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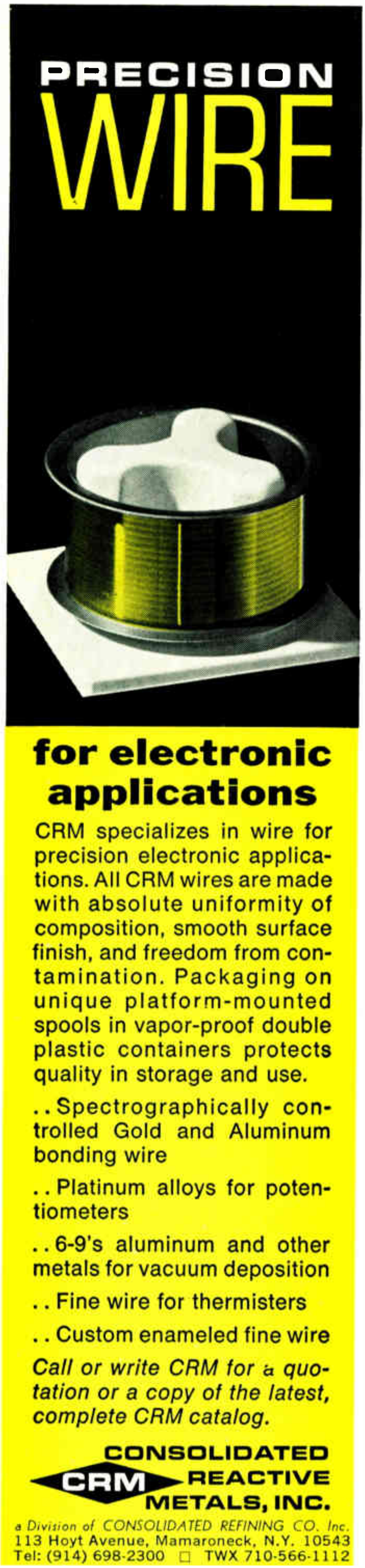
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World Radio History

Forum

Readers are invited to comment in this department on material previously published in IEEE SPECTRUM; on the policies and operations of the IEEE; and on technical, economic, or social matters of interest to the electrical and electronics engineering profession.

World population growth

The article by Austin and Brewer on "World Population Growth and Related Technical Problems" in the December 1970 issue is truly *profound!* It represents an outstanding job by the authors in researching many fields and distilling the resulting information into a highly readable form. It also represents a supremely worthwhile dedication on the part of *IEEE Spectrum* to the presentation of selected social issues to its audience.

It is clear that by 1985 there will be a world food crisis.

I think, therefore, that each engineer must consciously consider what role he will play in the impending collision between the quality of life and the population explosion.

Frank H. Myers
Bell Laboratories
Columbus, Ohio

Austin and Brewer provide one model of population growth that predicts a maximum population of about 30 to 50 billion. A different model will, of course, give a different prediction. Forrester's computer simulations ("Counterintuitive Behavior of Social Systems," *Tech. Rev.*, Jan. 1971) give maximums of five to six billion. And most of his simulations show a catastrophic decline to around two billion by the middle of the 21st century.

Both models need to be studied for their implications of societal consequences. Forrester's model seems to indicate that intensive study should be given immediately.

John K. Mickelsen
Syracuse University Research Corp., Syracuse, N.Y.

It is pleasing to find that Austin and Brewer's estimate of the energy use in 2100 A.D.—10 Q per year, from their Fig. 4—agrees exactly with the figure that I gave in 1967[1] for the same year, but it is difficult to reconcile their estimate of world population at that time with present trends. They show the population as continuing to have a positive second differential at least until 2050 A.D. (their Fig. 1), but assume that the energy consumption per capita per year suddenly becomes linear at about 1960. The available figures, from the *Statistical Abstract of the U.S., 1970,* would seem to warrant exactly the opposite conclusion: it is well known that the birthrate has dropped markedly in the more advanced countries, and this is borne out by Table 1250 in the Abstract, which shows that from 1960 to 1968 the U.S. population rose by 11.3 percent while the world population rose by 15.9 percent. During the same period the U.S. per capita energy use (from Table 774) rose by 25.5 percent (from 248 to 310 × 10^6 Btu), while the per capita world use rose slightly less, 23.5 percent, from 41 to 50.4 × 10^6 Btu. The evidence is, therefore, that a technically advanced nation does reduce its rate of population growth, but far from linearizing its use of energy per head, it increases it even faster than the rest of the world. I do not know of any reason to suppose that the underdeveloped countries will not go through the same process, as they reach and surpass our present standard. Consequently, I would project both curves A and B of Austin and Brewer's Fig. 3 with continuing increases of slope, but cut back the rate of increase in Fig. 1 so that the slope of the upper part conforms more closely to the established lower rate—11.3 percent in eight years, for example—of an advanced nation, and drops to about one percent per year after 2050, when the world standard will have passed the present U.S. standard of energy use per capita; this last part is about the same as the 60 percent per half century that I assumed (as a deliberately conservative estimate based on Putnam's) in 1967. Superimposed on the present higher rates, it leads to a population of about 16 billion by 2100. It may well be that before then the human race will have learned that it can cut the birthrate to the point of stabilizing the population, and can then indulge its passion for gadgets up to the limit I gave previously, the thermal balance of the earth.

A point that seems to escape notice in Austin and Brewer's discussion of thermal pollution by power stations is that it is not only the waste heat at the station that causes thermal pollution: substantially all of the electric energy sent out, however it is used, ends up as low-grade heat. The same is true of almost all the fuel we use: whatever the primary purpose of its conversion, and however efficiently this is done, nearly all the energy becomes heat at the surface of the earth. A little becomes chemical energy of synthesized compounds, and even less, I would guess, becomes potential energy or is radiated into space at a high temperature.

If I had any sure way of collecting, I would bet that the population will stabilize at about 12 billion by 2100 A.D., and that they will find about 3 Q per year is as much as they can handle without devastating the earth, even with the most heroic measures to distribute it uniformly. This would give them a material living standard about five times the present U.S. standard, nearly the same ratio as that between the U.S. and the world as a whole now. Obviously it is difficult to imagine how they will use this much energy—if we could imagine it, we would be doing it now—but presumably the Indian peasant cannot really imagine how the average American can have 15 horsepower working for him night and day, year in and year out. Yet, in a century, his great-grandson will wonder how one ever got along without that and more.

J. Rodney M. Vaughan
Redwood City, Calif.

REFERENCE

1. Vaughan, J. R. M., *IEEE Spectrum* (correspondence), p. 196, Mar. 1967.

Solid-state electronics

Mr. Barbe's letter in the September issue of *IEEE Spectrum* points to an actively felt need for an IEEE Group in the area of solid-state electronics. Electrical engineers cannot conduct research exclusively in devices without getting involved in electronic properties of materials from which devices are fabricated. A good example is the ovonic switching devices recently proposed, which require a good understanding of electronic processes in amorphous materials in order to understand the switching action of the device. Realization of this fact was the reason for the development of an area in universities variously called electronic materials, solid-state electronics, or simply solid state.

As Mr. Barbe mentions, researchers in this area have been forced to publish in journals and attend conferences of other scientific societies. Although there is an IEEE Group on Electrical Insulation, it is a pity there is no Group serving the general area of solid state, which includes areas like semiconductors, thin films, etc.—an area that has contributed so much, and will continue

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to do so in the future, to the development of solid-state devices.

Therefore, an IEEE Group in this area, which may be called either Solid-State Electronics or simply Solid State, would be most welcome.

T. S. Jayadevaiah
University of Wisconsin
Milwaukee, Wis.

We appreciate Mr. Jayadevaiah's comments concerning a Solid-State Electronics Group in IEEE, and the earlier letter from Mr. Barbe. As was pointed out in the summary of Technical Activities Board planning in the August *IEEE Spectrum* (pp. 113–116), this is one of the areas that TAB has considered this year.

It is generally the custom to start a new area through an *ad hoc* committee, which may sponsor conferences and publish newsletters, but whose primary responsibility is the recommendation as to whether a new IEEE Group or Council is needed, or if the subject can be taken care of by existing Groups.

The materials area presents a special problem in that materials papers are accepted in a number of the *IEEE Transactions*, including those on electron devices, insulation, and magnetics, and the *Journal of Quantum Electronics*. Many members also wish to publish their less applied papers in physics or chemical journals in order to keep a live interchange with the physicists and chemists working in related fields. We thus have many members who feel that it would be a mistake to set up a Group that would in any sense duplicate work and publication of the American Physical Society, the American Chemical Society, or the Electrochemical Society.

This is by no means a closed issue. The Technical Activities Board will welcome comments on any side of this issue. It will be helpful to have those favoring a new Group mention specific areas that are not taken care of by one of our *Transactions* or the physics and chemical journals. Letters may be sent to the Technical Activities Board, in care of R. M. Emberson at IEEE Headquarters.

J. R. Whinnery
1970 Vice Chairman, TAB

No unemployment solution?

I wish to take issue with M. Silverstein's letter to the President, which you published in "Forum" in the December 1970 issue. For years now, engineers in government-supported work have made substantially higher salaries than their counterparts in nongovernment-supported industry. Some, unfortunately, lost their engineering perspective and became highly skilled in a narrow field, thus limiting their future employment opportunities. Few have had the business sense to save part of their high salaries for that inevitable day when, as Mr. Silverstein so aptly puts it, "the axe fell."

Since the taxpayers have for years contributed to the higher salaries of these engineers, I do not feel that they should now be asked to finance re-education, new businesses, or any other programs that the engineers should have saved for over the years.

