-459. It can be seen from the equation sequence in paragraph 1-458 that the difference between the channel frequencies is equal to the difference between the discriminator frequencies. Thus,

$$\Delta f_d = f_{d2} - f_{d1} = \Delta f_o = f_2 - f_1 = 0.2 \text{ mc} \qquad 1\text{—}459\ (1)$$

Note in the equation sequence in paragraph 1-458, that for a given harmonic frequency, f_h, all the f_x crystals are used in sequence before the polarity of the discriminator outputs are reversed. In other words, f_{d1} and f_{d2} are first subtracted from all combinations of a particular harmonic with the X-crystal frequencies and then added to the same combinations. This process is repeated with each harmonic. If we subtract the equation for f_1 from the equation for f_3, we have

$$\Delta f_x = f_{x2} - f_{x1} = f_3 - f_1 = 0.4 \text{ mc} = 2\,\Delta f_d \qquad 1\text{—}459\ (2)$$

In the general case,

$$\Delta f_x = D\,\Delta f_d \qquad 1\text{—}459\ (3)$$

where D is the number of discriminators. The same value of Δf_x also holds between any other two consecutive values of f_x. Thus,

$$f_{x2} = f_{x1} + \Delta f_x$$
$$f_{x3} = f_{x1} + 2\,\Delta f_x$$
$$f_{xn} = f_{x1} + (n - 1)\,\Delta f_x = f_{x1} + (n - 1)\,D\,\Delta f_d \qquad 1\text{—}459\ (4)$$

The highest frequency of the second crystal oscillator is given by equation (4) when $n = X =$ the total number of crystals. Thus,

$$(\text{max})\ f_x = f_{x1} + (X - 1)\,D\,\Delta f_d \qquad 1\text{—}459\ (5)$$

With the use of equation (5) we can find the lowest discriminator frequency, f_{d1}. This is done by subtracting the equation for f_{20} from the equation for f_{21}, which gives

$$f_{21} - f_{20} = f_{x1} - f_{x10} + 2f_{d1} = \Delta f_d = 0.2 \text{ mc} \qquad 1\text{—}459\ (6)$$

Since f_{x10} is a particular case of (max) f_x, we can substitute equation (5) in equation (6) to obtain a general equation. We find

$$f_{x1} - f_{x1} - (X - 1)\,D\,\Delta f_d + 2\,f_{d1} = \Delta f_d$$

On rearranging after canceling out f_{x1},

$$f_{d1} = \Delta f_d \frac{(1 + DX - D)}{2} \qquad 1\text{—}459\ (7)$$

The general equation for any particular discriminator frequency, f_{dn}, is similar to that for f_{xn} given by equation (4). Thus,

$$f_{dn} = f_{d1} + (n - 1)\ \Delta f_d. \qquad 1\text{—}459\ (8)$$

and

$$(\max)\ f_d = f_{d1} + (D - 1)\ \Delta f_d \qquad 1\text{—}459\ (9)$$

The next problem is to obtain a general equation for f_c. This can be had by subtracting the equation for f_{40} from the equation for f_{41}. The remainder is

$$\Delta f_d = f_c + f_{x1} - f_{x10} - 2\ f_{d2}$$

where $f_{x10} =$ (max) f_x and $f_{d2} =$ (max) f_d. Thus,

$$f_c = 2\ f_{d2} - f_{x1} + f_{x1} + (X - 1)\ D\ \Delta f_d + \Delta f_d$$

or

$$f_c = 2\ f_{d2} + 2\ f_{d1}$$

$$f_c = 2\ [(\max)\ f_d + (\min)\ f_d] \qquad 1\text{—}459\ (10)$$

Finally, with f_c determined, we can use the equation for f_1 to find f_{x1}.

1-450. We are now in a position to express any of the circuit frequencies in terms of the parameters f_1, Δf_o, N, H, X, and D. For the nth channel,

$$f_n = f_1 + (n - 1)\ \Delta f_o \qquad 1\text{—}460\ (1)$$

For the highest channel,

$$(\max)\ f_o = f_1 + (N - 1)\ \Delta f_o \qquad 1\text{—}460\ (2)$$

For the lowest discriminator frequency,

$$f_{d1} = \frac{\Delta f_o\ (1 + DX - D)}{2} \qquad 1\text{—}460\ (3)$$

For the nth discriminator frequency,

$$f_{dn} = \frac{\Delta f_o\ (DX - D + 2n - 1)}{2} \qquad 1\text{—}460\ (4)$$

For the highest discriminator frequency,

$$(\max)\ f_d = \frac{\Delta f_o\ (DX + D - 1)}{2} \qquad 1\text{—}460\ (5)$$

For the fundamental of the 1st crystal oscillator,

$$f_c = 2\ \Delta f_o\ D\ X = \frac{\Delta f_o\ N}{H} \qquad 1\text{—}460\ (6)$$

For the nth harmonic frequency,

$$f_{hn} = 2\ \Delta f_o\ D\ X\ n \qquad 1\text{—}460\ (7)$$

For the highest harmonic,

$$(\max)\ f_h = \Delta f_o\ N \qquad 1\text{—}460\ (8)$$

For the lowest frequency of the 2nd crystal oscillator,

$$f_{x1} = \frac{2\ f_1 - \Delta f_o\ (3\ DX - D + 1)}{2} \qquad 1\text{—}460\ (9)$$

For the nth frequency of the 2nd crystal oscillator,

$$f_{xn} = \frac{2\ f_1 - \Delta f_o\ (3\ DX + D + 1 - 2\ D_n)}{2} \qquad 1\text{—}460\ (10)$$

For the highest frequency of the 2nd crystal oscillator,

$$(\max)\ f_x = \frac{2\ f_1 - \Delta f_o\ (DX + D + 1)}{2} \qquad 1\text{—}460\ (11)$$

and the difference between the consecutive values of f_x,

$$\Delta f_x = \Delta f_o\ D \qquad 1\text{—}460\ (12)$$

1-461. On applying the equations in paragraph 1-460 to the numerical example that has been assumed, where f_1 = 100.2 mc, Δf_o = 0.2 mc, N = 280, H = 7, X = 10, and D = 2, we find that

$$f_{d1} = \frac{0.2\ (1 + 20 - 2)}{2} = 1.9 \text{ mc}$$

$$f_{d2} = 1.9 + 0.2 = 2.1 \text{ mc}$$

$$f_c = \frac{0.2 \times 280}{7} = 8 \text{ mc}$$

f_{h1}, f_{h2}, etc. = 8, 16, 24, 32, 40, 48, 56 mc

$$f_{x1} = \frac{200.4 - 0.2\ (60 - 2 + 1)}{2} = 94.3 \text{ mc}$$

$$\Delta f_x = 0.4 \text{ mc}$$

f_{x1}, f_{x2}, etc = 94.3, 94.7, 95.1, 95.5, 95.9, 96.3, 96.7, 97.1, 97.5, 97.9 mc

Note that the highest harmonic frequency is equal to the bandwidth of the frequency range being covered.

1-462. From the equation, $f_o = f_h + f_x \pm f_d$, it can be seen that the frequency stability will be approximately the stability of the crystal oscillators. This is because f_d is so very much lower than f_h and f_x. Although the discriminators should be designed with low-loss, temperature-compensating materials, even a large percentage variation in f_d would be negligible in its percentage effect upon f_o. Since it is the sum of f_h and f_x that determines f_o, the maximum *percentage* frequency deviation of the total can be no greater than that of the crystal oscillators individually. Without oven control, the channel frequencies can be maintained within a tolerance of ±0.005 per cent, and better.

RADIO SET AN/ARC-33

1-463. Radio Set AN/ARC-33 is an airborne receiver-transmitter designed to operate in the v-h-f and u-h-f spectrum. This equipment, developed by the Bendix Corporation, employs a modified version of the Bendix frequency-synthesizing system discussed in the foregoing paragraphs. The frequency-control section (see figure 1-192 for block diagram) is designed to permit receive-transmit communication on any one of 1750 channels spaced 100 kc apart in the 225-to-399.9-mc band.

1-464. An important modification in the synthesizing system arises from the fact that the synthesized frequencies are utilized as heterodyne injection signals during reception and as carrier signals during transmission. The tuning controls are such that the receiver and transmitter circuits are always automatically tuned to the same channel, but since the desired receiver injection frequency must differ from the tuned channel frequency by an amount equal to the receiver intermediate frequency, the design engineers had to decide whether to let the variable-frequency oscillator be, in effect, a subharmonic local oscillator for the receiver or a subharmonic master oscillator for the transmitter. They decided in favor of the receiver. Thus, the stabilized output frequency, f_o, of the vfo, after being multiplied 12 times, is used directly as the injection voltage in the 1st mixer stage in the receiver. This provides a fixed intermediate frequency, f_i, of 15.325 mc for each of 1750 channels. Now, the v-f-o output, f_o, is also used in the synthesis of the channel frequency of the transmitter. Since the 12th harmonic of f_o always differs from the channel frequency by an amount equal to f_i, all that needs to be done in principle is to mix $12f_o$ with a fixed oscillator frequency equal to f_i and for the sum product to be isolated and amplified for use as the transmitter carrier. In the ARC-33 transceiver this effect is achieved by mixing the output, f_s, of a crystal-controlled oscillator (called the "sidestep" oscillator) with the 6th harmonic of f_o, then selecting the sum frequency ($6f_o$ plus f_s) and doubling it to form the carrier frequency, f_a, which, after amplification, is fed to the antenna. Summarizing these frequency relations in the form of equations, we have:

injection frequency $= 12\, f_o$

intermediate frequency $= f_i = f_a - 12f_o$

antenna (channel) frequency $= f_a = 12f_o + f_i$
$= 2\,(6f_o + f_s)$

sidestep frequency $= f_s = f_i/2$

Note the necessary harmonic relation between the sidestep output and the intermediate frequency. It is also important to note that the principle involved in the use of a sidestep oscillator permits, not only a Bendix synthesizing system, but any synthesizing system to be readily modified for the dual-purpose requirements of transceiver frequency control.

1-465. Comparison of figures 1-191 and 1-192 will reveal that the 2nd crystal oscillator in the basic Bendix synthesizer has been replaced by two crystal oscillators (the 2nd and 3rd in figure 1-192) in Radio Set AN/ARC-33. This has been done to permit a greater number of frequencies with a fewer number of crystals.

1-466. Another significant modification occurs in the ARC-33 discriminator circuit. As is explained in more detail in a subsequent paragraph, the ARC-33 discriminator is not a conventional type that employs a parallel tuned circuit to control the phase differences between the input voltage components. In that type of discriminator the output voltage always has a net d-c component unless the input frequency is equal to the antiresonant frequency of the tuned tank. The polarity of the d-c component is an index of whether the input frequency is higher or lower than that at which the tank is tuned, and the amplitude of the d-c component can be a measure of the amount of difference between the two frequencies. It is this type of discriminator that is assumed in the discussion of the basic Bendix synthesizing system, the d-c output of which is used to control the variable oscillator tuning. In the ARC-33, however, the discriminator does not employ a tuned circuit, but instead is fed a signal that is controlled by the 4th crystal oscillator. The discriminator also receives

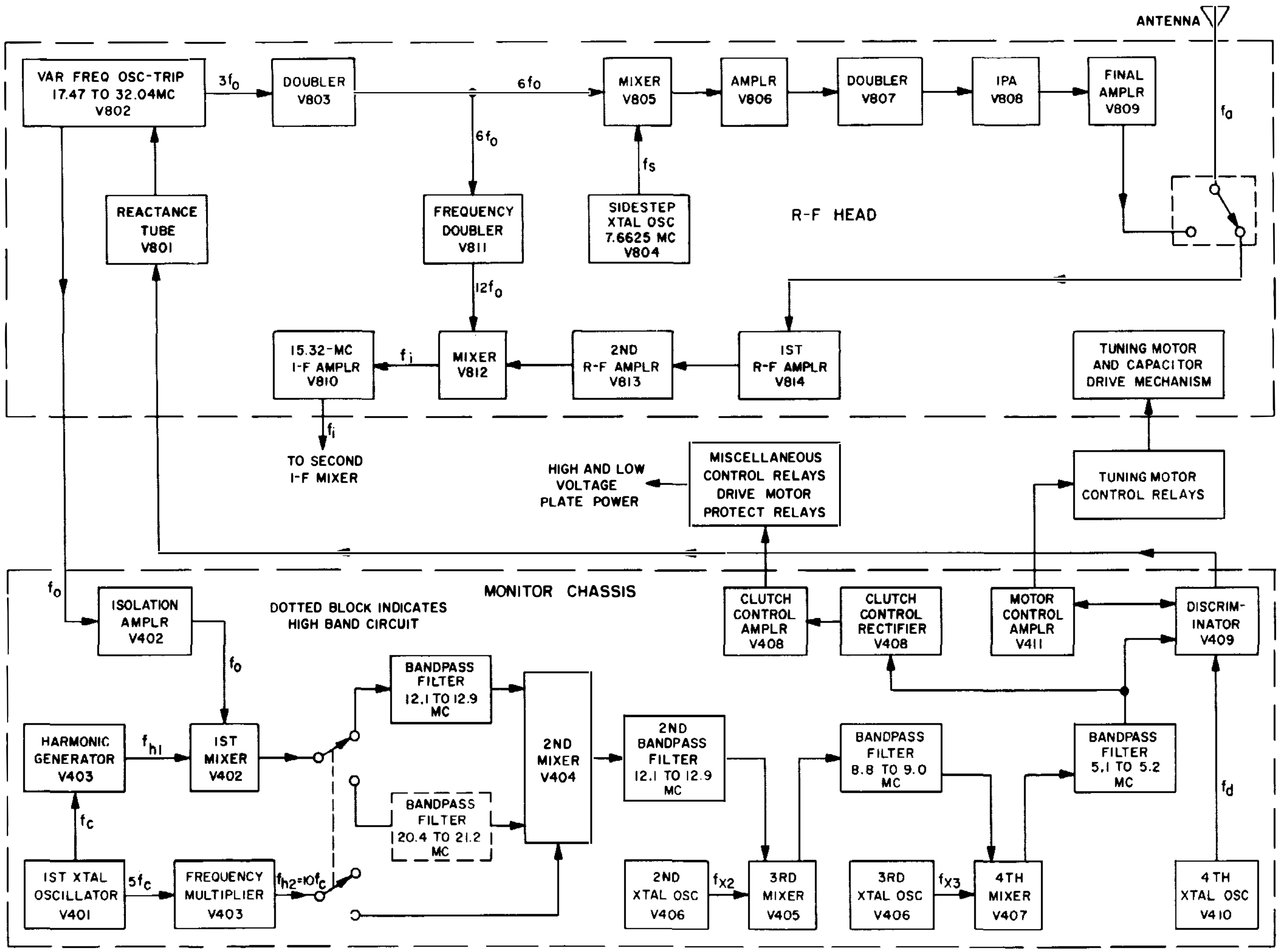


Figure 1-192. Block diagram of frequency control section of Radio Set AN/ARC-33 showing modified version of Bendix synthesizing system

an input signal from the 4th mixer. The two signals combine in the discriminator circuit to provide a net d-c output only when the two signals are of the same frequency. When the frequencies are identical, the behavior of the discriminator is quite similar to the behavior of one that employs a tuned circuit, the d-c output depending upon the differences in phase between the input voltages. Since the 4th oscillator can be controlled by either one of two crystals, this arrangement is equivalent to having two discriminators of the tuned-circuit type.

1-467. In paragraph 1-456 it is explained that for each discriminator, two values of f_o are possible. In Radio Set AN/ARC-33 only one of these values is used for each 4th-oscillator frequency, namely

$$f_o = f_h + f_x + f_d \qquad 1\text{-}467\ (1)$$

In the equation above, f_h is the selected *effective* harmonic of the 1st crystal oscillator, equal to f_{h1} on the low band and to $(f_{h1} + f_{h2})$ on the high band, f_x equals the sum of the frequencies of the 2nd and 3rd crystal oscillators $(f_{x2} + f_{x3})$, and f_d is the frequency of the 4th crystal oscillator fed to the discriminator.

1-468. Other modifications of the Bendix system as occur in the frequency-control circuits of Radio Set AN/ARC-33 are of an even less radical nature than those described above. There is the division of the 1st bandpass stage into low-band and high-band circuits, and there is the addition of an a-f-c reactance tube, which is actually more of an extension of the modification caused by the use of a crystal-controlled discriminator, but these and other special circuit arrangements are best explained in the more detailed analyses later. At this point it will be helpful to examine briefly the role each of the various oscillator frequencies plays in controlling the final antenna frequency.

1-469. First, we shall examine the simplified block diagram shown in figure 1-193. The frequency-control system indicated represents an imaginary synthesizing circuit that provides the same transmitter output frequencies, f_a, as does Radio Set AN/ARC-33. The principal difference between the

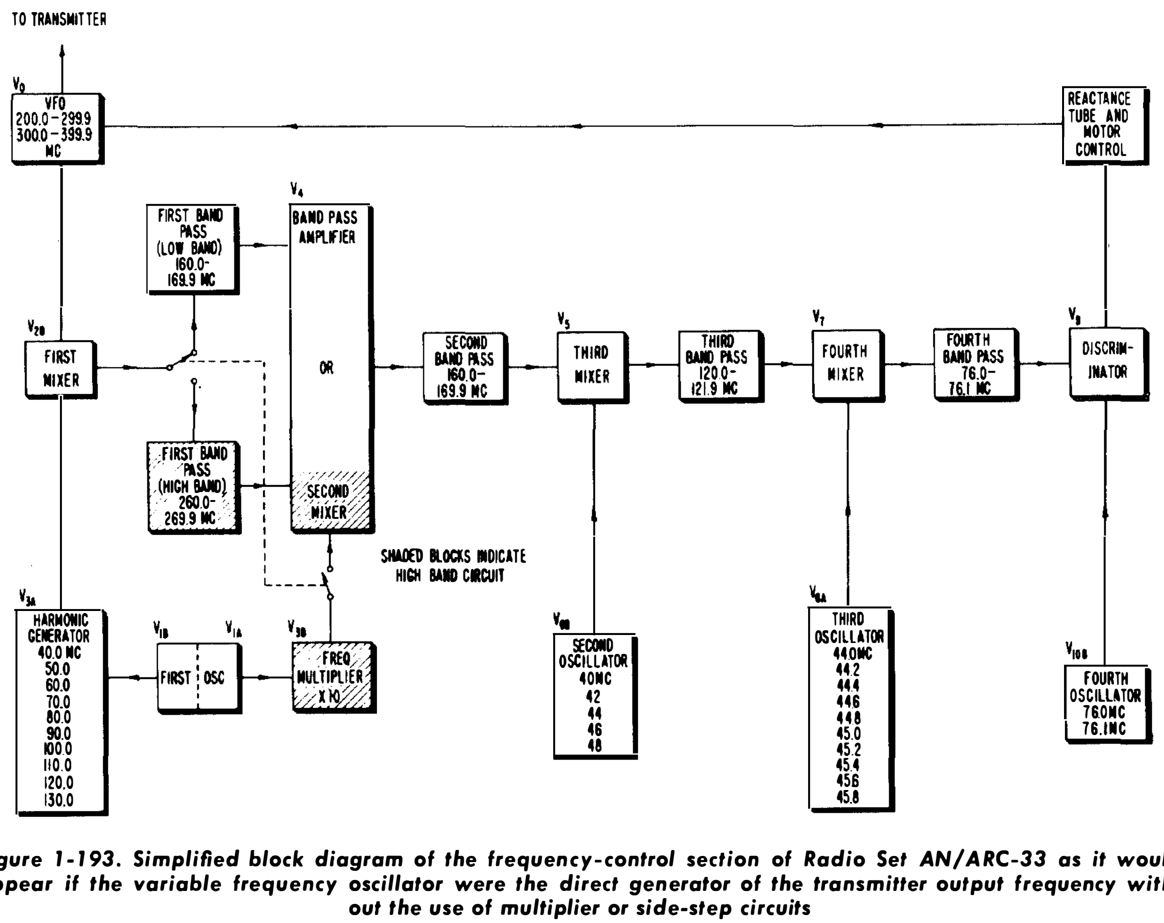


Figure 1-193. Simplified block diagram of the frequency-control section of Radio Set AN/ARC-33 as it would appear if the variable frequency oscillator were the direct generator of the transmitter output frequency without the use of multiplier or side-step circuits

imaginary and the actual systems is that the imaginary system is not required to provide an injection voltage for a receiver heterodyne circuit, nor is it required to employ multiplier stages following the variable oscillator. In other words, f_0 can be assumed to equal f_a. Under these conditions the various crystal and harmonic frequencies would assume the simple values shown in figure 1-193. The actual circuit frequencies are those indicated in the frequency diagram of figure 1-194. A comparison of the frequencies in the two systems will show that the injected frequencies in each mixer stage of the imaginary system vary from one to the next in steps that are 12 times greater than are the corresponding steps in the actual system. This does not mean, except in the case of the 1st crystal oscillator and its harmonics, that the imaginary crystal frequencies are 12 times the actual crystal frequencies; it is only the *differences* between adjacent frequencies that are related in the proportion of 12 to 1. Since the frequency of the variable frequency oscillator in the actual circuit is eventually multiplied 12 times, it can be seen that the frequency steps in the actual circuit are equivalent to those in the imaginary circuit insofar as they add or subtract in the control of the antenna frequency. Thus, we can say that the antenna frequency is effectively synthesized in 10-mc units by the 1st crystal oscillator and harmonic generator, 2-mc units by the 2nd crystal oscillator, 0.2-mc units by the 3rd crystal oscillator, and finally to the nearest 0.1-mc unit by the discriminator and 4th crystal oscillator.

Detailed Circuit Description

1-470. The principal component of Radio Set AN/ARC-33 is Receiver-Transmitter RT-173/ARC-33. The receiver-transmitter is divided into a number of sectional components, two of which are of importance to us:

a. The *monitor chassis,* which contains all the crystal circuits for controling the variable-frequency oscillator.

b. The *r-f head,* which contains the variable-frequency oscillator, the multiplier circuits, the 1st i-f mixer, the sidestep oscillator, as well as the r-f amplifiers of the receiver and the power amplifiers of the transmitter. Also of importance to us are the relays which control the tuning motor. These are mounted on the main frame. We shall discuss the monitor circuits first, and then those in the r-f head. Except for occasional insertions and editing, the descriptions to follow are largely extracts from USAF Technical Order No. 12R2-2ARC33-2.

Monitor Chassis

1-471. The monitor chassis in Radio Set AN/ARC-33 concerns only the frequency control of the variable-frequency oscillator and electronic control of the tuning-capacitor drive motor with its clutches. There are no other circuits involved.

The detailed circuit descriptions of the monitor chassis are made with reference to the component symbols employed in the block diagram of figure 1-195 and the schematic diagram of figure 1-196. With the exception of a coaxial connector for the r-f input from the variable-frequency oscillator, all external connections to the unit are made through a single connector, which is so arranged that connection automatically is made when the chassis is inserted in its proper place in the main frame.

1-472. *FIRST CRYSTAL OSCILLATOR.* The 1st crystal oscillator is a single-frequency, fundamental-mode, 833.333-kc oscillator of the cathode-coupled Butler type. A selected harmonic of the oscillator is mixed with the frequency of the variable frequency oscillator in the 1st mixer. A dual triode tube, V401, is employed as the oscillator tube, which has two output connections. Section A is tuned to the 5th harmonic of the crystal and feeds the grid of frequency multiplier V403B. The plate output of the grounded-grid oscillator section B is coupled through capacitor C407 to the harmonic generator grid. The crystal unit, which is of the type CR-28/U, is mounted in a type HD-54/U crystal oven. The oven employs two heaters and thermostats, one heater being used to bring the temperature quickly up to the operating level, whereas the other, which has a lower wattage, is used to maintain constant operating temperature. In order to check the oscillator for proper operation, a test connection for measuring rectified grid current is brought out to test socket X412. The stability of the final transmitter frequency is more dependent upon the stability of this oscillator than upon that of any of the others. The reason is that the 1st crystal oscillator controls a greater percentage of the final frequency, especially in the high band, than do the other oscillators. This can readily be seen if we visualize the final frequency as being synthesized by adding together the crystal-oscillator frequencies in the simplified block diagram of figure 1-193. The key function of this oscillator is the reason why the highly stable Butler circuit is employed.

1-473. *HARMONIC GENERATOR.* The function of harmonic generator V403A is to produce any selected harmonic of the first crystal oscillator

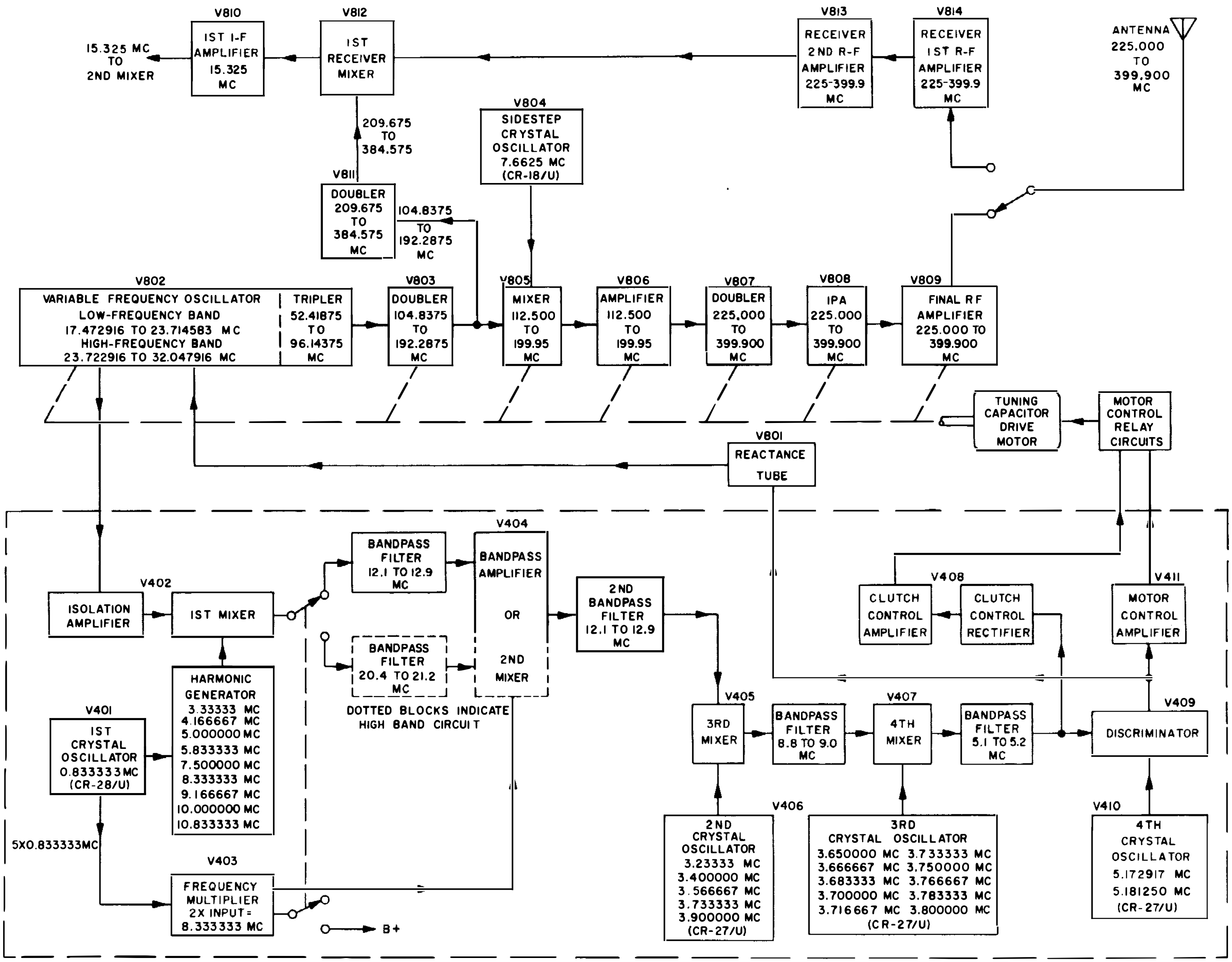


Figure 1-194. Frequency diagram of frequency-control system in Radio Set AN/ARC-33

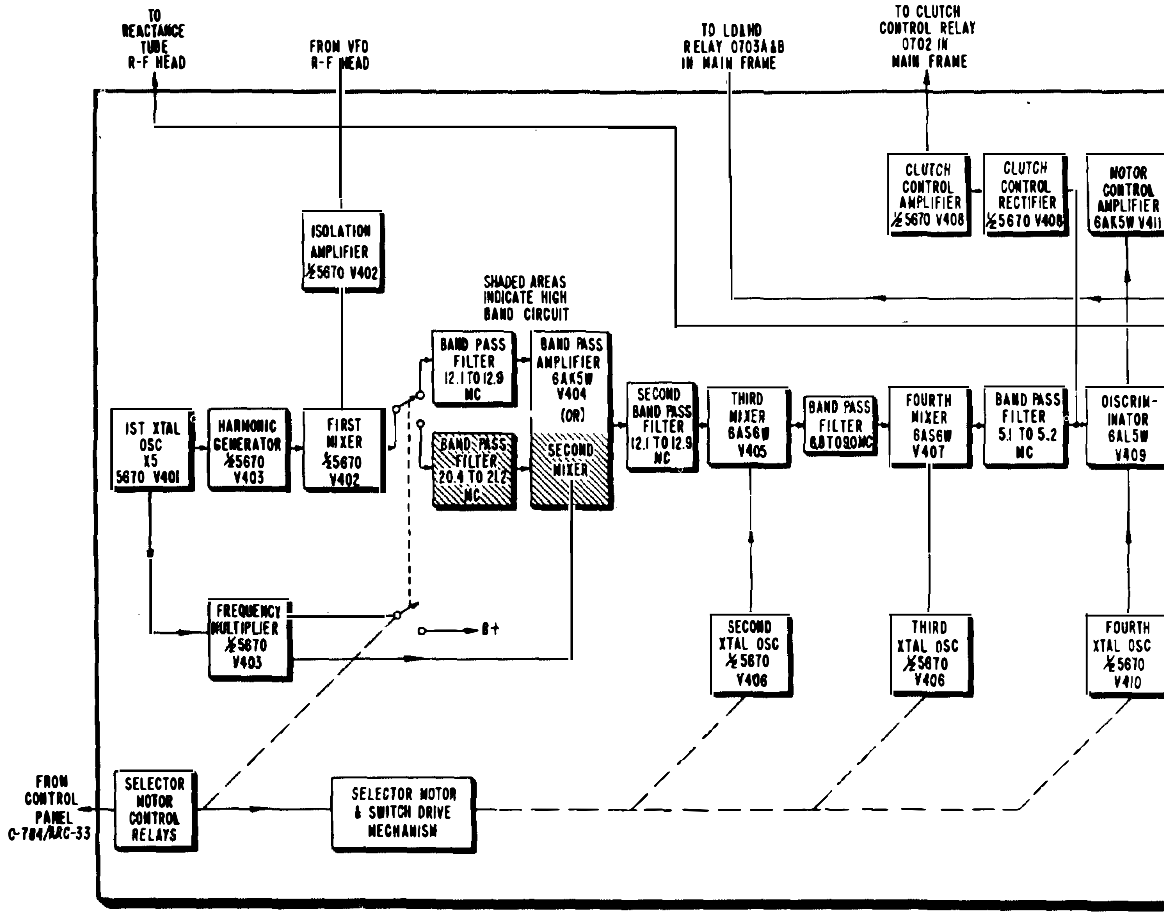


Figure 1-195. Monitor chassis. Block diagram of a-f-c system in Radio Set AN/ARC-33

output from the 4th to the 13th, inclusive. The selected harmonic is fed to the 1st mixer to be heterodyned with the frequency from the variable frequency oscillator. Harmonic selection is achieved by capacitance tuning the primary and secondary of the harmonic generator output transformer T401 to the desired harmonic frequency. The switching is accomplished by 10-position rotary switches S403A and S403B, which are driven by the selector motor through a harmonic generator clutch. These switches merely select the proper fixed capacitors for tuning the primary and secondary winding to the desired harmonic of the 833.333-kc fundamental.

1-474. R-f input from the 1st crystal oscillator is fed to the harmonic generator tube, V403A, through coupling capacitor C407. A grid bias far below cutoff is provided by grid resistor R407 in order to ensure an output rich in harmonics. The primary and secondary windings of transformer T401 are permeability tuned for alignment at the lowest (4th) harmonic. The highest frequency, which is the 13th harmonic, is determined by capacitors C418 and C419 across the primary and C420 and C421 across the secondary. Capacitors C418 and C421 are trimmers for alignment at the highest frequency. The selection of all harmonics, up to but not including the 13th, is accomplished by switching in the proper fixed capacitor C408 through C417 across the primary of transformer T401, and C222 through C431 across the secondary. For the lowest harmonic, C416 and C417 for the primary and C430 and C431 for the secondary are connected in parallel. For the highest harmonic, no auxiliary capacitor is switched into the circuit. For proper operation of the monitor, it is necessary that the input level of the 1st mixer be approximately the same for each selected harmonic. This is accomplished by selecting a grid bias for the harmonic generator which causes the output level to vary inversely with frequency under a constant plate load impedance. Then, by maintaining the "Q" of the transformer T401 windings constant at all selected frequencies, the resultant plate load impedance of tube V403 can be made to vary directly with frequency because of the increasing inductance/capacitance ratio. The resultant voltage across the T401 transformer windings, therefore, is essentially constant regardless of the harmonic selected.

1-475. *FIRST MIXER.* The 1st mixer tube, V402, combines the output of the harmonic generator with the output frequency of the variable frequency oscillator (vfo) whose frequency is to be controlled by the monitor circuits. The 1st mixer output consists of the difference frequency between the selected harmonic generator frequency and the v-f-o frequency, the latter frequency always being the higher. Input from the harmonic generator is fed directly to the control grid, whereas the input from the vfo, which enters the monitor chassis through connector J401, is fed from isolation amplifier V402A to the cathode of the 1st mixer through coupling capacitor C432. Output coupling is provided by bandpass coil assembly Z401, which consists of a low-band circuit and a high-band circuit. The bandpass selector relay selects the proper circuit of bandpass coil assembly Z401. The reason for using two bandpass circuits is that the broad frequency range of the radio set makes it necessary to divide the range into two smaller ranges. The band-determining factor is the selection of the 1st digit of the desired channel frequency at the control panel. The control panel switch energizes the band selector relay when the 1st digit is 2; that is, when the antenna frequency is to be less than 300 mc. For antenna frequencies of 300 mc and above, the band relay is unenergized. Coil assembly Z401 is designed to pass a band from approximately 12.1 to 12.9 mc and a band from approximately 20.4 to 21.2 mc. Because of the use of two bandpasses at the 1st mixer output, each of the harmonic generator selected output frequencies is used twice, once in each band.

1-476. *BANDPASS AMPLIFIER.* The bandpass amplifier, V404, is an amplifier for the 1st mixer output on the low-frequency band and is used to improve the bandpass characteristics of the circuit. This stage is unconventional inasmuch as it also functions as a 2nd mixer, the operation of which is described in the following paragraph. The plate load consists of transformer T402, which is tuned to the same frequency band as the low band of Z401. The bandpass amplifier is capacitance-coupled through capacitor C443 to the number one grid of the 3rd mixer, V405.

1-477. *SECOND MIXER.* The 2nd mixer stage utilizes the same tube elements of V404 as are used when the tube is operated as a bandpass amplifier. However, this dual usage does not occur simultaneously. Tube V404 functions as a 2nd mixer only when the radio set operates in the high band of the frequency range. The 2nd mixer combines the 1st mixer output frequency with the 10th harmonic of the 1st crystal oscillator frequency. The output consists of the difference frequency, where the 10th harmonic frequency is always lower than the output frequency of the 1st mixer. Note that the reduction of the 1st mixer output frequency by an amount equal to the 10th harmonic of the 1st crystal oscillator frequency is equivalent to extending the harmonic generator range from the 13th to the 23rd harmonic and eliminating the high band of bandpass coil assembly Z401. Input to the 2nd mixer from the high-band bandpass filter is fed to the control grid along with the input from frequency multiplier V403B. The plate load consists of transformer T402, which is tuned to pass a band from 12.1 to 12.9 mc. The desired band width for the transformer is obtained through the use of loading resistors R418 and R419. Transformer T402 is the same plate load circuit for the 2nd mixer as is used when the stage operates as a bandpass amplifier.

1-478. *FREQUENCY MULTIPLIER.* The frequency multiplier, V403B, is fed with the 5th harmonic output from the 1st crystal oscillator. The output transformer, T406, is tuned to 8.33333 mc, which is the 2nd harmonic of the multiplier input frequency. Thus, V403B is a frequency doubler and feeds its output, the 10th harmonic of the 1st crystal oscillator frequency, to the control grid of the 2nd mixer, V404. When the radio set is operated in the low band of its frequency range, no output is derived from the frequency multiplier because plate voltage is removed from tube V403B. The application of plate voltage is controlled by the band selector relay. When the band selector relay switches the high bandpass filter into the 1st mixer circuit, it also applies voltage to the plate of the frequency multiplier so that an 8.33333-mc signal is fed to the 2nd mixer. The frequency multiplier is capacitively coupled to the control grid of the 2nd mixer through capacitor C457.

1-479. *SECOND CRYSTAL OSCILLATOR.* The 2nd crystal oscillator provides a selection of any

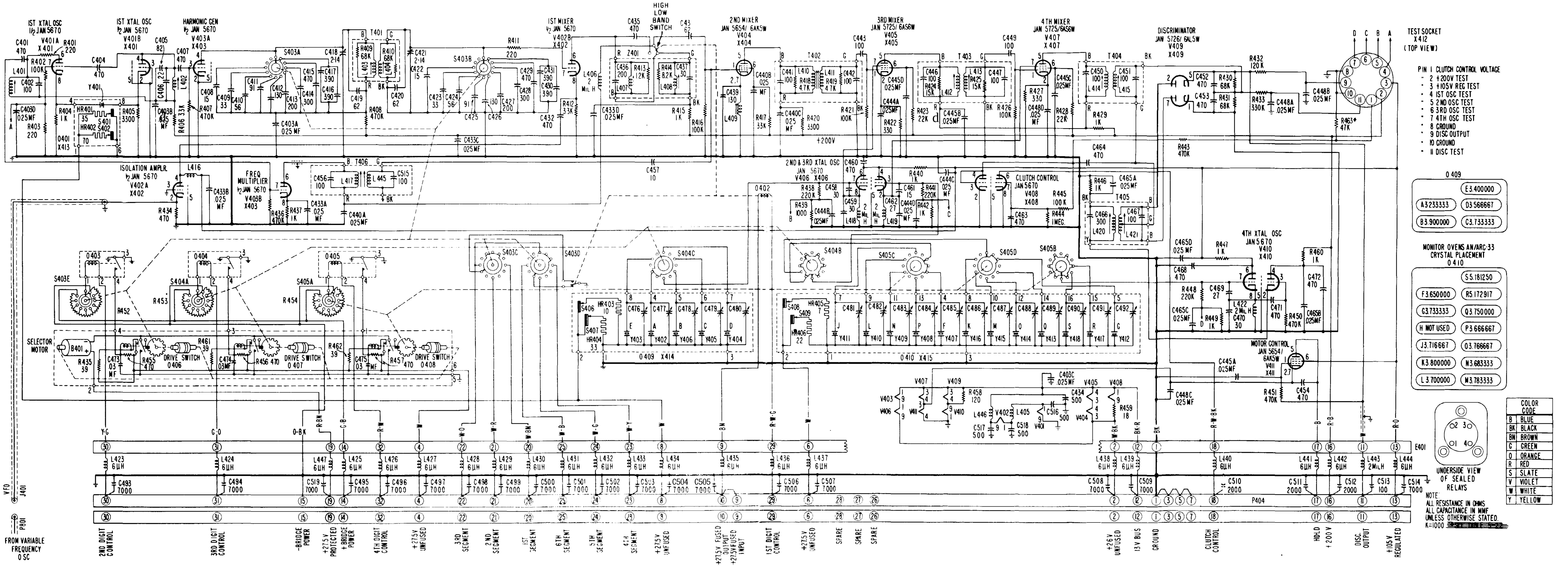


Figure 1-196. Monitor chassis. Schematic diagram of a-f-c system in Radio Set AN/ARC-33

one of five crystal-controlled frequencies by means of crystal switching. The crystal-controlled output frequency is fed to the suppressor grid of the 3rd mixer to be mixed with either the 2nd mixer or the bandpass amplifier output frequency. The oscillator circuit utilizes one half of the duplex triode tube V406 as a grounded-plate Pierce oscillator. R-f output is taken from the cathode. Each of the five crystal units, which are of the CR-27/U type, has a separate trimmer so that each selected 2nd crystal oscillator frequency can be adjusted exactly. All of the crystal units with their trimmers are enclosed in a single oven (Bendix Radio type L205628) which is kept at approximately 75°C (167°F) by a thermostat-controlled heater. A booster heater and thermostat are provided in addition for quick warmup. In order that the oscillator operation may be checked, a test connection from the grid circuit is brought out for checking rectified grid current at test socket X412.

1-480. *THIRD MIXER.* The circuit of the 3rd mixer, V405, is similar to that of the 2nd mixer except that the signal voltage from its heterodyne crystal oscillator is injected at the suppressor grid. The 3rd mixer output circuit is tuned to a center frequency of 8.9 mc and is designed to pass a band approximately from 8.8 to 9.0 mc. The 3rd mixer combines the output frequencies of either the 2nd mixer or bandpass amplifier and the 2nd crystal oscillator, the output frequency of the former two always being the higher. The 3rd mixer output, which is the difference frequency, is fed from the secondary of transformer T403 to the 4th mixer through capacitor C449.

1-481. *THIRD CRYSTAL OSCILLATOR.* The 3rd crystal oscillator, the V406B circuit, is of the same design as the 2nd crystal oscillator except for the operating frequencies. The crystal units, which are of the CR-27/U type, are divided into two groups of five each. All are used throughout the total frequency range of the radio set. The 3rd crystal oscillator employs the second half of the same tube that is used for the 2nd crystal oscillator. The 10 crystal units with their trimmers are housed in a 13-position oven (Bendix Radio type N205651), of which one position is not used. The other two positions are used to mount the two crystals of the 4th crystal oscillator. The oven is thermostatically controlled at 75°C (167°F). A separate booster heater and thermostat are provided for quick warmup.

1-482. *FOURTH MIXER.* The 4th mixer, V407, is identical to the 3rd mixer and operates in the same manner. The 4th mixer output circuit is designed to pass a band of approximately 5.1 to 5.2 mc. In the 4th mixer are combined the output frequency of the 3rd mixer and that of the 3rd crystal oscillator, the former frequency always being the higher. The difference frequency is selected by the tuned plate transformer and is inductively coupled into the discriminator circuit.

1-483. *DISCRIMINATOR.* The purpose of the discriminator, V409, is to indicate a deviation in phase or frequency between its two inputs. One of these inputs is obtained from the 4th crystal oscillator and is used as the reference signal. The other discriminator input is derived from the 4th mixer output, and it is the deviation in phase of this signal from that of the reference signal that is to be indicated by means of the output voltage across R430 and R431. The magnitude of this discriminator voltage indicates the extent of the deviation, whereas the polarity indicates the direction of the deviation. The circuit design is similar to the type employed for frequency-modulation (FM) receivers, except that in this case the reference voltage is obtained from a separate reference oscillator instead of a parallel resonant tank circuit. The voltages applied to the discriminator are shown in figure 1-197. E_I is the input from the 4th mixer and E_R is the reference voltage from the 4th oscillator. To simplify the discriminator explanation, E_R, insofar as it adds vectorially with E_I to form the r-f voltages across the two diodes, is best interpreted in terms of two equal and separate voltages 180 degrees out of phase. One is E_{2B}, the r-f voltage at pin 2 with respect to point B; the other is E_{7B}, the r-f voltage at pin 7 with respect to point B. Assuming that the bypass capacitors C452 and C453 offer zero impedances to the r-f signals, pins 1 and 5 and point A are all at r-f ground potential. Thus, the r-f voltage (E_2) of plate pin 2 with respect to cathode pin 5 is the same as the voltage of pin 2 with respect to point A. Similarly, E_7, the r-f plate voltage of the second diode, is equal to the voltage of pin 7 with respect to point A. Now, E_2 and E_7 have one voltage component in common, which is E_I, the voltage of point B with respect to point A. Thus, E_2 and E_7 are equal to the resultants, respectively, obtained by adding vectorially to the common voltage E_I the voltages E_{2B} and E_{7B}. If we assume that C452 charges to the peak of E_2 and that C453 charges to the peak of E_7, the d-c polarities will be as indicated in figure 1-197 (D), where point A is shown as negative with respect to the two cathodes. The d-c output equals ($E_2 - E_7$), where both voltage symbols represent the positive peak magnitudes only. The output is positive or negative according to whether E_2 is greater or less than E_7, respectively; and this is dependent, respectively, upon whether E_I lags or leads the zero-output position, which is the 90-degree phase displacement from E_R that is shown in figure 1-197(A). Upon examination of the discriminator diagram it can be seen that the application of either of the input voltages alone does not develop a discriminator output, since in each case E_2 and E_7 will equal each other, and hence equal currents will flow in opposite directions through the two halves of discriminator load R430 and R431. Also, it can be seen from the vector diagram (A) that, if the two discriminator input voltages are exactly 90 degrees out of phase, the resultant voltages applied to the diode plates are equal, thus producing zero discriminator output. If, however, the variable frequency oscillator should vary in phase, the resultant voltages applied to the discriminator plates are unequal, as shown in vector diagrams (B) and (C). This results in a discriminator output voltage whose value and polarity depend upon the magnitude and direction, respectively, of the phase deviation. If there is a frequency deviation, the phase of the 4th mixer output, E_I, rotates completely around the 4th crystal oscillator output, E_R, with a consequent a-c voltage appearing in the discriminator output. The frequency of this ac is equal to the difference between the two discriminator input frequencies.

1-484. Discriminator output is applied to the tuning-motor-control amplifier and to the reac-

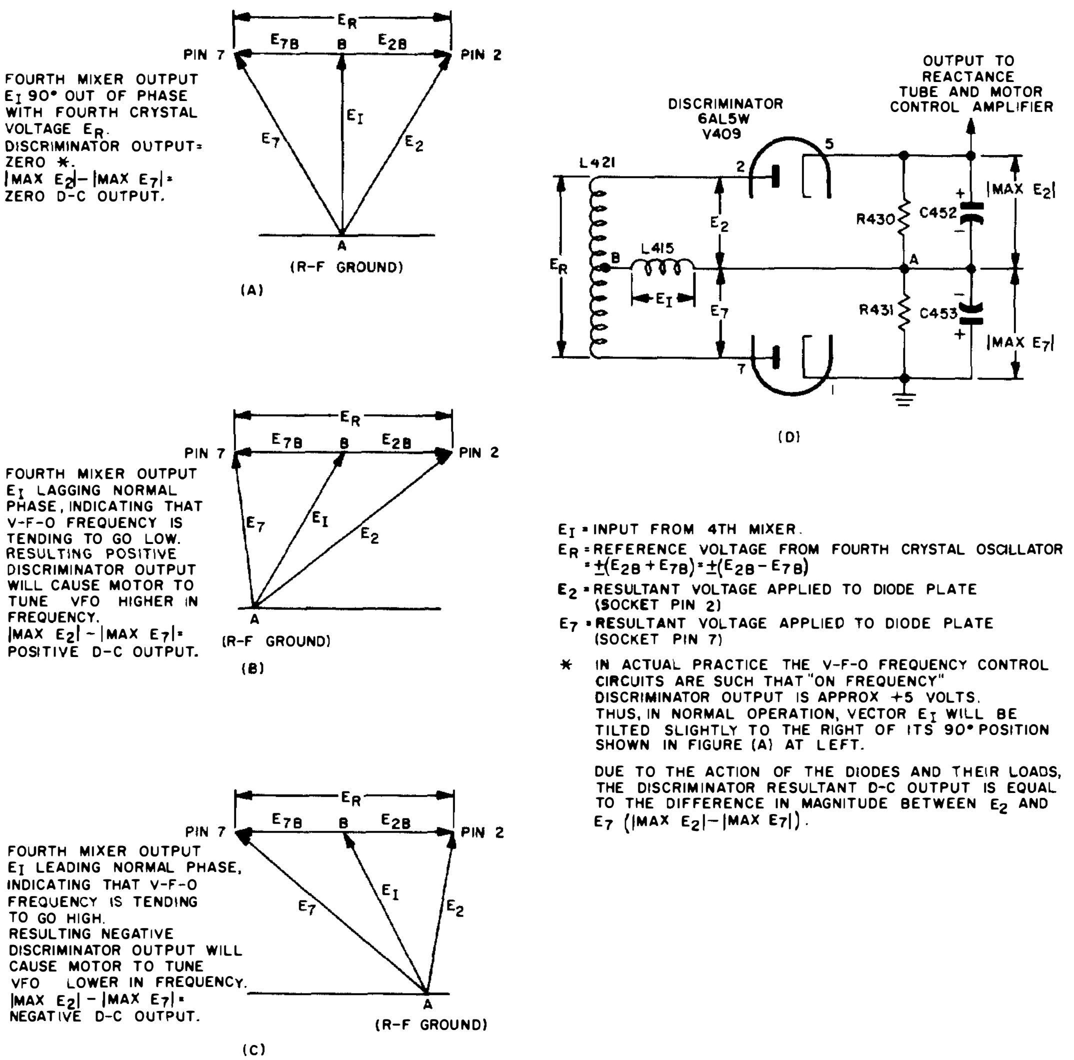


Figure 1-197. Discriminator operation as a function of the phase difference between the input voltages

tance tube shunted across the variable frequency oscillator. When the variable frequency oscillator approaches the desired operating frequency, a point is reached when the discriminator output frequency becomes equal to or less than the maximum frequency at which the reactance tube can respond. At this point the reactance tube immediately locks the variable frequency oscillator exactly to its correct frequency. When this occurs, the discriminator output is dc and is proportional to the phase difference between its two input voltages, which difference is, in turn, proportional to the amount of "pull" exerted by the reactance tube. This d-c control voltage is fed to the motor-control amplifier, V411, and causes the tuning motor to drive the variable frequency oscillator tuning capacitor to that point which eliminates excessive pull by the reactance tube. The circuit design is such that the equilibrium point corresponds to a discriminator output of approximately 5 volts positive. Normally it would be assumed that the "on-frequency" point would be reached when the discriminator output dropped to zero, which would occur when the two input signals were exactly 90 degrees out of phase. The zero-voltage state is not used, however, in order to avoid ambiguity in identifying the on-frequency condition and to simplify the control circuit by having an equilibrium control voltage of definite magnitude. For example, zero output not only occurs at each of the two (plus and minus) 90-degree phase conditions, but also occurs when the variable frequency oscillator is so far off frequency that there is no input from the 4th mixer, or if one or both of the input voltages fails, or when the two input frequencies are different and the discriminator records the best frequency. Thus, a plus 5-volt reference is used, and the control circuits are so biased that this control level establishes the on-frequency condition. The ac which the discriminator develops before the reactance tube "pull-in" point is reached, is prevented from affecting the motor-control amplifier by the low-pass resistance-capacitance filter, R451 and C445A.

1-485. By employing a phase-sensitive discriminator, it is possible to feed a correction voltage to the control circuits before an actual frequency deviation occurs, since a frequency deviation, unless it is an instantaneous, discontinuous jump, is first indicated as a phase deviation. Therefore, no frequency error occurs except that which may be due to the reference crystals themselves. All small and rapid frequency shifts of the variable frequency oscillator are corrected by the reactance tube. Larger and slower drifts are corrected by the motor-control amplifier and the tuning motor. An extremely large and sudden frequency jump of the variable frequency oscillator which takes it out of the range of the reactance tube causes the entire tuning sequence to recycle. However, this does not occur during normal operation. To aid in discriminator alignment and test, two test points are brought out to pins of test socket X412. One of these, pin 9, makes it possible to measure the total discriminator output. The other, pin 11, is connected to the load center-tap for checking the discriminator operation.

1-486. *FOURTH CRYSTAL OSCILLATOR.* The 4th crystal oscillator is used to control the operating frequency of the discriminator and thereby has final control of the exact frequency of the variable frequency oscillator. The 4th crystal oscillator can be switched to either one of two type CR-27/U crystal units, whose frequencies are spaced 8.333 kc apart. This spacing is equal to one-twelfth of the channel spacing of 100 kc. Thus, it is the spacing of the 4th crystal oscillator frequencies that determines the channel spacing. For maximum stability the crystal units are mounted in the same crystal oven that houses the crystals of the 3rd crystal oscillator. Crystal selection is controlled by a selector switch on the main control panel. The selection is determined by the choice of the 4th digit of the channel frequency.

1-487. The 4th crystal oscillator circuit employs one half of a duplex triode tube, V410, which is connected as a radio-frequency grounded-plate Pierce oscillator, with the circuit designed in a manner similar to those of the 2nd and 3rd crystal oscillators. The output is taken from the cathode and coupled through C471 to the grid of the other half of tube V410, which is operated as a cathode follower. The output of the cathode follower is inductively coupled into the discriminator circuit through transformer T405. The purpose of the cathode follower is to isolate the loading effects of the discriminator from the 4th crystal oscillator. As a check on the 4th crystal oscillator operation, a test connection for rectified grid current measurement is brought out to test socket X412, pin 7.

1-488. *CLUTCH CONTROL TUBE.* The purpose of the clutch control tube V408, is to shift the r-f head tuning drive from the medium-speed clutch to the low-speed clutch for fine tuning of the exact channel frequency. Whenever the frequency of the variable frequency oscillator, after being fed through the four mixers, comes within range of the discriminator input tuning at transformer T404, the tuning should be shifted into low speed. Thus, by coupling part of the discriminator input through capacitor C464 to clutch-control rectifier V408A and amplifying the rectifier output through V408B, a control voltage is obtained which is used to operate the clutch-control relay. The clutch-control relay is mounted on the main frame and has its coil connected in series with the plate of clutch-control amplifier V408B. The relay is *energized* when the tube grid receives *no* excitation. When a signal is fed to the discriminator from the 4th mixer, a portion of the signal is fed through capacitor C464 and rectified in diode-connected tube V408A. The resulting current flow in resistor R443 causes the ungrounded end to become more negative. This negative voltage is fed to the grid of the clutch-control amplifier through low-pass filter R445 and C463, cutting the tube off. This in turn *de-energizes* the clutch-control relay and places the low-speed clutch in operation. It should be noted here that the over-all tuning range of the radio set is divided into two nearly equal bands and that while the variable frequency oscillator tuning is being driven through the unused band a spurious frequency may pass through to the discriminator and the clutch-control tube. In this case, however, even though the clutch-control relay is de-energized, the tuning remains in high speed because of a lockup circuit.

1-489. *MOTOR-CONTROL AMPLIFIER.* The motor-control amplifier, V411, is a d-c amplifier whose purpose is the control of a reversible tuning motor in accordance with a d-c control voltage from the discriminator. The discriminator output voltage is fed to the grid through the low-pass resistance-capacitance filter R451 and C445A so that the ac, which appears in the discriminator output as the channel frequency is being tuned, does not influence the operation of the amplifier. As its plate load, the motor-control amplifier works into two relay coils in series. These are called the *high* and *low discriminator relays.* The two relays are mounted on the main frame. The high discriminator relay is designed to "pull in" at approximately 9 ma and to "drop out" at approximately 6 ma. The low discriminator relay is designed to pull in at approximately 5 ma and to drop out at approximately 3 ma. The plate current of the motor-control amplifier is so adjusted that, with a plus 5-volt "on-frequency" output from the discriminator, approximately 5.5 ma plate current flows. This is sufficient to energize the low discriminator relay but not the high one. The contacts of the low discriminator relay are so connected that when the relay is energized the tuning motor (mounted in the r-f head) is de-energized. If the frequency of the variable frequency oscillator should tend to drift too low, the discriminator voltage controlling the amplifier becomes more positive, increasing the amplifier plate current and thereby energizing both relays. This has the effect of causing the tuning motor to turn in a direction that diminishes the tuning capacitance and hence raises the frequency. If the variable frequency should tend to drift too high, the plate current of the motor-control amplifier decreases to a value below 3 ma and both relays are de-energized. This has the effect of causing the tuning motor to turn in the opposite direction, so that the frequency of the variable frequency oscillator is decreased. Note that in order to keep the motor-control amplifier bias constant, regardless of plate current flow, a positive 6-volt cathode bias is obtained from the d-c drop across the heater, rather than from the drop across a cathode biasing resistor.

R-F Head

1-490. The r-f head contains all the main channel transmitter and receiver r-f circuits including the reactance tube, the tuning capacitor drive mechanism, and certain tuning control circuits. A block diagram is shown in figure 1-198 and a schematic diagram is shown in figure 1-199. Beginning with the variable frequency oscillator, we shall discuss first the transmitter circuits and then the receiver circuits.

1-491. *VARIABLE FREQUENCY OSCILLATOR AND TRIPLER.* The variable frequency oscillator employs an electron-coupled Hartley circuit. The v-f-o frequency is given by the formula:

$$\text{v-f-o frequency} = \frac{\text{transmitter output freq} - \text{receiver 1st intermediate freq}}{12}$$

The screen grid of the v-f-o tube, V802, is the anode of the oscillatory circuit. The plate circuit is tuned to three times the grid frequency, thereby tripling in this tube. Coupling capacitor C907 from grid inductor 1801 feeds a small amount of r-f energy to the frequency-control circuits in the monitor. A reactance tube is shunted across the oscillator tank for fine frequency control. Oscillator-tripler output is capacitively coupled to the untuned grid of the doubler tube, V803.

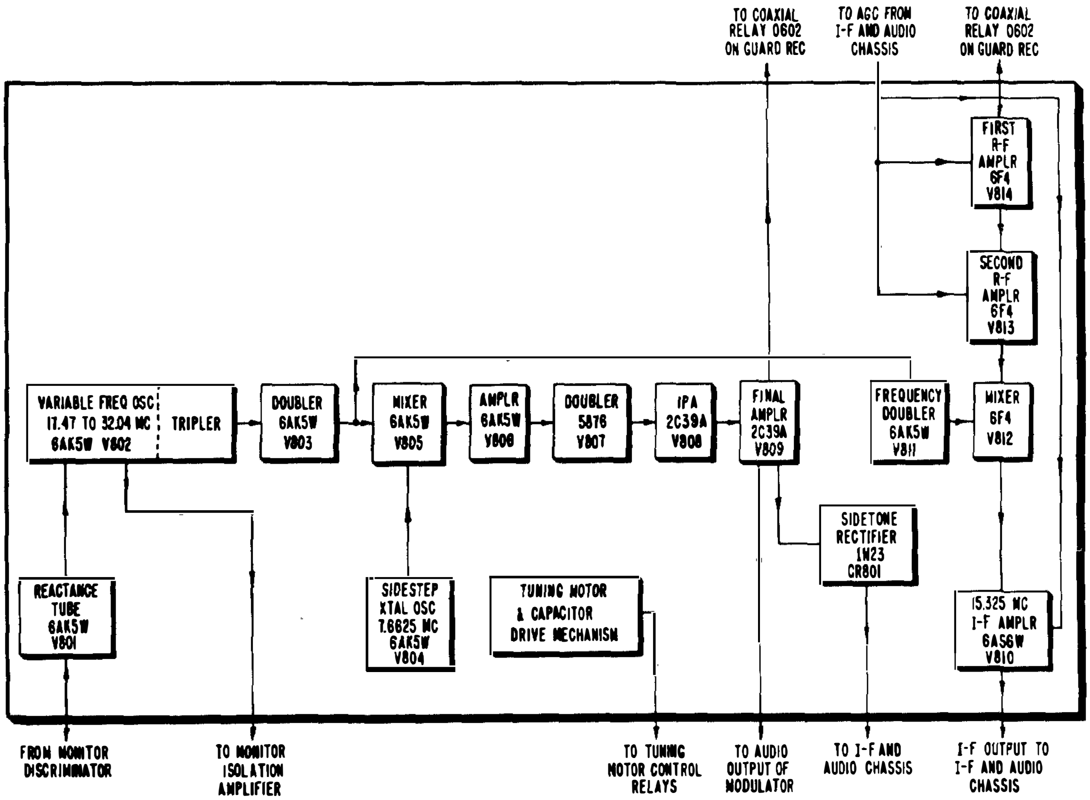


Figure 1-198. R-f head. Block diagram of main channel transmitter and receiver r-f circuits in Radio Set AN/ARC-33

1-492. *REACTANCE TUBE.* The reactance tube, V801, is a device that represents a variable and controllable reactance shunted across the tuned grid circuit of the variable frequency oscillator. Its purpose is to convert d-c control voltage from the monitor into a small frequency variation of the variable frequency oscillator. Radio-frequency voltages from the oscillator are coupled through capacitor C805 to the plate of the reactance tube and directly to a network made up of capacitors C804 and C818, resistors R801 and R804, and the grid-cathode capacitance of the tube, C_g. Capacitor C804 is a blocking capacitor, so that as far as the a-c functioning of the tube is concerned, it need not be considered. Cathode bypass capacitor C803 is sufficiently large for the cathode to be considered at r-f ground potential. Resistor R802, inductor L829, and capacitors C801 and C802 provide a de-coupling network and filter through which the d-c control voltage is applied to the control grid without shunting its r-f input impedance. Resistor R807 is the plate feed component. Resistors R801 and R804 and capacitor C818 in conjunction with the grid-cathode capacitance, C_g, form a phase-shifting network such that the r-f voltage applied to the grid lags the oscillator voltage input to the network by 90 degrees approximately. Since the plate current of reactance tube V801 is in phase with the grid voltage, the plate current flow lags the oscillator r-f voltage applied to the plate by approximately 90 degrees. This appears to the variable frequency oscillator as an inductive reactance since the current in an inductor lags the applied voltage by 90 degrees. By controlling the d-c bias applied to the grid, it is possible to control the amplitude of the current, and thus the magnitude of the effective inductive reactance shunting the oscillator tank. In this way, the frequency of the variable frequency oscillator can be controlled within a narrow range by means of a d-c control voltage. A positive control voltage increases the reactive current in the tube and thereby decreases the effective inductive reactance across the oscillator tank. The effect, therefore, is to cause an increase in the frequency. A negative-going control voltage has the opposite effect. The reactance tube is biased by cathode resistor R805 by an amount that permits normal "on-frequency" oscillator operation when the control voltage from the monitor is approximately plus 5 volts.

1-493. *FIRST DOUBLER.* The first doubler of the r-f head is a conventional grid-leak-biased frequency multiplier. The stage employs tube V803 with an output circuit tuned to twice the input frequency. The plate circuit consists of two inductors in parallel in order to increase their sizes and make it possible to employ coiled inductors rather than a more space-consuming linear line. The tuning is ganged with the other r-f tuned circuits as shown in the schematic diagram. Radio-frequency energy used for heterodyne oscillator injection into the main channel receiver 1st mixer is taken off at a tap of plate tank L802. The variable frequency oscillator-tripler, the reactance tube, and the doubler are continuously supplied with plate and screen voltages since these circuits are employed in both transmit and receive operation. From the latter stage to the antenna, however, plate and screen voltages are removed from the transmitter tubes during receive operation.

1-494. *TRANSMITTER MIXER.* It is in the mixer that a 7.6625-mc signal from the sidestep oscillator is added to the first doubler output to make the mixer output frequency exactly one half of the transmitting antenna frequency. (Remember that 7.6625 mc is one half of the 1st receiver intermediate frequency of 15.325 mc.) A conventional mixer circuit employing tube V805 is used here, with the side-step oscillator voltage being injected at the control grid through capacitor C821 and with the doubler output being fed to the control grid through capacitor C820. The mixer plate circuit is tuned to the sum frequency by means of a doubler-inductor arrangement. One of the inductors is wound with concentric cable, which allows the tuning capacitor rotor and the cold ends of the inductors to be returned to ground. The center conductor is connected to the plate on the hot end and is bypassed to ground on the cold end by capacitor C825. Plate voltage is supplied to tube V805 through resistor R819.

1-495. *SIDESTEP CRYSTAL OSCILLATOR.* The sidestep crystal oscillator employs pentode V804, connected as a triode and operating on a single frequency of 7.6625 mc. The circuit is arranged as a Pierce oscillator with the crystal connected directly from the plate to the grid of the oscillator tube. A type CR-18/U crystal unit is employed. Since this oscillator controls only a small percentage of the final transmitter frequency, its normal operating stability is sufficient without the use of thermostatically controlled temperature for the crystal unit. Note that an 18-$\mu\mu$f capacitor is shunted directly across the crystal unit and that no externally connected plate-to-ground capacitor is used. The 18 $\mu\mu$f represents 60 per cent of the required load capacitance for the type CR-18/U crystal unit. With the grid-to-ground capacitance probably 8 or 10 $\mu\mu$f greater than the 68 $\mu\mu$f of the externally connected grid-ground capacitor C822, it appears that the circuit has been designed to provide a maximum output voltage consistent with the drive-level and load-capacitance requirements of the crystal unit, rather than for maximum frequency stability. The oscillator is fed plate voltage through the voltage-dropping resistor R815. The oscillator output is capacitively coupled from the plate of the oscillator tube to the control grid of mixer tube V805 through C821. Plate voltage is derived from the plus 200-volt transmitter supply, so that the tube operates only when the radio set is turned to the transmit position.

1-496. *AMPLIFIER.* This is a conventional grid-leak-biased r-f amplifier employing pentode V806. The circuit acts as a direct amplifier with the output circuit tuned to the same frequency as the input circuit. The input is capacitance-coupled to the preceding mixer stage through capacitor C828 and the output is capacitance-coupled to doubler V807 through capacitor C832. The amplifier tuned output circuit is a double-inductor tank of the same type that is used in the plate circuit of the transmitter mixer.

1-497. *TRANSMITTER DOUBLER.* The doubler, V807, is a grid-leak-biased stage that employs a cavity-tuned plate circuit tuned to twice the input frequency. A pencil triode tube is used because its small size permits superior u-h-f operating characteristics.

1-498. *INTERMEDIATE POWER AMPLIFIER.* The intermediate power amplifier (IPA) employs a lighthouse tube, V808, in a grounded-grid circuit with cavity tuning of the plate. The cavity is tuned to the doubler output frequency, providing an IPA output frequency equal to the transmitter antenna frequency. Radio-frequency input is coupled magnetically from the output of the preceding doubler tube, V807, through coupling loop L809, which applies rf between ground and the IPA cathode through capacitor C838. Concentric inductor L810 is inserted in the heater circuit to make it at the same r-f potential as the cathode. Capacitor C838 also blocks the 12.6-volt d-c filament potential from shorting to ground through the ground coupling loop L809. One heater con-

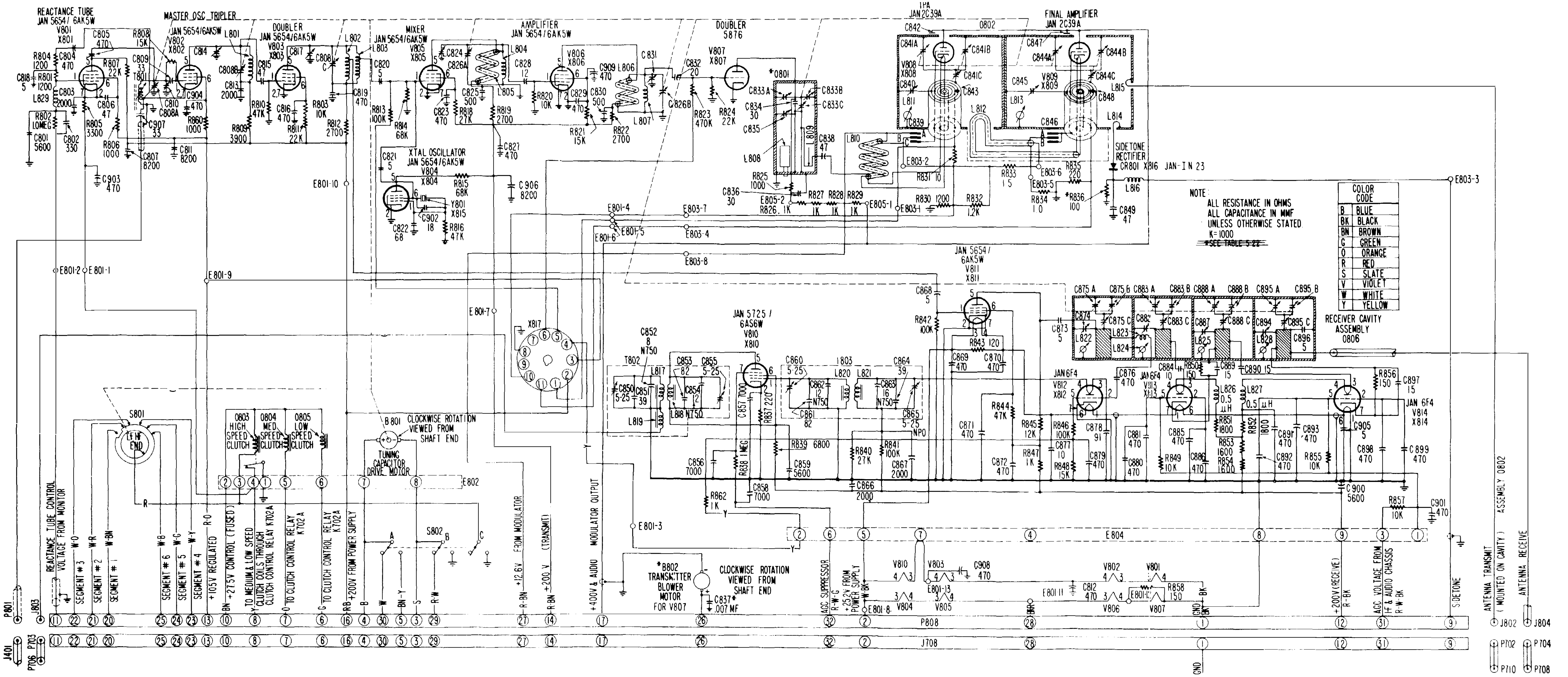


Figure 1-199. R-f head. Schematic diagram of main channel transmitter and receiver r-f circuits in Radio Set AN/ARC-33

nection and the cathode connection are common. The heater circuit is wired in series with the heater of the final amplifier so as to provide the required 12.6-volt total drop across the entire circuit after passing through dropping resistor R833. A voltage divider composed of resistors R830 and R832 is connected between one side of the filament and ground. The grid return is connected to the tap to provide the proper negative bias voltage. These resistors are high in value so that they do not influence the heater voltage and current. IPA output is coupled to the final power amplifier input through an inductive coupling loop, L812, in the cavity. Since the final power amplifier is a grounded-grid r-f amplifier, it cannot be plate modulated 100 per cent unless the output of the exciting stage also is modulated. Therefore, to be able to modulate fully the final amplifier output, the plates of doubler V807 and IPA V808 are modulated by the same audio voltage as the power amplifier.

1-499. *FINAL POWER AMPLIFIER.* The final r-f power amplifier, V809, is similar in design and operation to the intermediate power amplifier. Grid bias for the final amplifier is derived from resistor R835. Resistor R834 supplies a cathode bias. Radio-frequency output to the antenna is coupled inductively to the cavity magnetic field by pickup loop L815 and is fed to the antenna receive-transmit changeover relay through a coaxial cable. The coaxial connector is mounted directly on the cavity assembly, O802. To provide a sidetone voltage that is indicative of the transmission quality, the sidetone voltage is obtained from the final r-f output cavity through inductive coupling L814. This r-f voltage is rectified through the sidetone crystal rectifier, CR801, and applied to the i-f and audio chassis to be fed through the receiver audio system to the operator's headset.

1-500. *MAIN CHANNEL RECEIVER R-F CIRCUITS.* The main channel receiver includes two r-f amplifier stages, a mixer, a frequency doubler, and a 15.325-mc i-f amplifier. Both r-f amplifiers and the doubler are cavity-tuned. The doubler is included to multiply the output frequency of the first r-f-head doubler. By employing a heterodyne injection frequency controlled by the same variable frequency oscillator that controls the transmitter frequency, perfect tracking between the transmitter and receiver tuning is possible. The i-f output is fed by coaxial cable to the second mixer and the 2.8-mc intermediate frequency amplifier on the i-f and audio chassis.

1-501. *FIRST R-F AMPLIFIER.* The first r-f amplifier, V814, is a grounded-grid cavity-tuned amplifier with a shunt-fed plate. The cavity is similar to the cavities used in the transmitter section and operates in exactly the same way except that the amplifier tube is located outside of the cavity. Electrical connections are made through holes in the cavity walls. Tube V814 is tapped down on the cavity center column to prevent loading of the tuned circuit by the tube. The antenna input is also tapped down to match the characteristic impedance of the coaxial line. The grounded-grid circuit is effective in preventing coupling between the input and output circuits, thus preventing oscillations due to feedback through the tube capacitance. Also, the grounded-grid design with the cathode-injection input provides a more constant input impedance over the frequency range. Two plate and two grid connections are provided on the JAN-6F4 tubes to make it possible to use shorter leads and to obtain better bypassing balance, as is illustrated by the V814 grid connections to ground through capacitors C892, C893, C898, and C899. Examination of the tuned-cavity circuits will show that only the first input cavity employs a capacitance connection, represented by C896. Since this cavity is not loaded by the plate capacitance of a vacuum tube, capacitor C896 is added in order for all three cavities to have similar tuning characteristics. Automatic-gain-control voltage, which is obtained from the i-f and audio chassis, is applied to the 1st and 2nd r-f stages and to the 15.325-mc i-f stage.

1-502. *SECOND R-F AMPLIFIER.* The second r-f amplifier, V813, is similar to and operates in the same manner as the first r-f amplifier. Its input is capacitance-coupled from the preceding stage through capacitors C889 and C890. Its output is coupled to the mixer through capacitors C876 and C884 and inductor L823.

1-503. *RECEIVER MIXER.* The mixer, V812, employs the same type of tube as do the r-f amplifiers; however, in this case it is operated as a normal triode mixer with the r-f signal applied to the grid. Heterodyne oscillator injection voltage is coupled from the doubler cavity through capacitor C876 and coupling inductor L823 to the grid, in order to produce the desired 15.325-mc intermediate frequency. The plate circuit is bypassed at the plate connection by a small capacitor, C878, to ensure that no input signal r-f voltage appears in the plate circuit. As regards the much lower intermediate frequency, this capacitor merely forms part of the i-f transformer tuning capacitance. No automatic gain control voltage is applied to the mixer.

1-504. *RECEIVER DOUBLER.* In order to provide the proper heterodyne injection frequency at the receiver first mixer without frequency multiplication in the mixer, itself, it is necessary to include a doubler, V811, between the r-f head first doubler and the receiver first mixer. The doubler output is cavity-tuned to the received antenna signal frequency minus 15.325 mc, and is ganged to the main channel receiver r-f tuning.

1-505. *I-F AMPLIFIER.* The 15.325-mc i-f amplifier, V810, is conventional in all respects with the possible exception of the suppressor injection of the automatic gain control voltage. By applying automatic gain control voltage to the suppressor grid, it is possible to keep the impedance of the signal grid-return circuit low, thus preventing paralysis of the amplifier on strong noise pulses. The i-f output transformer, T802, is designed to work into the low-impedance (52-ohm) coaxial cable through which the i-f output is carried to the second receiver mixer, located on the i-f and audio chassis.

1-506. *TEST POINTS.* To facilitate testing of the r-f head, an 11-pin test socket is provided. Connections to the socket pins provide for measurement of plate supply voltages, mixer injection voltages, and grid drive.

1-507. *TUNING.* All of the tuned r-f circuits of the transmitter and receiver that must be varied when changing channels, are ganged together and driven by a reversible tuning motor, B801. The drive is through a system of gears and any one of three clutches 0803, 0804, and 0805. A detailed description of the mechanical and switching design of the tuning system is beyond our present assignment.

The Collins Synthesizing System

1-508. An interesting approach to the problem of controlling many r-f channels with the use of only a few crystals is afforded by the frequency-control system employed in a number of the multichannel radio sets developed by the Collins Radio Company. Fundamentally this system is a crystal-controlled multi-conversion superheterodyne circuit when employed in radio receivers, and is equivalent to the same circuit operated in reverse when employed in radio transmitters. The receiver system is quite similar to the Bendix synthesizing system, except that the signal from the variable frequency oscillator is replaced by the antenna signal to which the r-f circuits are tuned and the discriminator is replaced by a final i-f stage. When operated in reverse for transmitter use, the system resembles somewhat the Plessey synthesizing system, except that the frequency dividers and harmonic selectors of the Plessey system are replaced by crystal oscillators with banks of crystal units. We shall not discuss the Collins system further from a generalized point of view, but shall advance at once to the discussion of its application in two aircraft receiver circuits (Radio Receiver R-252A/ARN-14 and Radio Receiver R-278/GR) and in the aircraft transceiver, Radio Set AN/ARC-27. Since Radio Set AN/ARC-27 performs essentially the same function as does Radio Set AN/ARC-33, described in the foregoing paragraphs, a study of the two circuits provides an interesting comparison of the two systems of frequency control and how they can be applied to achieve the same end.