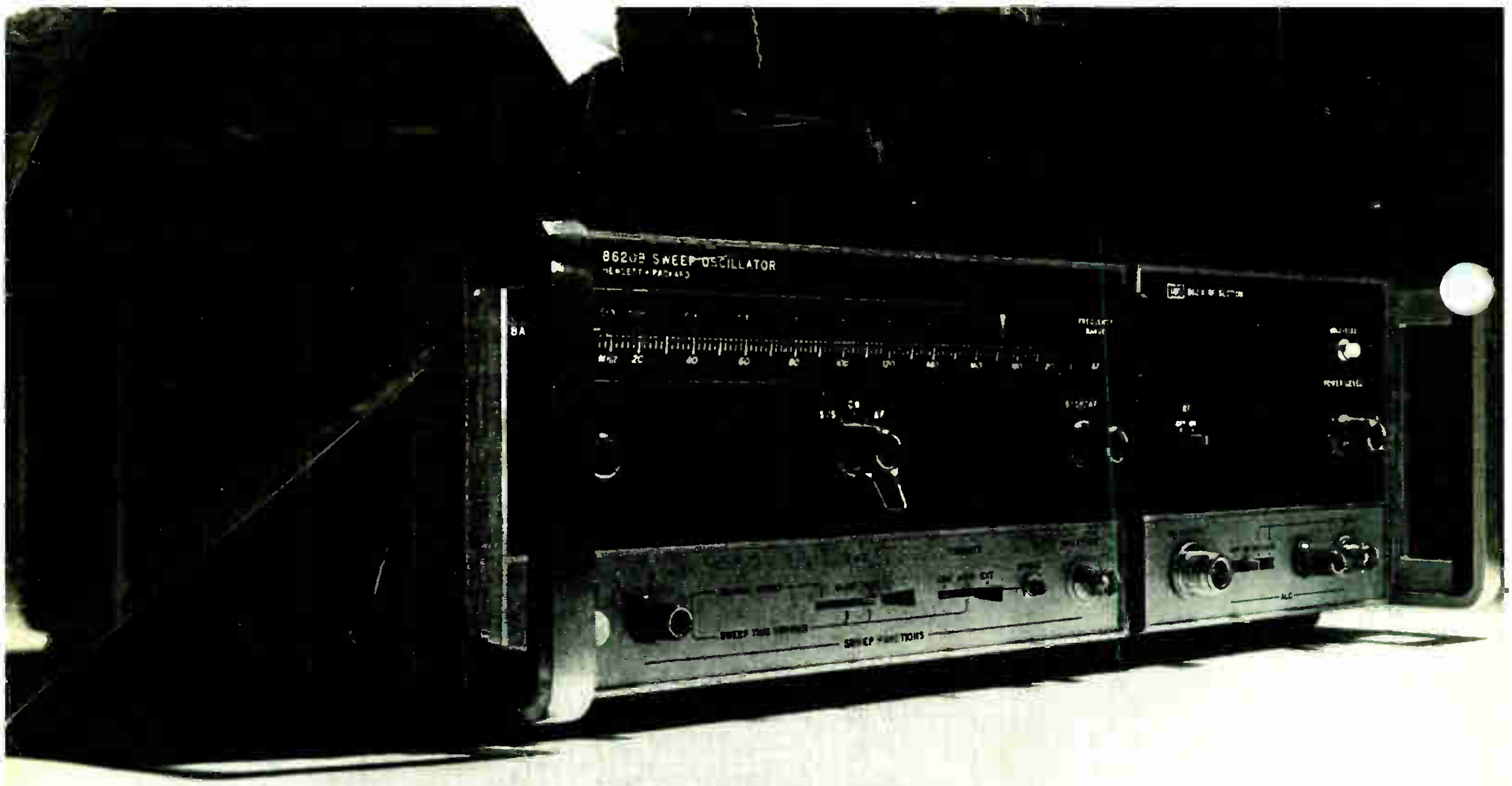
John R. Truitt
Chicago, Ill.

Correction

The article "World Population Growth and Related Technical Problems" by A. L. Austin and J. W. Brewer contains an error of omission in Fig. 1 (p. 46) that can be corrected by adding the numbers one and ten in increasing order to the right-hand ordinate (fertility rate r). In other words, the curve for Eq. (8) is asymptotic to the upper bound 10 percent.

In addition, the second paragraph in column one of page 52 is incorrectly stated. It should properly read (items italicized were omitted from the published article):

"Wheat is the only well-rounded bulk food that requires few changes in dietary habits, and is the only effective food that can be produced and distributed on a large scale. Canada, Australia, Argentina, and the U.S. are the only countries that produce more wheat than internally needed. For economic reasons, the U.S. has been the only country capable of supplying wheat free of charge to undernourished countries, and will likely remain the major supplier. Paddock and Paddock[35] have made an extensive study of world food needs and supplies, and conclude that serious shortages will develop sometime between 1975 and 1985. Figure 12 summarizes their results. *It should be noted that these curves do not include the needs of mainland China and Russia.* They are special cases since accurate data are not available, past politics may preclude U.S. food aid, and the U.S.S.R. is not usually considered an underdeveloped country in need of assistance. *Paddock and Paddock, however, point out that since 1964 Russia has changed from exporting wheat to importing it.* They point out that, even if fully cultivated, the U.S. will not have enough foodstuffs to satisfy future needs of all hungry nations, and may soon be in the position of selectivity providing food aid or rationing food surpluses. *It can only be speculated that political tensions between the three major powers may increase as food shortages increase.*"



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Spectral lines

Professional integrity. *Ralph Nader's Jan. 15 editorial in* The New York Times *is reprinted here because he has effectively used public presentation of cases of neglect of the public interest to promote legislation and the movement he represents may be even more successful in the future. His remedy, the ability to "turn the company in," is a clumsy means of last resort. The real remedy lies with ourselves, the practicing engineers. We make the technical proposals and formulate the nature of new products and programs. We propose the improvements in existing products. We must adopt the professional view that our efforts serve the public and that our companies deserve and can afford our services because what we propose and achieve is a profitable public service.—David DeWitt, Editor*

At what point should corporate or government scientists, engineers or other professionals dissent openly from their employer-organization's policy? If the professional does dissent, what is there to protect or defend his decision to place his professional conscience over what he believes is his organization's illegal, hazardous or unconscionable behavior?

These are important questions and they are rarely answered in the context of controversies such as the defoliation of Vietnam or the standards for constructing nuclear power plants. "Duty," said Alfred North Whitehead, "arises from our potential control over the course of events." Staying silent in the face of a professional duty, almost invariably articulated in the profession's canons of ethics, has direct impact on the level of consumer and environmental hazards. This awareness has done little to upset the slavish adherence to "following company orders."

Employed professionals are among the first to know about industrial dumping of mercury or fluoride sludge into waterways, defectively designed automobiles, undisclosed adverse effects of prescription drugs and pesticides. They are the first to grasp the technical capabilities to prevent existing product or pollution hazards. But they are very often the last to speak out, much less refuse to be recruited for acts of corporate or governmental negligence or predation.

The twenty-year collusion by the domestic automobile companies against development and marketing of exhaust control systems is a tragedy, among other things, for engineers who, minion-like, programmed the technical artifices of the industry's defiance. Settling the antitrust case brought by the Justice Department against such collusion did nothing to confront the question of subverted engineering integrity.

A prime foundation for professionalism is sufficient independence to pursue a mission that could save lives, secure rights, or preserve property unjustly imperiled by the employer-organization. The overriding ethic of the professional is to foresee and forestall the risks to which he is privy by his superior access and knowledge, regardless of vested interests. Physicians should strive first to prevent disease; lawyers should apply the law to prevent auto casualties; economists should try to clarify product and service characteristics in the context of quality competition; engineers should make technology more humane as a condition of its use; scientists should anticipate the harmful uses of their genius.

All these ideal missions unfortunately possess neither the outside career roles for their advancement nor the barest of independence for the organizationally employed professional to exert his conscience in practice beyond that of the employer's dictates. The multiple pressures and sanctions of corporate and government employers are very effective to daunt the application of professional integrity. When on occasion such integrity breaks through these restraints, the impact is powerful, which might explain the organization's determined policy of prior restraint.

During the past half dozen years of disclosures about corporate and government injustices, the initiators have largely been laymen or experts who were outsiders to the exposed. The list is legion—black lung, brown lung, DDT, mercury contamination, enzymes, phosphates and NTA in detergents, SST hazards... MER-29, and nerve gas storage and disposal. Inside the systems, however, mum's the word.

Three basic changes are needed as a start.

First, Congress should enact legislation providing for safeguards against arbitrary treatment by corporations against employes who exercise their constitutional rights in a lawful manner. At a minimum, such an act would help Congress obtain expert witnesses for its hearing and authorize the courts to protect a professional's "skill rights" in a far more defined manner.

Second, employed professionals should organize to provide a solid constituency for the adoption by management of the requisite due process procedures, which the professional can appeal to or enforce in the courts.

Third, professional societies should clearly stake out their readiness to defend their colleagues when they are arbitrarily treated for invoking their professional ethics toward the corporate or government activity in which they were involved. Most of the established professional societies or associations never challenge corporate or governmental treatment of lawyers, engineers, scientists, or physicians as the American Association of University Professors has done on occasion for university teachers denied academic freedom. And where there is no willingness to challenge, there is less willingness for the employee to dissent.

To require an act of courage for stating perceived truth is to foster a system of self-censorship and the demise of individual conscience against the organization.

The monolithic phase-locked loop—a versatile building block

Costly and complex in its discrete form, the phase-locked loop has developed into a low-cost monolithic package that has made available a wide range of new applications

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The basic concept of a phase-locked-loop (PLL) system, known since the early 1930s, has been widely used in a variety of applications in instrumentation and space telemetry. However, because of its cost and complexity in discrete (nonintegrated) system design, its application has been limited to precision measurements requiring a high degree of noise immunity and very narrow bandwidths. With the recent developments in integrated circuit design and processing technology, this situation is rapidly changing. The monolithic PLL is now emerging as a new and versatile building block, similar to the monolithic operational amplifier in the diversity of its applications. The availability of a low-cost PLL system in a monolithic circuit package opens up many new applications that cost and complexity previously precluded. The purpose of this article is to acquaint the reader with the basic principles and design parameters of an integrated PLL, as well as with the wide range of applications it offers in both analog and digital signal processing.

In the design of frequency-selective integrated circuits, the lack of integrated inductors is a very serious drawback. Active *RC* filters, in which one can use a number of resistors and capacitors in feedback around a gain block to obtain a frequency-selective response present a possible solution to this problem. A large number of design techniques have been developed for active *RC* filters, and these are well covered in the literature.[1–4] There are three basic limitations to integrated filters using active-*RC* techniques: (1) frequency range—stability considerations limit the use of most of these filters to below 100 kHz; (2) sensitivity—the center frequency, as well as the selectivity, are very sensitive functions of active gains or the absolute values of feedback components; (3) cost—typically, four precision components (*R*'s and *C*'s) are required for each complex pole pair. Since these components cannot, in general, be fabricated in a monolithic form, they have to be externally connected to the circuit package, adding to cost and complexity of design.