RADIO RECEIVER R-252A/ARN-14

1-509. Radio Receiver R-252A/ARN-14, a product of the Collins Radio Company, is a component of Radio Receiving Set AN/ARN-14. It is an airborne navigation receiver that provides reception on any one of 280 channels spaced 100 kc apart between 108.0 and 135.9 mc. A crystal-controlled double-conversion superheterodyne circuit is used to "de-synthesize" the incoming signal to a fixed intermediate frequency of 3.15 mc. The 280 channels are obtained from a total of 24 crystal units. Actually, this number of channels can be obtained from a Collins synthesizing system employing a fewer number of crystal units, if a triple- rather than a double-conversion superheterodyne circuit is used. That is, if an additional oscillator, mixer, and i-f amplifier stage is inserted, the required number of crystal units becomes less. In this equipment, however, maximum economy in space, weight, and the over-all number of component parts clearly is achieved by increasing slightly the number of crystal units rather than by the addition of a number of extra vacuum-tube circuits. The functional operation of the frequency-control system is indicated in the block diagram of figure 1-201. Figure 1-200 shows the schematic diagram of the radio-frequency control circuits. With the exception of minor editing changes and technical insertions the circuit descriptions to follow are primarily extracts from USAF Technical Order No. 12R5-2ARN14-12.

R-F Control Circuits

1-510. One stage of tuned r-f amplification is used in Radio Receiver R-252A/ARN-14. The grid and plate tank circuits of the r-f amplifier, V101, pass a 2-mc band of frequencies and cover the entire frequency range 108.0 to 135.9 mc in 14 increments. The output of the r-f amplifier is coupled to the grid of the 1st mixer V102. The 1st mixer receives its injection frequency from a three-stage

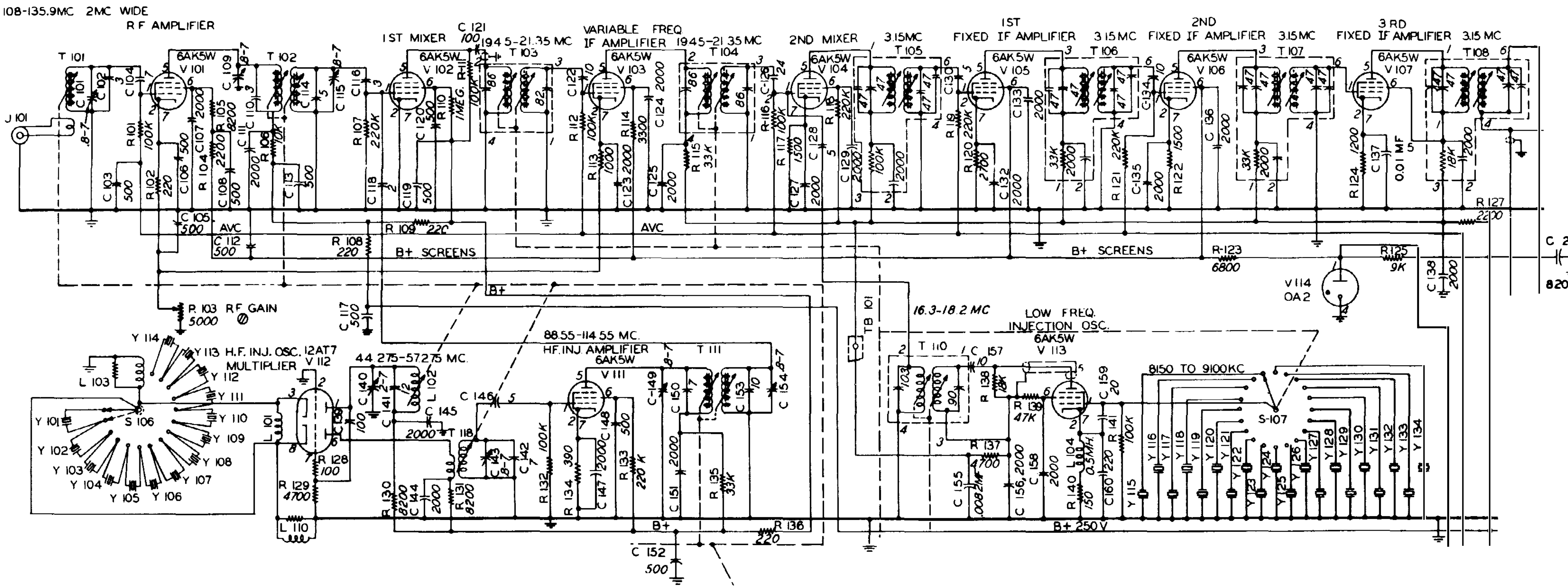


Figure 1-200. Block diagram of the Collins crystal-controlled, multichannel, frequency-control system as employed in the double-conversion superheterodyne circuit of Radio Receiver R-252A/ARN-14

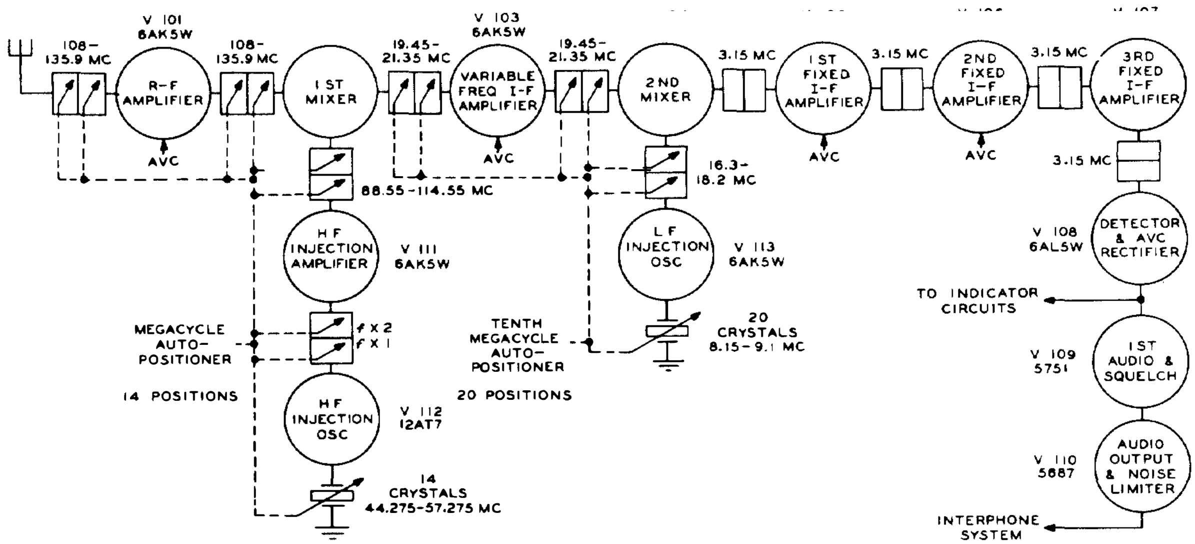


Figure 1-201. Schematic diagram of radio-frequency-control circuits of Radio Receiver R-252A/ARN-14. Only the 1st amplifier of the three-stage fixed (3.15 mc) i-f amplifier circuit is shown. The automatic switching system is not shown

exciter consisting of a crystal-controlled Butler oscillator-multiplier, V112, and an amplifier, V111. The oscillator employs a bank of 14 type CR-23/U crystal units whose frequencies cover the range 44.275 to 57.275 mc in one-megacycle steps. The plate tank of the grounded-grid tube is tuned to the desired crystal frequency. Output from the oscillator is inductively coupled from the plate circuit of the cathode follower to the input tank of amplifier V111. Both the input and output circuits of the amplifier are tuned to twice the oscillator frequency. Thus, the injection frequencies fed to the 1st mixer cover the range of 88.55 to 114.55 mc in 14 two-megacycle steps. These injection frequencies, when mixed with the r-f signals in the range of 108.0 to 135.9 mc, produce an intermediate frequency in the range of 19.45 to 21.35 mc. Note that the function of the 1st injection oscillator and multiplier is to generate 14 very high frequencies spaced 2 mc apart so that by selecting the proper injection frequency, the 280 possible antenna frequencies within a 28-mc range can be reduced to 20 possible intermediate frequencies within a 2-mc range. Each of the 20 i-f channels handles 14 of the r-f channels. The output of the 1st mixer is thus fed to the variable i-f amplifier, and thence to the 2nd mixer. The injection frequency for the 2nd mixer originates in an electron-coupled oscillator-doubler circuit. The oscillator is of the grounded-plate Pierce type and employs a bank of 20 type CR-18/U crystal units spaced 50 kc apart in the 8.15-to-9.10-mc range. The screen grid of tube V113 serves as the oscillator anode. The plate circuit is tuned to twice the oscillator fundamental, so that the actual injection frequencies for the 2nd mixer are spaced 100 kc apart and cover the range of 16.3 to 18.2 mc. Each of the 2nd injector frequencies can be matched with one of the 20 i-f channels to produce a difference frequency of 3.15 mc. This 3.15-mc frequency is thus used as the fixed intermediate-frequency channel. It passes through three amplifier stages of similar design except that the 3rd 3.15-mc i-f amplifier does not operate with a-v-c grid bias before being fed to the detector.

1-511. The tuning of the r-f and variable i-f circuits, as well as the selecting of the high- and low-frequency crystals, is performed by two Collins "Autopositioners." One Autopositioner controls the tuning in 14 2-mc steps and the other controls the fine tuning in 20 100-kc steps. The tuning elements of the grid and plate tank circuits of the r-f amplifiers and the frequency multiplier, and the high-frequency crystal selector switches are ganged and operated by the megacycle Autopositioner. The 0.1-megacycle Autopositioner tunes the grid and plate circuits of the variable i-f amplifier and the plate circuit of the low-frequency oscillator, and selects the low-frequency crystals. Because the crystal selector switches and tuning elements are ganged, the injection frequencies that result in the fixed intermediate frequency of 3.15 mc are always automatically selected. Table (1) shows the tuning of the double-conversion system for the first 22 channels. By extending the table in the same manner as shown, complete information for the entire range of received frequencies can be obtained.

Channel Selector Position		*Frequency of Received Signal*	*Frequency Band Passed by R-F Amplifier*	*Frequency of*			
Mc	*.1 Mc*			*1st Mixer Inj. Signal*	*Variable I-F Amplifier*	*2nd Mixer Inj. Signal*	*Fixed I-F Amplifier*
108	0.0	108.0 mc	108.0 to 110.0	85.55 mc	19.45 mc	16.3 mc	3.15 mc
	0.1	108.1	108.0 to 110.0	85.55	19.55	16.4	3.15
	0.2	108.2	108.0 to 110.0	85.55	19.65	16.5	3.15
	0.3	108.3	108.0 to 110.0	88.55	19.75	16.6	3.15
	0.4	108.4	108.0 to 110.0	88.55	19.85	16.7	3.15
	0.5	108.5	108.0 to 110.0	88.55	19.95	16.8	3.15
	0.6	108.6	108.0 to 110.0	88.55	20.05	16.9	3.15
	0.7	108.7	108.0 to 110.0	88.55	20.15	17.0	3.15
	0.8	108.8	108.0 to 110.0	88.55	20.25	17.1	3.15
	0.9	108.9	108.0 to 110.0	88.55	20.35	17.2	3.15
	1.0	109.0	108.0 to 110.0	88.55	20.45	17.3	3.15
	1.1	109.1	108.0 to 110.0	88.55	20.55	17.4	3.15
	1.2	109.2	108.0 to 110.0	88.55	20.65	17.5	3.15
	1.3	109.3	108.0 to 110.0	88.55	20.75	17.6	3.15
	1.4	109.4	108.0 to 110.0	88.55	20.85	17.7	3.15
	1.5	109.5	108.0 to 110.0	88.55	20.95	17.8	3.15
	1.6	109.6	108.0 to 110.0	88.55	21.05	17.9	3.15
	1.7	109.7	108.0 to 110.0	88.55	21.15	18.0	3.15
	1.8	109.8	108.0 to 110.0	88.55	21.25	18.1	3.15
	1.9	109.9	108.0 to 110.0	88.55	21.35	18.2	3.15
110	0.0	110.0	110.0 to 112.0	90.55	19.45	16.3	3.15
	0.1	110.1	110.0 to 112.0	90.55	19.55	16.4	3.15

Table 1-511(1). Tuning position for first 22 channels of Radio Receiver R-252A/ARN-14. Remaining channels are obtained in similar manner.

RADIO RECEIVER R-278/GR

1-512. Radio Receiver R-278/GR provides a second example of the Collins method of controlling many frequencies with a few crystals. In principle, the frequency-control system is the same type as that described above for Radio Receiver R-252A/ARN-14, except that in the R-278/GR receiver a triple-conversion superheterodyne circuit is employed instead of a double-conversion circuit. Functionally, the receiver is the same as the AN/ARC-33 receiver circuit in that it provides crystal-controlled reception of 1750 channels spaced 100 kc apart between 225 and 399.9 mc. This task is performed with the use of 38 crystal

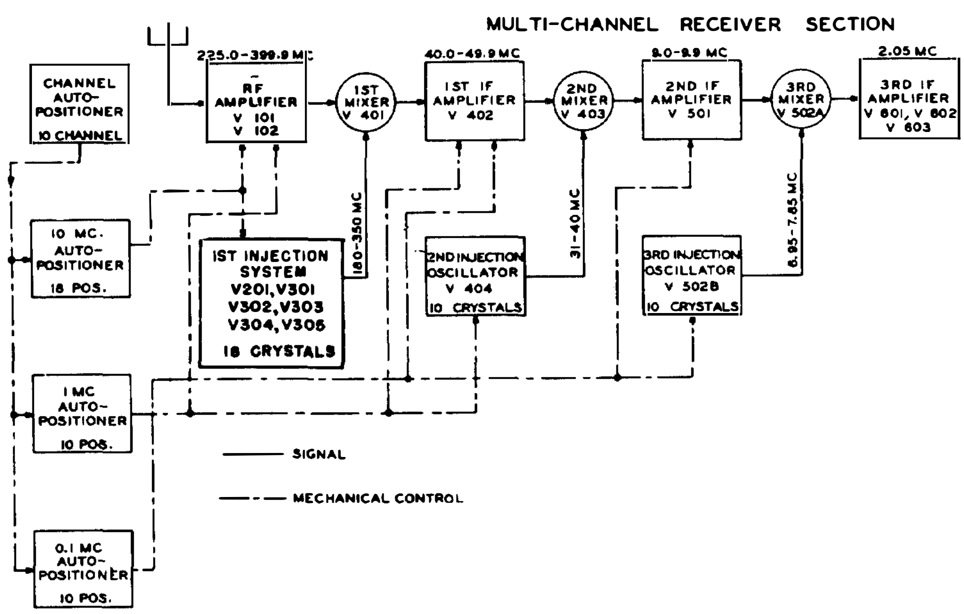


Figure 1-202. Block diagram and frequency chart of Collins frequency-control system as employed in Radio Receiver R-278/GR

10-MC Dial Reading	*R-F Band MC*	*Inj Freq MC*	*Crystal Freq MC*	*Mult Factor*
22	220.0 - 229.9	180.0	30.0000	6
23	230.0 - 239.9	190.0	31,6667	6
24	240.0 - 249.9	200.0	33.3333	6
25	250.0 - 259.9	210.0	35.0000	6
26	260.0 - 269.9	220.0	36.6667	6
27	270.0 - 279.9	230.0	38.3333	6
28	280.0 - 289.9	240.0	26.6667	9
29	290.0 - 299.9	250.0	27.7777	9
30	300.0 - 309.9	260.0	28.8888	9
31	310.0 - 319.9	270.0	30.0000	9
32	320.0 - 329.9	280.0	31.1111	9
33	330.0 - 339.9	290.0	32.2222	9
34	340.0 - 349.9	300.0	33.3333	9
35	350.0 - 359.9	310.0	34.4444	9
36	360.0 - 369.9	320.0	35.5555	9
37	370.0 - 379.9	330.0	36.6667	9
38	380.0 - 389.9	340.0	37.7778	9
39	390.0 - 399.9	350.0	38.8889	9

1-MC Dial Reading	*1st I-F Band MC*	*Crystal and Inj Freq MC*
0	40.0 - 40.9	31.0
1	41.0 - 41.9	32.0
2	42.0 - 42.9	33.0
3	43.0 - 43.9	34.0
4	44.0 - 44.9	35.0
5	45.0 - 45.9	36.0
6	46.0 - 46.9	37.0
7	47.0 - 47.9	38.0
8	48.0 - 48.9	39.0
9	49.0 - 49.9	40.0

0.1-MC Dial Reading	*2nd I-F Freq MC*	*Crystal and Inj Freq MC*
.0	9.0	6.95
.1	9.1	7.05
.2	9.2	7.15
.3	9.3	7.25
.4	9.4	7.35
.5	9.5	7.45
.6	9.6	7.55
.7	9.7	7.65
.8	9.8	7.75
.9	9.9	7.85

	Frequency Range	
R-f amplifier	220.0-399.9 mc*	Tuned in 180 1-mc steps
1st injection	180.0-350.0 mc	Tuned in 18 10-mc steps
1st i-f amplifier	40.0-49.9 mc	Tuned in 100 0.1-mc steps
2nd injection	31.0-40.0 mc	Tuned in 10 1-mc steps
2nd i-f amplifier	9.0-9.9 mc	Tuned in 10 0.1-mc steps
3rd injection	6.95-7.85 mc	Tuned in 10 0.1-mc steps
3rd i-f amplifier	2.05 mc	Fixed tuned

*Note that although the frequency range of the receiver is specified as being from 225.0 to 399.9 mc, the frequency system employed is inherently capable of providing 50 additional frequencies in the range 220.0 to 224.9 mc.

Figure 1-202 (Continued). Frequency chart of Collins frequency-control system

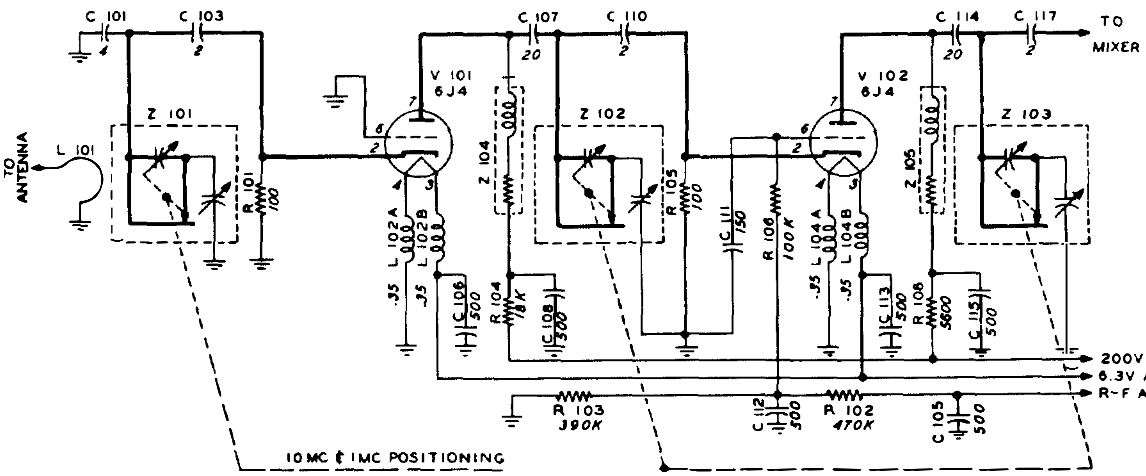


Figure 1-203. Radio-frequency amplifier unit of Radio Receiver R-278/GR. Schematic diagram

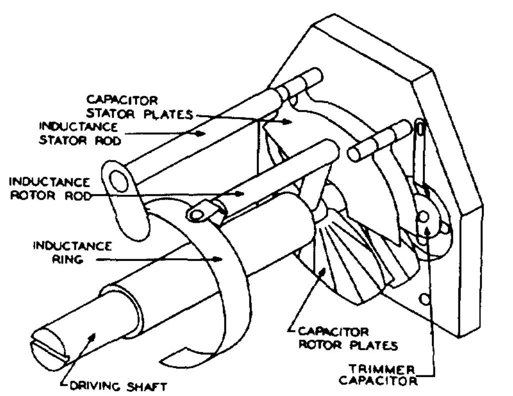


Figure 1-204. Construction of r-f tuner employed in Radio Receiver R-278/GR

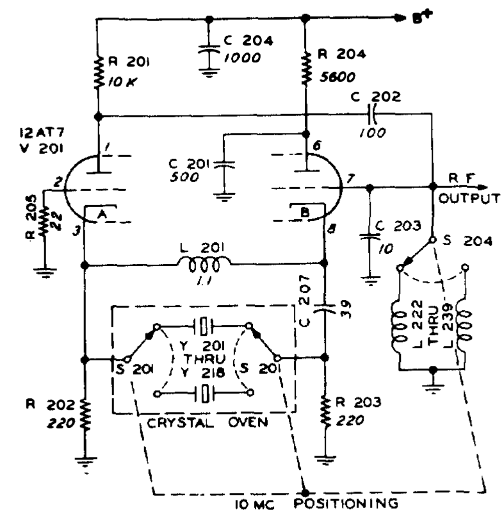


Figure 1-205. Oscillator unit of 1st injection system in Radio Receiver R-278/GR. Schematic diagram

units. For the reader who has followed the discussion of the frequency-control system in Radio Receiver R-252A/ARN-14, the functional operation of the system outlined for Radio Receiver R-278/GR in the frequency diagram of figure 1-202 should be largely self-explanatory.

1-513. Briefly, the r-f amplifier consists of two stages of grounded-grid vacuum-tube amplification, as shown in the schematic diagram of figure 1-203. The r-f amplifier is tuned in 180 one-mc steps covering the range 225 to 399.9 mc. A special type of r-f tuner is used, and is illustrated in figure 1-204. This tuner consists of a variable capacitor and a variable inductor which are rotated simultaneously so that the resonant frequency changes linearly 180 mc in 180 degrees of rotation, or 1 mc per degree. The variable capacitor consists of two stator plates and three rotor plates, the front and rear of which are radially slotted for the purpose of alignment adjustments. The inductive loop consists of the inductance stator rod, a ring segment, and the inductance rotor rod. The three r-f tuners are geared together and driven through a mechanical differential that combines the rotary motion of an 18-position 10-mc tuning shaft with that of a 10-position 1-mc tuning shaft to permit a total of 180 1-mc steps.

1-514. The 1st injection system consists of the main oscillator unit shown in figure 1-205 followed by the five-stage multiplier-amplifier unit shown in figure 1-206. The oscillator is a Butler

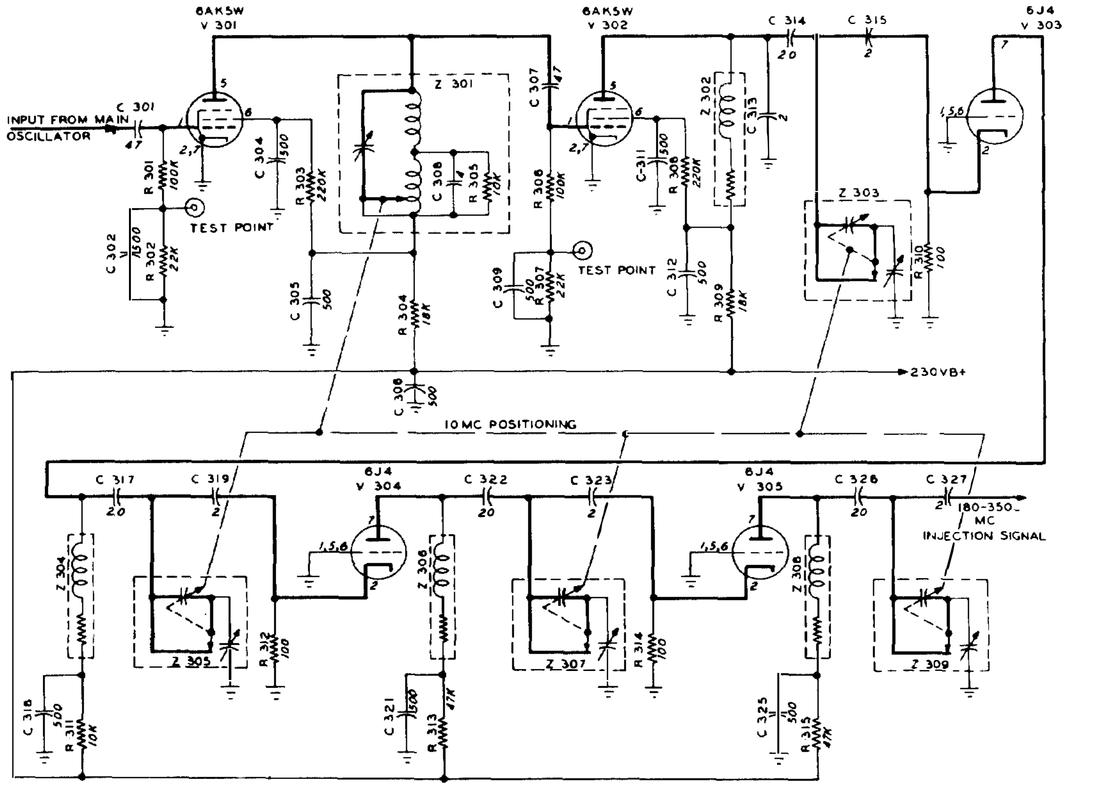


Figure 1-206. Multiplier-amplifier unit of 1st injection system in Radio Receiver R-278/GR. Schematic diagram

circuit employing a temperature-controlled bank of 18 crystal units of the CR-32/U type, and a bank of 18 tuning coils in the plate output circuit of the grounded-grid tube. The first tube, V301, of the multiplier-amplifier unit serves as a doubler for the six lowest *injection* frequencies (not the six lowest oscillator frequencies) and as a tripler for the 12 highest injection frequencies. The second tube, V302, is tuned to triple all its input frequencies. The next three stages are grounded-grid, shunt-fed amplifiers. The r-f tuners shown in the last four stages of the multiplier-amplifier unit are similar to the r-f tuners described in the foregoing paragraph, but differ in that the frequency range is different. The 1st injection system is tuned in eighteen 10-mc steps by the 10-mc tuning shaft, which also operates the main oscillator crystal selector switch.

1-515. The 1st i-f amplifier unit, see figure 1-207, consists of the 1st mixer, the 1st i-f amplifier, the 2nd oscillator, and the 2nd mixer. The 1st mixer stage reduces the 1750 possible antenna frequencies to 100 possible intermediate frequencies spaced 0.1 mc apart in the 40.0-to-49.9-mc range. The i-f interstage transformers are therefore tunable in 0.1-mc increments over a 10-mc i-f band. Permeability tuning is employed, with the tuning controlled by a cam-driven tuning rack. The second oscillator is another Butler circuit, this one employing a bank of 10 type CR-23/U crystal units operating at frequencies 1 mc apart from 31 to 40 mc. The 2nd mixer serves to reduce the 100 first i-f channels to 10 second i-f channels spaced 0.1 mc apart in the 9.0-to-9.9-mc range.

1-516. The 2nd i-f amplifier unit, see figure 1-208, consists of the 2nd i-f amplifier, the 3rd injection oscillator, and the 3rd mixer. The 2nd i-f amplifier signal is conducted over a type RG-58/U coaxial line from the plate of the 2nd mixer to the primary of the input transformer, T501, of the 2nd i-f amplifier. The 2nd i-f transformers are permeability-tuned in ten 0.1-mc increments over the range 9.0 to 9.9 mc by a cam-driven tuning rack. The 3rd oscillator, which uses one half of a twin triode tube (12AU7), is a grounded-plate Pierce circuit that uses a bank of 10 type CR-18/U crystal units. The crystal frequencies are spaced 0.1 mc apart in the range 6.95 to 7.85 mc. The 3rd

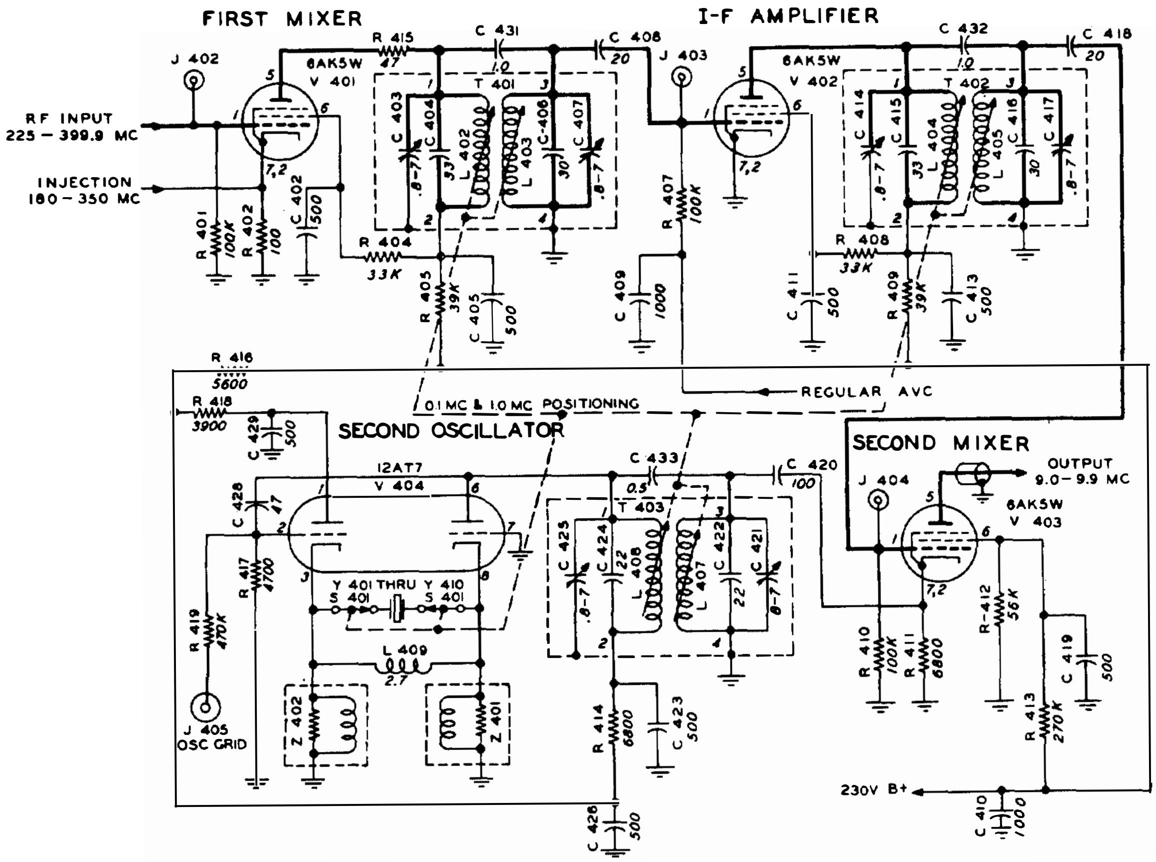


Figure 1-207. First i-f amplifier unit in Radio Receiver R-278/GR. Schematic diagram

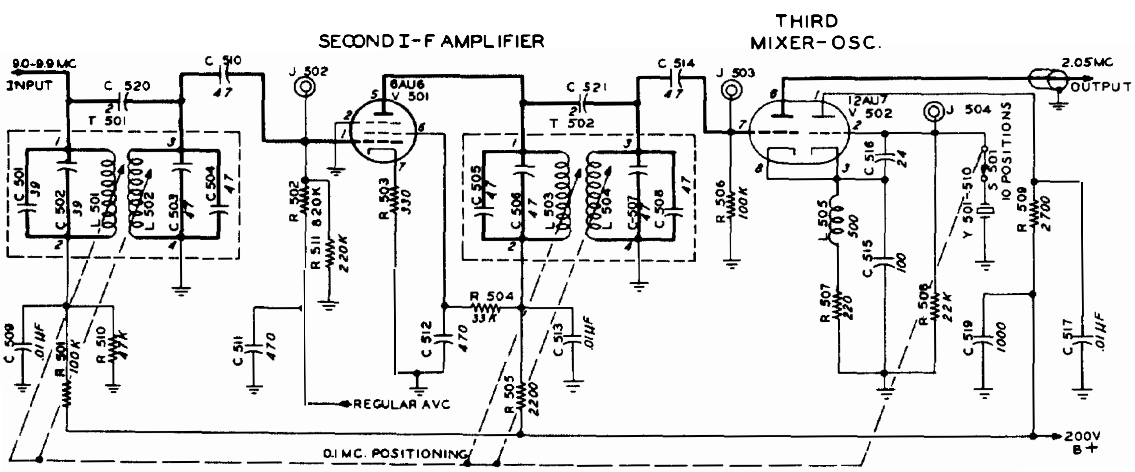


Figure 1-208. Second i-f amplifier unit in Radio Receiver R-278/GR. Schematic diagram

oscillator signal is fed to the cathode of the 3rd mixer, where it is heterodyned with the incoming 2nd i-f signal to form the 3rd and fixed intermediate frequency of 2.05 mc. The 2.05-mc output of the 3rd mixer is carried over a type RG-58/U coaxial line to the input transformer, T601, of the 3rd i-f amplifier, which consists of three fixed-tuned amplifier stages. See figure 1-209.

RADIO SET AN/ARC-27

1-517. Perhaps the most interesting example of the Collins system of frequency synthesis is provided by Radio Set AN/ARC-27. This is a v-h-f and u-h-f airborne transceiver set developed by the Collins Radio Company; it is the functional equivalent of Radio Set AN/ARC-33, described earlier as an example of the Bendix synthesizing method. For transmitter operation of Radio Set AN/ARC-27, four crystal oscillators employing a total of 22 crystal units are interrelated by means of harmonic generators and frequency mixers to cover the range between 225 and 399.9 mc in 0.1-mc increments. The operator thus has a choice of any one of 1750 crystal-controlled channels of 100-kc width. If transmission were the only function of this radio set, at least one of the four crystal oscillators (the 3.45-mc oscillator) could be eliminated and no doubt much of the remaining network could be simplified. As it is, however, the ARC-27 is also designed to receive the same frequencies that it transmits. The same operation that tunes the receiver to a particular channel automatically tunes the transmitter to the same channel. This is achieved by using many of the same tuned circuits that pass the received signal in one direction during receiver operation, to pass the transmitted signal in the opposite direction during transmitter operation. For example, a transformer primary in the plate circuit of a receiver vacuum tube may become a transformer secondary in the grid circuit of a transmitter tube when the set is switched to transmit operation. The transmitter frequency synthesizer is basically the receiver superheterodyne system operated in reverse. It is the set of modifications imposed upon the frequency synthesizing network by this

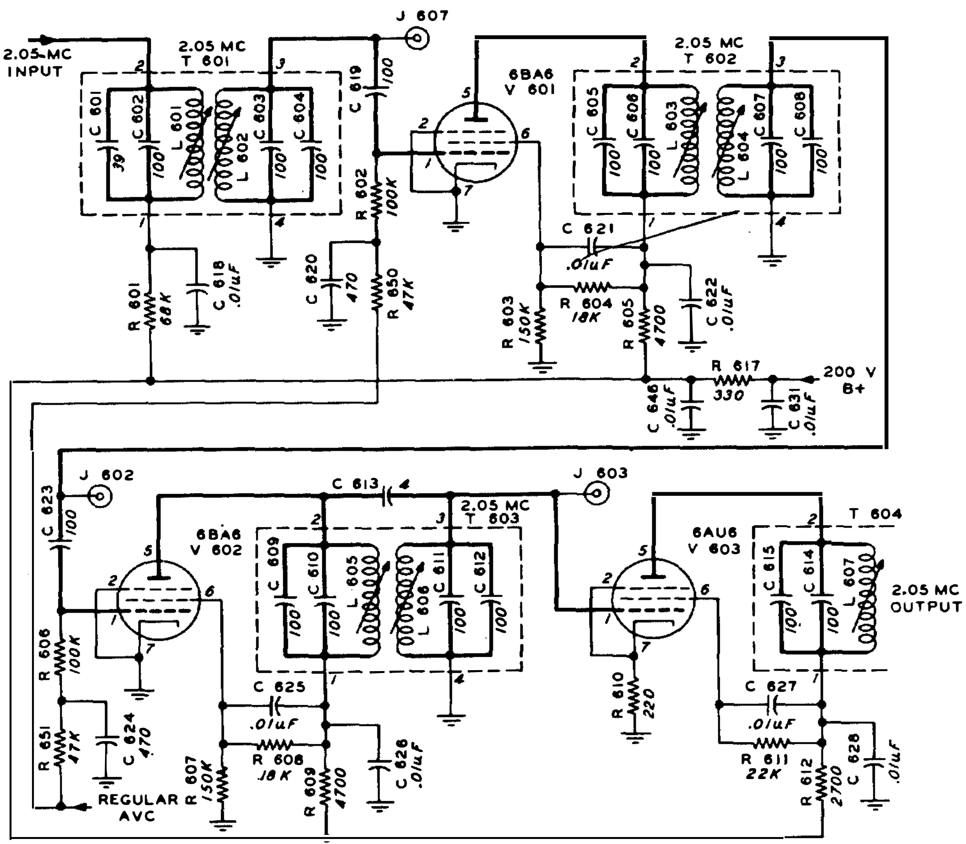


Figure 1-209. Third i-f amplifier in Radio Receiver R-278/GR. Schematic diagram

bi-directional feature that makes the ARC-27 circuits of special interest.

1-518. The network system incorporated in the design of Receiver-Transmitter RT-178/ARC-27 (the principal unit of Radio Set AN/ARC-27) suggests that, in approaching the problem of frequency control, the designers probably placed first emphasis upon the superheterodyne circuit of the receiver; then, to this basic design introduced the switching arrangements necessary for the same circuit to operate in the opposite direction. Where the receiver reduces all incoming signals to a fixed crystal-controlled intermediate frequency of 3.45 mc, the transmitter starts with a fixed crystal-controlled frequency of 3.45 mc and, by reversing the direction of each of the receiver heterodyne processes, is able to generate for transmission the same channel frequency to which the receiver is tuned. The reversed superheterodyne network thus forms a principal feature of the Collins synthesizing system, a feature uniquely applicable in the design of two-way multichannel radio sets.

1-519. The ARC-27 synthesizing circuit is more readily explained when we examine the network first from the receiver point of view. A block diagram of the receiver circuit is shown in figure 1-210. This circuit operates in exactly the same manner as the triple-conversion superheterodyne circuit in Radio Receiver R-278/GR discussed previously. The only difference is in the choice of the values of the intermediate frequencies and in the fact that the ARC-27 set employs a fixed-frequency 10-mc crystal oscillator and a spectrum system that permits a selection of any one of eighteen 10-mc harmonics for the 1st injection frequency, whereas the R-278 receiver employs an 18-crystal oscillator and five-stage multiplier-amplifier system for the same purpose, which is to reduce the 1750 antenna channels to 100 first i-f channels. The 2nd and 3rd oscillators of the ARC-27 receiver circuit are of the same design as the 2nd and 3rd oscillators, respectively, of the R-278 receiver except for the actual values of the crystal frequencies. The 1-mc and 0.1-mc spacing of the 2nd and 3rd injection frequencies are, nevertheless, common to both receivers.

1-520. If we now examine the block diagram of the ARC-27 transmitter synthesizer shown in figure 1-211, we see that it essentially is a reversal of what might be described as the "frequency de-synthesizer" network of the ARC-27 receiver. Let f_a, f_b, f_c, and f_d equal the antenna frequency and the 1st, 2nd, and 3rd intermediate frequencies, respectively; and let f(10), f(1), and f(0.1) represent the 10-mc, 1-mc, and 0.1-mc injection frequencies respectively, as indicated in the block

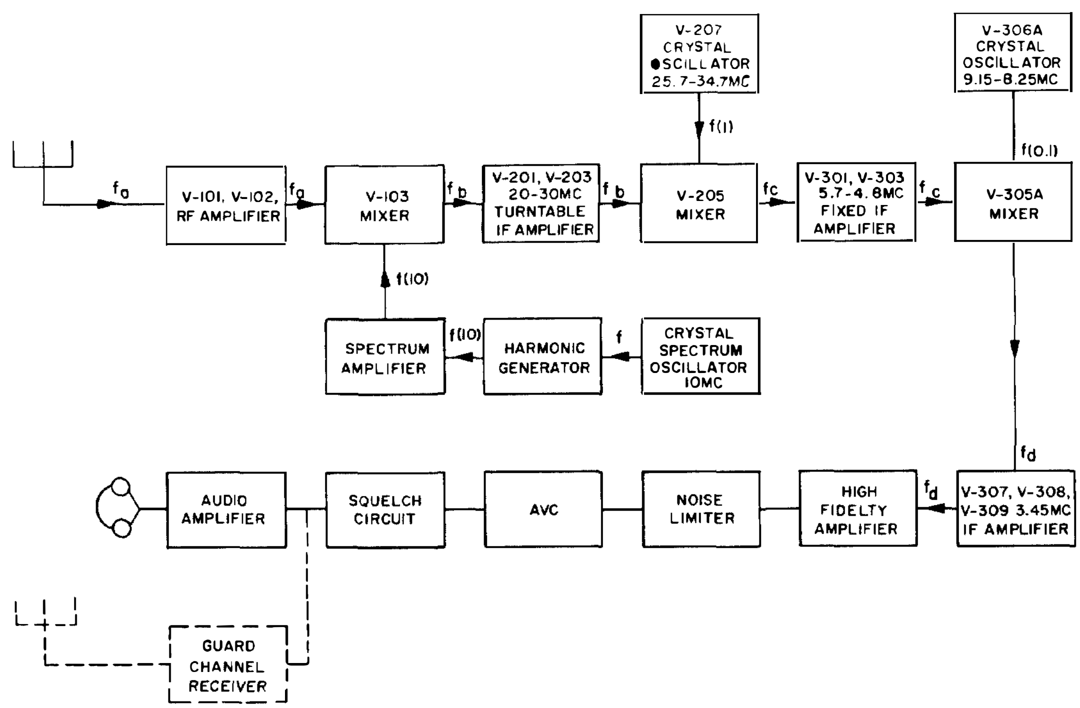


Figure 1-210. Block diagram of main receiver network of Receiver-Transmitter RT-178/ARC-27

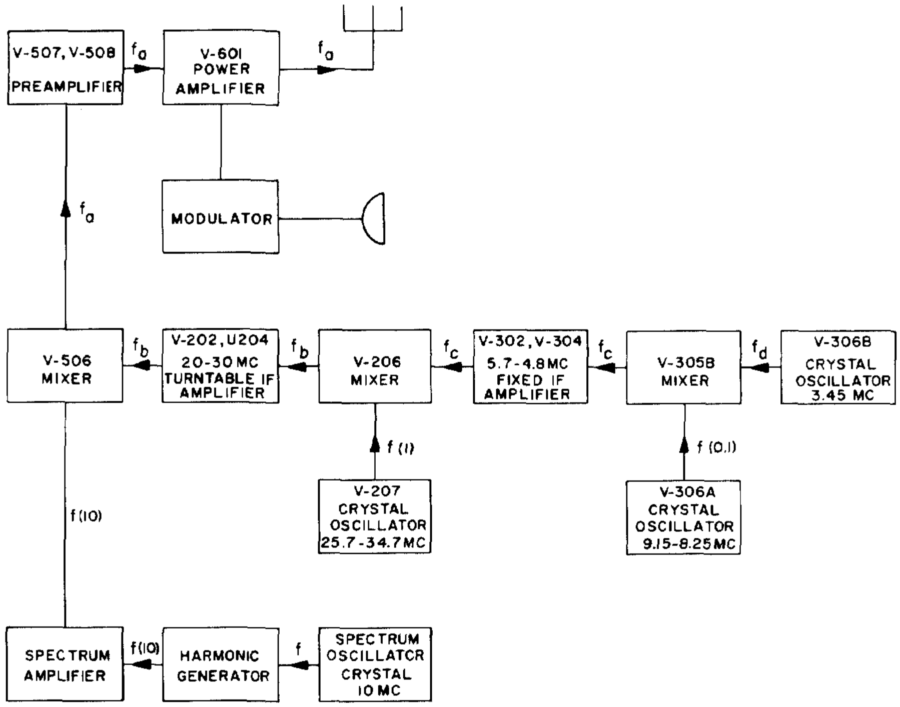


Figure 1-211. Block diagram of transmitter network of Receiver-Transmitter RT-178/ARC-27

diagrams. For both receiver and transmitter operation the frequency of the antenna signal can be expressed by the equation

$$f_a = f(10) + f_b \qquad 1\text{-}520\ (1)$$

where

$$f_b = f(1) - f_c \qquad 1\text{-}520\ (2)$$

and

$$f_c = f(0.1) - f_d \qquad 1\text{-}520\ (3)$$

so by substitution

$$f_a = f(10) + f(1) - f(0.1) + f_d \qquad 1\text{-}520\ (4)$$

Ordinarily for frequency synthesis in a decade system we would suppose that the final frequency would be given by the formula

$$f_a = f(10) \pm f(1) \pm f(0.1) \qquad 1\text{-}520\ (5)$$

where f(10) determines the frequency to the nearest 10-mc unit, f(1) to the nearest 1-mc unit, and f(0.1) to the nearest 0.1-mc unit. Consequently, the employment of a constant frequency, f_d, which in the ARC-27 transmitter is a 3.45-mc oscillator signal, would normally be superfluous. But in the ARC-27 set, the 3.45-mc crystal oscillator is inserted with advantage since it enables the same crystal units that control the injection frequencies in the receiver to control the injection frequencies in the transmitter.

1-521. With the aid of equations 1-520 (1) to (4) and the block diagram in figure 1-211, the frequency synthesizing system employed in the ARC-27 transmitter is largely self-explanatory. The desired transmitter signal is synthesized by heterodyning the 3.45-mc and 4.8—5.7-mc oscillator outputs and selecting their difference product, f_c, for amplification. In turn, f_c is heterodyned with the 25.7—34.7-mc oscillator output and again the difference product, this time f_b, is selected. Next, f_b is amplified and mixed with the f(10) output of the spectrum amplifier. The sum product, f_a, which is the desired final frequency, is selected and amplified for transmission.

1-522. The various injection frequencies used in Radio Set AN/ARC-27 are generated by conventional crystal oscillators, except for the output of the spectrum generator. The function of the spectrum generator is to convert an initial 10-mc output from an electron-coupled Pierce oscillator into any desired harmonic of 10 mc between 200 and 370 mc. Although the function of the spectrum generator is effectively that of a harmonic generator, the output frequency, f(10), is not simply the product of a straightforward sequence of frequency-multiplier and harmonic-selector stages.

Rather, f(10) is a synthesized frequency produced by selecting two appropriate harmonics of the 10-mc fundamental and mixing them to form the desired product. Referring to the block diagram in figure 1-212, it can be seen that the 10-mc oscillator is isolated from the multiplier stages by a buffer amplifier. The output of the buffer amplifier is used to excite two different multiplier circuits. One multiplier is fixed-tuned to select and amplify the 9th harmonic (90 mc); the second multiplier can be switched to select and amplify either the 1st, 2nd, 3rd, or 4th harmonic (10, 20, 30, or 40 mc). The outputs of the two multiplier circuits are then mixed. Selection of the desired frequency product is made by the ganged tuning in 10-mc steps of two selective vacuum-tube amplifier stages. In table (1) below are given the harmonic multiplier combinations by which the desired spectrum frequency, f(10), is obtained. Note that the 90-mc signal is not employed, but only its 2nd, 3rd, and 4th harmonics.

1-523. Figures 1-213 to 1-217 are schematic diagrams of the frequency-control circuits in Receiver-Transmitter RT-178/ARC-27. Minor differences exist among the various models of the transceiver which have been produced since the original. The schematics shown apply to Collins model No. 6 of the RT-178/ARC-27 series. The positions of the different switches and tuning adjustments are controlled automatically by a Collins Autotune system. Any of the 1750 channels can be selected by using the decade frequency selectors on the local control panel. In addition, any one of 18 preset channels or a guard receiver can be selected from both remote and local control panels. When the radio set is being operated in the receive position, plate voltage is automatically removed from those tubes that are used only during transmitter operation; likewise, when the radio set is being operated in the transmit position, plate voltage is removed from those tubes used only during receiver operation. With the aid of the block dia-

$f(10) = m \times 9f \pm nf$	$f(10) = m \times 9f \pm nf$
$200 = 2 \times 90 + 20$	$290 = 3 \times 90 + 20$
$210 = 2 \times 90 + 30$	$300 = 3 \times 90 + 30$
$220 = 2 \times 90 + 40$	$310 = 3 \times 90 + 40$
$230 = 3 \times 90 - 40$	$320 = 4 \times 90 - 40$
$240 = 3 \times 90 - 30$	$330 = 4 \times 90 - 30$
$250 = 3 \times 90 - 20$	$340 = 4 \times 90 - 20$
$260 = 3 \times 90 - 10$	$350 = 4 \times 90 - 10$
$270 = 3 \times 90 + 0$	$360 = 4 \times 90 + 0$
$280 = 3 \times 90 + 10$	$370 = 4 \times 90 + 10$

Table 1-522 (1). Harmonic generator combinations employed in the synthesis of f(10), the output frequency of the spectrum generator in Radio Set AN/ARC-27. All frequencies are in megacycles per second.

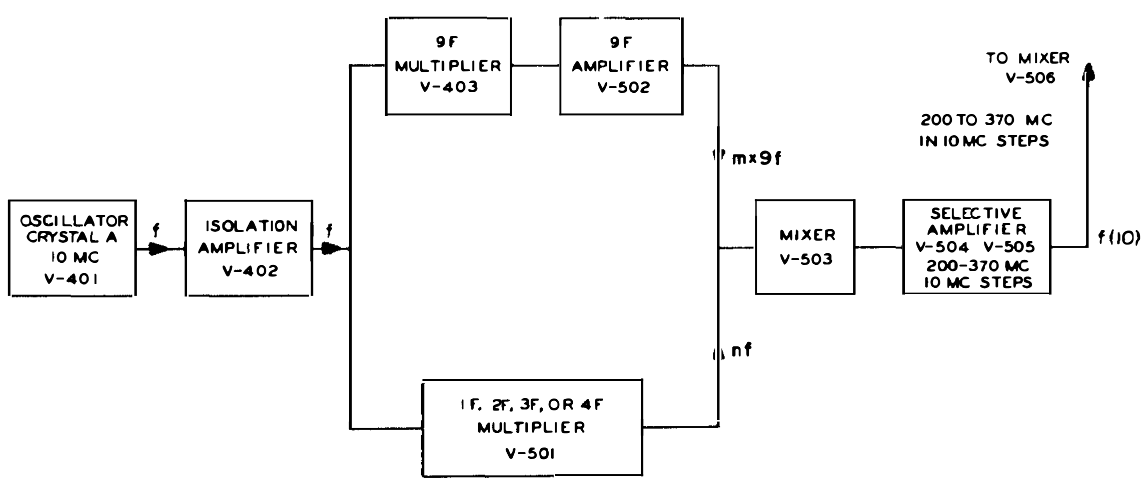


Figure 1-212. Block diagram of spectrum generator system in Radio Set AN/ARC-27

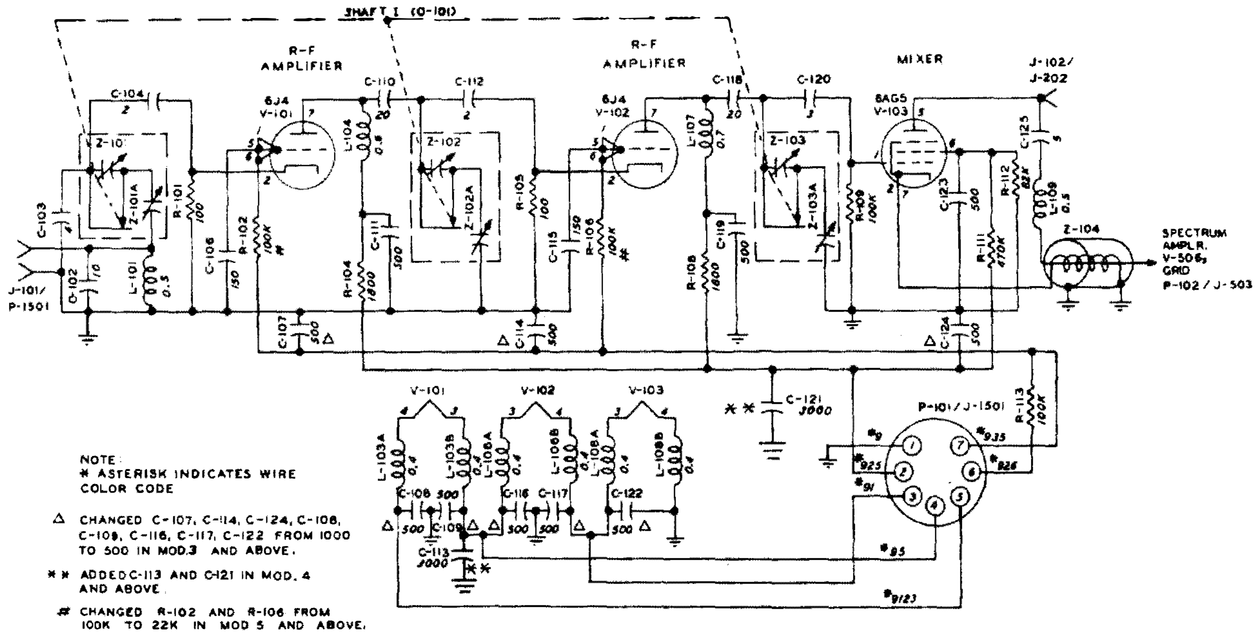


Figure 1-213. Main receiver r-f amplifier subassembly, Receiver-Transmitter RT-178/ARC-27. Includes receiver circuits V101, V102, and V103 (r-f amplifier and 1st mixer stages). See figure 1-210 for relation of the above stages to the rest of the receiver frequency-control system

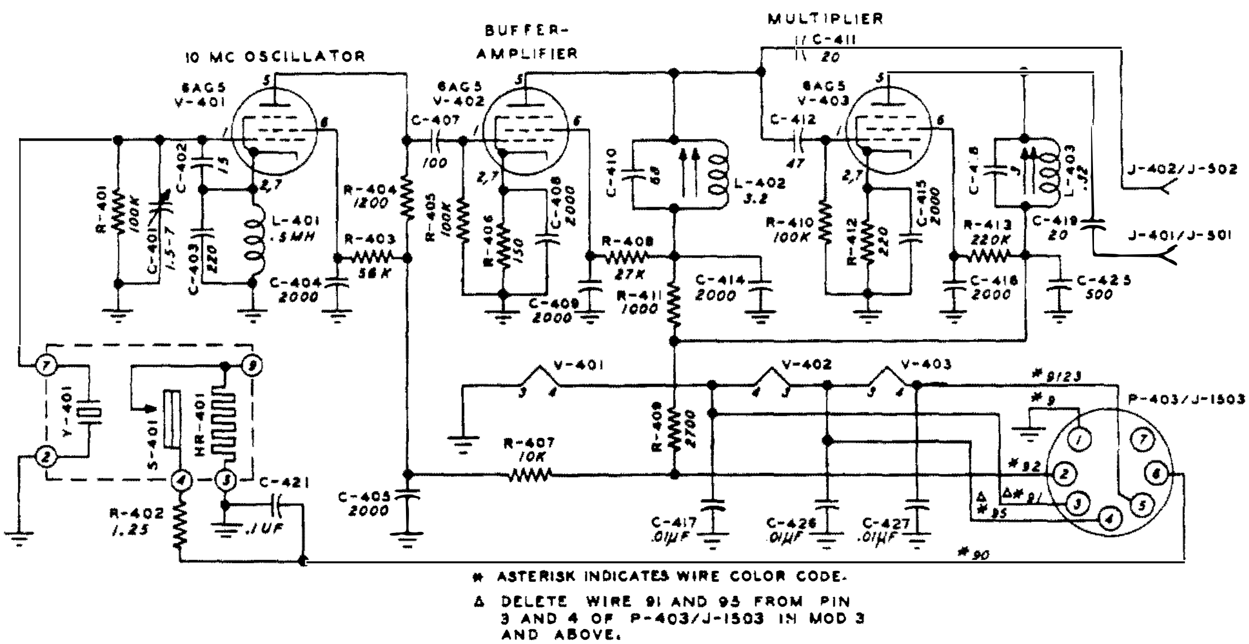


Figure 1-214. 10-mc spectrum oscillator, V401, buffer amplifier, V402, and 90-mc multiplier, V403, of Receiver-Transmitter RT-178/ARC-27. See figure 1-212 for function of the above circuits in the spectrum generator system

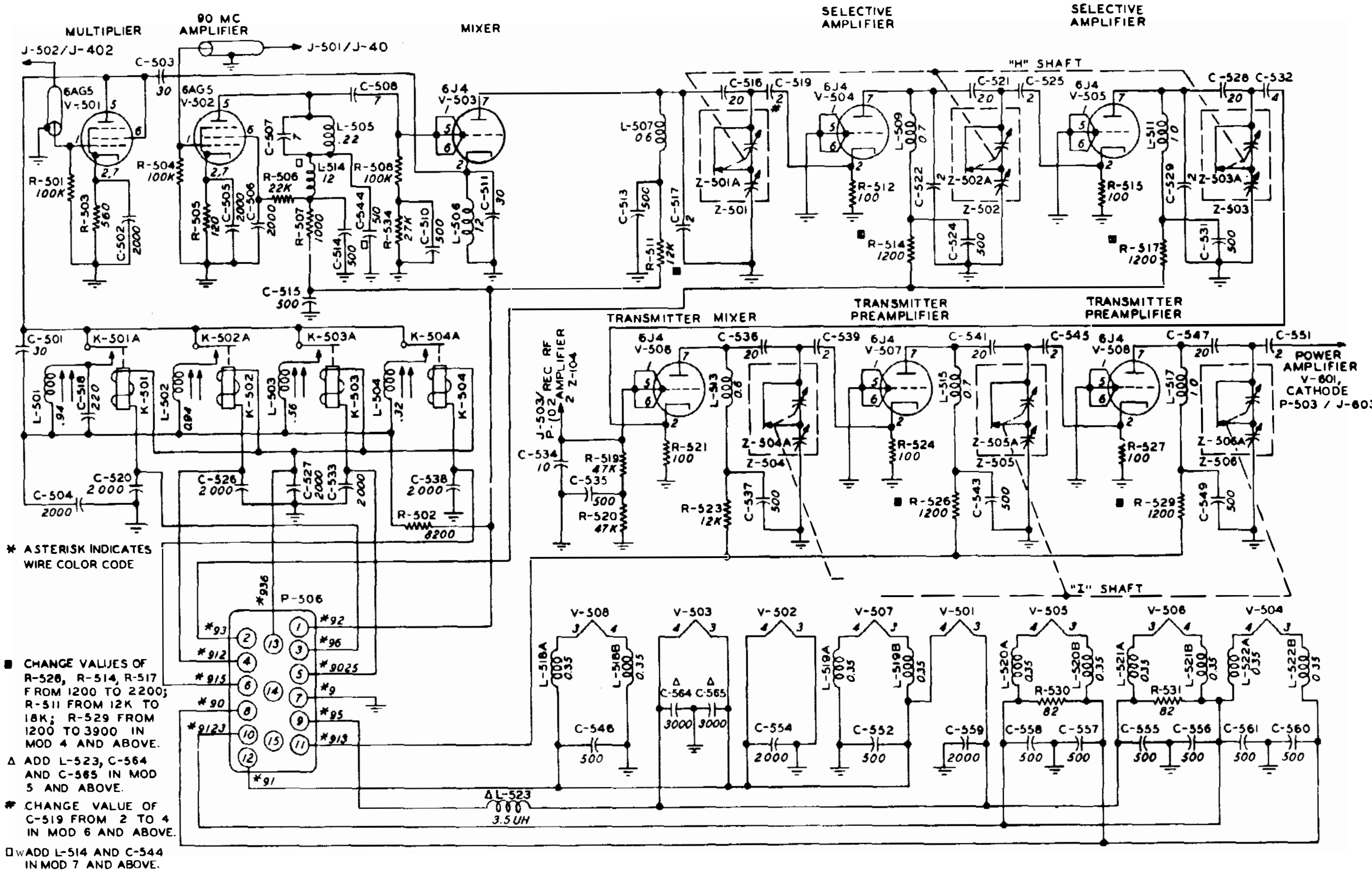


Figure 1-215. Spectrum amplifier and transmitter preamplifier subassembly, Receiver-Transmitter RT-178/ARC-27. See figure 1-212 for functional operation of spectrum generator stages V501, V502, V503, V504, and V505. See figure 1-211 for functional operation of transmitter stages V506, V507, and V508

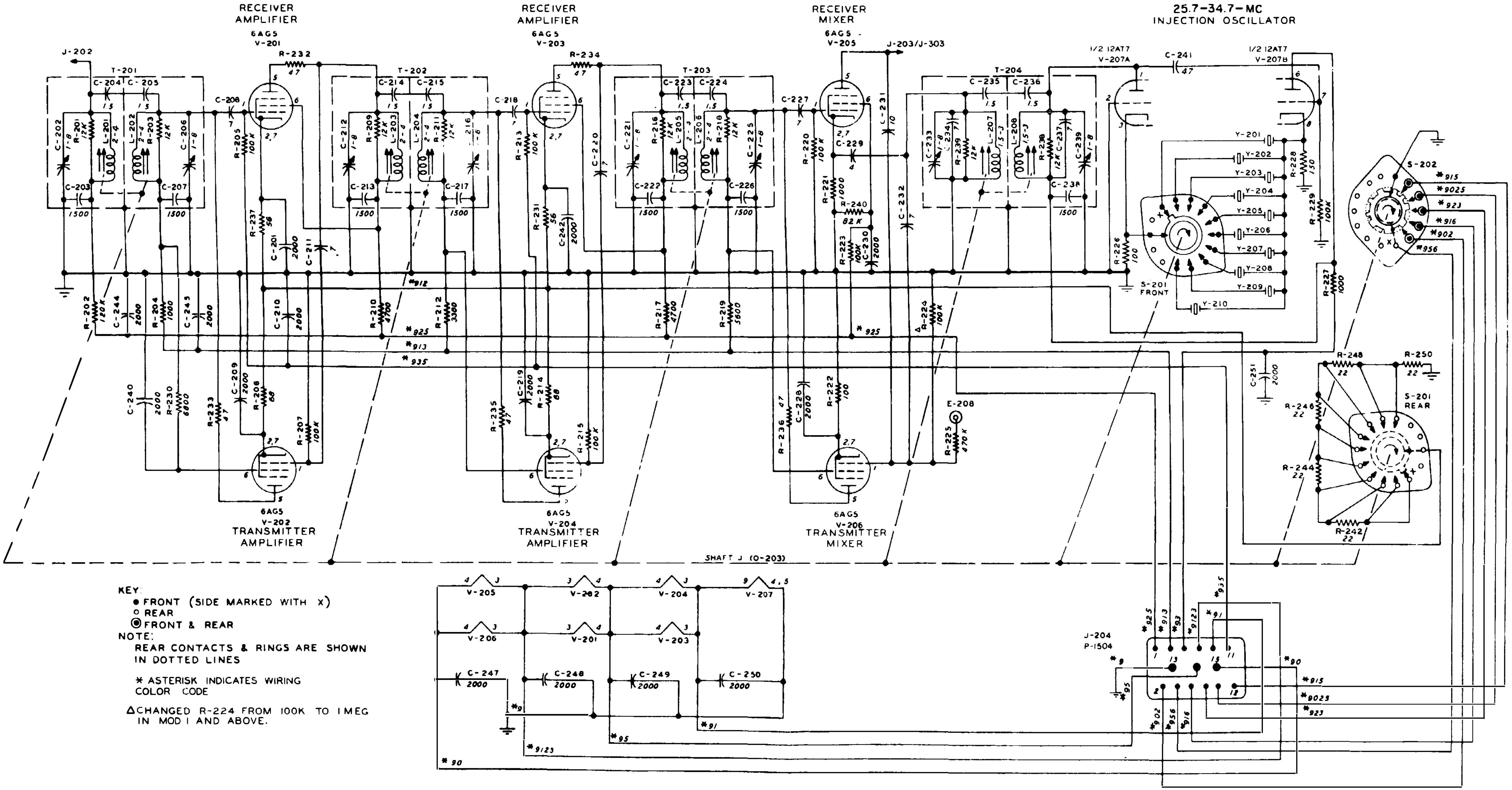


Figure 1-216. 20—30-mc i-f amplifier subassembly, Receiver-Transmitter RT-178/ARC-27. See figure 1-210 for functional operation of receiver circuits V201, V203, V205, and V207. See figure 1-211 for functional operation of transmitter circuits V202, V204, V206, and V207

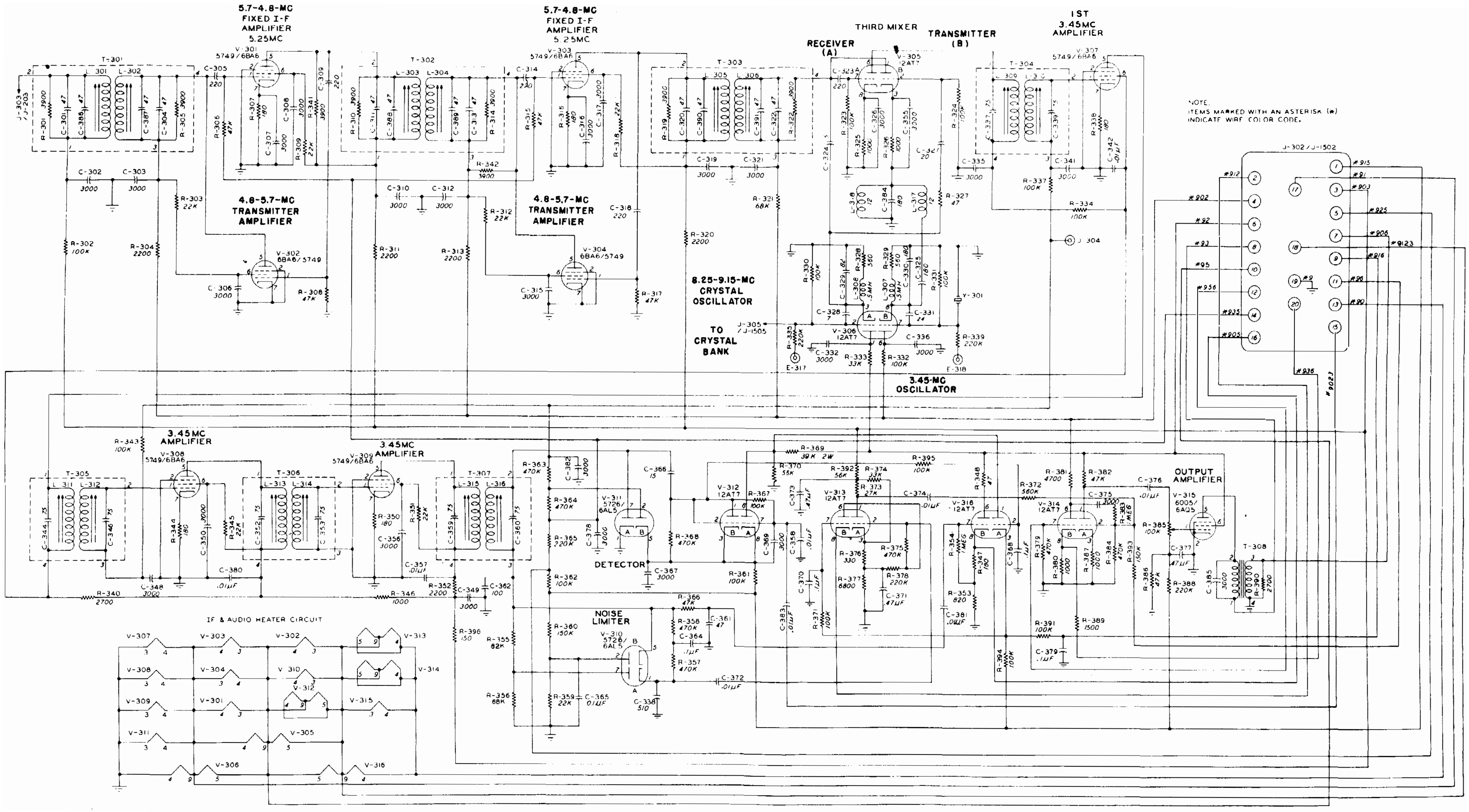


Figure 1-217. Fixed i-f amplifier subassembly, Receiver-Transmitter RT-178/ARC-27. See figure 1-210 for function of receiver circuits V301, V303, V305A, V306A, and V307. (V308 and V309, the 2nd and 3rd 3.45-mc amplifiers, which are similar in design to the V307 stage, are not shown. Also omitted are the vacuum-tube heater circuits.) See figure 1-211 for function of transmitter circuits V302, V304, V305B, V306A, and V306B

grams in figures 1-210, 1-211, and 1-212, the functions of the various circuit components shown in the schematic diagrams should be reasonably apparent in most instances. The designs of the individual stages are for the most part conventional. The r-f tuners used in the receiver r-f amplifier, in the spectrum amplifier, and in the transmitter preamplifier are not of a conventional type, but their design has been described in connection with similar circuits in Radio Receiver R-278/GR. See figure 1-204. Also unconventional is the coupling between the 1st mixer in the receiver and the spectrum amplifier. In figure 1-215, note that the V505 output of the spectrum generator is fed to the cathode of the 6J4 transmitter mixer, V506. Capacitor C534, which is connected between the grid of V506 and ground, is large enough to effectively ground the grid at the spectrum generator frequencies, but is small enough to present a fairly high impedance to the input from the 20—30-mc transmitter amplifier, V202. Thus, V506 operates as a grounded-grid circuit insofar as the spectrum injection frequencies are concerned, so that these frequencies tend to be attenuated somewhat in the mixer output. Attenuation of the 20—30-mc signal in the mixer output is less of a problem since the 20—30-mc band is separated by such a large percentage difference from the frequencies at which the transmitter preamplifier circuits are tuned. Although C534 effectively grounds the grid of V506 at the spectrum frequencies during transmitter operation, it does not completely do so. During receiver operation, when plate voltage is removed from V506, the output of the spectrum generator is coupled through the grid-cathode capacitance of the tube to the grid circuit, and the voltage developed across C534 is sufficient for injection excitation of the 1st receiver mixer. This injection voltage for receiver operation is fed through the same cable, Z104 (see figure 1-213), that is used to transmit the 20—30-mc signal to V506 during transmitter operation. The receiver mixer, V103, employs cathode injection by means of capacitance coupling between two wires in the specially constructed cable, Z104. From the schematic it can be seen that the wire that is connected directly to the grid of V506 is connected to the plate of V103 through C125 and L109. L109 is an r-f choke at the spectrum generator frequencies, so that for all practical purposes this circuit can be assumed to be open during receiver operation. For transmitter operation, L109 and C125 in series present a sufficiently low reactance for the 20—30-mc signal from the transmitter amplifier, V202, to be fed to the transmitter mixer, V506. The reason that the 20—30-mc transmitter signal is fed to the transmitter mixer via the plate circuit of the receiver mixer instead of being fed directly is simply one of economy—to make use of the J102/J202 connection from the plate of the receiver mixer to transformer T201 during transmitter operation as well as during receiver operation. In figure 1-216 it can be seen that during receiver operation, transformer T201 couples the output of receiver mixer V103 to the input of receiver amplifier V201; during transmitter operation, T201 is operated in the reverse direction, coupling the output of V202 to the plate of V103 (which tube is inoperative since its plate voltage is removed) and from there through C125, L109, Z104, and connection P102/J503 to the grid of the transmitter mixer, V506.

1-524. Since the greater part of the final transmitter frequency is controlled by the 10-mc spectrum generator oscillator, it is primarily the stability of this oscillator that determines the stability of the transmitted frequency. The 10-mc oscillator is a grounded-plate (r-f grounded-screen-anode) electron-coupled Pierce circuit that employs an oven-mounted type CR-27/U crystal unit. The remaining oscillators do not employ thermostatically controlled ovens for mounting their respective crystal units, since the normal frequency deviations of these crystals with temperature changes can have but an insignificant percentage effect upon the final frequency. R-f grounded-plate oscillators of the Pierce type employing CR-18/U crystal units are used for the 3.45-mc and the 8.25—9.15-mc oscillators. The 25.7—34.7-mc injection oscillator is of the cathode-coupled Butler type and employs a 10-position bank of type CR-23/U crystal units. For the reader interested in a more detailed discussion of the over-all design of Radio Set AN/ARC-27, reference can be made to USAF Technical Order No. 12R2-2ARC27-2.

Attenuation of Unwanted Products in Frequency Synthesis

(This discussion is primarily an abstract of portions of "Developments in Frequency Synthesis" by Mr. H. J. Finden of the Plessey Company, Ltd., England, a paper delivered before the Conference on High Frequency Measurements, Washington, D.C., 1953 and published in revised form in *Electronic Engineering*, May, 1953.)

1-525. Among the more difficult problems that face the designer of frequency synthesizing circuits are those that concern the elimination of all frequencies except the one desired signal. When two frequencies, f_1 and f_2, are mixed, it is customary to regard the output of the mixer stage as being composed of four frequencies, namely, f_1, f_2, f_1 plus f_2, and f_1 minus f_2. In practice, we know that due to non-linearities of both the input and output stages, higher harmonics of each of the fundamental frequencies are also present, as well as all possible combinations of their sum and difference products. Thus, whenever a particular output frequency is to be selected from a synthesized mixture, especially if a relatively pure sine wave output is desired, many more than the four principal fundamental frequencies must be considered. The conventional method of selecting one frequency from a mixture of several is, of course, to employ filters that readily pass a narrow band of frequencies in the neighborhood of the desired frequency, but greatly attenuate all frequencies above and below the passband. In the special case of where the desired frequency is the lowest frequency present, a low-pass filter can be used to reject all frequencies higher than the one desired. (Where all harmonics are present, high-pass filters are not applicable for the rejection of all but one frequency.) Where a relatively pure output is desired, satisfactory filtering can become quite difficult and expensive to achieve in a frequency synthesizer unless two preventive steps are taken: one step is to ensure that the input frequencies of the mixer are attenuated in the mixer stage itself; a second step is to ensure that none of the significant unwanted frequencies lies close to the desired frequency.

THE BALANCED MODULATOR IN FREQUENCY SYNTHESIS

1-526. When two or more frequencies are to be mixed to obtain a sum or difference product, the normal approach is to employ a mixer stage of maximum efficiency as measured by the ratio of the heterodyne output level to the input signal level. However, where greater-than-normal purity of the desired heterodyne product is the goal, as well may be the requirement of a frequency synthesizer, the more important signal ratio to consider in the mixer is the ratio of the heterodyne output level to the level of the input frequencies as they are measured in the *output*. For this reason the balanced modulator can be used to advantage as a mixer stage in a frequency synthesizer of pure sine waves, even though its conversion efficiency is less than that of other types of mixers. Figure 1-218 illustrates the circuit of a balanced linear modulator of the same design as the mixer stages that are used in the Plessey frequency synthesizer. Note that since the two balanced amplifiers are connected in parallel, but are excited 180 degrees out of phase, the input signals are virtually eliminated in the output. In practice, it is not possible to balance the circuit perfectly. Optimum performance is obtained by adjusting the circuit for maximum attenuation of whichever input frequency differs from the desired heterodyne product by the smallest percentage. Generally, this is the higher of the two input frequencies, except when the desired frequency is the difference product, where the higher of the two input frequencies is not greater than 2.6 (approximately) times the lower.

1-527. Measurements made at the Plessey Company of the relative power levels of the frequencies in the output of an experimental balanced modulator are shown in table (1). For this experiment f_1 equaled 100 kc and f_2 equaled 740 kc. The modulator was balanced for maximum attenuation of f_2. The output strength of the fundamental difference product, 640 kc, was taken as the zero db reference level. Note that the sum product, 840 kc, was measured at the same level, but that all other frequencies in the output were at negative db levels. The sum and difference products involving the 5th harmonic of f_1 (500 kc) were negligible. Where the strength of the wanted frequency must be made 60 or 70 db greater than that of any other frequency, it can be seen that a mixer stage which does not, itself, attenuate the input frequencies in its output circuit will make the problem of eliminating all the unwanted frequencies by filters and/or selective amplifiers alone much more formidable.

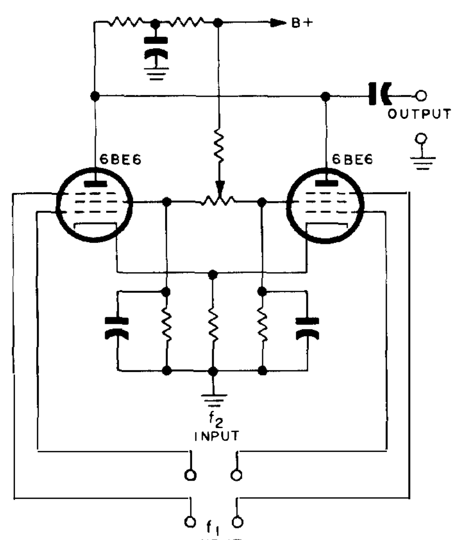


Figure 1-218. Balanced modulator circuit, which is useful as a mixer stage when maximum attenuation of the input frequencies in the output is desired

	$f_2 = 740$ kc — 30 db	$2f_2 = 1480$ kc — 30 db	$3f_2 = 2220$ kc — 33 db	$4f_2 = 2960$ kc — 58 db
$f_1 = 100$ kc — 8 db	$f_2 \pm f_1$	$2f_2 \pm f_1$	$3f_2 \pm f_1$	$4f_2 \pm f_1$
	640 kc 0 db	1380 kc —37 db	2120 kc —49 db	2860 kc —68 db
	840 kc 0 db	1580 kc —37 db	2320 kc —49 db	3060 kc —68 db
$2f_1 = 200$ kc — 35 db	$f_2 \pm 2f_1$	$2f_2 \pm 2f_1$	$3f_2 \pm 2f_1$	$4f_2 \pm 2f_1$
	540 kc —38 db	1280 kc —46 db	2020 kc —69 db	2760 kc —72 db
	940 kc —38 db	1680 kc —47 db	2420 kc —69 db	3160 kc —72 db
$3f_1 = 300$ kc — 42 db	$f_2 \pm 3f_1$	$2f_2 \pm 3f_1$	$3f_2 \pm 3f_1$	$4f_2 \pm 3f_1$
	440 kc —65 db	1180 kc —66 db	1920 kc —73 db	2660 kc —78 db
	1040 kc —63 db	1780 kc —68 db	2520 kc —72 db	3260 kc —79 db
$4f_1 = 400$ kc — 57 db	$f_2 \pm 4f_1$	$2f_2 \pm 4f_1$	$3f_2 \pm 4f_1$	$4f_2 \pm 4f_1$
	340 kc —58 db	1080 kc —70 db	1820 kc —77 db	2560 kc —78 db
	1140 kc —60 db	1880 kc —72 db	2620 kc —77 db	3360 kc —78 db
$5f_1 = 500$ kc — 71 db				

Table 1-527(1). Frequencies present in the output of an experimental balanced modulator when the input fundamental frequencies are 100 kc (f_1) and 740 kc (f_2). All output levels in db are given with respect to a 0-db level assumed for the difference product ($f_2 — f_1$) of 640 kc. The modulator was balanced for maximum attenuation of f_2.