The purpose of this article is to survey an alternate design approach to monolithic frequency-selective circuits, based on the phase-locked-loop principle. In many applications, phase-locked integrated circuits overcome the basic drawbacks of active filters, and offer significant cost and performance advantages in digital or analog signal processing.

The basic concept of a phase-locked loop is by no means new. It has been known since the early 1930s,[5] and has been used for a variety of applications in instrumentation and space telemetry. Cost and complexity in discrete (nonintegrated) system design, however, have limited its use to precision measurements requiring very narrow bandwidths or a high degree of noise immunity. Therefore, what is "new" at this point is not the idea of a PLL but its availability in the form of a low-cost, self-contained monolithic circuit package. In many ways, this is similar to the case of the operational amplifier, which until less than a decade ago was a fairly expensive building block used predominantly in analog computation. With the advent of monolithic circuit technology, however, it has become a basic building block in many system designs. The monolithic phase-locked loop, although not as versatile as an op amp, offers a similar potential. In fact, many of the applications for a PLL described in this article become economically feasible only because the PLL is now available as a low-cost monolithic building block.

The phase-locked loop

Basic principles of operation. The phase-locked loop is a frequency feedback system comprised of a phase com-

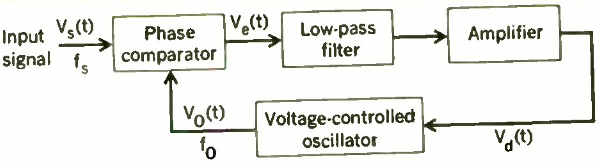


FIGURE 1. Block diagram of a phase-locked loop.

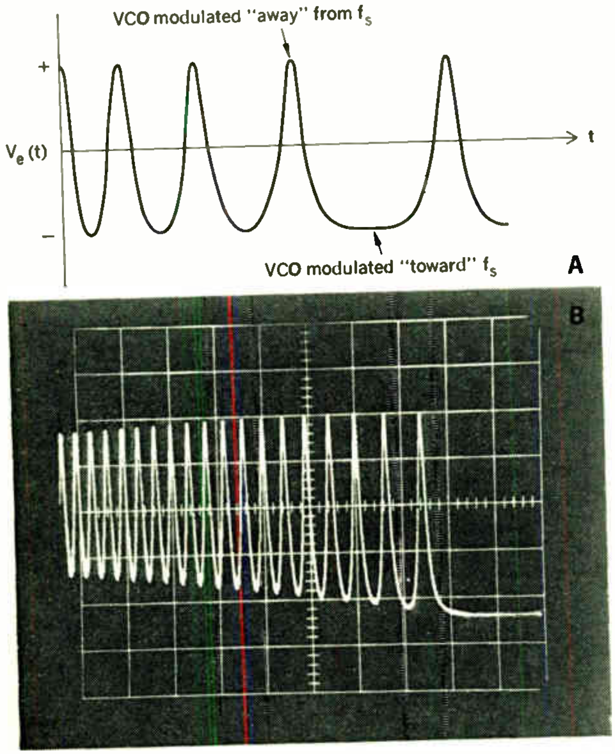


FIGURE 2. Asynchronous error beat note during capture process. A—Typical beat-note waveform. B—Oscillogram of capture transient.

parator, a low-pass filter, and an error amplifier in the forward signal path, and a voltage-controlled oscillator (VCO) in the feedback path. The block diagram of a basic PLL system is shown in Fig. 1. Detailed analysis of the PLL as a nonlinear control system has been discussed in the literature.[6-9]

A rigorous mathematical analysis of the system is quite cumbersome and will not be repeated here. However, from a qualitative point of view, the basic principle of operation of the PLL can be briefly explained as follows: With no signal input applied to the system, the error voltage V_d is equal to zero. The VCO operates at a set frequency f_0, which is known as the "free-running" frequency. If an input signal is applied to the system, the phase comparator compares the phase and the frequency of the input with the VCO frequency and generates an error voltage $V_e(t)$ that is related to the phase and the frequency difference between the two signals. This error voltage is then filtered, amplified, and applied to the control terminal of the VCO. In this manner, the control voltage $V_d(t)$ forces the VCO frequency to vary in a direction that reduces the frequency difference between f_0 and the input signal. If the input frequency f_s is sufficiently close to f_0, the feedback nature of the PLL causes the VCO to synchronize or "lock" with the incoming signal. Once in lock, the VCO frequency is identical to the input signal, except for a finite phase difference. This net phase difference ϕ_0 is necessary to generate the corrective error voltage V_d to shift the VCO frequency from its free-running value to the input signal frequency f_s, thus keeping the PLL in lock. This self-correcting ability of the system also allows the PLL to "track" the frequency changes of the input signal once it is locked. The range of frequencies over which the PLL can maintain lock with an input signal is defined as the "lock range" of the system. This is always larger than the band of frequencies over which the PLL can acquire lock with an incoming signal. This latter range of frequencies is known as the "capture range" of the system.

The capture process is highly complex, and does not lend itself to simple mathematical analysis. However, a heuristic and highly qualitative description of the capture mechanism may be given as follows: Since frequency is the time derivative of phase, the frequency and the phase errors in the loop can be related as

$$2\pi\Delta f = \frac{d\phi_0}{dt}$$

where Δf is the instantaneous frequency separation between the signal and VCO frequencies.

If the feedback loop of the PLL were opened, say between the low-pass filter and the VCO control input, then for a given setting of f_0 and f_s the phase-comparator output will be a sinusoidal beat note at a fixed frequency Δf. If f_s and f_0 were sufficiently close in frequency, then this beat note would appear at the filter output with negligible attenuation. Now suppose that the feedback loop is closed by connecting the low-pass filter output to the VCO control terminal. Then the VCO frequency would be modulated by the beat note; and when this happens, Δf itself becomes a function of time. If, during this modulation process, the VCO frequency moves closer to f_s (i.e., decreasing Δf), then $d\phi_0/dt$ decreases and the output of the phase comparator becomes a slowly varying function of time. Similarly, if the VCO is modulated away from f_s, $d\phi_0/dt$ increases and the error voltage becomes a rapidly varying function of time. Therefore, under this condition, the beat-note waveform no longer looks sinusoidal; instead it looks like a series of aperiodic "cusps," depicted schematically in Fig. 2(A). Because of its asymmetry, the beat-note waveform contains a finite dc component that pushes the "average" value of the VCO toward f_s, thus decreasing Δf. In this manner, the beat-note frequency rapidly decreases toward zero, the VCO frequency drifts toward f_s, and the lock is established. When the system is in lock, Δf is equal to zero and only a steady-state dc error voltage remains.

Figure 2(B) displays an oscillogram of the loop error voltage V_d in an actual PLL system during the capture process. Note that, as lock is approached, Δf is reduced, the low-pass filter attenuation becomes less, and the amplitude of the beat note increases.

The total time taken by the PLL to establish lock is called the "pull-in" time. Pull-in time depends on the initial frequency and phase differences between the two signals, as well as on the overall loop gain and the low-pass-filter bandwidth. Under certain conditions, the pull-in time may be shorter than the period of the beat note, and the loop can lock without an oscillatory error transient.

In the operation of the loop, the low-pass filter serves a dual function. First, by attenuating the high-frequency error components at the output of the phase comparator, it enhances the interference-rejection characteristics; second, it provides a short-term memory for the PLL and ensures a rapid recapture of the signal if the system is thrown out of lock due to a noise transient. Since the low-pass filter attenuates the high-frequency error voltage within the loop, it directly controls the capture and the transient-response characteristics of the PLL. The reduction of the filter bandwidth has the following effects on system performance:

1. The capture process becomes slower, and the pull-in time increases.
2. The capture range decreases.
3. Interference-rejection properties of the PLL improve since the error voltage caused by an interfering frequency is attenuated further by the low-pass filter.
4. The transient response of the loop, i.e., the response of the PLL to sudden changes of the input frequency within the capture range, becomes underdamped.

The last effect also brings about a practical limitation on the low-pass loop filter bandwidth and roll-off characteristics from stability considerations. These points will be explained further in the following section.

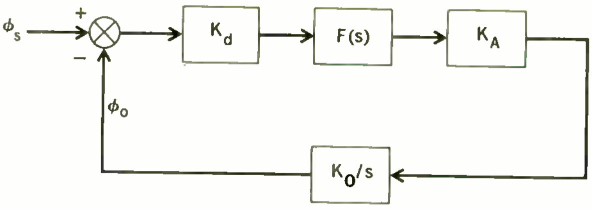


FIGURE 3. Linearized model of the PLL as a negative feedback system.

FIGURE 4. PLL root locus for a simple lag filter ($\tau_1 = R_1C_1$).

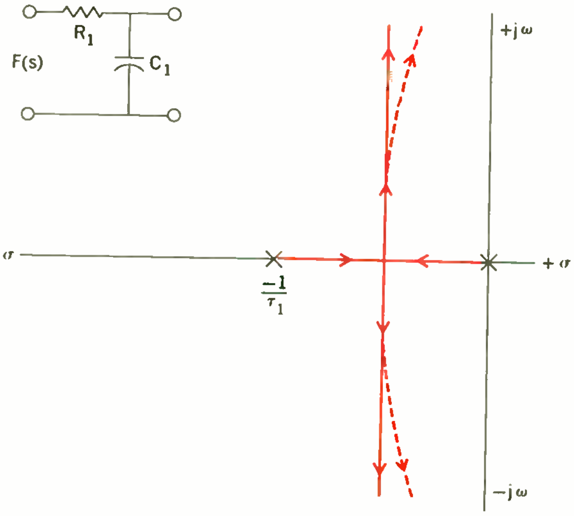
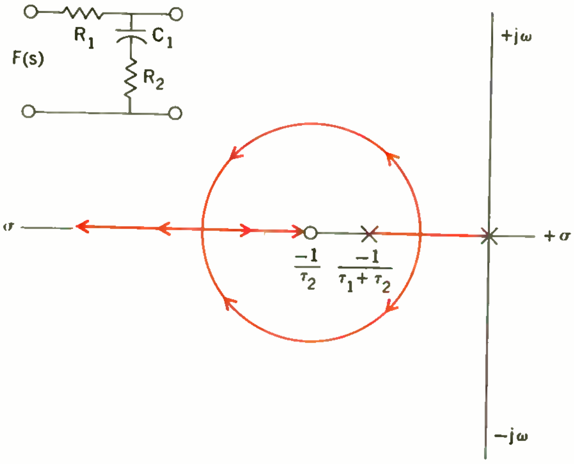


FIGURE 5. PLL root locus for lag-lead-type filter ($\tau_1 = R_1C_1$, $\tau_2 = R_2C_1$).