SELECTION OF INPUT FREQUENCIES FOR SYNTHESIZING STAGE

1-528. Where the output strength of the desired synthesized frequency must be at least 60 db greater than that of any unwanted frequency, it is of utmost importance that none of the unwanted frequencies (that are not already 60 db or more below the desired frequency) in the mixer output approach the frequency of the desired signal. To avoid such a possibility may require a very careful selection of the original frequency elements to be mixed. Assume, for example, that the desired signal is to be the difference product, $(f_2 — f_1)$, where f_2 is greater than f_1, and that no accompanying frequency is to have a filter or selective amplifier output power greater than —70 db relative to that of the desired signal. The question arises: What rules-of-thumb can guide the radio engineer in avoiding unwanted frequencies so close to the wanted frequency that they cannot be easily separated by conventional filter circuits? The answer to this question is to avoid all selections of f_1 and f_2 that cause the ratio $f_2/(f_2 — f_1)$ to approach any of the values shown in table (1). How these forbidden ratios have been derived is explained in the following paragraph.

Forbidden Values of $f_2/(f_2 — f_1)$	
5 4 3 5/2 2 5/3 3/2 4/3 5/4	Note: Insofar as a 70-db separation in power levels is concerned, this table applies only to the use of balanced-modulator mixers adjusted for maximum attenuation of f_2. However, the forbidden values should generally be avoided regardless of the type of mixer circuit.

Table 1-528(1). Forbidden values of the ratio $f_2/(f_2 — f_1)$, where f_2 is greater than f_1 and the desired frequency, $(f_2 — f_1)$, is to be made at least 70 db greater in signal strength than any accompanying frequency.

1-529. The values of the forbidden ratios given in the table of the preceding paragraph were originally derived by H. J. Finden with the aid of the empirical data in table 1-527(1). The assumption was made that this empirical data for f_1 of 100 kc and f_2 of 740 kc could be accepted as generally representative of any values of f_1 and f_2, where f_2

is the higher frequency. The manner of derivation of the forbidden values is as follows:

a. First, let us keep in mind that our principal purpose is to avoid the presence of unwanted frequencies that lie close to the desired frequency, which we here assume to be the difference product, $(f_2 - f_1)$.

b. Let it be assumed that f_2 is greater than f_1, but that f_1 is large enough that f_2 is at least 5 per cent (or better yet, at least 10 per cent) greater than the difference frequency $(f_2 - f_1)$.

c. With assurances of at least a 5 per cent difference between f_2 and $(f_2 - f_1)$, and with the modulator specifically balanced to eliminate f_2 [note the —30 db level of f_2 in table 1-527(1)], we can assume that the remaining attenuation of f_2 can be achieved by elementary filter design.

d. Likewise we can assume that all overtones of f_2 ($2f_2$, $3f_2$, etc) can be eliminated by elementary filter design, as well as all overtones of $(f_2 - f_1)$; i.e. $(2f_2 - 2f_1)$, $(3f_2 - 3f_1)$, etc.

e. Similarly, we can disregard all sum frequencies, $(f_1 + f_2)$, $(2f_1 + f_2)$, $(f_1 + 2f_2)$, etc., since all will be more than 10 per cent greater than the desired frequency, $(f_2 - f_1)$. This group also includes all sums of $(f_2 - f_1)$, or its overtones, with other frequencies. For example, $(f_2 - f_1) + f_2$, which is the same as $(2f_2 - f_1)$.

f. With the above frequencies eliminated from consideration, the remaining frequencies which might prove troublesome, as indicated in table 1-527(1), are generalized below in table 1-529(1).

g. We equate each of these frequencies with the desired frequency, $(f_2 - f_1)$, and solve for f_2 in terms of f_1. These equations are shown below, with the corresponding solutions for f_2 and for $f_2/(f_2 - f_1)$.

h. Since the numerical values of the ratios derived above correspond to values of f_1 and f_2 that produce harmonic products equal to the desired frequency, it can be seen that any close approach to such ratios should be avoided when selecting the input frequencies to a mixer stage.

f_1	$f_2 - f_1$			
$2f_1$	$f_2 - 2f_1$			
$3f_1$	$f_2 - 3f_1$	$2f_2 - 3f_1$		
$4f_1$	$f_2 - 4f_1$	$2f_2 - 4f_1$	$3f_2 - 4f_1$	

Table 1-529(1). Frequencies present in the output of a mixer that can approach in value the desired difference frequency $(f_2 - f_1)$ and be of a sufficient power level to make elimination by conventional filter circuits difficult.

When	*Then*	*And*
$f_2 - f_1 = f_1$	$f_2 = 2f_1$	$f_2/(f_2 - f_1) = 2$
$= 2f_1$	$= 3f_1$	$= 3/2$
$= 3f_1$	$= 4f_1$	$= 4/3$
$= 4f_1$	$= 5f_1$	$= 5/4$
$= 2f_1 - f_2$*	$= 3/2f_1$	$= 3$
$= 3f_1 - f_2$*	$= 2f_1$	$= 2$
$= 4f_1 - f_2$*	$= 5/2f_1$	$= 5/3$
$= 2f_2 - 3f_1$	$= 2f_1$	$= 2$
$= 3f_1 - 2f_2$	$= 4/3f_1$	$= 4$
$= 2f_2 - 4f_1$	$= 3f_1$	$= 3/2$
$= 4f_1 - 2f_2$	$= 5/3f_1$	$= 5/2$
$= 3f_2 - 4f_1$	$= 5/2f_1$	$= 5/3$
$= 4f_1 - 3f_2$	$= 5/4f_1$	$= 5$

* Note that only the equation where f_2 is assumed to be less than the f_1 harmonic is used. Otherwise f_1 would necessarily be zero. For example, if

$$f_2 - f_1 = f_2 - 2f_1$$

then f_1 must be zero. In the last six equations, however, the possibility exists for the f_2 harmonic to be either above or below the f_1 harmonic.

Crystal-Phase-Controlled Harmonic Multipliers

1-530. Of great promise in the field of frequency synthesis is the possible future application of harmonic generators of the crystal-phase-controlled type. With such a generator a single low-, medium-, or high-frequency crystal can be used to control a wide band of radio channels extending well into the u-h-f range. The basic circuitry is quite simple. Where maximum economy in parts is necessary, a single triode stage is capable of producing any one of a sequence of crystal overtone frequencies, with the selected signal being 40 db above the level of the two adjacent harmonics. Such a circuit was demonstrated during the first investigations of phase-controlled multipliers reported in the United States. The basic research, reported by Dr. A. Hahnel*, and a developmental project, reported by L. R. Battersby and E. A. Conover†, were both undertaken at the Signal Corps Engineering Laboratories at Fort Monmouth, N. J.

1-531. The principal operational features of a crystal-phase-controlled harmonic multiplier are illustrated in figure 1-219. As shown in the block diagram of (A), the basic circuit consists of a variable-frequency oscillator arranged to be keyed at a crystal-controlled rate, f_c; which, of course, should be a much lower frequency than the v-f-o frequency, f_o. If the circuit is properly designed, the variable frequency oscillator, when tuned through its band, will generate a succession of output frequencies equal to those overtones of the crystal frequency which lie within the v-f-o tuning range. For example, let us suppose that when the variable frequency oscillator is operated in a steady, constant-amplitude state (a steady key-on operation, with the crystal-controlled key shorted across), f_o is found to vary continuously from 9990 kc to 20,000 kc. Then, when crystal-controlled keying is applied, say, at a frequency, f_c equal to 50 kc, it will be found that all the tuned frequencies are suppressed except those that are harmonics of the 50-kc keying voltage. Thus, with the oscillator tuned to its original 9990-kc position, the primary output is no longer 9990 kc, but the nearest 50-kc harmonic, 10,000 kc. As the oscillator is tuned through its natural 10,000-kc position, the output at this frequency reaches a maximum. As the oscillator tuning is varied farther in the high-frequency direction, a 10,050-kc signal begins to increase in amplitude. When the oscillator is tuned to its original 10,025-kc position, the 10,000-kc and the 10,050-kc signals will appear in the output at approximately equal amplitudes. Finally, when the tuning passes through the oscillator 10,050-kc maximum position, the 10,000-kc signal will have dropped to a level 40 db or more below the 10,050-kc output. In such a manner this particular circuit can generate any desired overtone of the crystal frequency between the 200th and 400th harmonics—200 crystal-controlled frequencies in all.

1-532. If all frequencies in the v-f-o output are to be suppressed except the single desired crystal-controlled harmonic, the output waveform, see figure 1-219(B), must meet certain conditions. First, the rise time must be as short as possible.

* Hahnel, Alwin. "Multichannel Crystal Control of VHF and UHF Oscillators," *Proc. I. R. E.*, Vol. 41, Pages 79-81, January 1953. See also Bibliography Nos. 898 to 901.

† Battersby, Lyle R. and Conover, E. A. "A Single Crystal Multi-Channel Oscillator," *Technical Memorandum No. M-1567*, Signal Corps Project No. 132A, March 1954.

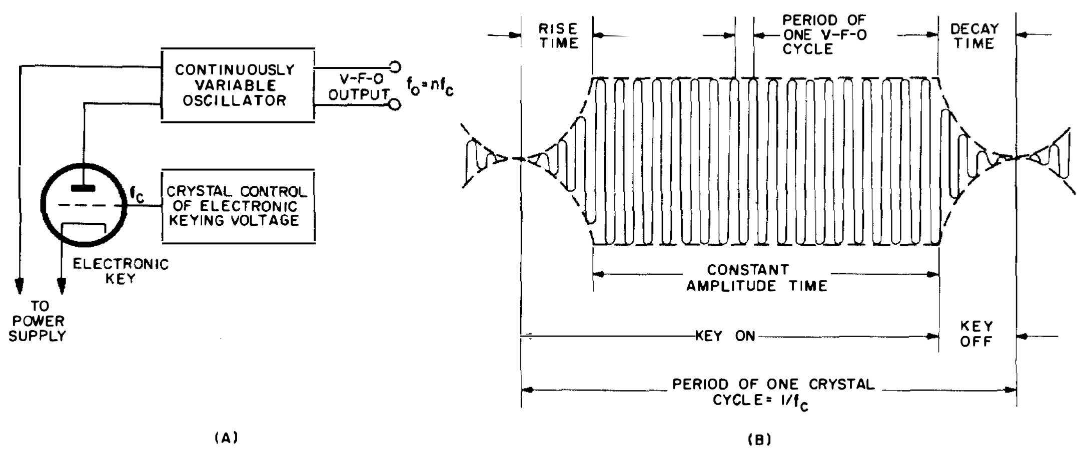


Figure 1-219. Simplified block diagram (A) and output waveform (B) of crystal-phase-controlled harmonic multiplier

This is the time it takes the v-f oscillations to build up to their equilibrium amplitude. Second, the decay time must also be as short as possible. However, the total key-off period must be extended sufficiently to ensure that the oscillations die down to noise-level proportions. This is necessary in order for oscillations at the beginning of each keying pulse to start in the same phase. Third, as implied by the first two conditions, the constant-amplitude time must be long relative to the rest of the keying cycle — three-fourths or more of the total. If these conditions are not met, the output will always contain a multiplicity of different harmonic frequencies at approximately the same amplitude level.

1-533. That the operational conditions described above are necessary in order to obtain a single-frequency output has been confirmed empirically. Such conditions were originally indicated theoretically when the waveform was subjected to a Fourier analysis. Intuitively, from a qualitative examination of the waveform, it would appear that those conditions of maximum constant-amplitude time, which indicate minimum distortion of the output from a pure unmodulated continuous wave, could be expected to be the conditions approaching most closely single-frequency phase-controlled operation. What may not be qualitatively obvious is why the constant-amplitude portion of the output pulse, which supplies practically all of the output energy, should not cause the dominant frequency to be the actual frequency at which the variable frequency oscillator is tuned to operate. Let us assume that the keying voltage is a square-wave pulse. Since the v-f oscillations must build up from thermal level, we can assume that they contain no "memory" of the preceding keying pulse or of the pulse rate. In other words, once the oscillator is keyed, the buildup waveform and the constant-amplitude frequency are exactly the same whether the keying pulse is to last 1 microsecond or 1 day. So qualitatively we must conclude that the steady-state oscillations have a period and corresponding frequency that are independent of the keying frequency. The steady-state period of a v-f-o cycle during crystal control is thus no different from that of the same tuned circuit without crystal control. How then are we to account for the fact that when crystal control is applied, the measured output frequency immediately shifts from the tuned-circuit resonance value to the nearest harmonic of the crystal frequency? The answer appears to lie in the fact that the generated phase-controlled output physically has the same period and cycle-to-cycle frequency as the unmodulated tuned oscillator, but because of the periodic phase shift at the crystal frequency, a tuned receiving circuit can indicate a maximum resonance absorption of energy only if it is tuned to a harmonic (approximately) of the periodic phase shift.

1-534. Imagine an ideal phase-controlled waveform in which the build-up time and the decay time are instantaneous, so that the periodic phase shift can also occur instantaneously without the intermission of a key-off interval. In this event, if the v-f-o tuned frequency is exactly equal to a harmonic of the phase-control frequency, no phase adjustment occurs, and the output does not differ in form from a steady-state continuous wave. Now suppose that the v-f-o tuned frequency is not a harmonic of the phase-control frequency, and that we attempt to couple electronically the output to a tuned receiving circuit. With the receiving circuit tuned to the v-f-o frequency, we can imagine that during the first phase cycle the received resonance energy builds up to a certain level. To idealize further, let us also imagine that the resistance of the receiver tuned circuit is effectively zero, so that when the first phase cycle is completed, the energy absorbed during that period continues undiminished as a free-running oscillation without a change in phase. During the succeeding phase period a corresponding component of oscillation of equal amplitude but different phase is fed the receiving circuit. After several such periods, in which the phase of each succeeding oscillation component is rotated in the same direction an equal amount from the phase of the preceding component, it can be assumed that the accumulation of opposite-phased oscillations tend to cancel each other (i.e., the tuned circuit returns energy to its power source as fast as it is supplied). Thus, even if the circuit ohmic resistance is zero, a resonance condition cannot be indicated by a continuous buildup of oscillations. On the other hand, if the receiving circuit is tuned to the near harmonic of the phase cycle, we can imagine that the energy absorbed during each phase period continues as a free-running oscillation at the harmonic frequency, nf_c. Since there are always exactly n of these cycles during each phase period, the imaginary succession of oscillation components will all have the same phase, and hence will tend to add to each other rather than cancel. Remember that the beginning of each phase cycle does not constitute a phase correction of its harmonic cycle, but of the v-f-o cycles, which we here assume are feeding energy to the semi-free-running harmonic-tuned receiving circuits. Before the phase of the input cycle can gradually shift to a point where it is in phase opposition to the hypothetical harmonic oscillations, it is abruptly returned to its starting position and a new phase cycle begins. In this way we can see how the phase-controlled v-f-o oscillations are able to continually feed energy to oscillations that are multiples of the phase-control frequency, and yet tend to suppress oscillations that have the same period as the v-f-o steady-state output. Thus, where the frequency-measuring technique involves the absorption of energy in a variable circuit of calibrated tuning range, we would conclude that the output frequency of the phase-controlled vfo is apparently a harmonic of the crystal frequency. Similarly, if the phase-controlled frequency were being tested by matching with the phase of an oscilloscope sweep, the sweep would have to be synchronized by the crystal harmonic frequency to hold the pattern still. If synchronized at the v-f-o tuned frequency, the pattern of a single wave will take a hop in phase at the beginning of each phase correction, and thus appear to move across the screen. When synchronized by a crystal harmonic, the wave may begin a change of phase, but immediately jumps back to its original position at the beginning of each phase correction. Again, the phase-controlled harmonic, nf_c, becomes the measured frequency of the vfo when the measuring technique involves the beat-frequency method, since basically this method measures the unknown frequency by determining the rate at which its phase changes with respect to that of a known standard. Thus, purely from qualitative considerations, we are led to suppose that any conventional frequency-measuring method would indicate that the output of a phase-controlled oscillator has an apparent frequency, f_o, equal, not to the v-f-o tuned operating frequency, but to nf_c.

1-535. The reader should accept the qualitative explanation of the principle of phase-controlled multiplication, as given in the foregoing paragraph, as being somewhat on the speculative side. In the practical oscillator, it may be that sufficient coupling can be expected to exist between the v-f-o oscillator and the crystal oscillator to pull them into a mutual synchronization with each other. Certainly, such synchronization is to be expected when the natural v-f-o frequency approaches very closely a harmonic of the crystal fundamental. Also, the explanation given above implies that a tuned receiving circuit, which, perhaps because of periodic clamping or quenching, lacks a sufficient "memory" to store an oscillation above noise level for the duration of the key-off interval, could not be used by itself to detect a phase-controlled output. If such a circuit were used as a coupling stage, maximum transfer of energy would occur if the stage were tuned to pass the tuned v-f-o frequency rather than the crystal harmonic. Since the discrimination of an effectively low-Q circuit between nearly equal frequencies is negligible in practical circuits, raising the question at this point is academic, except to remind the reader that our qualitative explanation of the phase-controlled signal implies that the value of f_o is not independent of the detecting system. On the other hand, the Fourier analysis of the output waveform, leads us to consider the crystal-controlled harmonics as being the actual frequencies of the v-f-o output, inherent in the waveform.

HAHNEL SPECTRUM GENERATOR

1-536. Figure 1-220 (A) is the schematic diagram of a one-tube spectrum generator of the phase-controlled v-f-o type developed by Hahnel and associates. Since the v-f-o frequency is several times higher than the crystal frequency, inductors that present very low impedances at the crystal frequency, but high impedances at the v-f-o frequencies, can be used to separate the crystal feedback path from that of the variable oscillator, thus permitting a single triode to serve as amplifier for both oscillators. In figure (B) is shown the equivalent circuit of the crystal oscillator, which we see is more or less a conventional Miller circuit. In figure (C) is shown the equivalent v-f-o circuit, which we see is a Colpitts circuit. This basic spectrum generator is operable at all lower frequencies, but of special interest is the fact that a crystal-controlled output was obtained even in the u-h-f range when an L_4C_4 butterfly circuit tunable from 250 to 900 mc was used.

1-537. As explained earlier, a single-frequency output from the phase-controlled oscillator in figure 1-220 requires that the constant-amplitude interval of the phase cycle be as long as possible consistent with a sufficient decay time. To achieve this R_g must be varied to a value consistent with the resistance of the crystal unit so that the crystal oscillator is biased in the class A region; that is, the tube is cut off (for v-f operation) only a small portion of each crystal cycle. In order to make sure that the decay time is very short, the f_o tuned circuit should be rather heavily damped, and low ratios of L/C should be used in the tuned circuit. Since these design features will tend to keep the energy storage in the v-f-o system low, they will also serve to shorten the buildup time.

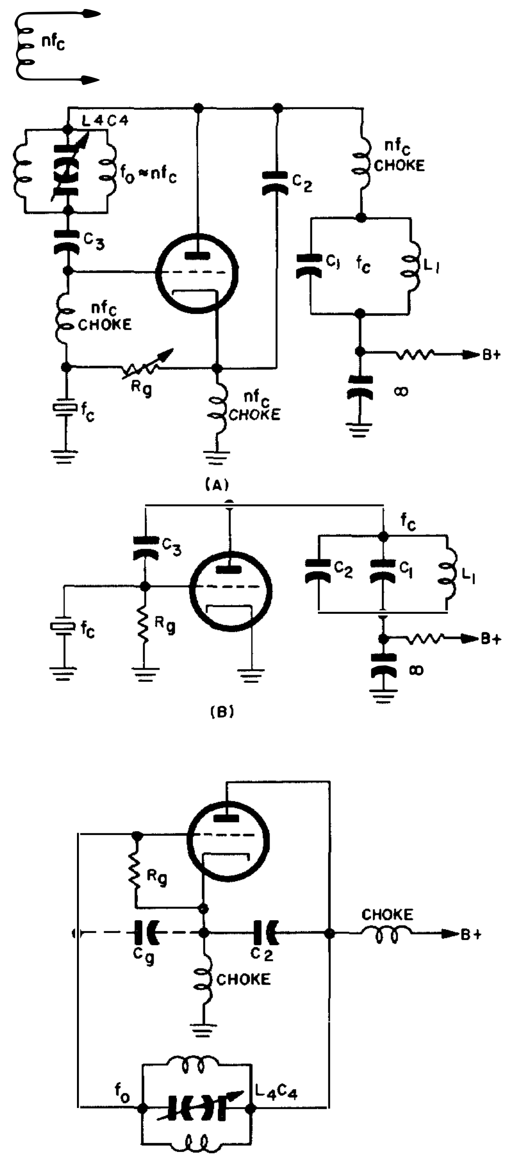


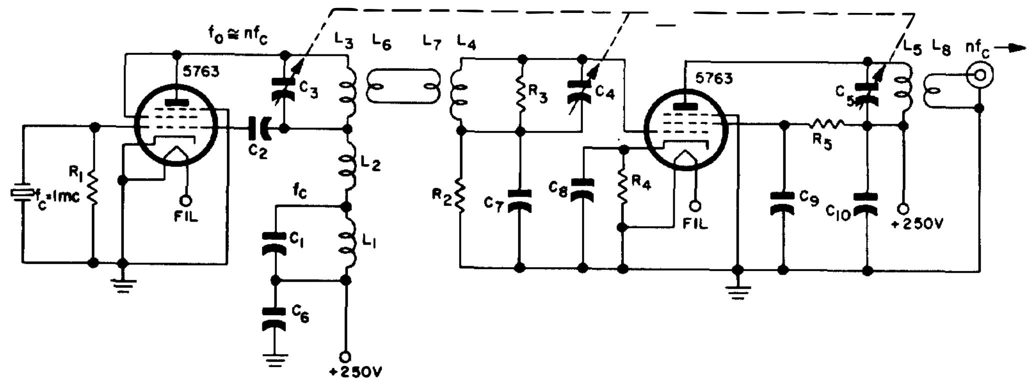
Figure 1-220. (A) Schematic diagram of Hahnel spectrum generator in which two oscillator circuits are combined in one stage; (B) Simplified schematic diagram of the crystal oscillator circuit of the spectrum generator; (C) Simplified schematic diagram of the v-f-o circuit of the spectrum generator

It is also important in this latter respect to use a tube of high transconductance, and to ensure that this transconductance is effective during the constant-amplitude interval. Short rise and decay times also require that the crystal keying voltage is changing at a fast rate at the v-f-o on and off points. For this purpose, the amplitude of crystal oscillation should be as large as practical. Of course, the gridleak resistance will need to be relatively small in order to keep the bias in the class-A region.

1-538. Figure 1-221 shows the complete schematic of a Hahnel circuit followed by an amplifier. This circuit was developed by Battersby and Conover and designed to cover the frequency range of 20 to 30 mc in 1-mc steps. With the additional amplifier stage, a 50-db selectivity is obtained for the desired harmonic channel relative to the adjacent channels. An output of 3 volts is possible across a 50-ohm load. The oscillator circuit is essentially the same as that in figure 1-220. The amplifier stage is quite conventional. The link coupling to the amplifier is employed to minimize the leakage of the 1-mc signal from the oscillator circuit to the amplifier input. R_3 serves to shorten the decay time. Its damping effect also improves the tracking performance of the ganged tuning.

TRANSISTOR OSCILLATORS

Our assignment in transistor oscillators is not to provide detailed mathematical analyses of the various types of circuits, but to describe and discuss semi-quantitatively experimental single-stage circuits that are representative of basic methods for obtaining the loop feedback required to maintain stable oscillations. The basic methods discussed are those that, to obtain feedback, employ series-mode crystal units in the feedback arm, negative-resistance circuitry, and transformer coupling. Modifications of each type of basic circuit are given, but unfortunately the test data available is insufficient to permit definitive comparisons of the characteristics and limitations of the different circuits. The design engineer, of course, would prefer concrete recommendations in choosing a circuit and in guiding its design to provide optimum characteristics for the particular needs at hand. However, crystal oscillators employing transistors in lieu of vacuum-tube amplifiers are at the present writing still more or less in the trial-and-error experimental stage. Although technical information is being accumulated in a number of laboratories, the detailed test data for the most part represent investments in competitive enterprises unavailable at this time for public communication. The reader should understand that the particular circuits described are primarily experimental in nature, being given as illustrations of basic oscillator types and not as endorsements for general use.



R_1 = 30,000 ohms	C_1 = 112 $\mu\mu f$	L_1 = 114 μh	f_c = 1 mc
R_2 = 15,000 ohms	C_2 = 10 $\mu\mu f$	L_2 = 37 μh	f_0 = 20-30 mc
R_3 = 2,200 ohms	C_3, C_4, C_5 = 6-80 $\mu\mu f$	L_3, L_4, L_5 = 0.8 μf	
R_4 = 150 ohms	C_6, C_7, C_8, C_9, C_{10} = .01 μf	L_6, L_7 = •.12 μh	
R_5 = 68,000 ohms		L — 0.17 μh	

Figure 1-221. Schematic diagram 20—30-mc spectrum generator with filter-amplifier stage added for greater adjacent channel selectivity

Transistor Crystal-Feedback Oscillators

POINT-CONTACT TRANSISTOR CRYSTAL-FEEDBACK OSCILLATOR

1-539. The circuit parameters indicated are those of an experimental low-frequency oscillator that was investigated at Bell Telephone Laboratories as a possible future replacement of the present vacuum-tube circuits now used to control telephone carrier frequencies. The information given here is based upon a discussion of the experimental oscillator by R. S. Caruthers.

1-540. A point-contact transistor of the 1729 type was used in the experiments. Superior performance characteristics can be obtained with the more recently developed junction transistor. The extreme simplicity of the basic circuit is possible because the r-f emitter current, I_e, is in phase with the r-f collector current, I_c. All that is necessary for the oscillations to be maintained is that a series resonant circuit, which in this case is the series-mode crystal, feed back a sufficient current to supply I_e. The total feed-back current, however, must be somewhat greater than I_e. Amplitude equilibrium is reached when

$$\alpha = 1 + \frac{R}{R_L} + \frac{R}{r_c} \qquad \text{1—540 (1)}$$

where α is the effective r-f current amplification factor of the transistor, equal to I_c/I_e, R is the crystal resistance, R_L is the load resistance, and r_c is a resistance parameter of the transistor collector circuit, somewhat the analogue of R_p of a vacuum tube. For the point-contact transistor, r_c is generally between 15,000 and 20,000 ohms, so that the term R/r_c is usually quite small compared with the other terms in the equation. For oscillations to build up, the initial value of α must be greater than that given by equation (1). The larger the difference between the initial α and the equilibrium α, the greater is the final activity. In figure 1-223(A), it can be seen that the values of α are not large, so that R_L cannot be made much smaller than the maximum permissible R of the crystal unit if equation (1) is to hold for all crystal units of a given type. Limiting occurs when the current peaks of I_e extend into the low-amplification regions indicated in figure 1-223(A). The a-c load line in figure 1-223(B) indicates the approximate operating region during oscillations. The point Q, which is at the middle of the operating range is predetermined experimentally by adjustments of the d-c supply voltages and the series dropping resistances. Because of the small margin of excess gain in the transistor oscillator, the variations in collector characteristics from one transistor to another result in large percentage variations in oscillator activity. Note the large differences in collector current for the same value of

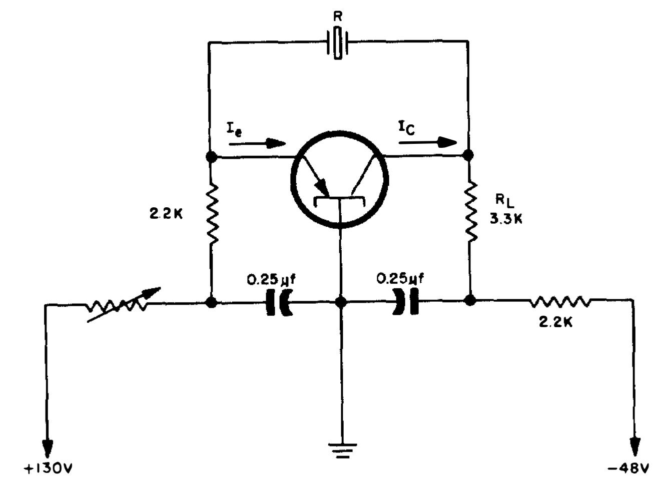


Figure 1-222. Basic circuit for crystal-controlled transistor oscillator

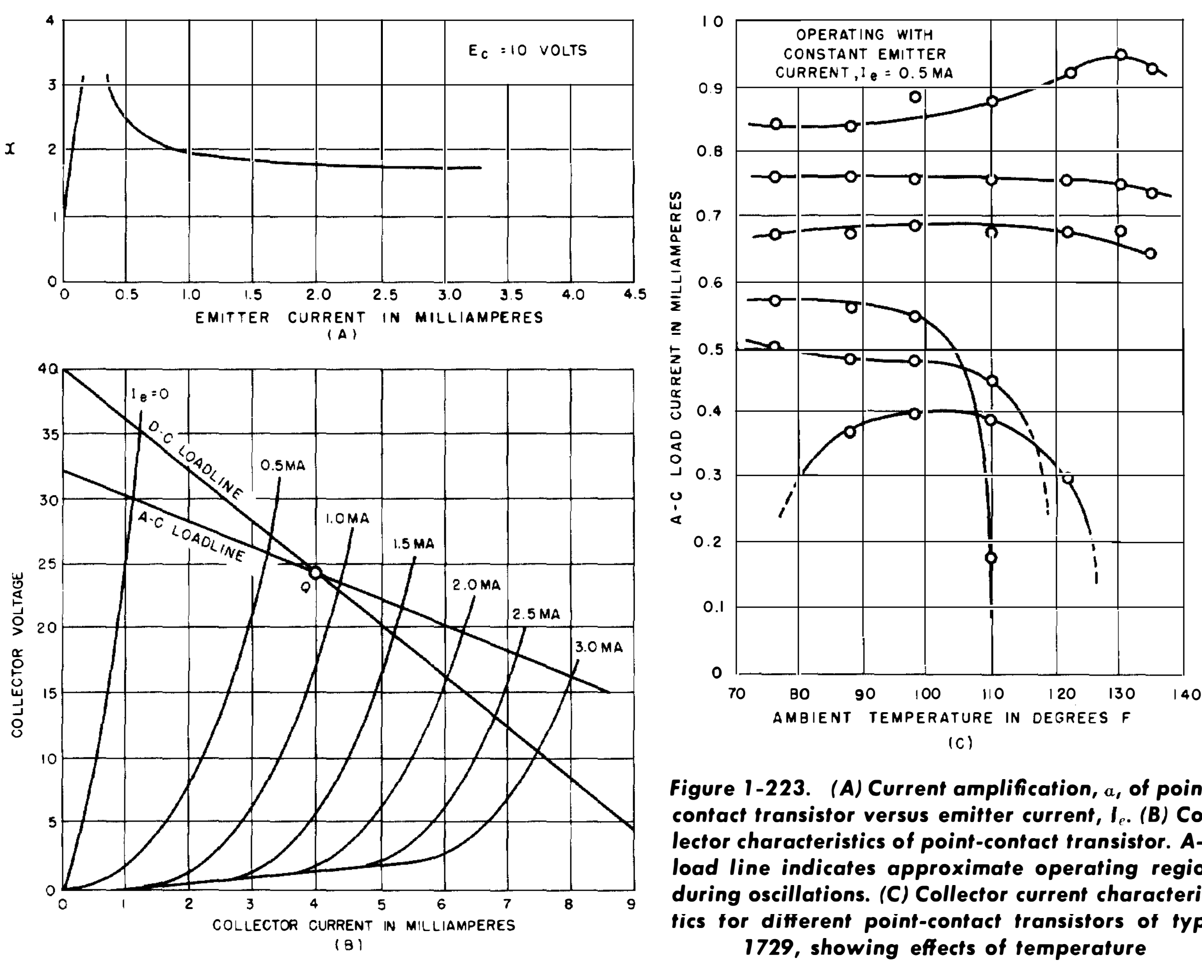


Figure 1-223. (A) Current amplification, a, of point-contact transistor versus emitter current, I_e. (B) Collector characteristics of point-contact transistor. A-C load line indicates approximate operating region during oscillations. (C) Collector current characteristics for different point-contact transistors of type 1729, showing effects of temperature

I_e when different transistors are used, as shown in figure 1-223(C). The temperature effects also require consideration. An increase in temperature causes R_c to become lower. In the case of "low-activity" transistors, this effect can easily be sufficient to stop oscillations.

1-541. The frequency stability and sine-wave output of the transistor oscillator can be comparable to that of vacuum-tube oscillators. The oscillator in figure 1-222 was able to supply as much as 30 mw to the load at a frequency of 184 kc. It was found that 0.8 watt less power was required to operate a transistor oscillator and amplifier than is required for an equivalent vacuum-tube circuit, and at least a 50-per cent greater saving should be possible. The consequent reduction in compartment heating, as well as in requirements of weight, space, etc, can be quite advantageous in small, compact units where several power-dissipating elements are packed closely together.

JUNCTION TRANSISTOR CRYSTAL-FEEDBACK OSCILLATOR

1-542. The basic collector-to-emitter crystal feedback circuit shown in figure 1-222 requires a current amplification greater than unity, and therefore is restricted to transistors of the point-contact type and to frequencies generally under 500 kc. When a crystal-feedback oscillator is desired using a junction transistor, or using a point-contact transistor at a frequency too high for alpha to be greater than unity, modifications must be introduced in order to obtain the required current amplification. Perhaps the simplest solution is to employ a voltage-step-down (current-step-up) transformer. However, to simplify the discussion we shall treat the transformer-coupled transistor oscillator in a separate category. A more direct solution, and one that affords a more dependable oscillator insofar as the elimination of free-running oscillations is concerned, is to exploit the current amplifying characteristic of the parallel-tuned circuit. One branch of a tuned tank with a series-mode crystal inserted can be used as a feedback circuit. One such circuit, where the collector output operates into the tuned tank, is shown in figure 1-224. This circuit was designed and tested during a Signal Corps research project* directed by B. J. Dasher at the Georgia Institute of Technology. A simplified schematic illustrating the necessary loop gain and loop phase conditions for oscillation is shown in figure 1-225. Assuming that the r-f emitter-to-base input impedance of the transistor is small compared with the external 10-kilohm shunting resistance, it can be seen that the feedback current ($I_1\angle\theta$) through the crystal and the tank tuning capacitor must be approximately equal to I_e, the r-f emitter current, whose phase is taken as the zero reference. The exact phase conditions depend upon the alpha characteristics of the transistor.

1-543. Figure 1-226 shows the alpha characteristics versus frequency of a typical CK-720 junction transistor as measured by Dasher et al at the Georgia Institute of Technology. Note the large lag (ϕ) in the phase of I_c with respect to I_e as the frequency is increased. Alpha is the current gain under external collector-to-base short-circuit conditions. Its value is dependent upon the frequency but is otherwise independent of the particular circuit external to the transistor. The collector circuit shown in figure 1-225 is, of course, not a short circuit, but is assumed to have an impedance very small compared with the internal collector-to-base impedance of the transistor, so that to a first approximation the phase of the operating I_c is equal to the phase, ϕ, of the short-circuit alpha. The phase lag, ϕ, is due to the transit time of the transistor current carriers. It can be seen that if $I_1 \angle\theta$ of the feedback arm of the collector tank in figure 1-225 is to have the same phase as $I_e \angle\theta^\circ$ ($\theta = 0$), how nearly the tank is to be tuned to parallel resonance and whether the tuning is above or below resonance depend upon the phase of ϕ, and that this in turn depends upon the frequency. At frequencies in the neighborhood of 400 kc, ϕ has a value of approximately 80°. If it can be assumed that the reactance of the capacitor in the feedback arm is sufficient to cause the feedback current to lead the tank voltage by 80°, then $\theta \approx 0$ when the tank is operated at resonance. At small values of ϕ (at low frequencies), the tank circuit tuning must deviate considerably from the parallel resonant condition in order for θ to equal zero. At very large values of ϕ the loop phase relations approach the 180-degree inversion characteristic of vacuum-tube oscillators, with the circuit in figure 1-224 assuming certain similarities to a Pierce circuit. Since the magnitude of alpha becomes very small as ϕ becomes large, it may be necessary to introduce additional phase-shifting elements in order to ensure that the tank is operated sufficiently close to resonance. (Actually, the practical solution would be to employ a transistor

* *Transistor Oscillators of Extended Frequency Range*, Quarterly Report No. 5, by B. J. Dasher, D. L. Finn, S. N. Witt, Jr., W. B. Warren, Jr., and T. N. Lowry of the Engineering Experiment Station, Georgia Institute of Technology, Atlanta, Georgia—Department of the Army Contract No. DA-36-039-sc-42712.

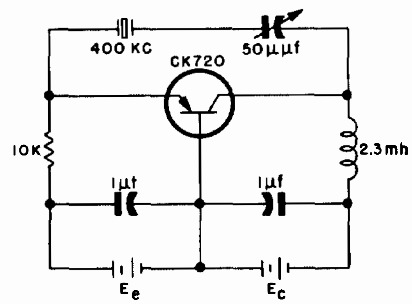


Figure 1-224. Junction transistor crystal-feedback oscillator

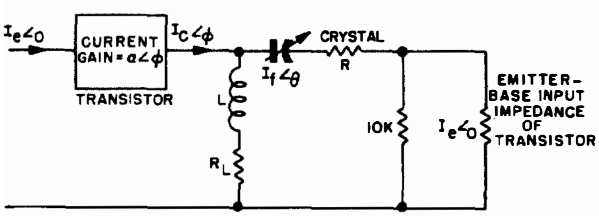


Figure 1-225. Simplified equivalent circuit of junction transistor crystal-feedback oscillator, showing tuned tank in collector circuit for obtaining desired current amplification and loop phase shift for feedback to emitter input. For stable oscillations to be maintained

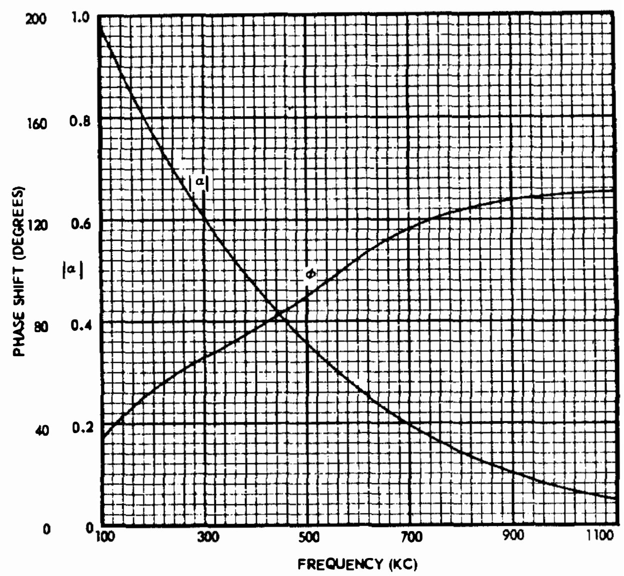


Figure 1-226. Typical alpha characteristics versus frequency of a type CK-720 junction transistor

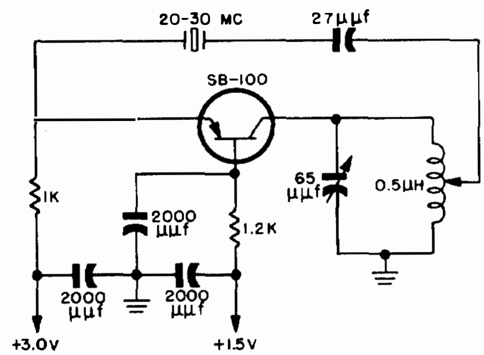


Figure 1-227. High-frequency transistor oscillator

especially fabricated for use at higher frequencies, as is discussed in the following paragraph.) The value of L in figure 1-224 is purposely made large so that the tank Q is sufficient to permit the necessary current amplification over as wide a tuning range as is possible. With the crystal shorted out, the circuit was found to oscillate in a free-running state at frequencies as high as 2 mc, which is several times the alpha cutoff frequency. (The alpha cut off frequency is that frequency at which the short-circuit current gain is 3 db below the d-c value of alpha.) With a crystal inserted, the upper frequency limit depends upon the crystal's effective resonance resistance. But the upper dependable frequency when employing Military Standard crystal units having element C or D characteristics appears to lie between 300 and 400 kc. Frequency Control Branch engineers of the Fort Monmouth Signal Corps Engineering Laboratories do not recommend this type of oscillator for general use because the operating stability is critically dependent upon the stability of alpha, whose phase and magnitude can be quite difficult to maintain constant under variations of temperature and voltage. Nevertheless, the progressive development of h-f and v-h-f transistors of ever closer tolerances are rapidly increasing the frequency range over which junction transistor oscillators of all types can be operated below alpha cutoff—in which region unstable transit time effects are less pronounced. For example, a modification of the circuit shown in figure 1-227 (a type that employs a parallel-tuned circuit for amplification of the feedback current), without crystal control, is already in commercial use. For a modified version of the oscillator in figure 1-224 designed to provide maximum frequency stability, see paragraph 1-555.

H-F TRANSISTOR OSCILLATORS

1-544. The comparatively long transit time of the transistor has been a great handicap in applying the semiconductor amplifier in h-f and v-h-f circuits. Although it is quite possible to demonstrate that transistors can operate satisfactorily in h-f crystal oscillators, it is quite difficult to guarantee that such circuits will operate dependably when the demonstration transistors and crystal units are replaced by others of the same nominal characteristics, unless close tolerances are assured. One of the few types of transistors for h-f and v-h-f application that is commercially available at the time of this writing is the Philco Surface Barrier transistor, SB-100, which is basically a special type of pnp junction transistor designed for low-power, wide-band, h-f applications. Its unique fabrication process permits an above-average degree of uniformity in its low-power, h-f operating characteristics. The maximum dependable oscillation frequency of the SB-100 is rated at 30 mc. This is the highest frequency at which any randomly selected transistor of this type can be expected to show a power gain of at least unity. The highest frequency at which the *average* SB-100 transistor can be expected to show a power gain of unity is 45 mc. With selected SB-100 transistors, oscillators at limit frequencies of 70 mc are possible. If these transistors are to be employed as amplifiers in oscillator circuits at frequencies above 20 mc, the design engineer should specify closer tolerances than those described for randomly selected units. Figure 1-227 shows an oscillator circuit designed for operation in the 20—30-mc range. This oscillator will operate as a free-running circuit at its tuned frequency if the crystal is shorted across. The ratio of the useful oscillator output to the power supplied from the SB-100 collector voltage supply is given by the empirical equation

$$\text{collector efficiency} = 0.8 \log_{10} \frac{f_{max}}{f} \ \% \qquad 1\text{-}544\ (1)$$

where f_{max} is the highest frequency at which the oscillator in figure 1-227 will oscillate (the frequency at which the circuit power gain is unity at starting amplitudes) and f is the operating frequency. Equation (1) assumes that the losses in the crystal unit are negligible; otherwise these losses should be interpreted as part of the "useful" oscillator output.

Negative-Resistance Transistor Oscillators

1-545. Although the driving element of any electronic oscillator can be described as a negative-resistance circuit, it is not customary to classify the oscillator, itself, as being of the negative-resistance type unless the negative resistance is an inherent d-c, as well as an a-c, characteristic of the driving element. Among negative-resistance vacuum-tube oscillators, the transitron circuit is an example. Among transistor oscillators, the series-tuned-emitter circuit, such as that shown in figure 1-228, is an example.

1-546. A negative-resistance characteristic can be obtained between any two terminals of a point-contact transistor by allowing the amplified current to flow through a feedback resistance of such magnitude that the feedback voltage produced is more than sufficient to compensate for any change in the input voltage. Alpha must be greater than one, so that the negative-resistance circuit is not applicable for a single-stage junction type of transistor, nor, for that matter, for a point-contact transistor except at the lower frequencies. The negative-resistance characteristics of the emitter input in a common-base circuit, such as that shown in figure 1-228, is generally superior to a tuned-collector negative-resistance

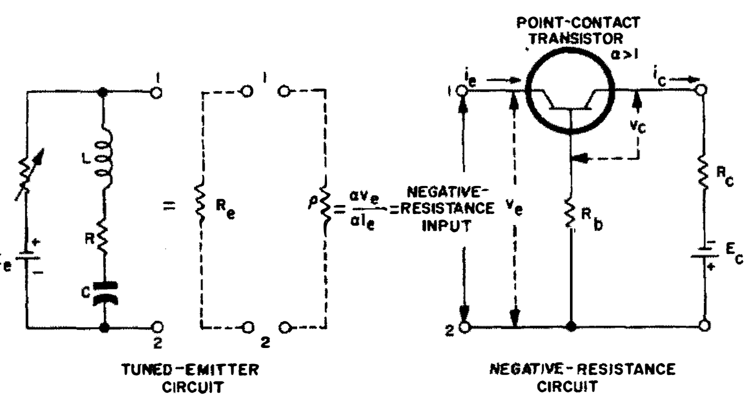


Figure 1-228. Tuned-emitter negative-resistance circuit. Essential features are that alpha be greater than unity, the insertion of an external resistance, R_b, in the common-base circuit to increase the collector-to-emitter energy feedback; and for oscillations to build up, the effective positive resistance, R_e, of the tuned-emitter circuit must be less, numerically, than the negative resistance, at the emitter input

mode of operation insofar as oscillator stability is concerned. Figure 1-229 shows a typical emitter-to-base characteristic curve for the circuit in figure 1-228, where the emitter voltage is plotted as a function of emitter current. The slope of this curve equals the dynamic resistance of the input. (By "dynamic resistance" we mean the instantaneous resistance, dv_e/di_e, for an infinitesimal change in voltage and current at any given point on the curve, as opposed to the "static resistance," equal to v_e/i_e, the total voltage divided by the total current at any given point on the curve.) The negative slope between points A and B represents the negative-resistance zone of the emitter circuit. If the tuned circuit shown in figure 1-228 is connected to the emitter input and the rheostat is adjusted so that the transistor is biased to operate at point 0 on the v_e-i_e curve, oscillations at the resonance frequency of the emitter LC circuit will build up provided that the magnitude of the positive resistance, R_e, of the external circuit connected between terminals 1 and 2 is less than that of the negative resistance presented by the transistor at the same terminals. As oscillations about point 0 build up, the positive peaks of I_e swing toward the positive-going slope region at the bend of the curve, so that the effective negative dynamic resistance decreases, and amplitude limiting is achieved. Constant-amplitude oscillations are reached when the effective dynamic impedance looking into the emitter-base terminals is exactly equal and opposite in sign to the impedance of the external emitter-base circuit as faced by the input terminals, 1 and 2. If we assume that the bias-control resistance can be neglected and that the current and voltage at the fundamental frequency are approximately in phase, then we can say that the condition for stable oscillations is that

$$R + \rho \approx R_e + \rho = 0 \qquad 1\text{-}546\ (1)$$

1-547. Figure 1-230 shows the values obtained for ρ for a particular 1768 type transistor when the externally connected resistors R_b and R_c are 800 ohms and 2700 ohms, respectively. The emitter bias current is constant. The horizontal scale represents the rms a-c value of the emitter current. Since the magnitudes of ρ and R_e are equal after oscillation equilibrium is reached, it can be seen that the curves in figure 1-230 can be used to predict the limiting amplitude of oscillation as the external emitter-to-base resistance R_e ($= -\rho$) is varied. The curves indicate that if a low-frequency series-mode crystal is connected across the transistor input, oscillations can be expected to build up as long as the resonance resistance of the crystal unit is not greater than a few hundred ohms. Not many types of low-frequency Military Standard quartz crystal units have such small values of resonance resistance. Exceptions are the precision G-element crystal units CR-39/U and

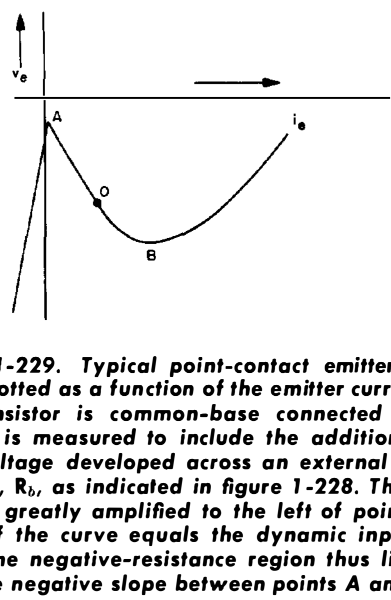
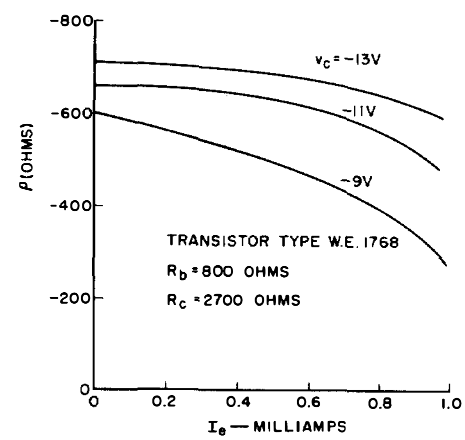


Figure 1-229. Typical point-contact emitter voltage curve plotted as a function of the emitter current when the transistor is common-base connected and the voltage is measured to include the additional feedback voltage developed across an external base resistance, R_b, as indicated in figure 1-228. The current scale is greatly amplified to the left of point A. The slope of the curve equals the dynamic input resistance. The negative-resistance region thus lies along the negative slope between points A and B



Note: The Western Electric type transistor shown above and the one following are experimental prototypes whose production has been discontinued. Similar electrical characteristics can be obtained by using Western Electric transistor types 2N21, 2N110, and GA-52837, which are commercially available, but solely for the use of the U.S. Armed Services and their contractors.

Figure 1-230. Negative input resistance of the circuit shown in figure 1-228, plotted as a function of the a-c component of the emitter current for three representative values of collector-to-base d-c voltage. The emitter is biased at approximately point 0 as shown in figure 1-229. Measurements were made by Dasher, Finn, and Jones at the Georgia Institute of Technology

CR-40/U, which have maximum effective resonance resistances of 150 ohms in the 160—250-kc range and of 600 ohms in the 250—330-kc range. The resistance deviations of these crystal units are rated at not greater than 33 per cent of the maximum permissible values—that is, maximum deviations of 50 ohms and 200 ohms, respectively, for the 160—250-kc range and the 250—330-kc range. No data is available concerning their performance in negative-resistance transistor circuits, but from the curves of figure 1-230 wide variations in the amplitude must be expected when replacing one crystal unit with another unless special steps are taken to control the gain of the transistor circuit. Whereas a 33 per cent change in crystal resistance can generally be considered negligible in the average vacuum-tube oscillator, this much resistance deviation in the negative-resistance tuned-emitter oscillator can cause a proportionally much larger percentage change in the output amplitude.

1-548. Figure 1-231 is a schematic diagram of the equivalent a-c circuit of the tuned-emitter circuit. The parameters r_b, r_c, and r_e represent the impedances of the equivalent T network of the transistor at low frequencies and small amplitudes, conditions under which the equivalent impedances can be assumed to be approximately linear and purely resistive. The amplifying properties are represented by the constant-current generator, αI_e. In analyzing this circuit two equations suffice for defining the state of operation at oscillation equilibrium. Following the analysis developed by B. J. Dasher et al in *Transistor Oscillators of Extended Frequency Range*, we find that one equation can equate the over-all voltage drop around the emitter-base loop to zero; the second equation can equate the over-all voltage drop around the collector-base loop to zero. On combining these equations and separating the real and complex parts, two additional equations are obtained. The complex part gives the loop phase requirement, which serves to determine the frequency:

$$f = \frac{1}{2\pi\sqrt{LC}} \qquad \text{1-548 (1)}$$

(That the complex part would reduce to the form shown in equation (1) could virtually have been predicted simply from a qualitative inspection of the circuit.) The real part gives the loop gain requirements and can be expressed as an equilibrium equation for alpha:

$$\alpha = \left(\frac{r_b + R_b + r_e + R_e}{r_b + R_b}\right)\left(\frac{r_c + R_c}{r_c}\right) + \frac{r_e + R_e}{r_c} \qquad \text{1-548 (2)}$$

If we assume that r_c is much greater than all of the other resistance parameters, equation (2) can be reduced to the approximate equation

$$\alpha \approx \text{but} > \frac{r_b + R_b + r_e + R_e}{r_b + R_b} \qquad \text{1-548 (3)}$$

Note that equation (3) requires that the transistor current gain at equilibrium be slightly greater than the ratio of the total resistance of the emitter-base loop to the total feedback resistance of the base as illustrated in figure 1-231. The reader should not confuse r_e, r_b, or any of the other positive resistance values in figure 1-231 with the effective negative resistance of the input circuit. The equivalent effect of the negative-resistance characteristic is provided by the current generator.

1-549. If desired, the operating voltages of the base-feedback point-contact transistor can be so adjusted that the negative resistance appears in the collector rather than in the emitter circuit, and oscillations can be controlled by connecting a series-mode crystal in the collector circuit. The equivalent tuned-collector circuit is shown in figure 1-232(A). A sample collector-controlled circuit which was tested at 50 kc is shown in figure 1-232(B)*. This circuit, as reported by B. J. Dasher, exhibited a negative resistance sufficient to drive a 40,000-ohm crystal unit. Al-

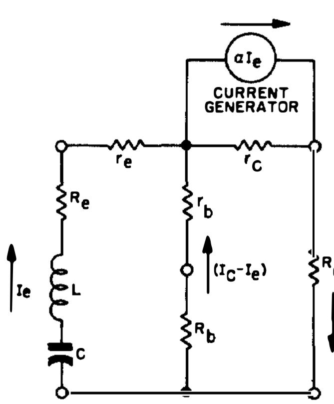


Figure 1-231. Equivalent a-c circuit of tuned-emitter negative-resistance oscillator. For crystal control, L, C, and R_e can be interpreted as being the series-arm parameters of a crystal unit

* To avoid confusing free-running oscillators with their crystal-controlled counterparts, it has been suggested by Frequency Control Branch engineers at SCEL that when a crystal unit replaces a "tuned-emitter" or a "tuned-collector" LC circuit, the oscillator be designated, respectively, as being "emitter-controlled" or "collector-controlled."

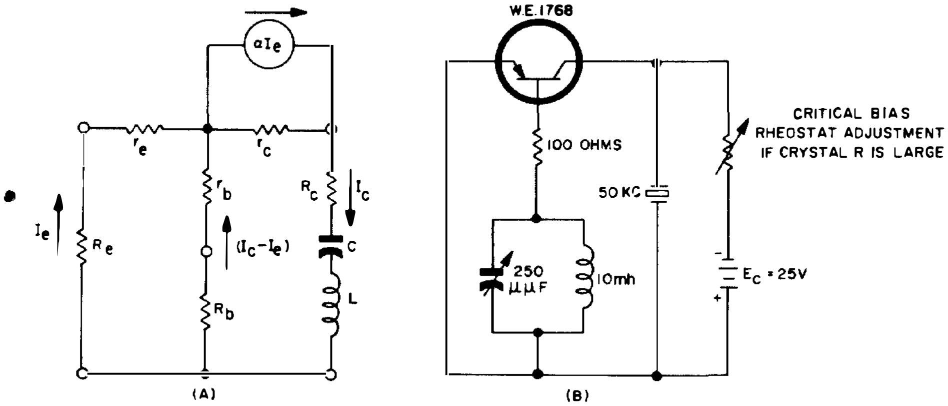


Figure 1-232. (A) Equivalent a-c circuit of tuned-collector negative-resistance oscillator. For crystal control, L, C, and R_c can be interpreted as being the series-arm parameters of a crystal unit; (B) Crystal-controlled tuned-collector negative-resistance oscillator. The parallel tuned tank is inserted in the base circuit in lieu of a resistor in order to obtain negative resistance values in excess of 40,000 ohms

though negative resistance values of 50 kilohms are possible at d-c potentials, these values drop rapidly with increasing frequency, and may be only a few hundred ohms at 100 kc. The tuned-collector (and collector-controlled) circuit is governed by exactly the same loop equations that apply in the case of the tuned-emitter (and emitter-controlled) circuit, which are given in the preceding paragraph (equations 1-548 (1) and (2)). However, the degradation of the crystal Q in the collector-controlled circuit is generally greater than in the emitter-controlled circuit, so that the latter permits the higher frequency stability. Because the tolerances required for good stability are difficult to obtain with point-contact transistors neither circuit, to Signal Corps knowledge, has been developed to the point that it could be recommended for general use.

Transistor Oscillators Using Parallel-Mode or High-Impedance Series-Mode Crystal Units

1-550. Since transistor negative-resistance oscillators require values of alpha greater than unity, they are generally limited to frequencies less than 500 kc. The negative-resistance circuits previously discussed place the crystal directly across the negative-resistance terminals. Yet at low operating frequencies it is difficult to fabricate crystal units having resonance resistances of magnitudes small enough to permit oscillations to start. For this reason, it may be more practical to connect the crystal unit across a high-impedance pair of point-contact transistor terminals and to insert a series-resonant LC circuit, or other low impedance, across the low-impedance negative-resistance terminals. Such a circuit will not exhibit (to the crystal circuit) the negative-resistance characteristic under d-c conditions—only in the neighborhood of the crystal frequency—and so is not conventionally classified as a negative-resistance oscillator. However, the operating principle is the same. Six oscillators of the high-impedance type are shown in figure 1-233. All the circuits shown (but not including the crystal load capacitances indicated by dotted lines) are the inventions of Everett Eberhard and Richard Endres of RCA, U. S. Patent 2,570,-436. A common feature of these oscillators is that the bias currents are so selected that the electrode to which the piezoelectric crystal is connected is made to present a high dynamic impedance, to match the high impedance expected of the crystal unit. A resonant LC circuit is connected to another electrode of the transistor, the tuned LC resonant frequency being approximately equal to the crystal frequency. Each circuit is designed to provide positive feedback sufficient to maintain oscillations. A high impedance can be obtained at the crystal-connected electrode of the transistor by adjustment of the bias current at any one of the three electrodes. Figure 1-233 (G) shows typical curves of the resistance that would be faced by the crystal unit between each terminal and ground as a function of the collector bias current, i_c. In circuits (A), (B), (C), and (D) of figure 1-233, where either the base or the collector impedance must match that of the crystal unit, the transistor is operated near point a on the i_c scale. In circuits (E) and (F) of figure 1-233, where the emitter input

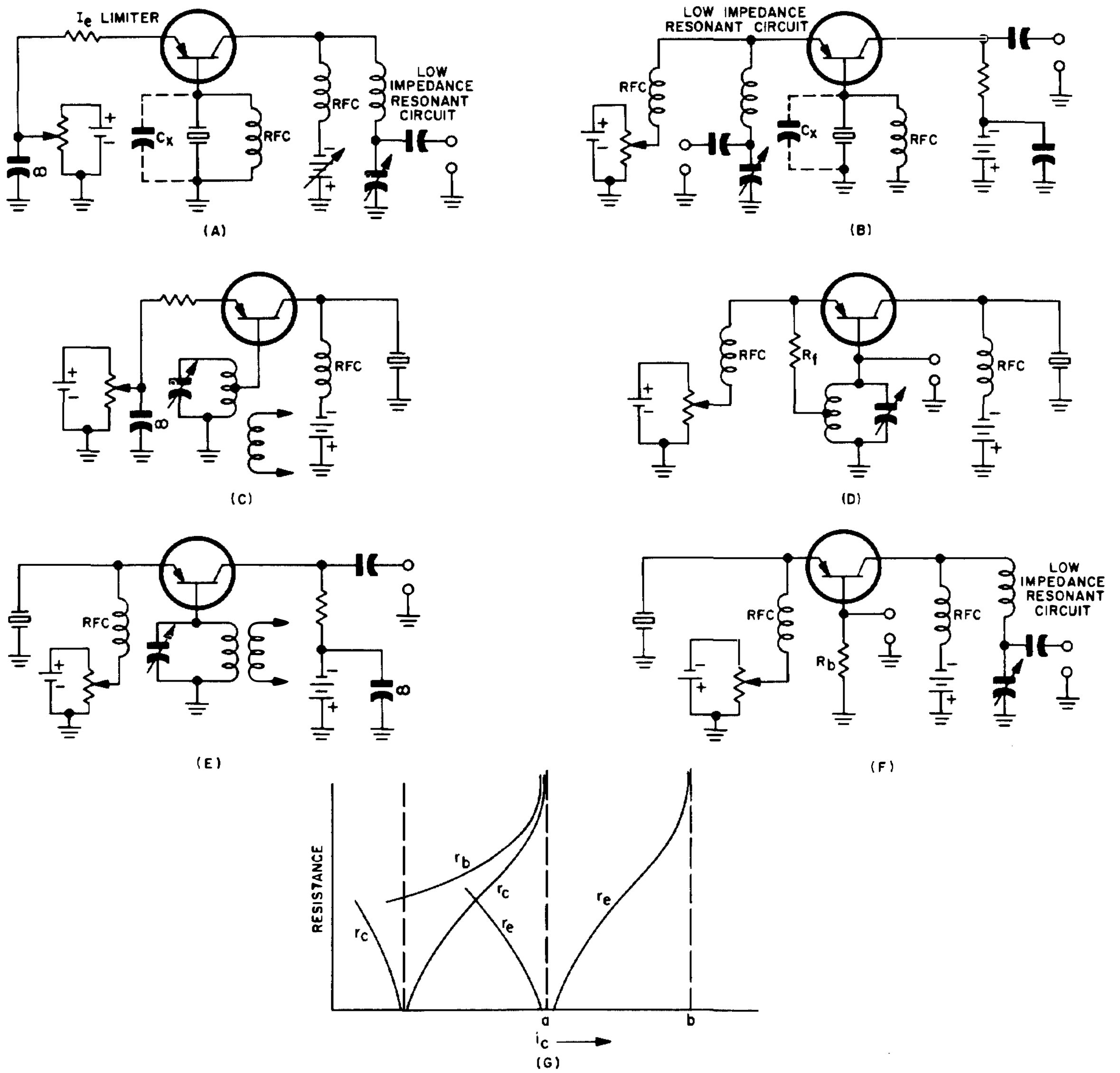


Figure 1-233. (A) to (F) Transistor oscillator circuits designed for high-impedance crystal control; (G) Variation of equivalent network resistances of point-contact transistor as a function of collector current. Point A is the high-impedance region of the base and collector electrodes, and point B that of the emitter electrode, r_b, r_c, and r_e are not drawn to the same scale

impedance must be high, i_c is adjusted to a value near point b in figure 1-233 (G). In all the circuits of figure 1-233, the collector is operated with a relatively large reverse bias, and the emitter with a relatively small forward bias.

1-551. The exact operating point of the crystal unit in the Eberhard-Endres oscillators is not specified in the patent claims. The inference is that the crystal unit is operated near antiresonance, since it is assumed to be a high-impedance element. However, the resonance resistance of v-l-f crystal units may well be greater than 100,000 ohms, and even the resonance resistance of an m-f crystal may be several kilohms; hence, it appears that at the low frequencies at which the point-contact transistor oscillators normally operate, the crystal units in the circuits of figure 1-233 are not restricted to a particular mode of resonance, but operate near series resonance as long as they show a sufficiently high impedance. If it can be assumed that the crystal unit and each of the tuned circuits are operating approximately as pure resistances, the gain requirements for sustained oscillations, except in circuit (D) which has an additional feedback element, are the same as those given by equation 1-548 (2) for the negative-resistance oscillators. This can be seen intuitively by noting that all the circuits obtain their necessary positive feedback by virtue of a high-impedance common-base circuit. In circuits (A) and (B) of figure 1-233, the impedance of the crystal unit can be taken as the approximate value of R_b in equation 1-548 (2). Since R_b can assume a resonance or an antiresonance value, either of which could conceivably satisfy the loop gain equation, it would appear that these two circuits might have tendencies to jump frequencies. No experimental evidence is available, but very possibly the dependability of such oscillators could be improved by employing parallel-mode crystal units directly shunted by suitable load capacitances, as indicated by the dotted-line connections. In practice, a proper selection of the reactive elements in each circuit, such as the r-f choke across the crystal unit, may well be entirely sufficient to ensure that a high-impedance crystal will operate at resonance rather than at antiresonance. In circuits (C) to (F) of figure 1-233 maximum feedback occurs when the crystal resistance is a minimum, so that only the series-resonance mode of the crystal unit need be considered. In other words, the loop gain conditions could not be satisfied with practical values of alpha if the extremely high antiresonance impedances of a low-frequency crystal unit were substituted for the values of R_c or R_e in equation 1-548 (2). Nevertheless, it is conceivable that a parallel-mode crystal unit, shunted by its rated load capacitance, could also be applicable in these circuits if desired. *For a given set of operating voltages and circuit parameters, however, only one mode of crystal operation can sustain stable oscillations. Whether the stable mode is series or parallel depends upon whether the limiting action is current- or voltage-controlled, respectively. (See paragraph 1-590.)* An output from any of the circuits in figure 1-233 can be obtained at any of the transistor electrodes. Where maximum sine-wave purity in the output is desired, the circuit parameters should be selected so that the equilibrium value of alpha is only slightly less than the starting value of alpha, so that the amplitude remains as small as possible. Also, the output harmonic content is greatly reduced if the output is taken from a parallel-tuned circuit having a small L/C ratio, as is done in circuits (C), (D), and (E). In circuit (D), additional feedback is obtained by the insertion of R_f, which may have any value from zero on up. For optimum operation the d-c resistance of the r-f choke in the emitter circuit should be less than R_f. Circuit (D) can be considered a transistor equivalent of a crystal-controlled Hartley oscillator by viewing the emitter, collector, and base as the cathode, plate, and grid, respectively, of the vacuum tube in the Hartley circuit. Note in circuit (F) that the emitter bias battery is shown with its polarity in the reverse direction. This does not mean that the emitter is operated with reverse bias, but simply that the large forward bias provided by the d-c voltage drop across R_b must be partially cancelled to obtain the proper operating bias.

Transformer-Coupled Transistor Oscillators

1-552. Figure 1-234 shows schematic diagrams for three experimental transformer-coupled transistor oscillators designed and tested by Dasher, Jones, and Witt at the Georgia Institute of Technology. The average frequency deviation of each of the three circuits for variations in the power supply is given below in table 1-552 (1).

1-553. An elementary transformer-coupled circuit is shown in figure 1-234 (A). Circuit (B) is a modified version of (A) that was found empirically to provide better frequency stability during changes in the applied voltage. The turns ratio of the transformers in circuits (A) and (B) are not specified, but presumably the ratios are 1:1. The transistor used is a point-contact type with an alpha greater than unity at the operating frequency of 50 kc. Circuit (C) is a 400-kc oscillator

Circuit	*Frequency Deviation (Parts Per Million Per 1 Per Cent Change in Voltage)*	*Bias Variable*	*Bias Constant*
(A)	1.2	E_c	i_e
(A)	0.7	E_e	i_c
(B)	0.25 to 0.8	E_c	
(C)	0.0013*	E_c	

Table 1-552 (1). Frequency stability data for circuits shown in figure 1-234.

* This small frequency deviation is obtainable only when R_b is adjusted for optimum bias compensation. See paragraph 1-555.

that can be adjusted to provide excellent frequency stability against changes in the supply voltage. Also, by the proper adjustment of the variable capacitor C, the current through the shunt capacitance of the crystal unit can be effectively cancelled insofar as the transistor input is concerned. The circuit is effectively a capacitance-bridge oscillator in which all the transformer feedback signals are annulled in the input circuit except those that pass through the series arm of the crystal unit, which of course is effectively an open circuit except at resonance. The principal adjustment for minimizing the deviation in frequency during fluctuations in the voltage supply is that of the base resistance, R_b. For each crystal unit and transistor a particular value of R_b will provide maximum frequency stability. The reason for this appears to be due to the compensating influences of simultaneous changes in emitter and collector bias currents. See paragraph 1-555 for a more detailed discussion of this bias compensation effect.

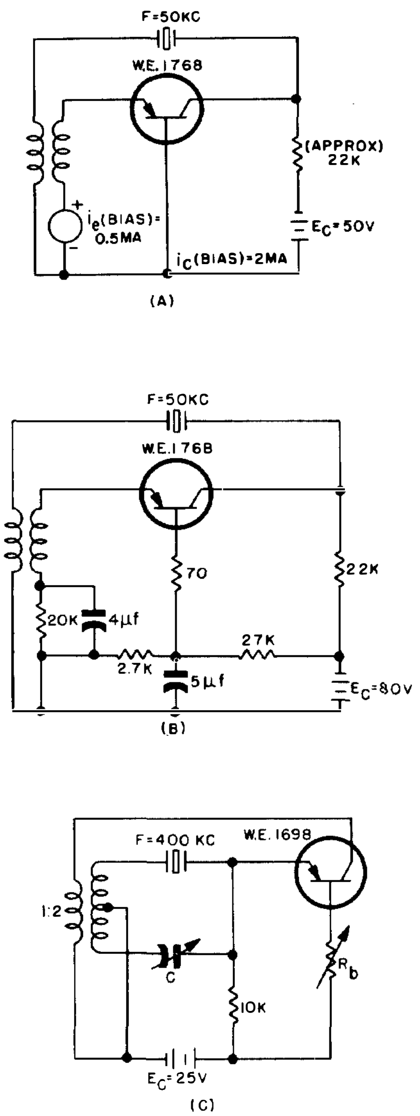


Figure 1-234. Transformer-coupled transistor oscillators

1-554. Each of the circuits shown in figure 1-234 exhibits a fair, but not exceptional, frequency stability under changes in temperature, with circuits (B) and (C) being somewhat the superior in this respect. The transformer-coupled transistor oscillators are generally less frequency stable under changes in temperature than are the resistive transistor circuits. An additional disadvantage is that the transformer-coupled circuit has greater-than-average tendencies to break into free-running oscillations, particularly if the shunt capacitance of the crystal unit is large. On the other hand, an important advantage of the transformer coupling is the increase possible in feedback gain. This makes it possible to operate at lower voltages, at higher frequencies, with junction transistors, and generally with smaller values of alpha than otherwise.

Frequency Stabilization In Transistor Oscillators By Bias Compensation

1-555. Figure 1-235 shows the frequency characteristics of the oscillator in figure 1-224 when the collector and emitter voltages are varied. Note that where an increase in collector voltage causes an increase in frequency, an increase in the emitter voltage causes a decrease in frequency. This property of the circuit suggested to its designers that by a proper choice of emitter and collector voltages the circuit could be frequency-stabilized against fluctuations in the voltage source. From figure 1-235 it can be seen that if the emitter voltage is to vary linearly with the collector voltage, the voltage-compensating circuit must operate in a region where the intersections of the emitter voltage curves with the horizontal constant-frequency lines are evenly spaced. A circuit designed for this mode of operation is shown in figure 1-236. Figure 1-237 shows the frequency-voltage characteristics of the compensated circuit for two values of the resistance R. Note that this mode of operation is a transistor analogue of the "class D" mode of operation described for vacuum-tube oscillators in paragraphs 1-298 and 1-342.

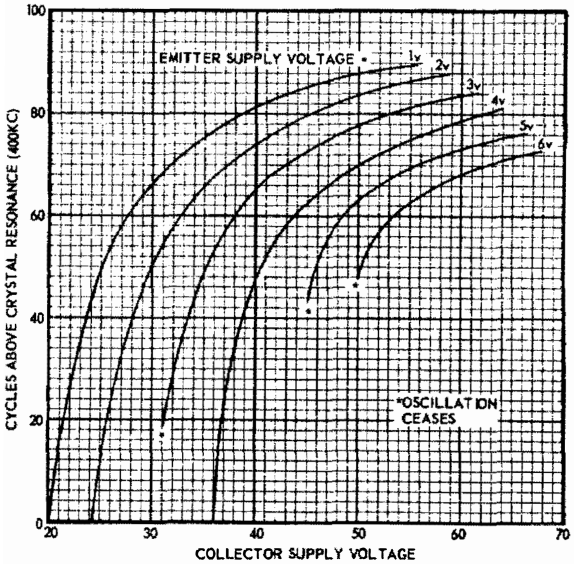


Figure 1-235. Deviation of frequency with the collector supply voltage for various values of emitter supply voltage. Test oscillator circuit is the same as that shown in figure 1-224

1-556. The adjustment of the frequency-stabilized oscillator depends upon the resistance of the crystal unit. A circuit adjusted for maximum stability with one crystal unit may be relatively unstable when a replacement crystal of the same type and nominal frequency, but of different resistance, is inserted. This might be expected, since the resistance of the crystal affects the amplitude of oscillations as well as the phase-shifting Q of the feedback circuit. A complete theoretical analysis of the relations among the stability parameters has not been attempted, and no doubt would prove quite complex. But one of the more important factors affecting the frequency stability is the harmonic content of the oscillations. Any change in the wave distortion, such as would occur with a change in amplitude due to a change in crystal resistance, would be sufficient to shift the fundamental frequency (see paragraph 1-596) to a point where the circuit may no longer be self-compensating.

1-557. Figure 1-238 shows a simple feedback oscillator tested by Dasher and Witt at the Georgia Institute of Technology. The values of the parameters were selected to provide maximum frequency stability for a particular transistor and crystal unit. Figure 1-239 shows how the frequency of this oscillator changed with a 20-volt change in supply voltage. That a number of crystal units having approximately the same resistance caused different effects in the frequency

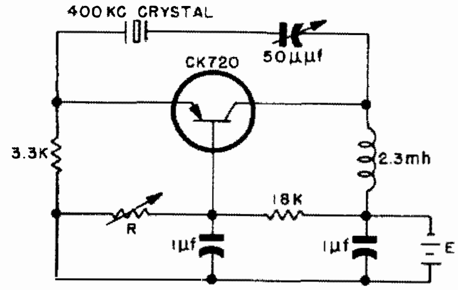


Figure 1-236. Frequency-stabilized circuit for single battery supply

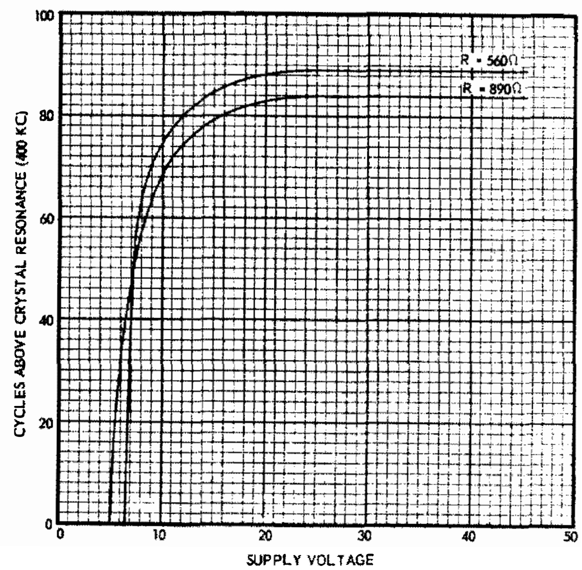


Figure 1-237. Deviation of frequency with changes in supply voltage for two values of R in circuit shown in figure 1-236

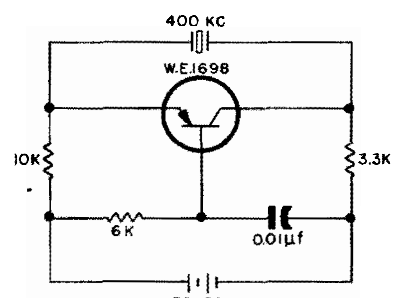


Figure 1-238. Crystal-feedback oscillator in which the resistance parameters have been selected empirically to provide maximum frequency stabilization for the particular 400-kc crystal unit and point-contact transistor being used

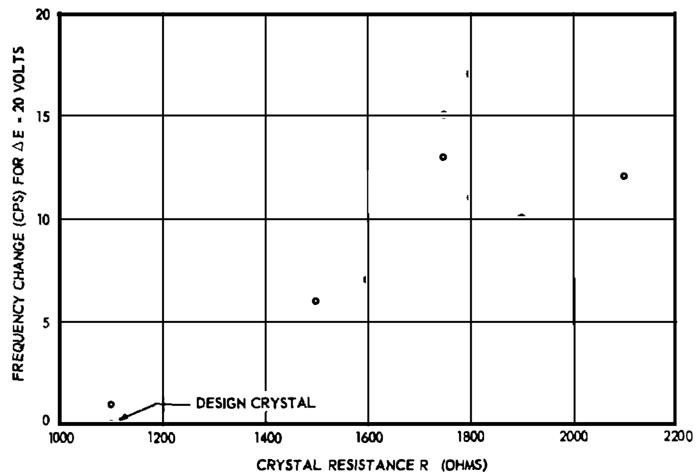


Figure 1-239. Frequency deviation of circuit shown in figure 1-238 for a given change in supply voltage at various values of crystal resistance

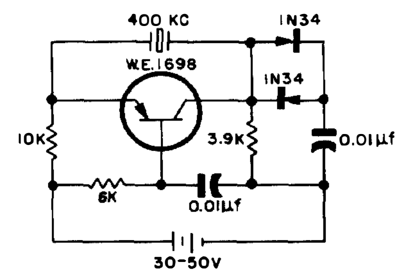


Figure 1-240. Modification of the crystal-feedback oscillator of figure 1-238 to provide greater frequency stability by providing greater amplitude stability with diode clipping

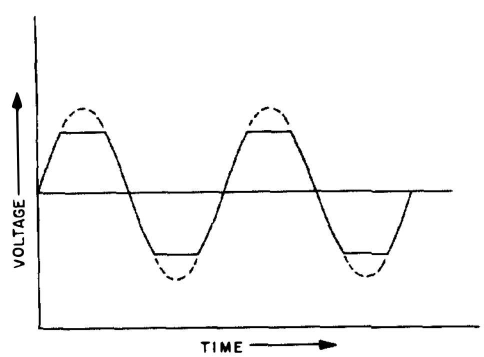


Figure 1-241. Voltage waveform after clipping of both peaks by diode rectifiers

stability apparently is due to differences in crystal resonance frequency, shunt capacitance, and L/C ratios from one crystal to the next. By different adjustments of the circuit resistances, it was nevertheless possible to obtain a maximum-stability point for each crystal.