System parameters. When the PLL is in lock, the nonlinear capture transients are no longer present. Therefore, under lock condition, the PLL can often be approximated as a linear control system (see Fig. 3), and can be analyzed using Laplace transform techniques. In this case, it is convenient to use the net phase error in the loop ($\phi_s - \phi_0$) as the system variable. Then, each of the gain terms associated with the blocks can be defined as follows:

K_d = conversion gain of phase detector (V/rad)
$F(s)$ = transfer characteristic
K_A = amplifier voltage gain
K_0 = VCO conversion gain (rad/V · s)

Note that, since the VCO converts a voltage to a frequency and since phase is the integral of frequency, the VCO functions as an integrator in the feedback loop.

The open-loop transfer function for the PLL can be written as

$$T(s) = \frac{K_T F(s)}{s}$$

where K_T is the total loop gain, i.e., $K_T = K_d K_A K_0$. Using linear feedback analysis techniques, the closed-loop transfer characteristics $H(s)$ can be related to the open-loop performance as

$$H(s) = \frac{T(s)}{1 + T(s)}$$

and the roots of the characteristic system polynomial can be readily determined by root-locus techniques. Figure 4 shows the PLL root loci as a function of the total loop gain K_T for a single-pole low-pass filter $F(s)$ of the form

$$F(s) = \frac{1}{1 + \tau_1 s}$$

where $\tau_1 = R_1C_1$. In the illustration, the open-loop pole at the origin is due to the integrating action of the VCO.

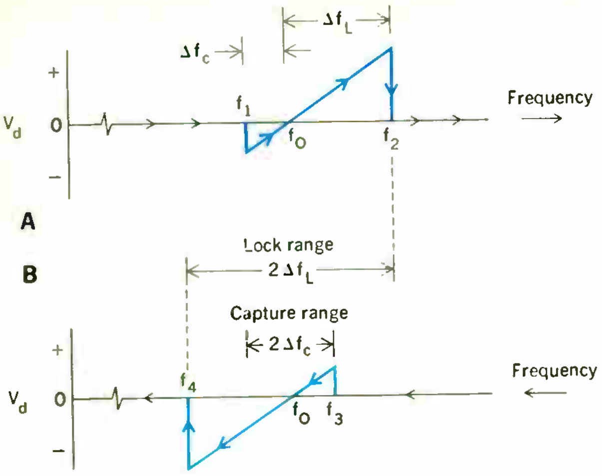


FIGURE 6. Typical PLL frequency-to-voltage transfer characteristics for increasing (A) and decreasing (B) input frequency.

One can make the following observations from the root-locus characteristics of the diagram:

1. As the loop gain K_T increases for a given choice of τ_1, the imaginary part of the closed-loop pole increases; thus, the natural frequency of the loop increases and the loop becomes more and more underdamped.
2. If the filter time constant τ_1 is increased, the real part of the closed-loop poles becomes smaller, and the loop damping is reduced.

As in any practical feedback system, excess shifts or nondominant poles associated with the blocks within the PLL can cause the root loci to bend toward the right-half plane, as shown by the dashed line in Fig. 4. This is likely to happen if either the loop gain or the filter time constant is too large, causing the loop to break into sustained oscillations.

The stability problem can be eliminated by using a lag–lead type of filter, as indicated in Fig. 5. This type of a filter has the transfer function

$$F(s) = \frac{1 + \tau_2 s}{1 + (\tau_1 + \tau_2)s}$$

where $\tau_2 = R_2C_1$ and $\tau_1 = R_1C_1$. By proper choice of R_2, this type of filter confines the root locus to the left-half plane and ensures stability. However, it also increases the noise bandwidth of the system and decreases its interference rejection since the high-frequency error components in the loop are now attenuated to a lesser degree.

In terms of the basic gain expressions in the system, the lock range of the PLL, $\Delta\omega_L$, can be shown to be numerically equal to the dc loop gain[6]

$$\Delta\omega_L = 2\pi\Delta f_L = K_T$$

Since the capture range $\Delta\omega_c$ denotes a transient condition, it is not as readily derived as the lock range. However, following Moschytz' analysis,[8] an approximate expression for the capture range can be written as

$$\Delta\omega_c = 2\pi\Delta f_c \approx K_T|F(j\Delta\omega_c)|$$

where $F(j\Delta\omega_c)$ is the low-pass-filter amplitude response at $\omega = \Delta\omega_c$. It should be noted that, since at all times $|F(j\Delta\omega_c)| \leqslant 1$, the capture range is always smaller than the lock range. If the simple lag filter of Fig. 4 is used, the capture-range equation can be approximated as

$$\Delta\omega_c \approx \sqrt{\frac{\Delta\omega_L}{\tau_1}} = \sqrt{\frac{K_T}{\tau_1}}$$

Thus, the capture range decreases as the low-pass-filter time constant is decreased whereas the lock range is unaffected by the filter and is determined solely by the loop gain.

Figure 6 shows the typical frequency-to-voltage transfer characteristics of the PLL. The input is assumed to be a sine wave whose frequency is swept slowly over a broad frequency range; the vertical scale is the corresponding loop error voltage. In Fig. 6(A), the input frequency is being gradually increased. The loop does not respond to the signal until it reaches a frequency f_1, corresponding to the lower edge of the capture range. Then, the loop suddenly locks on the input, causing a negative jump of the loop error voltage. Next, V_d varies with frequency with a slope equal to the reciprocal of VCO gain ($1/K_0$) and goes through zero as $f_s = f_0$. The loop tracks the input until the input frequency reaches f_2, corresponding to the upper edge of the lock range. The PLL then loses lock, and the error voltage drops to zero. If the input frequency is now swept slowly back, the cycle repeats itself as shown in Fig. 6(B). The loop recaptures the signal at f_3 and traces it down to f_4. The frequency spread between (f_1,f_3) and (f_2,f_4) corresponds to the total capture and lock ranges of the system; that is,

$$f_3 - f_1 = 2\Delta f_c \qquad \text{and} \qquad f_2 - f_4 = 2\Delta f_L$$

Note that, as indicated by the transfer characteristics of Fig. 6, the PLL system has an inherent selectivity about the center frequency set by the VCO free-running frequency f_0; and it will respond only to the input signal frequencies that are separated from f_0 by less than Δf_c or Δf_L, depending on whether the loop starts with or without an initial lock condition. It is also worth noting that the linearity of the frequency-to-voltage conversion characteristics for the PLL are determined solely by the VCO conversion gain. Therefore, in most PLL applications, the VCO is required to have a highly linear voltage-to-frequency transfer.

Basic applications. As a versatile building block, the phase-locked loop is suitable for a wide variety of frequency-selective demodulation, signal-conditioning, or frequency-synthesis applications. Some of these basic applications are listed in Table I. The operation of the PLL in each of these numerous applications will now be briefly described.

1\. *FM demodulation.* If the PLL is locked on a frequency-modulated (FM) signal, the VCO tracks the in-

I. Applications of a phase-locked loop

1. FM demodulation
   (a) Broadcast FM detection
   (b) AM/FM telemetry decoding
   (c) FSK demodulation
2. Frequency synchronization
3. Signal conditioning
4. Frequency multiplication/division
5. Frequency translation
6. AM detection

stantaneous frequency of the input. The filtered error voltage, $V_d(t)$, which forces the VCO to maintain lock with the input signal, corresponds to the demodulated output. In this case, the linearity of the demodulated output is determined solely by the VCO voltage-to-frequency conversion characteristics (see Fig. 6). The PLL can be used for detecting either wide-band (high-deviation) or narrow-band FM signals with a higher degree of linearity than can be obtained by other detection means.[10] It is worth mentioning that, for FM deviation purposes, the PLL functions as a self-contained receiver system since it combines the functions of frequency selection and demodulation.

It should be noted that since the PLL is in lock during the FM demodulation process, the frequency response as well as the rise time of the demodulated output can be readily predicted from the root-locus plots of Fig. 4 or 5.

In the case of frequency-shift-keyed (FSK) data transmission, the digital information is transmitted by switching the input frequency between any one of the two discrete input frequencies, corresponding to a digital "one" and a digital "zero," respectively. When the PLL is locked on an FSK input signal, the error voltage $V_d(t)$, which is in the form of discrete voltage steps, corresponds to the demodulated binary output.

2. *Frequency synchronization.* Using the phase-locked-loop system, the frequency of a relatively poor oscillator such as the VCO can be phase-locked with a low-level but highly stable reference signal. Moreover, the VCO output reproduces the input signal frequency at the same per-unit accuracy as the input reference, but at a much higher power level. In some applications, the synchronizing signal can be in the form of a low-duty-cycle burst at a specific frequency. The PLL can then be used to regenerate a coherent CW reference frequency by locking onto this short synchronizing pulse. A typical example of such an application is seen in the phase-locked chroma-reference generators of color television receivers.[11]

In digital systems, the PLL can be used for a variety of synchronization functions. For example, two system clocks can be phase-locked to each other such that one can function as a backup for the other; or used in synchronizing disk or tape drive mechanisms in information storage and retrieval systems. In pulse-code modulation (PCM) telemetry receivers or in repeater systems, the PLL is used for bit synchronization.[6,12]

3. *Signal conditioning.* By proper choice of the VCO free-running frequency, the PLL can be made to lock onto any one of a number of signals present at the input. Hence the VCO output reproduces the frequency of the desired signal while greatly attenuating the undesired frequencies of sidebands present at the input.[13] If the loop bandwidth is sufficiently narrow, the signal-to-noise ratio at the VCO output can be much better than the input. Thus, the PLL can be used as a noise filter for regenerating weak signals buried in noise.

4. *Frequency multiplication and division.* By inserting a frequency divider into the feedback loop, between the VCO output and the phase-comparator input, the PLL system can function as a frequency-selective frequency multiplier. A block diagram of this configuration is outlined in Fig. 7, where N is the frequency-divider modulus. Now, when the system is in lock, the two inputs to the phase comparator are at the same frequency and $f_0 = Nf_s$.