Transistor Oscillators of Stabilized Harmonic Content

1-558. As explained in paragraphs 1-596 to 1-598, a distorted waveform tends to lower the fundamental frequency of an oscillator. Since the distortion is normally due to the nonlinear amplitude-limiting parameters, any change in the other resistance losses of the circuit will cause the amplitude to change, and hence the amount of distortion to change. For, example, if the resistance of the crystal unit decreases in the oscillator in figure 1-238, the feedback increases and the transistor will be operated farther into the bend of its volt-ampere input curve. This means greater distortion and harmonic content, and hence a shift to a lower fundamental frequency, where the stability characteristics of the circuit will probably be changed. To reduce this change in circuit performance when replacing transistors or crystal units, the double-diode clipping modification shown in figure 1-240 has been designed and tested by Dasher and Witt. The diodes clip both peaks of the oscillator output as indicated in figure 1-241, and thus, as long as the amplitude is sufficiently high, the distortion of the waveform remains relatively constant. The circuit requires an alpha of at least 2 in order to oscillate. Generally, what changes that do occur in the frequency stability are more significant when a crystal unit is replaced than when a transistor is replaced. Although the unadjusted performance of the circuit is inferior to the performance of the circuit in figure 1-238 when the latter is adjusted for a particular transistor and crystal unit, the diode clipping does provide superior frequency and amplitude stability as compared with the average performance of the unclipped circuit when the latter is adjusted only for optimum performance with an average crystal unit. The measured frequency deviations of the diode-clipped circuit for 10 different crystal units ranged from 0.1 to 0.05 parts per million per 1 per cent change in voltage.

1-559. Where maximum amplitude and frequency stability is desired to be relatively independent of the particular transistor and crystal unit, it may be necessary to operate the transistor on the straight portions of its characteristic curves and to obtain the limiting from slowly responding devices such as thermistors or a-g-c circuits. Such methods permit the oscillator to operate with approximately linear rather than harmonically-varying circuit parameters and do not require distortion for limiting. Nevertheless, these methods involve the introduction of additional components that usually deprive the transistor circuit of its principal advantages of compactness and low power and hence should be avoided if possible.

PACKAGED CRYSTAL OSCILLATORS

1-560. The great strides made in the field of electronics during the past two decades have so rapidly multiplied the diversity and complexity of electronic applications in commercial and military equipments that the problem of adequate maintenance has become a major national concern. One aspect of the problem has been the steadily growing trend to eliminate maintenance on a part-by-part basis; that is, to eliminate the time-consuming type of trouble-shooting that traces a faulty system to the exact defective part, such as to a particular resistor or capacitor. For this purpose it is necessary that electronic equipments be designed as an assemblage of packaged units, in which each unit can be as readily removed, tested, and replaced as is now possible in the case of vacuum tubes. This cannot be achieved unless the circuit being separately packaged actually operates as a functional unit. It is also paramount that the packaged unit be standardized; otherwise, by definition, there can be no standard nor rated tolerances by which the unit can be tested. The standardization is equally important as a means to encourage production of commercially available packaged units of known operational characteristics, so that design engineers can concentrate more of their attention upon new system design and application instead of repeatedly retracing the basic circuit problems for each new equipment. In the special field of aircraft and guided-missile equipment, an additional requirement of the packaged unit is that, as much as possible, it be miniaturized. As a consequence of these factors, the crystal industry is faced with a growing demand for standardized, miniature packaged crystal oscillators—each of which can be treated simply as a black box, so that when plugged into a circuit connecting to a specified power supply and output impedance, it will supply a radio-frequency signal of a given nominal frequency, purity, and power level within specified tolerances.

1-561. The standardized packaged oscillator is still in the developmental stage. Indeed, it is probably more accurate to say that it is still in the stage of basic research, insofar as the standardization is concerned. Here and there one may find an oscillator fabricated as a packaged unit which has been designed for use in a particular equipment, but with one or two exceptions (see paragraphs 1-571 and 1-572), packaged crystal oscillators are not generally available in the commercial market as standard components for general-purpose applications. Current research in the field of packaged crystal oscillator standardization has primarily been spurred by USAF Wright Air Development Center, in large part due to the urging and initiative of E. H. Borgelt of the Frequency Control Group, and an initial investigatory stage has been centered at the Illinois Institute of Technology Armour Research Foundation. When eventually packaged oscillators are fully available, we can expect that the interests of radio equipment designers will be more attracted to the problems of designing systems around standard crystal oscillator units rather than designing oscillators around standard crystal units.

The Gruen Packet-Oscillator Series

1-562. In 1954-1955 initial steps of a USAF research project were taken at the Armour Research Foundation under the direction of H. E. Gruen* to develop a series of miniature, plug-in, crystal-controlled, vacuum-tube oscillators which could be standardized to cover the frequency range of 0.8 to 75 mc at operating tolerances suitable to meet a majority of the USAF frequency-control requirements. The first phase of

* *Development of Packet Oscillator Series*, H. E. Gruen, Armour Research Foundation of Illinois Institute of Technology, USAF Contract No. AF 33 (616)—2125, 1954-55.

the project was a comprehensive investigation of the frequency-control applications of a large sample of radio sets currently in use by the USAF. It was found that 80 to 90 percent of the frequency-control requirements within the 0.8-75-mc range could be met economically by three conventional circuits, each designed for three bands, or a series of nine standard oscillators in all. It does not require this many separate oscillator circuits simply to cover the necessary frequency range, but it is desirable to have the recommended operating band of each circuit sufficiently narrow so as to avoid having to make critical tuning adjustments. For the same reason the circuits are not designed for specialized optimum operating characteristics within each band, but are intended only to provide a useful output consistent with the minimum requirements of accuracy and stability of the average frequency-control circuit used by the military.

1-563. The primary purpose of the packet-oscillator project has been to develop a sequence of plug-in circuits that can serve as models for establishing an initial set of oscillator Military Standards having the broadest practical tolerances. Once a beginning set of standards is agreed upon, it is to be expected that electronic manufacturers of packaged oscillators will introduce their own design modifications in meeting these standards—modifications which may lead to more rigid standards to supplement the broader standards for special applications. For example, the need may arise for an oscillator having a minimum deviation in frequency with ambient temperature, or with time, or in having a minimum tolerance in output amplitude, or in having a maximum possible output for given load conditions, or in having a minimum harmonic content, and so on ad infinitum with all possible combinations of electronic, physical, and operational special requirements. From a Military Standards point of view it is not the interior design of the oscillator that need or should be officially specified, but only what the unit can do or be or require in relation to the mechanic-thermodynamic-electronic system in which it is to be mounted. In this way a manufacturer is not discouraged from developing an improved packaged oscillator that can do an equivalent job more economically or better than the original standards model.

1-564. On the other hand, from an economically practical point of view, the Gruen packet series can profitably serve as unofficial standards insofar as the designs of the various oscillators are concerned—economical since the circuits used in these units have known performance characteristics, which, of course, would meet any Military Standards based upon them. The circuits themselves are conventional, designed to be used with Military Standard crystal units. Any standards established for packaged oscillators based upon the Gruen series will, of course, be no more rigorous than those already established for the crystal units being used.

1-565. The Gruen packet series employs three basic oscillator circuits: the grounded-plate Pierce, the electron-coupled grounded-plate Pierce, and the cathode-coupled Butler circuit. The two Pierce-type circuits are both used to cover the same frequency range of 0.8 to 16 mc. The Butler circuit is designed to cover the higher range of 10 to 75 mc. The electron-coupled circuit is recommended as an alternate in the lower frequency range when the input to the following stage is a tuned circuit, or whenever a minimum degree of coupling is desired between the crystal oscillator and the succeeding stage.

GROUNDED-PLATE PIERCE PACKET (0.8 TO 16 MC)

1-566. Figure 1-242 is a picture of the miniature packaged Pierce oscillator designed as a plug-in unit by Gruen to cover the 0.8—16-mc range. This range is covered in three bands. A schematic diagram of the basic circuit and the values of the parameters for each of the three bands is

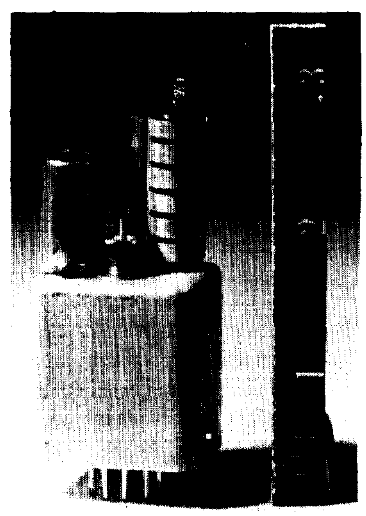
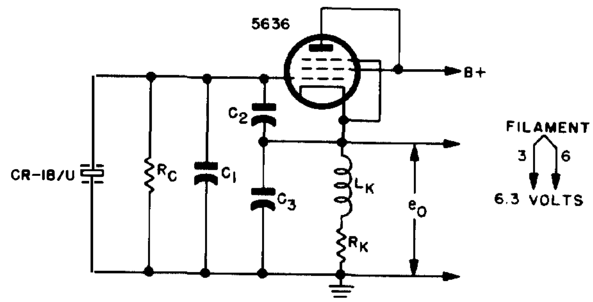


Figure 1-242. Miniature plug-in (standard octal base) Pierce oscillator (0.8 to 16 mc) developed at Armour Research Foundation by H. E. Gruen et al., as an initial basic packaged oscillator unit to be standardized for military use

shown in figure 1-243. Figure 1-244 shows the output voltage versus frequency for each of the three bands, and figure 1-245 shows the crystal drive in milliwatts versus frequency. These performance charts, prepared by J. S. Kurinsky of Armour Research Foundation, are based upon measurements made under no-load conditions. The range of values at each frequency is representative of the variations in output and crystal drive to be expected due to typical variations in crystal resistance. Thus, the bottom (lowest output and lowest crystal drive) curves of each of



Frequency Band (mc)	R_C (kilohms)	R_K (kilohms)	C_1 ($\mu\mu f$)	C_2 ($\mu\mu f$)	C_3 ($\mu\mu f$)	L_K (mh)	B+ (volts)
0.8 — 5	1000	3.9	15	10	47	7.0	75
3 — 11	100	2.0	15	10	47	0.8	100
5 — 16	47	1.5	15	10	36	0.3	100

Note: The value of C_3 is chosen so that proper operation is obtained when this circuit operates into a load of 15$\mu\mu$f. Wiring capacitance increases the values of C_1, C_2, and C_3 by the following amounts:

C_1: 5.5 $\mu\mu f$
C_2: 3.5 $\mu\mu f$
C_3: 10 $\mu\mu f$

Figure 1-243. Schematic diagram and parameters of the grounded-plate Pierce circuit of Gruen packet series

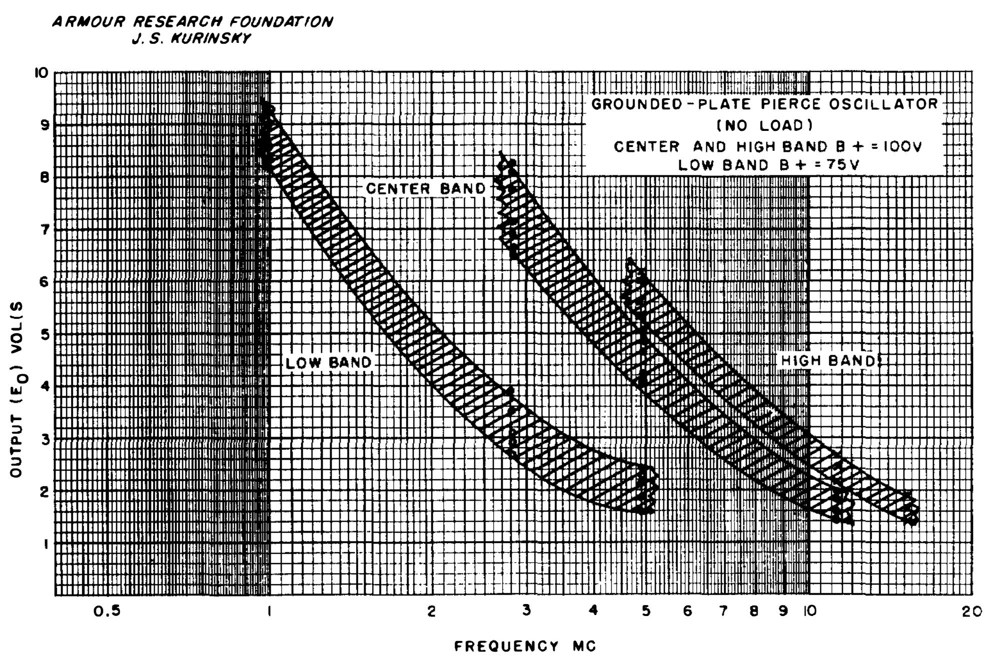


Figure 1-244. Typical output voltage and voltage tolerances of the grounded-plate Pierce circuit of Gruen packet series. These voltages were measured under no load conditions except for the connection of a 15-$\mu\mu$f capacitance across the output to insure a correct effective load capacitance for the crystal unit

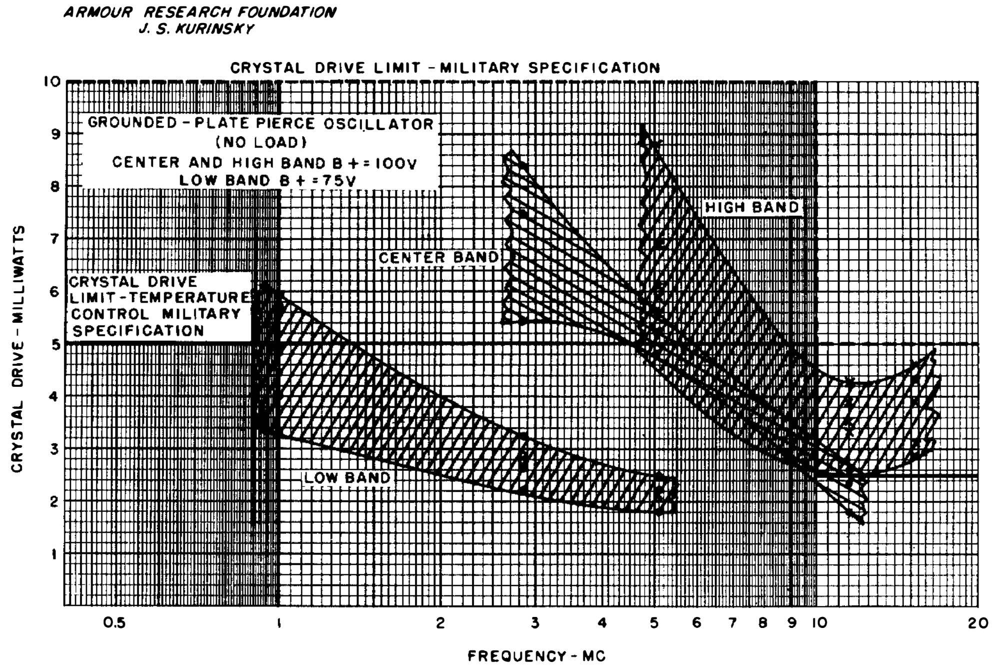


Figure 1-245. Typical drive levels and operating tolerances to be expected for crystal units connected in the grounded-plate Pierce circuit of Gruen packet series. The values shown assume no resistance losses in the output load

the bands should not be interpreted as representing the output and crystal drive when crystal units of maximum permissible resistance are inserted in the circuits. The performance characteristics for maximum-resistance crystals are fairly closely approximated by the bottom curve of the high band; but the highest-resistance crystal units used in the low-band measurements had resistance values only one-half to one-fourth the permissible limits. Note in figure 1-242 that the external mounting of the crystal unit and vacuum tube permits a ready replacement of either without diminishing the advantages of the packaged unit. The metallic clasp appearing to shield or support the vacuum tube is actually inserted as a thermal conductor to prevent overheating. A tuning element, in the form of a screw adjustment for C_3 (in figure 1-243), is provided to ensure that the correct load capacitance (32 $\mu\mu$f) is across the crystal unit. The parameters indicated for C_3 are based on the assumption that the external load will introduce an additional 15 $\mu\mu$f in parallel with C_3. If not, C_3 must be adjusted to give the proper total. C_3 can also be used as a trimmer adjustment in the event of crystal aging. The frequency stability of the Pierce packet as a function of the plate voltage

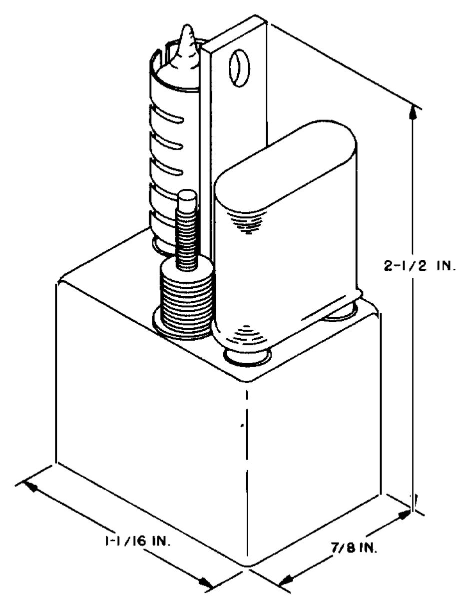
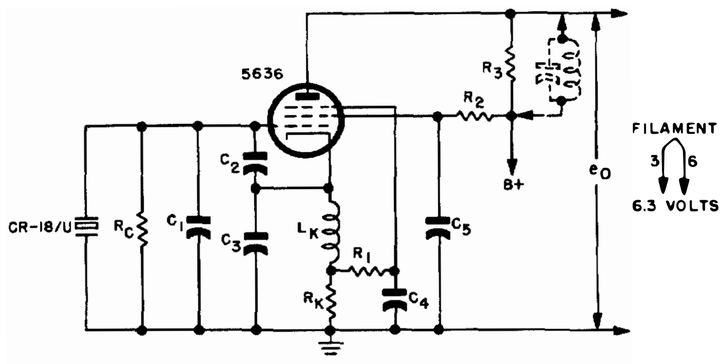


Figure 1-246. Drawing of miniature plug-in (standard octal base) electron-coupled Pierce oscillator (0.8 to 16 mc) developed at Armour Research Foundation by H. E. Gruen et al., as an initial basic packaged oscillator unit to be standardized for military use



Frequency Band (mc)	R_C (kilohms)	R_K (kilohms)	R_1 (kilohms)	R_2 (kilohms)	R_3 (kilohms)	C_1 ($\mu\mu f$)	C_2 ($\mu\mu f$)	C_3 ($\mu\mu f$)	C_4 ($\mu\mu f$)	C_5 ($\mu\mu f$)	L_K (mh)	B+ (volts)
0.8 — 5	1000	3.9	100	30	10	15	10	130	1000	1000	7.0	125
3 — 11	100	2.0	100	30	10	15	10	62	1000	1000	0.8	150
5 — 16	47	1.5	100	30	10	15	10	51	1000	1000	0.3	150

Wiring capacitance increases the values of C_1, C_2, and C_3 by the following amounts:
C_1: 5.5 $\mu\mu f$
C_2: 3.5 $\mu\mu f$
C_3: 10 $\mu\mu f$

Figure 1-247. Schematic diagram and parameters of the electron-coupled, grounded-plate Pierce circuit of Gruen packet series

supply is rated at 2 to 3 parts per million per 10 per cent voltage change.

ELECTRON-COUPLED PIERCE PACKET (0.8 TO 16 MC)

1-567. Figures 1-246, 1-247 and 1-248, respectively, show a dimensional drawing, the schematic diagram, and the output-voltage characteristics of the electron-coupled packaged unit. The output curves are representative of a typical group of crystal units, the same group that was used in plotting the performance of the Pierce oscillator as shown in figures 1-244 and 1-245, and, hence, are not intended to represent the entire output tolerance to be expected when crystal units of minimum resistance are replaced by those of maximum resistance. The tuned-output curves were obtained by replacing R_3 with a plate tank that was then tuned for maximum output at the fundamental. The fundamental frequency remains stable within 2 to 3 parts per million as the plate circuit is tuned throughout its range. The frequency stability as a function of plate supply voltage is 2 to 3 parts per million per 10 per cent voltage change. Note in the schematic diagram that the suppressor grid is connected to be operated at the same bias as the cathode, but is r-f bypassed to ground through C_4. This arrangement permits much greater independence of the frequency-control circuit from the plate load tuning, and therefore, smaller frequency-deviation limits.

CATHODE-COUPLED BUTLER PACKET (10 TO 75 MC)

1-568. Figure 1-249 shows a dimensional drawing of the packet design of a Butler circuit which has been tested in three bands to cover the over-all frequency range of 10 to 75 mc. Figure 1-250 is a schematic diagram of the circuit, with the parameters given for each of the three bands. Figure 1-251 shows the output voltage deviation to be expected from a typical group of crystal units. No data is available at this writing of the exact range of crystal resistance that is represented by the output curves, nor by the crystal-drive curves shown in figure 1-252. Very probably the bottom curves in each of the figures,

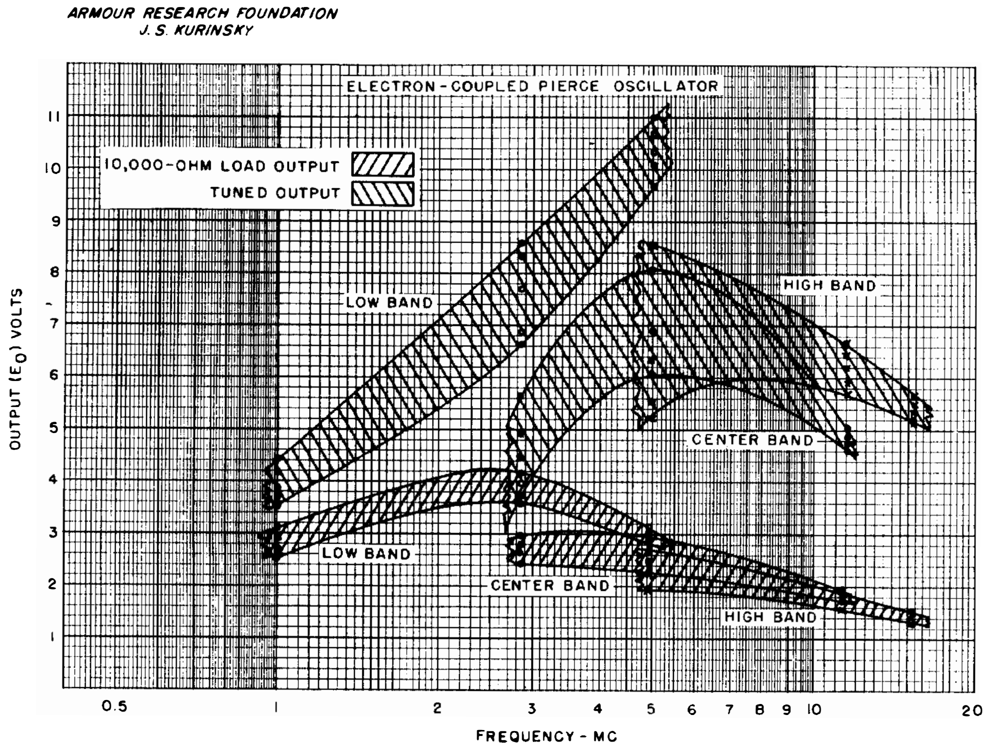


Figure 1-248. Typical output voltage and voltage tolerances of the electron-coupled, grounded-plate Pierce circuit of Gruen packet series

particularly in the high band, are approximately representative of crystal units having the maximum permissible values of resistance. Figure 1-253 shows the difference, in parts per million, between the series-resonance frequency of the crystal unit as measured in a CI meter and the operating frequency of the same crystal unit when connected in the Butler circuit, as measured by different operators at repeated tunings to peak output voltage. This circuit-frequency deviation should be added to the nominal frequency tolerance of the crystal unit in order to evaluate the over-all frequency tolerance of the packaged oscillator. For a 10 per cent change in plate supply voltage the stability of the oscillator unit is within 1 to 2 parts per million in the 10—20-mc range, 3 to 4 parts per million in the 20—40-mc range, and 4 to 5 parts per million in the 40—75-mc range.

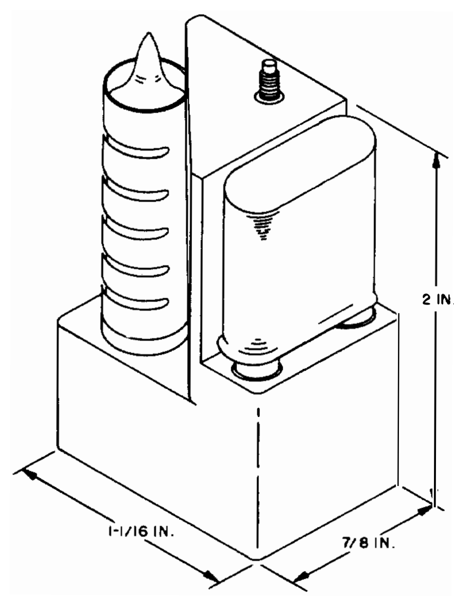
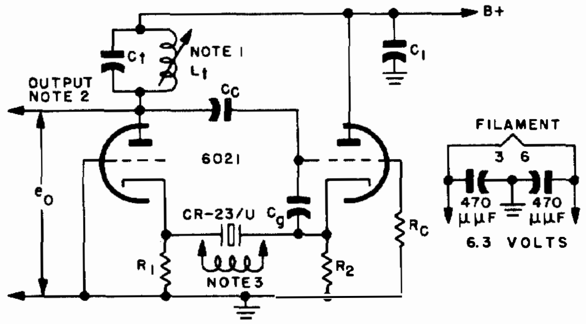


Figure 1-249. Miniature plug-in (standard octal base) cathode-coupled Butler oscillator (10 to 75 mc) developed at Armour Research Foundation by H. E. Gruen et al., as an initial basic packaged oscillator unit to be standardized for military use



Frequency Range (mc)	R_1 (ohms)	R_2 (ohms)	R_C (kilohms)	C_1 ($\mu\mu f$)	C_O ($\mu\mu f$)	C_g ($\mu\mu f$)	C_t ($\mu\mu f$) Note 4	B+ Volts
10 — 20	1000	1000	10	470	100	4.7	15	100
20 — 40	1000	560	10	470	100	4.7	15	100
40 — 75 Note 3	1000	270	10	470	100	1.5	15	100

Note 1: Inductance L_t is tuned to resonate with C_t at the operating frequency.

Note 2: All test results were obtained with the oscillator working into a 5000-ohm load.

Note 3: At frequencies of 40 mc and higher, an inductance which resonates with the static capacitance of the crystal should be used to prevent unwanted oscillation. A low value of Q is desirable; coils wound on 1500-ohm composition resistors perform well.

Note 4: The value of 15 $\mu\mu f$ for C_t includes stray wiring capacitance.

Figure 1-250. Schematic diagram and parameters of the Butler circuit of Gruen packet series

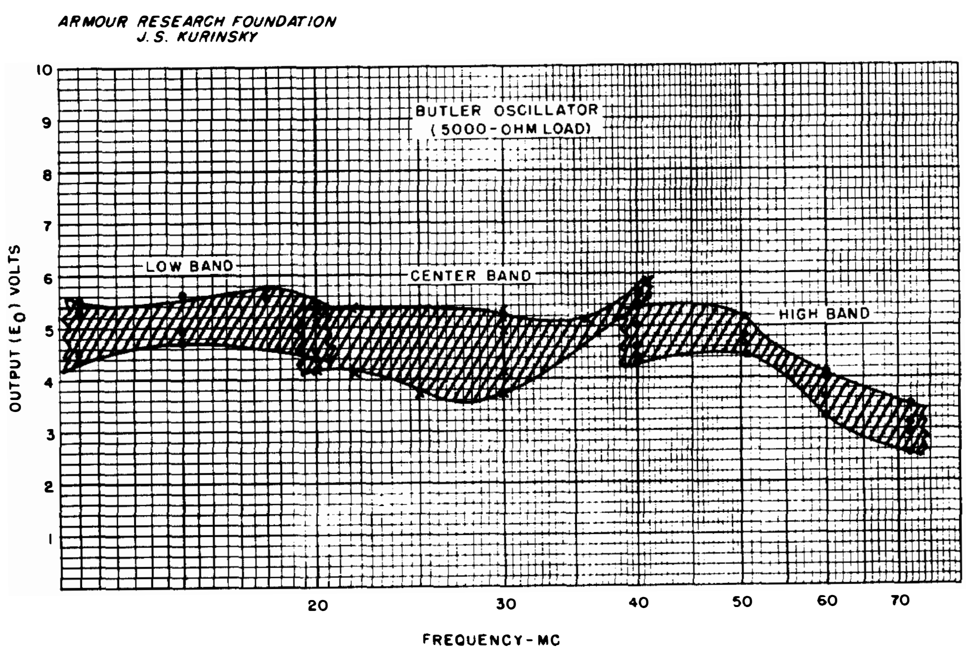


Figure 1-251. Typical output voltage and voltage tolerances of the Butler circuit of Gruen packet series when operating into a 5000-ohm load

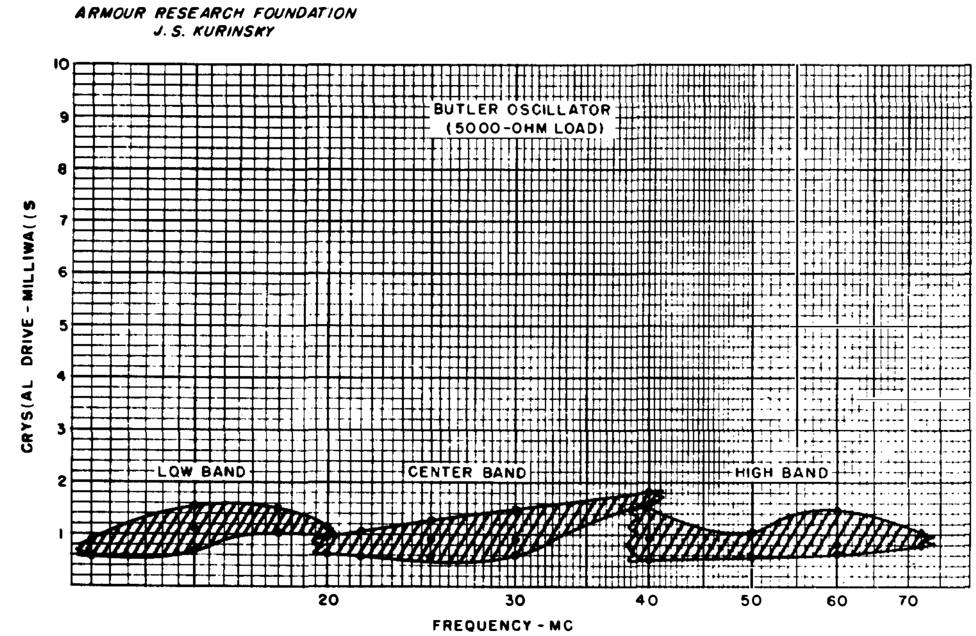


Figure 1-252. Typical drive levels and operating tolerances to be expected for crystal units connected in the Butler circuit of Gruen packet series. The values shown assume a 5000-ohm load resistance

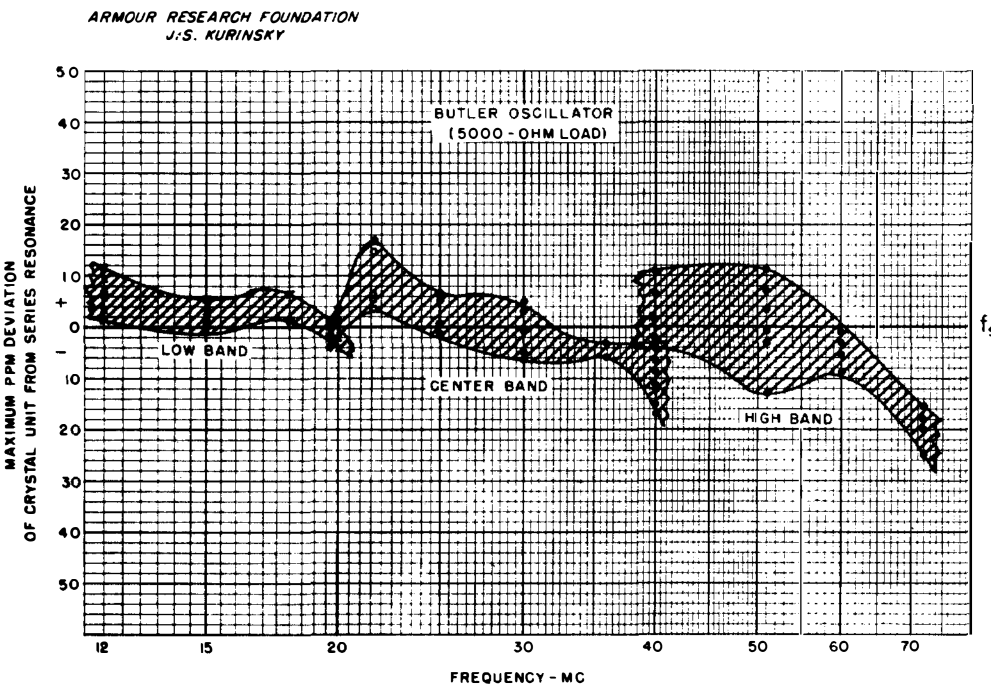


Figure 1-253. Typical frequency tolerances to be expected of crystal units connected in Butler circuit of Gruen packet series. The deviations plotted are deviations of the peak-output operating frequency from the series-resonance frequency of the individual crystal unit as measured with a CI meter, and are not deviations from the rated nominal frequency of the crystal unit. Thus, the overall tolerance to be expected of such an oscillator must be the total of the crystal unit nominal frequency tolerance plus the expected operating resonance tolerance, the latter being indicated by the measured deviation above

Packaged Crystal-Controlled Transistor Oscillators

1-569. Transistor circuits are ideally suited for miniaturized packaged units. Eventually, when transistors, especially those in the h-f range, are being manufactured with closer tolerances, it is to be expected that a majority of the packaged oscillators will be of transistor design. Almost certainly will this be true of those units designed to be temperature controlled since the heat dissipation of the transistor is so much smaller than that of the vacuum tube, and hence less of a factor in causing unwanted temperature gradients in an oven container. Research in the direction of developing standardized transistor packaged oscillators is being initiated by the U. S. Air Force. At the present writing the only information available concerning the development of any type of transistor packaged oscillator is the experimental model developed at the National Bureau of Standards, discussed in the following paragraph, and the temperature-controlled miniature units already being manufactured by the James Knights Company, discussed in paragraphs 1-571 and 1-572.

SULZER MINIATURE PACKAGED TRANSISTOR FREQUENCY STANDARD

1-570. A complete packaged 100-kc frequency standard with a self-contained power supply is shown in figure 1-254. This oscillator unit was developed by Peter G. Sulzer of the National Bureau of Standards. Although the primary purpose of the NBS project was to develop a pocket-size, self-sufficient frequency standard, the result provides an admirable illustration of the possibilities in packaged design. A G-element is employed for frequency control, and as can be seen, the dimensions of the crystal unit comprise about half of the total volume. Power is supplied from a mercury cell, which, of course, would not be a necessary feature if the oscillator unit were adapted for use as a general-purpose plug-in component for precision frequency control. A grounded-emitter junction transistor is used. Capacitors C_2 and C_3 form a voltage attenuator in the feedback circuit to reduce the voltage applied across the crystal unit. The relatively large values of the capacitors connecting to the terminals of the crystal unit serve to stabilize the feedback phase. All the circuit elements except the crystal unit and the mercury cell are supported in casting resin on a plastic frame. The mercury cell is at the base. The oscillator is packaged in a metal container (not shown), 7 inches by 1¾ inches. The oscillator has a frequency deviation no greater than 1 part in 10^8 per degree C change in temperature, or 1 part in 10^8 per 0.1 volt change in battery supply.

COMMERCIALLY AVAILABLE PACKAGED CRYSTAL-CONTROLLED TRANSISTOR OSCILLATORS

1-571. A pioneer in the field of miniaturized packaged oscillators has been the James Knights Company. Their research team composed of R. Ives, C. Reynolds, R. Beetham, R. Berge, C. Eickeberge,

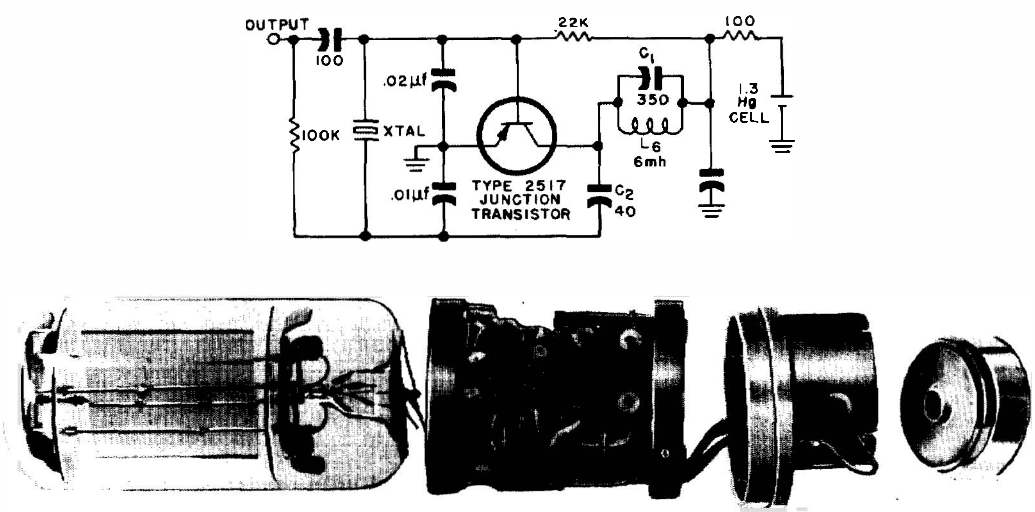
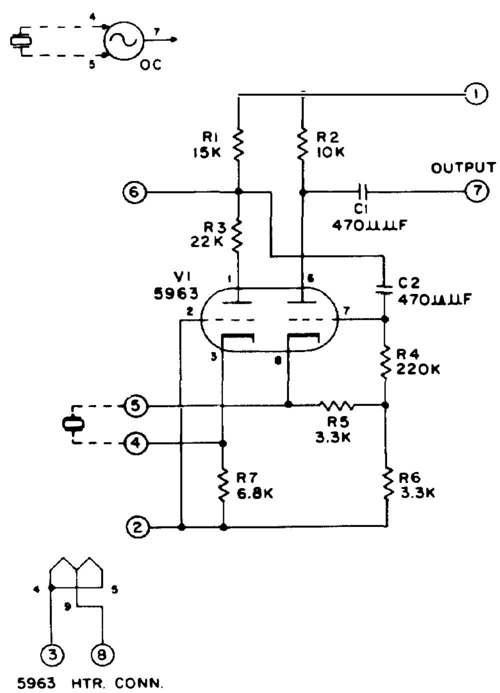


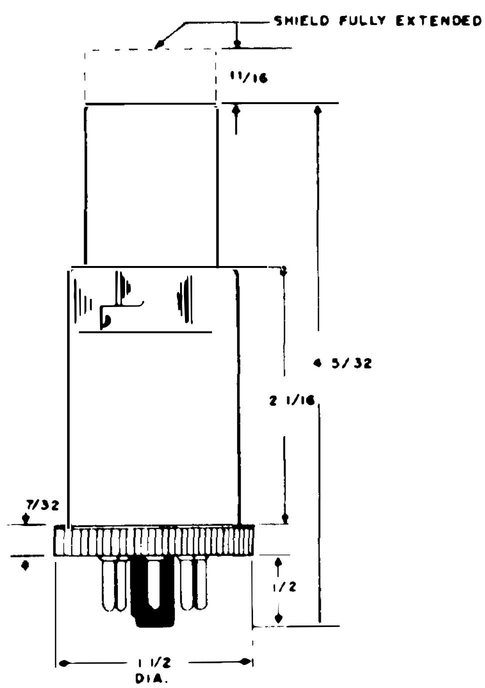
Figure 1-254. Small, portable, packaged transistor crystal oscillator developed by P. G. Sulzer of National Bureau of Standards. Unit is provided with mercury-cell power supply shown at right. Metal container is not shown

et al appears to have been the first in this country's electronics industry to have developed miniature, temperature-controlled, plug-in crystal oscillators available for general-purpose use.* Three oscillator units have been developed at this date which employ transistor circuits and are designed for mounting within small crystal ovens. One unit is a multichannel h-f oscillator which can be provided with a bank of up to 11 h-f crystal units mounted, together with its oscillator circuit, within a rotary switch. This assembly can plug into a crystal oven, or into an external octal socket, or it can be panel-mounted. When oven-mounted, the entire package occupies a cylindrical volume approximately 3¾ inches in diameter and 5 inches high. The crystal units are of a subminiature, evacuated, glass envelope type, which operates at frequencies of 6 mc and higher. Insufficient data exists to provide a complete technical description of the performance of the oscillator.

1-572. Of greater immediate interest are the two miniature oscillators that are mounted within a cylindrical plug-in oven of unusually high stability, the dimensions of which are shown in figure 1-255. There are no switching arrangements, there is only one crystal unit to a package, and each is soldered to its circuit. The low-frequency circuit is shown in figure 1-256, and the high-frequency circuit, in figure 1-257. Evacuated, glass-envelope crystal units are employed to improve the aging characteristics. The l-f circuit is essentially a crystal-controlled multivibrator. Its output is a square wave which can have a 4-to 5-volt peak. The h-f circuit has approximately a sine-wave output, 0.8-volt peak with a 6-volt supply. At 1 mc the measured deviation in frequency

* The Electronic Engineering Company of California has also pioneered a packaging technique finding application in small plug-in oscillator circuits. The EECO Production Company fabricates a plug-in, vacuum-tube, packaged oscillator circuit (see accompanying figure) designed for crystal control in the 90-to-250-kc range. A crystal unit is not supplied as a component part of the packaged unit, nor is temperature control provided. The circuit is a cathode-coupled Butler type. No tuned circuits are employed, so that an externally connected crystal unit will operate at its fundamental mode. The output under no-load conditions is rated at approximately 14 volts rms. The recommended load-impedance range is 100,000 to 250,000 ohms. The power requirements are a plate supply of 200 volts dc between pins 1 and 2 at 3.5 ma., and a heater supply of 6.3 volts at 300 ma. The d-c potential of the heaters relative to pin 2 must be between plus 90 and minus 70 volts. The circuitry is mounted in the bottom of the unit, the vacuum tube at the top. The weight is approximately 3.25 oz. A standard octal base is provided. A screw cap at the bottom permits the unit to be readily taken apart and assembled. The tube shield is removable to permit easy replacement of the vacuum tube. The finish is grey baked enamel. No information can be given concerning the tolerances and stability of the oscillator since it has not been designed for use with a specific crystal unit. For use with some Military Standard units it might become necessary to reduce the plate supply voltage in order to avoid operating the crystal beyond its rated drive level.





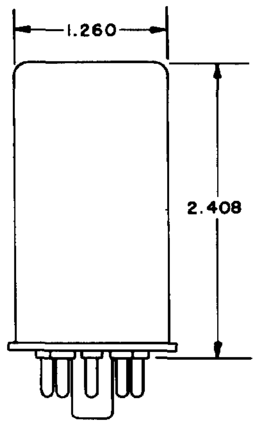


Figure 1-255. Dimensions of miniature, cylindrical, temperature-controlled, transistor plug-in oscillator developed by James Knights Company. Packaged unit is provided with a standard octal base

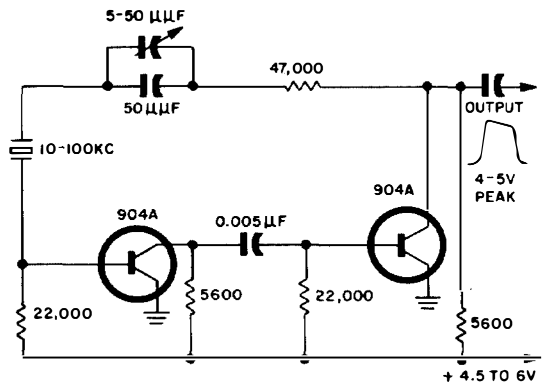


Figure 1-256. Schematic diagram of crystal-controlled transistor circuit employed in James Knights packaged low-frequency (10 to 100 kc) oscillators. Square wave output is rich in harmonics

was 9 cps when the supply voltage was changed from 6 to 4.5 volts. A 2-mc test oscillator has shown an operating stability of 1 ppm over a period of several weeks, but sufficient time has not elapsed for these oscillators, so recently developed, to be tested thoroughly for operating life and aging characteristics. The ovens in which the oscillators are mounted require a maximum of 1.5 amperes at 6.3 volts. The crystal temperature is maintained within plus or minus 1°C over an ambient range of —55° to +70° C. The assembled package is equipped with a standard octal base.

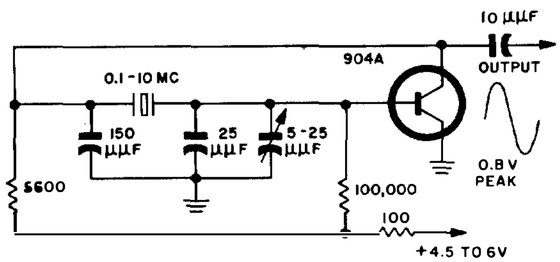


Figure 1-257. Schematic diagram of crystal-controlled transistor circuit employed in James Knights m-f and h-f (0.1 to 10 mc) oscillators. Relatively pure sine-wave output

Standards for Packaged Oscillators

Many of the problems associated with the standardization of electronic components appear to defy solutions that satisfy designer, fabricator, distributor, consumer, and repairer alike. As a result, the subject of standardization, itself, is often a controversial issue. After discussing a number of the conflicting points of view with different associates in the crystal industry, it seemed to the writer that perhaps some indirect good might result if these points of view, as applicable in the standardization of packaged oscillators, were assembled on paper where they might more easily be seen in relation to one another. For this reason, the discussion from paragraphs 1-573 to 1-581 detours somewhat from the domain of technical fact into an area more representative of individual opinion. The reader interested only in the technical aspects of packaged oscillators can disregard these paragraphs.

ADVANTAGES OF STANDARDIZATION

1-573. The primary advantages of standardized packaged oscillators, as of other standard packaged units, can be described as those of economy:

a. It is an economical saving for research and design departments to avoid having continually to assign and train engineers in the special design theory and developmental techniques of oscillator circuits; this saving can be achieved if there are commercially available oscillator units of known operational characteristics capable of meeting all normal frequency-control requirements.

b. It is more economical for equipment manufacturers to assemble a given oscillator circuit in a fixed manner as a separate unit, regardless of the different types of equipment in which it is to be used, than by fabricating the oscillator as an integral part of a much larger network where special production procedures are required to make the electrical and physical connections conform to each different chassis layout.

c. It is more economical for maintenance departments to replace defective oscillators as plug-in units than to find, educate, break in, and pay highly trained technicians to troubleshoot and repair them; to say nothing of the additional man hours saved by avoiding the necessity of having to store and distribute each oscillator part separately.

d. Consequently, it is an economical saving to the consumer, who ultimately must bear the costs of the research, design, development, fabrication, and maintenance. In addition, the consumer can also profit from the longer life of the oscillator due to the extra protection provided by the package design, particularly if the container is hermetically sealed.

PROBLEMS OF STANDARDIZATION

1-574. Standardization has its dangers as well as its advantages. The dangers arise both from too much standardization, and from not enough standardization—from freezing the design of an item so thoroughly that the development of improved models is discouraged, and from having such ambiguous and unsystematic standards that many of the advantages are lost. Over-standardization is most apt to result from the understandable resistance of maintenance and repair organizations to having endlessly to contend with an ever expanding jumble of different models of every description. Under-standardization is most apt to result from the equally understandable resistance of the manufacturer in committing himself to rigid technical standards that slow his production line, or limit his freedom to experiment with new techniques and designs. Both too much and too little standardization can suffocate technical progress. In the particular case of packaged oscillators, much thought is being given for the establishing of a system of standards that avoids the pitfalls of both extremes.

If Standards Are Too High

1-575. Over-standardization appears to exist at three different levels. When the standards are too high; when they are over-emphasized; and when they are too restrictive. The least offensive are the standards that are too high. Indeed, the effort to meet exceptionally high standards is usually a spur to the development of new inventions and new techniques. Where the product cannot be fabricated with profit, it simply is not produced; so that the problem has a way of working out its own solution.

If Current Standards Are Overemphasized

1-576. Over-emphasis of the importance of keeping the number of available models of a given component to a minimum is probably the most frustrating bugaboo that a pioneering manufacturer can face when he seeks official acceptance of a new design. It is true that standardization does and should discourage the development of new models that offer no improvement over the currently available standards, since such developments are effectively only a waste of engineering talent. But it should not discourage the development of improved models, which unfortunately does happen when the immediate problems arising from a new adjustment are emphasized ahead of the long-range advantages of a maximum rate of technical progress. It would seem that, if from the beginning, a standards program were planned that would anticipate the systematic development of improved models and higher standards, even to the extent of reserving pigeon holes for its eventual achievement, the advantages of standardization could be increased. For it appears that standardization of engineering products, in essence, is only a means of making known to the engineer *exactly* what is available. If the standards for packaged oscillators can be systematically organized at the start so that the development of new products can readily be placed in their anticipated categories, without disturbing the old standards or changing the cataloging system, then only the duplication of physical and functional characteristics need be discouraged in the development of new models.

If Standards Are Too Restrictive

1-577. Generally it is better to risk too much standardization than not enough, and because of this, probably the most difficult type of overstandardization to pinpoint is that in which too many of a component's characteristics are standardized. The problems that arise become particularly harassing when the committee controlling the standards are under strong pressure to keep the number of standard components to an absolute minimum. Under these conditions every newly developed component is considered primarily as a replacement of an older model, rather than as a new model, to compete with the old *only* for use in equipments of new design. For example, in the matter of standardizing crystal holders, experience has shown that considerable resistance is met when it is desired to gain official acceptance for any new design which in the future could replace an existing standard design, even though the new design is of proven superiority. If a plastic or metal container is standard for a given size of holder, reasons arise for delaying new standards for a superior metal or glass container, respectively. If an odd-shaped base has been standardized, it is very difficult even to gain acceptance for a standard octal base in an otherwise equivalent holder.

1-578. If a long delay is generally to be expected before new models can be standardized, or if the number of standard items must be kept to an absolute minimum, some of the difficulties facing production engineers are often avoidable when care is taken in establishing only those standards that specify the *function* that a particular item or unit is to perform in relation to any system in which it is to be used—not how the function is to be achieved, unless the manner of achievement in some way affects the design of the external system or the cost of maintenance and repair of the system as a whole. Why, for example, should the manner in which a crystal is mounted inside its holder be standardized, if there is no intention to repair defective units? If a crystal unit can meet all the required physical, electrical, and operational relations to the external system, what difference does it make how this is accomplished?

1-579. In regard to packaged oscillators, it is to be hoped that sufficient performance tests can be devised to eliminate the need for standards governing the internal design of the package. Unless the crystal unit is to be replaceable in the field, or is to have an outside mounting, as in the case of the Gruen package, it should not be necessary to specify the type of crystal unit employed. Indeed, what difference can it make to the user of two packaged units, identical insofar as their external characteristics are concerned, if one contains a crystal-controlled oscillator and the other a little imp turning a crank? As long as either controlling source is confined completely within its container and works dependably within the rated tolerances, the performance standards need not be concerned with whether the control is piezoelectric or metabolic. In the case of the crystal-controlled circuit, if the crystal unit is to be mounted inside the package container, as would be the case in a temperature-controlled unit, a manufacturer should not be required to use a Military Standard crystal unit if he can meet the oscillator requirements more satisfactorily with another type of crystal mounting or holder. (We assume that the oscillator standards will not require that the crystal unit be replaceable in the field. Such a specification for a completely packaged unit would mean that one of the primary advantages of packaging is being wasted—that in the case of breakdown, instead of replacing the packaged unit, the intention is to service it.) Also, it may be preferable to solder the crystal unit in place, than to depend upon a plug-in connection, since the former method usually permits a more stable circuit. A properly fabricated crystal unit can generally be depended upon to last as long as the useful life of the equipment in which it is to be used, so no objection to the soldered connection should arise except in special applications. For example, the specification for a plug-in crystal unit could be important when mounted outside the package, as in figure 1-242, where it can readily be replaced without disturbing the packaged unit. Such an arrangement should prove desirable if the same oscillator is intended to operate at more than one frequency where space does not permit a switching arrangement capable of mounting and connecting all the required crystals, so that these must be inserted manually in the field as the need arises. Also, the plug-in arrangement becomes important when the operating life of the crystal unit is expected to be much shorter than that of any other part of the packaged unit; or if the operating life of the entire unit is to be of relatively short duration. In any event, it would seem that progress in the development of packaged oscillators will be promoted better if the standards initially established are confined to specifying what the packaged oscillators are to be able to do under specified external conditions and do not tend to limit the ways in which the functional specifications are to be achieved, even if at the time the standards are established there appears to be only one way in which the desired function can be achieved.

Avoidance of Ambiguous Standards

1-580. If too much standardization vexes primarily the manufacturer of the standardized item, too little standardization scatters its headaches chiefly among those that must use the item. Particularly sensitive to the uselessness of ambiguous specifications or generalized advertising claims concerning a component is the design or developmental engineer who wishes to approach his circuit problems scientifically, yet who cannot do so unless he knows quantitatively exactly what he is putting into his circuit. To the engineer, a set of standards sufficiently rigid and dependable for predictable design and replacement purposes is equivalent to a set of rigid production line tests, since a unit cannot be guaranteed to pass certain specifications unless at least a representative sample has been tested to meet the specifications. Furthermore, since the performance of a unit depends upon the conditions under which it is operated, it is often of little help to a design engineer to specify exactly what a unit can do unless accompanied by equally exact data concerning the conditions under which the rated performance takes place.

1-581. As applied to packaged oscillator units, it is highly desirable that a systematized set of standard operating conditions be established simultaneously with a systematized set of performance standards. Otherwise, each oscillator may have to be adjusted for optimum performance in the particular equipment in which it is to be used. The ideal standard oscillator need not and should not be adjustable by either the equipment production line technician or by the equipment operator in the field. Too many variables are present. If adjustable, all tolerances must be rated as the maximum to be encountered over the adjustable range. If the equipment in which the unit is used can permit such tolerances, then there is no need to adjust the unit in the first place. A plausible exception would be the case of a small trimmer capacitor to offset the effects of aging in high-precision systems. Another exception would occur if it were practicable to design a unit with marked adjustment points at which the tolerances could be separately specified. For example, one adjustment setting could indicate a point of maximum output amplitude, another a point of maximum amplitude stability, another a point of maximum frequency stability, etc. But to the extent that the adjustments made by the user are not definitely predictable, the advantages of standardization are lost. It would be preferable for the manufacturer of the packaged units to make the adjustments and assign different model numbers where the same circuit had significant performance differences.

PRINCIPAL OSCILLATOR CHARACTERISTICS TO BE STANDARDIZED

1-582. Packaged oscillator characteristics that should be standardized can be grouped in two categories—physical and electrical. The former concerns primarily the mounting requirements and limitations; the latter covers the operating requirements and performance characteristics. Of the physical characteristics the most important are the conventional items such as dimensions, weight, shape, provisions for mounting, ambient temperature range, mechanical ruggedness as measured by well-defined tests, moisture resistance, seal test, operating life of thermostat, if any, and so forth. Of the more important electrical characteristics there are the operating requirements and permissible tolerances of such items as input voltages, currents, and power, the test load impedance, ground connections, and the like; also, there are the performance characteristics to be specified for the standard test conditions. These performance characteristics include the following: output fundamental frequency; db level of nearest harmonics relative to fundamental; equivalent emf and impedance of oscillator, when viewed as an impedance in series with a generator, at the fundamental frequency and at any useful overtones; output waveform; frequency and amplitude tolerances; frequency and amplitude deviations over input voltage range and over operating temperature range; frequency deviation over permissible load tolerances; and frequency and amplitude deviations permissible after the unit is subjected to the various aging and mechanical tests.

EFFECT OF CRYSTAL RESISTANCE ON PACKAGED OSCILLATOR STANDARDS

1-583. Even in those packaged oscillators in which a Military Standard crystal unit is not to be specified, the most economical fabrication process will probably require the use of such a unit, or at least one that has electrical tolerances equivalent to those of a standard unit, since the oscillator standards will undoubtedly be originally established on the basis of oscillators of known operational characteristics employing Military Standard crystals. In establishing the oscillator standards it is important that account be taken of the full variation in output level and frequency to be expected due to the range of the equivalent resistance of the crystal unit between its maximum permissible and minimum expected values. Likewise, it is important that the developers of the packaged oscillators test their circuits over the expected crystal resistance range to ensure that the oscillator standards are met throughout the range. Generally it is only the maximum resistance conditions that must be specifically tested. If the design engineer does not have maximum-resistance crystal units available, these can be simulated in making the laboratory tests.

Simulating Maximum-Resistance Crystal Units

1-584. First, the equivalent resistance of the crystal unit being used in the test circuit must be measured with the aid of the appropriate CI meter. If the test circuit employs a series-mode crystal unit, simulating a maximum-resistance crystal is readily achieved by connecting in series with the crystal a resistance of such a value that the total resistance is equal to the maximum permissible. If the test circuit employs a parallel-mode crystal unit, the simplest method is to employ a strategem devised by John W. Sherman, Jr.

of General Electric Co.* A high resistance is connected in parallel with the crystal unit. The value of this resistance is chosen so that the power losses of the actual crystal unit and of the resistance in parallel have a combined value equal to the losses that would occur if the crystal unit were of maximum resistance. Assuming that the impedance of the crystal unit is approximately equal to its effective reactance, X_e, which in turn is approximately equal to $\frac{1}{\omega C_x}$, where C_x is the rated load capacitance and ω is the angular frequency, then the current through the crystal unit in both the actual and the maximum-resistance case would be approximately equal to $E_c\omega C_x$, E_c being the voltage across the crystal unit. Letting R_{em} equal the maximum permissible resistance, R_e equal the actual effective resistance of the crystal unit, and R_x equal the value of the shunting resistance, we set the power losses in the hypothetical crystal unit of maximum resistance equal to those due to the actual R_e and R_x:

$$(E_c\omega C_x)^2 R_{em} = (E_c\omega C_x)_2 R_e + E_c^2/R_x$$

From which the equation for R_x is obtained.

$$R_x = \frac{1}{\omega^2 C_x^2 (R_{cm} - R_e)}$$

FACTORS INVOLVED IN OSCILLATOR LIMITING

1-585. The characteristics of any oscillator are dependent to a large extent upon the degree and type of amplitude limiting employed. The factors involved are especially significant in the case of crystal oscillators because the variations in crystal-unit effective resistance tend to introduce wide variations in the amplitude of oscillation when different crystal units of the same type and frequency are substituted in the same circuit. The method of limiting also affects the type of crystal unit to be used, whether it is to be a series- or a parallel-mode unit. Directly dependent upon the limiting method will be the harmonic content in the oscillations. Upon the degree of harmonic distortion will depend the tuning of the circuit, so that any deviations in the limiting characteristics are reflected as instabilities of frequency as well as of the amplitude. In the following paragraphs we shall discuss these factors briefly; but first, the reader may find it helpful to review paragraph 1-232, in which the crystal oscillator, as a generalized negative-resistance circuit, is discussed.

Negative-Resistance Limiting

1-586. In paragraph 1-232 we found that when a crystal oscillator is represented as a series-resonant negative-resistance circuit (figure 1-108 (B)), oscillations continue to build up as long as the numerical ratio $\left|\frac{\rho_s}{R_e}\right|$ is greater than 1. When the oscillator is represented as a parallel-resonant negative-resistance circuit (figure 1-108 (C)), oscillations build up as long as the numerical ratio $\left|\frac{Z_p}{\rho}\right|$ is greater than 1. Now, in figure 1-108, both the series-resonant and the parallel-resonant negative-resistance circuits are intended to represent the same crystal oscillator, one that employs a parallel-mode crystal unit. Let us modify our interpretation of the series-resonant circuit to let it represent any series-resonant circuit containing a crystal element. That is, we shall treat X_l as the reactance of any load capacitance actually connected in series with the crystal unit, as is the case, for instance, in a CI-meter circuit connected for parallel-mode crystal control. We shall also be free to assume that X_e and X_l each equals zero, in which case the circuit is equivalent to a series-mode crystal unit connected across the terminals of a negative resistance equal numerically to the series-resonance resistance of the crystal unit. As far as the parallel-resonant negative-resistance circuits shown in figure 1-108 are concerned, we shall continue to consider them as representing an actual parallel connection of a crystal unit and its load capacitor driven at parallel resonance by a negative resistance, ρ. In other words, we wish simply to make it clear that the generalized series-resonant circuit represents an actual series-resonant circuit, and that the generalized parallel-

* This method for effectively increasing the resistance of a parallel-mode crystal unit, although independently devised and tested successfully over a range of 1.5 to 51 mc by J. W. Sherman for the express purpose of testing the design fitness of particular oscillator circuits, is also reported by engineers of the Signal Corps Engineering Laboratories to have been used in their Frequency Control Branch for several years for measuring crystal activity in various test sets. During World War II a jig was devised that provided the crystal unit with an adapter and a variable shunt resistance. Effectively, the device was a means of readily degrading the activity of a test crystal unit to some predetermined marginal value. For example, by adjusting the shunt resistance, a given effective marginal resistance could be indicated when the drive level of the test circuit reached a certain value. Knowing the total effective resistance, the load reactance, and the shunt resistance, it would be a simple matter to compute the effective resistance of the crystal unit alone. The use of such a jig and test circuit can offer an advantage over the method described above for simulating maximum-resistance crystal units, if the process is to be repeated frequently enough to justify their construction. With the test-circuit activity indicator calibrated in terms of maximum-resistance values, the shunt resistance of the jig can be adjusted immediately to its desired value without computation or having to actually measure the crystal unit resistance.

resonant circuit represents an actual parallel-resonant circuit.

1-587. In order for oscillations to build up and reach a stable equilibrium in the series circuit, we see that the negative-resistance device must have the property of automatically decreasing its negative resistance as the amplitude of oscillations increases. The negative-resistance device of the parallel circuit must have the opposite property; that is, its negative resistance must increase as the amplitude increases. Literally, these opposing requirements mean that the same device used as a negative resistance in the series circuit cannot be used in the parallel circuit, or at least not with the same operating characteristics, and vice versa. By the term "negative resistance" used in this context, let it be plain that we mean a two-terminal network which, when connected across the crystal circuit, behaves exactly as an automatically variable resistor does, except that the negative resistance supplies energy rather than dissipates it. The act of connecting a negative resistance across a crystal circuit does not in itself generate a voltage and current for driving the circuit. Of course, a signal generator used to force-drive the circuit can be treated as a virtual negative resistance mathematically to simplify a network analysis, but that is not a "passive" type of resistance such as we are concerned with here. The voltage across, or the current through, a passive resistance (negative or positive) is not assumed to be self-generated by that element, but rather is the response of the resistance to the actions in the rest of the circuit. In an oscillator circuit, the initial action is the random thermal motions of the free electrons; the negative resistance reacts to this at the same energy level, the remainder of the circuit reacts to the negative-resistance reaction, the negative resistance reacts, in turn, to this, and so on until the energy builds up. At all times the frequency, amplitude, and phase of the negative-resistance voltages and currents are under the control of the external circuit.

1-588. As long as the oscillations are building up, part of the energy being supplied by the negative resistance is being stored by the circuit. As far as the negative resistance is concerned, the stored energy is equivalent to lost energy, so that process can be represented by the addition of equivalent resistances in the generalized circuits. Figure 1-258 shows the generalized negative-resistance circuits with the equivalent "energy-storing" positive resistances. In the series circuit, the instantaneous value of R_{es} is defined by the equation:

$$R_{es} = |\rho_s| - R_e$$

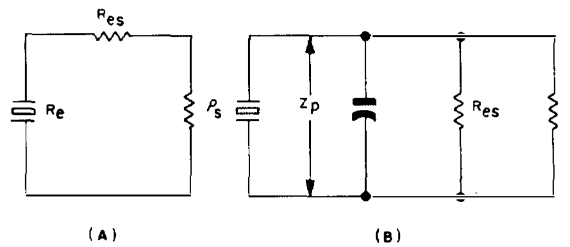


Figure 1-258. Equivalent circuit during build-up period of generalized negative-resistance crystal oscillator. (A) Series-mode circuit (B) Parallel-mode circuit

In the parallel circuit, the instantaneous value of R_{es} is defined by the equation:

$$\frac{1}{R_{es}} = \left|\frac{1}{\rho}\right| - \frac{1}{Z_p}$$

Note that for the oscillations to build up in the series circuit, the total effective positive resistance $(R_e + R_{es})$ must be greater at the start and during the build up than when equilibrium is reached. How much greater, of course, will depend upon the difference at each instant between the negative resistance and the true circuit resistance. For oscillations to build up in the parallel circuit, the total effective positive conductance $\left(\frac{1}{Z_p} + \frac{1}{R_{es}}\right)$ must be greater at the start and during the build up when the equilibrium is reached. The value of the additional conductance will at each instant equal the difference between the negative conductance and the actual tuned-circuit conductance.

1-589. Clearly, for oscillations to build up in the two circuits, the respective resistance ratios $\left|\frac{\rho_s}{R_e}\right|$ and $\left|\frac{Z_p}{\rho}\right|$ must be greater than unity at the start of oscillations. Also apparent should be the fact that these ratios will equal unity when equilibrium is reached. This means that the respective negative-resistance devices must permit ρ_s to decrease and ρ to increase as the amplitude of crystal oscillations increases. But suppose that the devices which produce the negative resistances in the two circuits are switched. What then? If the new starting ratios, $\left|\frac{\rho}{R_e}\right|$ and $\left|\frac{Z_p}{\rho_s}\right|$ are less than unity, no oscillations can begin; but if these ratios are equal to or greater than unity, oscillations can start, but they cannot attain a stable equilibrium.

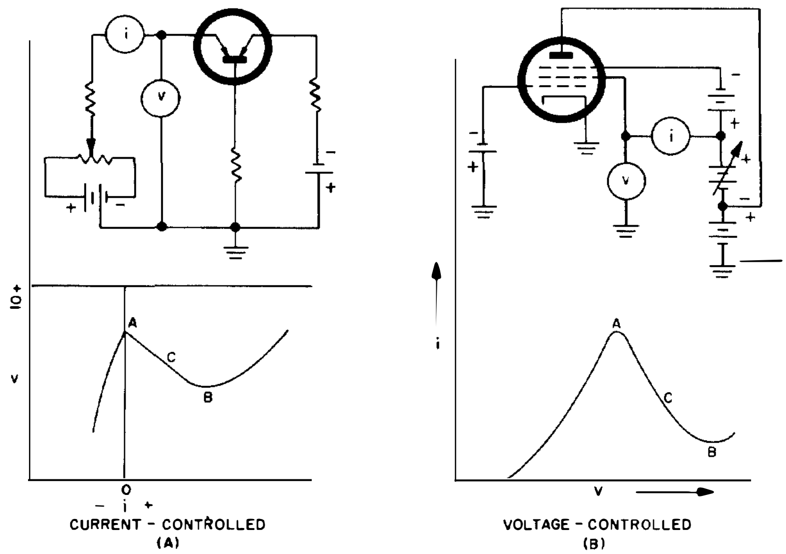


Figure 1-259. Typical negative-resistance devices and characteristics. (A) Transistor emitter current-controlled negative resistance (B) Transitron voltage-controlled negative resistance

Exactly what actions might result from the unstable conditions would depend upon the particular circuit and negative-resistance device.

Current- and Voltage-Controlled Negative-Resistance Characteristics

1-590. Figure 1-259 shows volt-ampere characteristic curves of two representative negative-resistance devices. (A) shows a typical transistor emitter characteristic, whereas (B) shows a typical transitron characteristic. The negative-resistance property is confined to those regions of the curves with a negative slope, as labeled between points A and B. When either of the devices is operated in the negative-slope region, an incremental increase in one of the variables will be accompanied by an incremental decrease in the other variable. Thus, in (A) at a point C between A and B, if the current increases by a very small amount, di, the voltage will decrease by an amount, dv. The ratio, dv/di, is the *dynamic resistance* at the point in question, which in this case is negative since dv and di are opposite in sign. This does not mean that the total resistance, v/i, is necessarily negative, but simply that the resistance to a small variation in current is negative. It is the resistance to be met by a small a-c signal superimposed on the direct current indicated at point C. In curve (A), the slope at each point is equal to the dynamic resistance at that point; in curve (B) the slope is equal to the dynamic conductance at each point.

1-591. Note that for curve (A), one and only one value of voltage corresponds to each value of current. That is, the voltage is a single-valued function of the current. However, note that for a given value of voltage in curve (A), there exist three possible values of current. The current in this case is thus a multivalued function of the voltage. In curve (B), the current is a single-valued function of the voltage, but the voltage is a multivalued function of the current. Where a negative-resistance characteristic is present, at least the voltage or the current must be a multivalued function of the other. Theoretically both can be, but in practical devices one of the parameters will be a single-valued function of the other. That the other must be multivalued is self-evident when we consider that a volt-ampere characteristic having a continuous negative slope would imply an infinite capacity to supply energy. For some value of current and voltage any electronic device can be made to show a positive resistance. But wherever a characteristic curve would show a change from negative to positive resistance, the bend in the curve will require that one of the parameters, v or i, repeat its values as the other continues to increase.

1-592. The characteristic curve, the emitter circuit, and the negative resistance indicated in figure 1-259 (A) are called *current-controlled,* since the voltage is uniquely determined by the current. The transitron circuit and its characteristics in figure 1-259 (B) are called *voltage-controlled,* since the current is uniquely determined by the voltage.

1-593. Imagine that the devices illustrated in figure 1-259 are to be used in crystal oscillator circuits. The no-signal operating currents and voltages are to be such that each device operates at point C. The initial negative resistance is thus the dynamic resistance at that point. As oscillations build up, the average a-c negative resistance in (A) decreases, that in (B) increases. Thus we see that the current-controlled negative resistance is suitable for use with series-mode oscillator circuits, whereas the voltage-controlled negative resistance is suitable for parallel-mode circuits. On the other hand, if a current-controlled negative resistance were to be used to drive a parallel-mode circuit, the oscillations might build up to a point at which the oscillating circuit would suddenly be shorted out, with the sudden loss of stored energy resembling the action of a relaxation oscillator. Or perhaps the circuit will have a tendency to jump from one mode of operation to another.

1-594. The oscillating current of the parallel-mode circuit is essentially self-contained, the only current required of the external circuit is that necessary to re-supply the energy lost each cycle. This is such a small fraction of the total current in a high-Q circuit, that its wave shape can be quite distorted without significantly affecting the sinusoidal shape of the voltage wave. A momentary variation in the impedance of the energy source can produce a large variation in the current wave through a negative-resistance device but only a small variation in the voltage wave. In other words, the energy stored in an antiresonant circuit is primarily a voltage-controlling source rather than a current-controlling source. If used with a device whose current is a multivalued function of the voltage, instability can be expected to result if the voltage amplitude cannot be limited within a single-valued region of the current. A resonant series-mode circuit, on the other hand, is essentially a current-controlling source. Extremely high voltages would be required to produce a significant distortion of the current wave shape of a high-Q resonant circuit connected in series with a negative resistance device. For stable oscillations to be sustained in a series circuit, it is important not to employ a voltage-controlled negative resistance, since its operating characteristics are not uniquely determined by the current.

1-595. If the limiting action is provided by an external circuit component, such as a thermistor, the fact that the negative-resistance is current- or voltage-controlled may not be critical, provided the limiter is certain to act before the oscillations enter an unstable region; nevertheless, the oscillating circuit should be matched to an inherently stable negative resistance for the particular resonance mode. The principal advantage of a thermistor-type limiter is that the limiting action is relatively slow, so that once the equilibrium point is reached, the non-linearity introduced by the limiter is relatively insignificant. Under these circumstances all the circuit elements, in particular the negative-resistance source, can be operated as linear parameters, thereby permitting the oscillations (both current and voltage waveforms) to be sinusoidal.

Frequency Deviation Due to Nonlinear Limiting

1-596. As is normally the case, the limiting is provided by the negative-resistance device. The oscillations build up until the amplitude swings into one or both of the bends of the negative-resistance characteristic. The farther that the current or voltage (depending upon whether current- or voltage-controlled, respectively) swings past the straight portion of the negative-resistance curve, the greater the distortion in the respective fundamental wave of the dependent parameter. If the swings are extended sufficiently, a point will be reached at which any further increase in the amplitude of the controlling parameter results in a decrease in the amplitude of the dependent parameter at the fundamental frequency. If the amplitude of the controlling parameter is increased further, the amplitude of the dependent parameter at the fundamental frequency will diminish until it disappears entirely, at that point where the effective negative resistance, itself, disappears. Beyond this critical point the fundamental of the dependent parameter reappears, but with its phase reversed, reflecting the fact that the average a-c resistance at the fundamental frequency has changed from negative to positive. Although the fundamental frequency of the dependent parameter may decrease in amplitude as the fundamental frequency of the controlling parameter increases in amplitude, the higher harmonics of the dependent parameter continue to increase, so that it is quite possible for the overtones to exceed the level of the fundamental in high-amplitude systems. On this account, where the fundamental frequency is of principal concern, it is generally preferable to obtain the output in a way that it depends upon the wave shape of the parameter, voltage or current, that controls the negative resistance. On the other hand, if harmonic outputs are desired, let them be obtained primarily with the aid of the dependent parameter.

1-597. When harmonics are being generated, the distortion causes the fundamental frequency to assume a slightly different value than would otherwise be the case. This property can be the source of considerable frequency instability in crystal oscillators, since the harmonic content, and hence the fundamental frequency, will vary with the crystal-unit effective resistance, the operating voltages, and the aging and temperature characteristics of the negative-resistance device.

1-598. Briefly, the reason that the fundamental frequency is affected by the presence of harmonics can be explained as follows: First, the mixing of all the harmonics in the nonlinear negative-resistance circuit provides a secondary source of the fundamental frequency, since the fundamental will result as the different frequency between any two successive harmonics. For example,

$$f = 3f - 2f = 4f - 3f, \text{ etc.}$$

Now, if the external crystal circuit is tuned to be approximately resonant at the fundamental frequency, it will generally appear capacitive to the higher harmonic currents. This means that the fundamental frequency obtained from the difference products of the harmonic currents will be displaced approximately 90 degrees in phase from the main fundamental, thereby causing a displacement in the phase of the total fundamental current. Thus, the nonlinear resistance serves to introduce a difference in phase between the current and voltage at the fundamental frequency, and in this manner behaves as if its impedance were partly reactive. If the negative-resistance device is effectively partly reactive, the external circuit must present an equal and opposite reactance in order for the net impedance around the circuit to be zero. The fact that the harmonic currents pass through the capacitive shunts for the most part means that the external circuit must appear inductive at the operating frequency. For a simple parallel-tuned circuit, this requires that the circuit operate at a frequency below resonance. The frequency displacement to be expected can be approximately obtained from the following equation. The equation was derived by J. Groszkowski.* It is based upon the postulate that a negative-resistance device cannot store energy.

$$\sum_{n=1}^{\infty} n I_n^2 X_n = 0 \qquad \text{1-598 (1)}$$

Where:

n = number of harmonic

I_n = current of nth harmonic

X_n = reactance of circuit to nth harmonic

This equation says that the sum of all the products, $nI_n^2X_n$, must equal zero. Since X_n for all values of n greater than 1 can generally be assumed to be negative, the crystal-circuit reactance facing the negative resistance terminals at the fundamental frequency must be positive (inductive) for the equation to hold.

* Note: See Bibliography No. 305. (Also W. A. Edson, No. 211.)