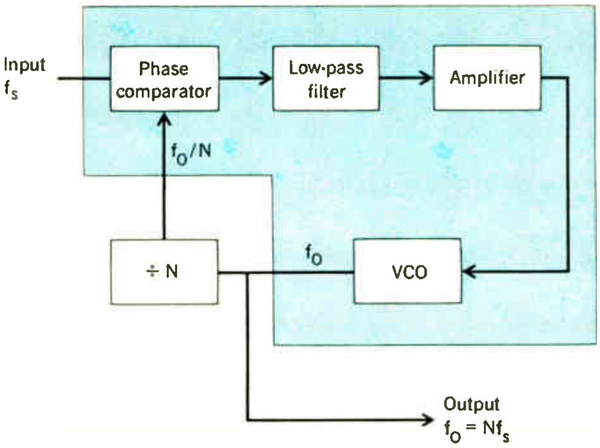


FIGURE 7. Frequency multiplication by using a frequency divider in a PLL.

Under certain conditions, frequency multiplication can also be achieved without the use of a frequency-divider network by operating the PLL in its "harmonic-locking" mode. If a harmonic-rich input signal is used (such as a pulse train), the VCO can be made to lock on the nth harmonic of the input; the VCO fundamental is n times the input frequency, or $f_0 = nf_s$. Similarly, if the VCO produces a harmonic-rich output waveform, then the mth harmonic of the VCO output can be synchronized with the input fundamental. Then, under this condition, the VCO fundamental is a subharmonic of the input frequency; i.e., $f_0 = f_s/m$. When the PLL is operated in the harmonic-locking mode, the spacing between the adjacent harmonics in the frequency spectrum decreases rapidly as the harmonic order n or m is increased. This, in turn, increases frequency-stability requirements for the VCO free-running frequency, to enable the system to differentiate between adjacent harmonics. In integrated phase-locked-loop systems, which use multivibrator-type oscillators,[13,14] thermal drifts of VCO frequency usually restrict the harmonic-lock operation of the system to values of n or $m < 10$. An additional disadvantage of harmonic locking at large values of n or m is that the phase-detector gain K_d decreases inversely with the harmonic order, thus decreasing both the lock and the capture ranges of the system at higher harmonics.

5. *Frequency translation.* The PLL system can be used to translate the frequency of a highly stable but fixed-frequency reference oscillator by a small amount in frequency. This can be achieved by adding a mixer and a low-pass-filter stage to the basic PLL, as shown in Fig. 8. In this case, the reference input f_R and the VCO output f_0 are applied to the inputs of the mixer stage. The mixer output is made up of the sum and the difference components of f_R and f_0. The sum component is filtered by the first low-pass filter. The translation or offset frequency f_1 is applied to the phase comparator, along with the $f_R - f_0$ component of the mixer output. When the system is in lock, the two inputs of the phase comparator are at identical frequency; that is,

$$f_0 - f_R = f_1 \qquad \text{or} \qquad f_0 = f_R + f_1$$

6. *AM detection.* The PLL can be used as a coherent detector for demodulating AM signals. In this mode of

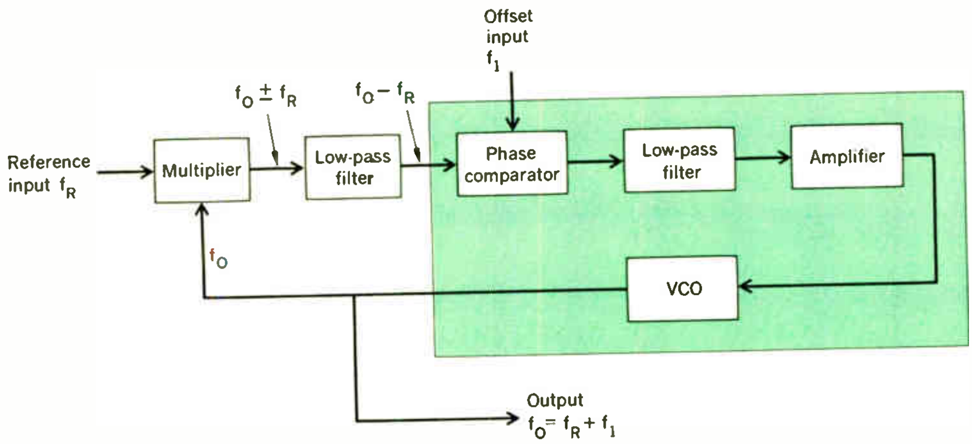


FIGURE 8. Frequency translation or "offset" loop.

FIGURE 9. Coherent amplitude-modulation detection using a phase-locked loop.

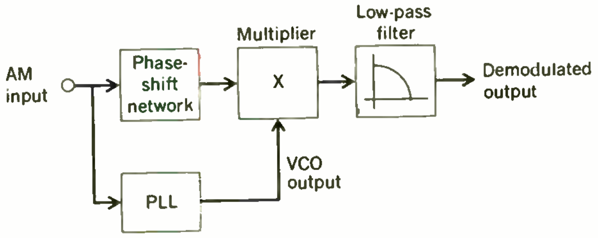
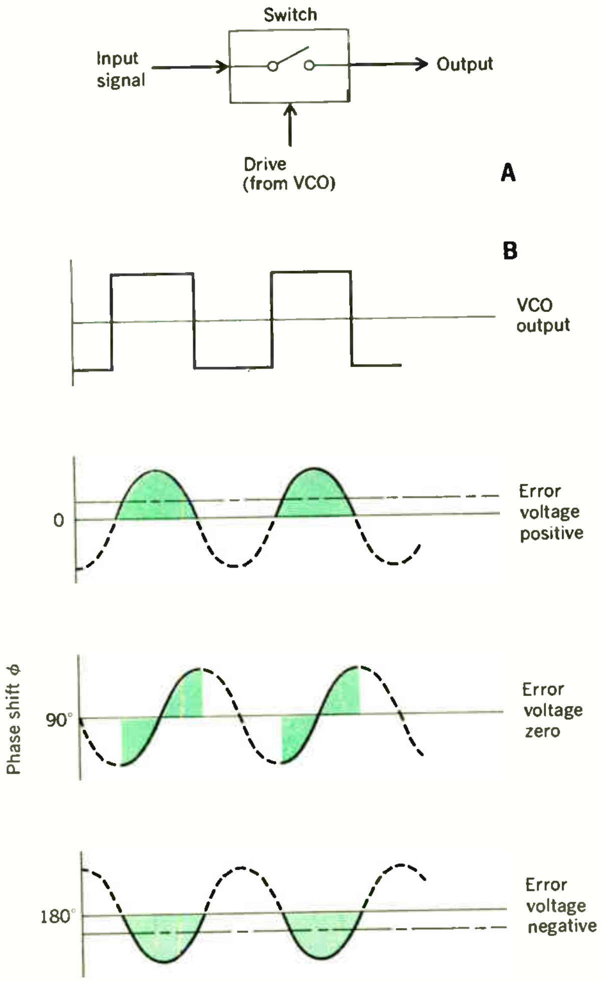


FIGURE 10. Operation of a switch-type phase detector. A—Basic block diagram. B—Output waveforms.



operation, the PLL locks on the carrier of the AM signal and produces a reference signal at the output of the VCO that has the same frequency as the AM carrier, but no amplitude modulation. Then, by multiplying this coherent reference signal with the modulated input signal and low-pass filtering the output of the multiplier, one can obtain the demodulated information. A block diagram of such a system is shown in Fig. 9. Since the PLL only responds to carrier frequencies very close to VCO frequency f_0, the phase-locked AM detector system also exhibits a high degree of selectivity, centered about f_0.[15] The phase-shift network of Fig. 9 is a noncritical *RC* network that is used to offset the 90-degree phase shift introduced by the PLL. The reason for this phase shift will be described in the following section.

The phase-locked AM detection method of Fig. 9 is a coherent detection technique; therefore, it offers a higher degree of noise immunity than conventional peak-detector-type AM demodulators.[16]

Building blocks for a monolithic PLL

With present-day IC technology, the PLL system can be easily and economically fabricated in a monolithic form. In this section, some basic circuit configurations will be described for the VCO and the phase-comparator sections of a monolithic PLL system. In the design of a monolithic circuit, a designer is normally faced with the following constraints: limited types of active devices, restricted ranges of component values, poor absolute-value tolerances, and lack of inductors. On the other hand, integrated circuit fabrication methods offer a number of unique advantages to the circuit designer. The most significant are the availability of large numbers of active devices, and close matching and thermal coupling between components due to their proximity in the silicon chip.[17] The particular circuit topologies that will be described in this section are chosen because their design and proper operation rely mainly on the matching and thermal tracking of monolithic components rather than on tight control of absolute-value tolerances.

The phase comparator. The simplest phase-comparator circuit suitable for monolithic integration is the switch-type phase detector that appears schematically in Fig. 10(A). This type of detector operates as a synchronous switch that is opened and closed by the reference input, and effectively "chops" the signal input at the same repetition rate as the reference drive. Normally, the reference drive is supplied by the VCO output. Figure 10(B) displays the typical output waveforms of the switch-type phase comparator for a sinusoidal input and a square-wave drive signal. The filtered error voltage V_d corresponds to the average value of the output waveform, shown as the tinted areas in the waveforms. The error voltage is zero when the net phase shift ϕ between the two inputs is 90 degrees. This 90-degree phase shift is a common property of all switch-type phase-detector circuits, and results in a detector gain expression of the form

$$K_d = K_A \cos \phi$$

where K_A is a constant of proportionality at a given input signal level. In a PLL system using this type of phase detector, when the PLL is in perfect lock condition [i.e., $V_d(t) = 0$], the VCO output is at quadrature phase with the input.

Figure 11 describes a more refined version of a switch-type phase-comparator circuit that is suitable for integration.[18] In this circuit, the signal input is applied to the bases of transistors T_1 and T_2 and controls the partitioning of the bias current I_T between these two devices. The cross-coupled transistor pairs (T_3,T_4) and (T_5,T_6) function as two sets of single-pole double-throw switches, actuated by the VCO waveform. The low-frequency output voltage V_e is related to the phase difference ϕ between the signals V_s and V_0 and can be expressed as

$$V_e = \frac{g_m R_L E_s}{\pi} \cos \phi$$

where g_m is the transconductance of T_1 or T_2, and E_s is the amplitude of the input signal V_s. The differential configuration of the phase-comparator circuit of Fig. 11 makes it particularly well suited to integration since the proper operation of the circuit depends mainly on the matching of the circuit components.

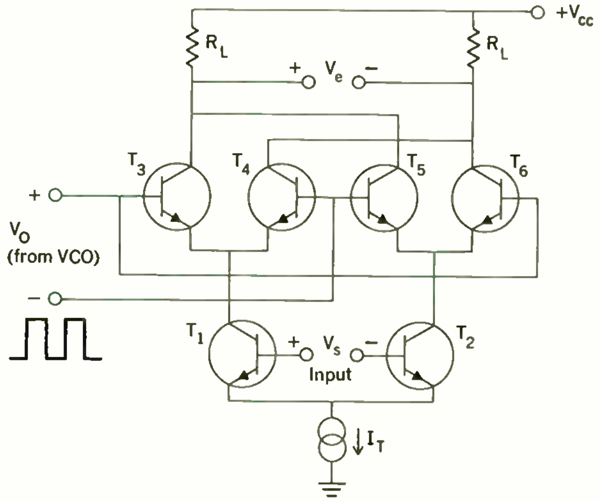


FIGURE 11. A phase-comparator circuit suitable for monolithic integration.

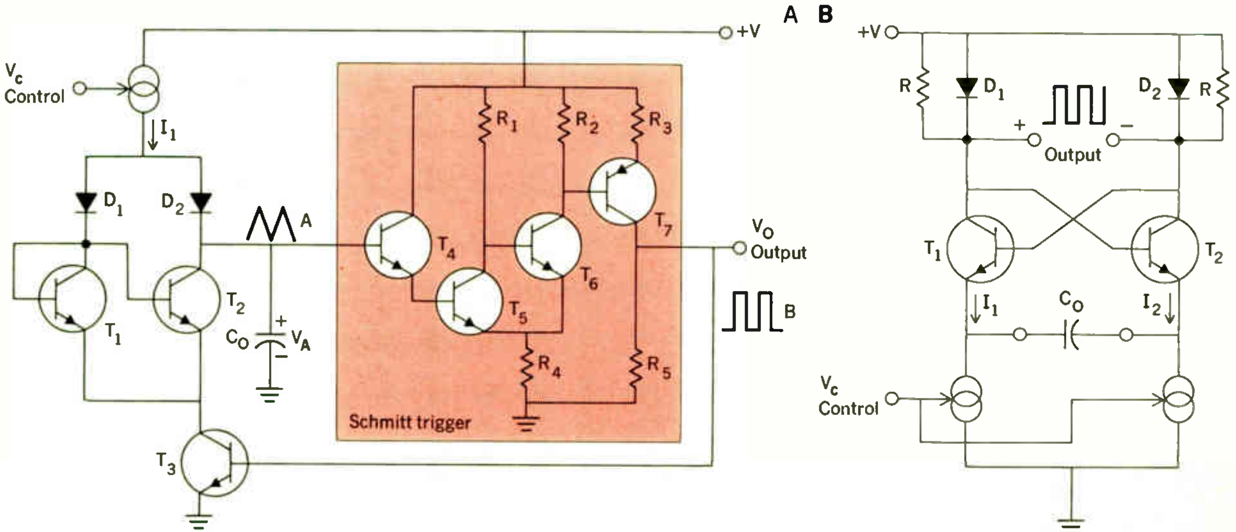
The VCO. In the design of a PLL, the voltage-controlled oscillator is usually the most critical block, since the frequency stability and the FM demodulation characteristics of the system are normally determined by the VCO performance. For maximum versatility, the VCO is required to have the the following desirable properties:

1. Linear voltage-to-frequency conversion.
2. Good frequency stability (low thermal and long-term drift).
3. High-frequency capability.
4. High conversion gain.
5. Wide tracking range.
6. Ease of tuning (frequency of oscillation determined by a minimum number of circuit components).

In addition to these requirements, to be suitable for monolithic integration, the VCO circuit should contain no inductors.

Figure 12 offers two possible oscillator circuits that fulfill most of the requirements just listed. The circuit of Fig. 12(A) is an integrator–Schmitt-trigger combination, where the timing capacitor C_0 is alternately charged and discharged by a voltage-controlled current source I_1.[19] The Schmitt trigger senses the voltage level V_A across C_0 and turns the switch transistor T_3 off or on to initiate the charge and discharge cycles respectively. The frequency of oscillation f_0 can be expressed as

FIGURE 12. Some basic VCO configurations for monolithic PLL design. A—Integrator–Schmitt-trigger circuit. B—Emitter-coupled multivibrator.



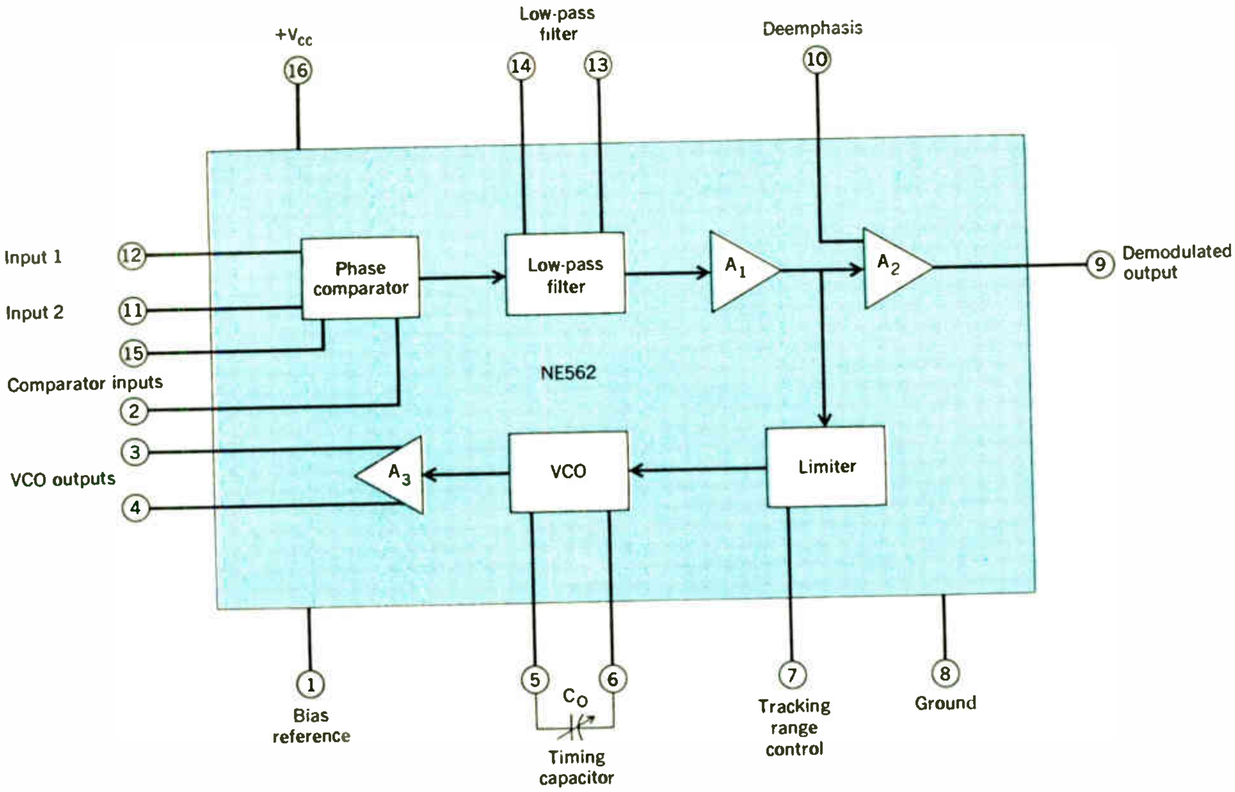


FIGURE 13. Functional block diagram of a monolithic PLL system (Signetics NE562).

$$f_0 = \frac{V_c g_m}{2C_0(V_2 - V_1)}$$

where g_m is the transconductance of the voltage-controlled current source, and V_2 and V_1 are the upper and lower trip levels for the Schmitt trigger. This type of oscillator can provide either a triangular-wave (at node A) or a square-wave (at node B) output.

The circuit of Fig. 12(B) is an emitter-coupled multivibrator where the cross-coupled transistors T_1 and T_2 form the positive-feedback gain stage.[13,15] At any one time, either T_1 or T_2 is on, and the timing capacitor C_0 is alternately charged and discharged by the voltage-controlled current sources I_1 and I_2. Normally, I_1 and I_2 are chosen to be equal. The frequency of oscillation can be expressed as

$$f_0 = \frac{V_c g_m}{4C_0 V_{BE}}$$

where V_{BE} is the transistor base–emitter drop and g_m is again the transconductance of the voltage-controlled current sources. The circuit provides a balanced square-wave output across the diodes D_1 and D_2. Since the VCO of Fig. 12(B) is a nonsaturating switching circuit, it offers a higher-frequency capability than that of Fig. 12(A). As will be described in the following section, a monolithic PLL system using this type of VCO has been made that operates up to 30 MHz. Note that both of the circuits in Fig. 12 offer linear control characteristics (i.e., f_0 proportional to V_c) and can be readily tuned to a desired frequency through the use of a single timing capacitor C_0.

A monolithic PLL system

The basic building blocks described in the previous section can be readily fabricated and interconnected in monolithic form to build a self-contained phase-locked-loop system. Figure 13 shows the functional block diagram of such an integrated PLL (Signetics NE562) in terms of its monolithic system package. The numbers given on the external system terminals correspond to the actual pin numbers in the case of a 16-pin monolithic IC package. This particular monolithic PLL is designed as a general-purpose building block, suitable for any one or a combination of the circuit functions listed in Table I. For maximum versatility, the phase-locked feedback loop is not internally connected. Instead, the loop is broken between the VCO output and the phase-comparator input so that either a divide-by-N circuit or a mixer circuit can be inserted into the loop for frequency-multiplication or frequency-translation applications (see Figs. 7 and 8). In addition to the basic blocks necessary to form the PLL (see Fig. 1), the monolithic system also contains two additional amplifiers, A_2 and A_3. A_2 is used as a buffer amplifier to increase the demodulated output level for FM or FSK detection; A_3 amplifies the VCO output signal and provides digital interface capability with conventional logic circuits. A limiter block is incorporated into the loop to control the maximum amplitude of the loop error voltage applied to the VCO. The threshold of this limiter block can be adjusted by means of an external bias, thus allowing one to control the total tracking range of the VCO. In the present design, the tracking range of the VCO can be externally adjusted from ±1 percent to ±25 percent of the VCO center frequency.

Circuit design. Figure 14 gives the complete circuit diagram for the monolithic PLL system of Fig. 13. For simplification, the basic blocks within the circuit are identified by the tinted areas. Note that the externally accessible terminals of the circuit are designated with the numbers corresponding to the package diagram of Fig. 13. In this particular design, the phase-comparator section is formed by a balanced-modulator circuit, similar to that shown in Fig. 11. The VCO is designed as an emitter-coupled multivibrator, similar to that described in Fig. 12(B). However, in this case, a differential control-voltage input is used to minimize the dc offset or drift problems. The frequency of the VCO is determined by means of an external capacitor C_0 connected across

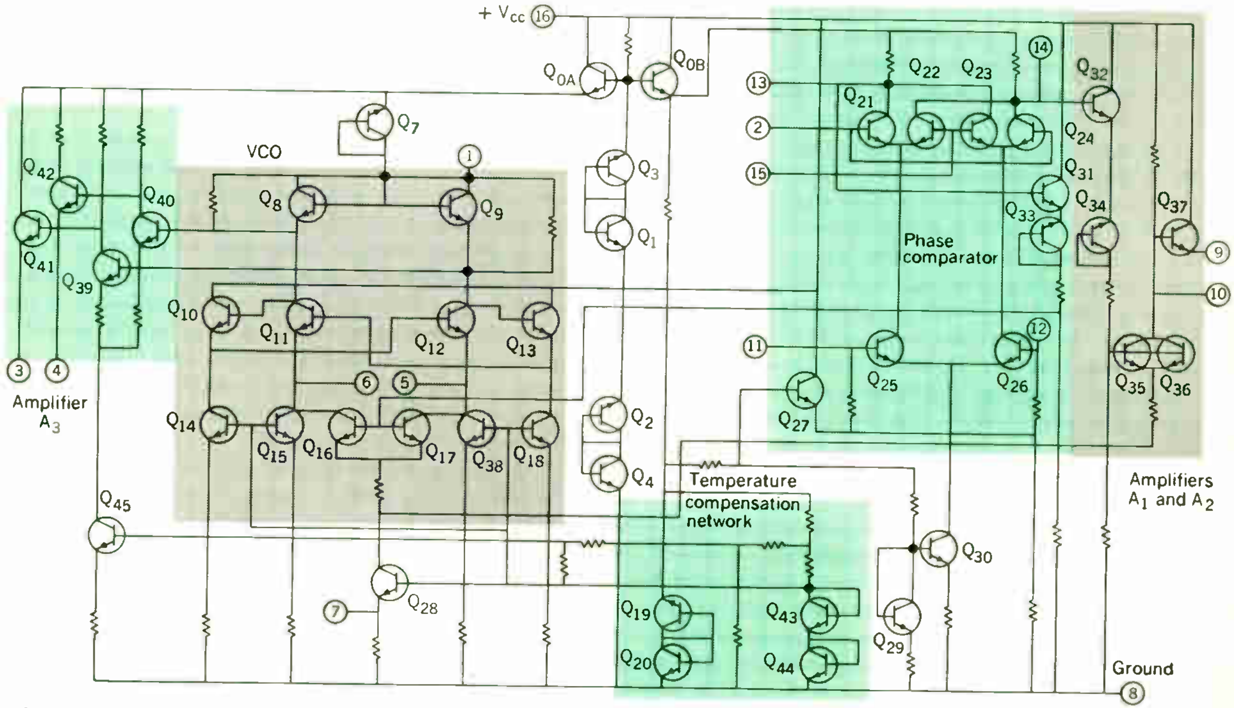


FIGURE 14. Circuit schematic of the integrated PLL system shown in Fig. 13.

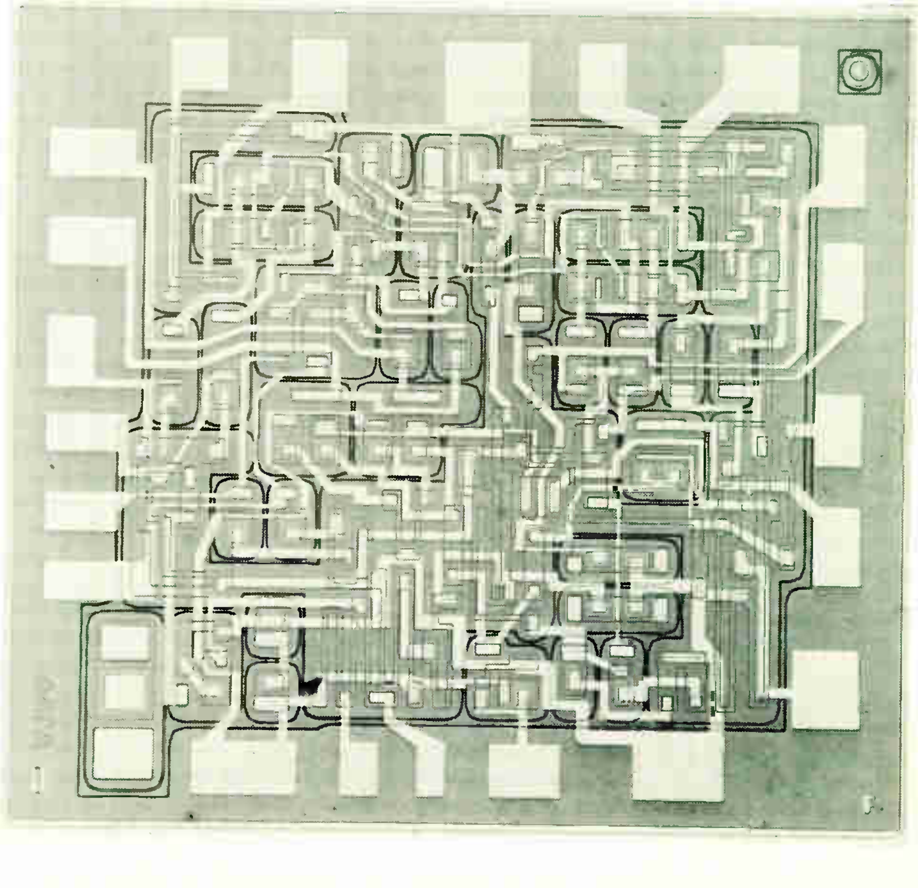


FIGURE 15. Photomicrograph of the monolithic circuit. Chip size is 1.7 mm by 1.83 mm.

terminals 5 and 6. In this particular design, the free-running frequency can be expressed as

$$f_0 = \frac{(3)(10^8)}{C_0} \text{ Hz}$$

where C_0 is expressed in picofarads. Thus, the VCO frequency can be varied from a fraction of a cycle to in excess of 30 MHz by a proper choice of C_0.

The limiter block is incorporated into the VCO by means of a current-sharing scheme between the common-collector transistor pairs (Q_{15},Q_{16}) and (Q_{17},Q_{38}). A temperature-compensating bias network is also included on the chip to minimize the temperature drift of f_0. This compensation network varies the current levels in the circuit to compensate for the transistor V_{BE} changes with temperature, and thus keeps the temperature drift of f_0 to less than 0.06 percent/°C.

The low-pass filter can be formed by connecting a

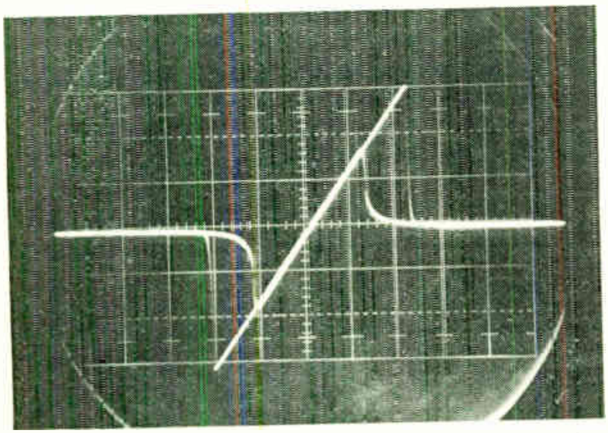


FIGURE 16. Integrated PLL frequency-to-voltage transfer characteristics at f_0 = 10 MHz. Scale—vertical: 0.5 V/div; horizontal: 500 kHz/div.

FIGURE 17. Interference rejection characteristics of the integrated PLL for FM demodulation.

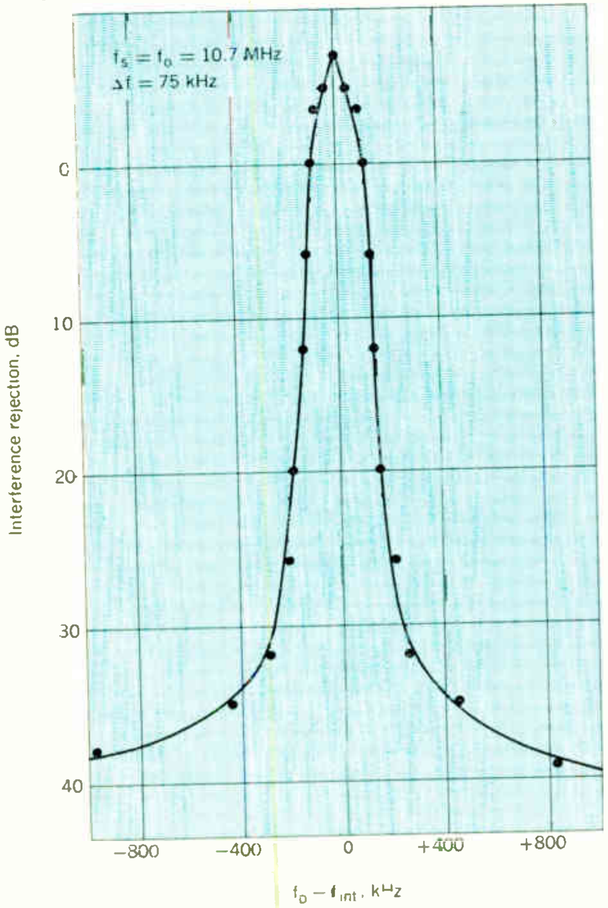
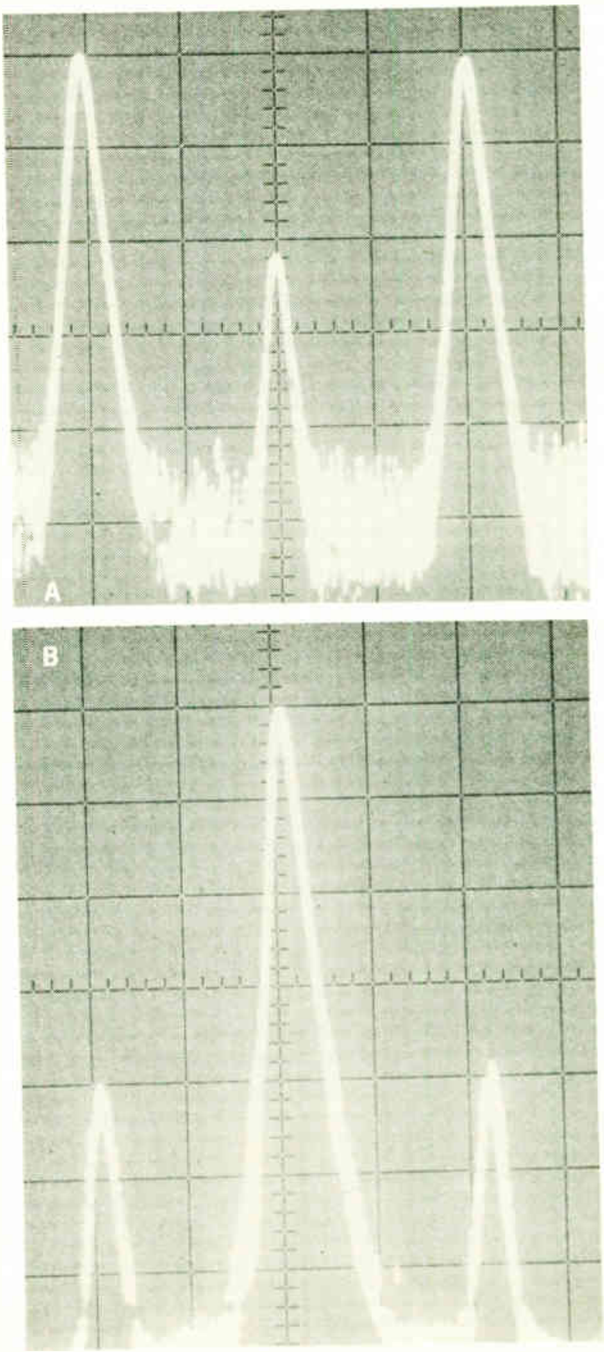


FIGURE 18. Performance of the integrated PLL as a signal conditioner. A—Input spectrum with a desired signal (center) at 1.0 MHz, and two undesired signals at 0.96 MHz and 1.04 MHz. B—VCO output spectrum. Scale—center frequency = 1.0 MHz; horizontal: 20 kHz/div; vertical: 10 dB/div.



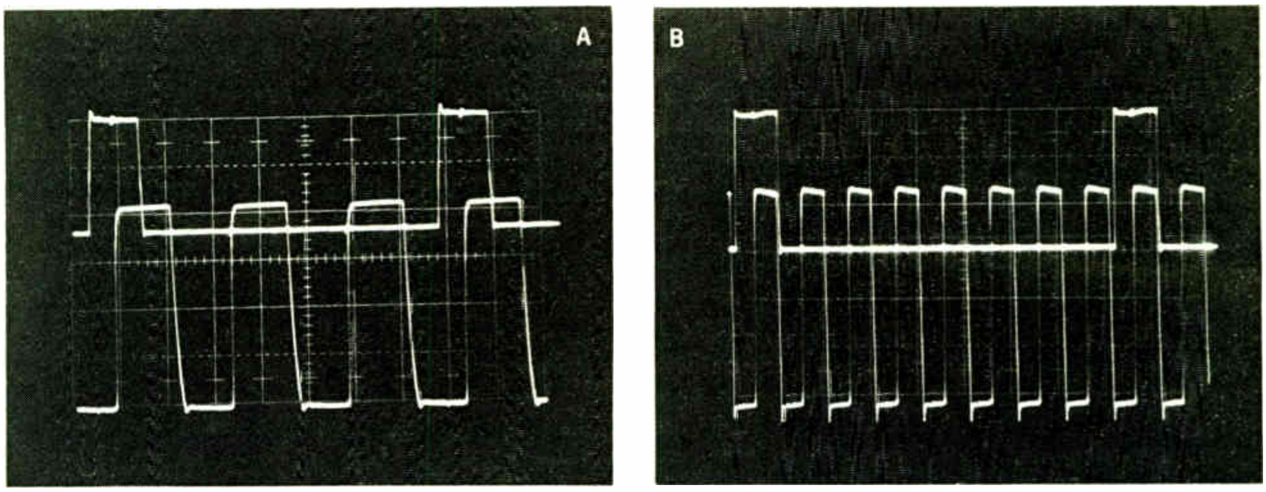


FIGURE 19. PLL as a frequency multiplier. Upper trace represents signal input and lower trace represents VCO output waveforms. Vertical scale: 1 V/div. A—VCO locked on 3rd harmonic of the input. Input: 333 kHz; VCO output: 1 MHz. B—Frequency divider in feedback (N = 8). Input: 125 kHz; VCO output: 1 MHz.

series resistor and capacitor combination, R_1 and C_1, across the output terminals 13 and 14 of the phase comparator. This results in a lag-lead type of filter with the transfer function

$$F(s) = \frac{1 + (R_1C_1)s}{1 + (R_1 + R)Cs}$$

where $R = 12$ kΩ is the internal impedance between nodes 13 and 14 in the illustration. If R_1 is set equal to zero, this low-pass filter reduces to a simple lag filter.

The monolithic PLL circuit was fabricated on a 1.7- by 1.83-mm chip. The photomicrograph of the finished circuit chip is reproduced in Fig. 15. The circuit is designed to operate with a single power supply, in the 15–24-volt range, and has a nominal power dissipation of 160 mW at $V_{cc} = 18$ volts. The system can operate over the full military temperature range (-55 to $+125$°C) with a typical VCO frequency drift of approximately 600 ppm/°C, and can handle input signals over a 80-dB dynamic range (from 200 μV to 2 volts).

Performance characteristics. Figure 16 shows the typical frequency-to-voltage conversion characteristics of the monolithic PLL circuit (NE562) of Figs. 13 and 14 at 10-MHz center frequency. The inner and the outer traces correspond to the capture and the lock range of system (i.e., a superposition of the transfer characteristics sketched in Fig. 6). The typical interference rejection characteristics of the PLL as an FM demodulator are given in Fig. 17 for a 15-kHz low-pass loop filter at 10.7-MHz center frequency. The skirt selectivity of the interference-rejection characteristics exhibits a form factor* of 2.2 for 3–30-dB roll-off, which compares favorably with the skirt selectivity of a three-stage conventionally tuned FM IF strip.

Figure 18 displays the input and output spectrum for the integrated PLL in signal-conditioning applications. The input signal system shown in Fig. 18(A) is made up of a desired signal at 1 MHz with two undesired sidebands at ±40 kHz away from it. In this case, the undesired signal levels are 20 dB higher than the desired signal. Figure 18(B) shows the VCO output spectrum with the PLL locked on the desired signal. Note that, at the VCO output, the undesired sidebands are 40 dB below the desired signal level.

Figure 19 describes the input and output waveforms for the PLL when used as a frequency multiplier. In Fig. 19(A), the system is operated in its harmonic-lock mode, with the VCO locked on the third harmonic of the input ($n = 3$). In the case of Fig. 19(B), the monolithic PLL is operated with a divide-by-N circuit in the feedback path (see Fig. 6) with $N = 8$.

PLL versus active RC

At this point, having examined the basic properties and the performance characteristics of the monolithic PLL, it is worthwhile to pause briefly and compare the pros and cons of the PLL approach with conventional active *RC* filters. From the preceding discussion, the following advantages of the monolithic PLL have become apparent:

1. *High-frequency capability.* The monolithic PLL can operate at frequencies in excess of 30 MHz; most active-*RC* techniques using integrated components are limited to a frequency range below 100 kHz.
2. *Independent control of selectivity and center frequency.* The center frequency is set by the VCO free-running frequency; and the selectivity is determined by the low-pass loop filter. This eliminates the "alignment" problem associated with cascading conventional filter stages.
3. *Fewer external components.* Compared with monolithic active filters, the integrated PLL generally requires fewer external energy-storage or precision components. For example, to obtain a skirt selectivity comparable to that given in Fig. 16 using active *RC* filters would require at least eight capacitors and as many precision resistors whose absolute values must be controlled to better than 0.5 percent.
4. *Ease of tuning.* The PLL can be tuned to any desired

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* Form factor is the ratio of the bandwidth of a filter measured at a high attenuation to the bandwidth at a low attenuation.

frequency by proper choice of the VCO free-running frequency. This frequency is normally set by a single external component, and can be continuously adjusted from a fraction of a cycle to a value in excess of 30 MHz.

There is, however, one significant point that must be considered. Active *RC* filters, at least in theory, form a direct replacement for conventional *LC*-tuned filters. The PLL, on the other hand, is a somewhat specialized filter; therefore, it is not a direct replacement for conventional filters in all applications. Compared with *LC* or active *RC* filters, the PLL has the following drawbacks:

1. *Lack of amplitude information.* The PLL responds only to the frequency of the input signal and not the amplitude, as long as the amplitude of the input is high enough to maintain lock. Thus, it filters out only the frequency information and not the amplitude.

2. *Response to harmonics.* Most PLL systems suited to integration also tend to respond to the harmonics and the subharmonics of the input through the "harmonic-lock" effect described earlier. This is useful for frequency multiplication and division, but degrades the interference rejection of the system for parasitic signals that are harmonically related to the desired input.

3. *Difficult to provide AGC.* Since the PLL responds to frequency and not amplitude of a given input, it is difficult to derive a control signal from the PLL for automatic-gain-control (AGC) applications. The AM detection capability of the PLL (see Fig. 8) can be used for "squelch"- or "muting"-type AGC, but is not suitable for conventional gain-control applications.

4. *Synthesis techniques not yet developed.* Because of nonlinear capture characteristics, no detailed synthesis procedures exist for a PLL filter design with a predetermined interference-rejection characteristic.

Conclusion

The design of frequency-selective monolithic circuits has long been hampered by lack of integrated inductors. Active *RC* filters have provided only a partial solution to this problem, because of their frequency limitations and component tolerance requirements. Phase-lock techniques provide an alternate and versatile design approach for frequency-selective ICs that overcome most of the drawbacks of other inductorless filter techniques.

The phase-locked loop is not a direct replacement for active *RC* or *LC*-tuned filters in all applications. However, in a wide variety of applications, particularly in the communications field, it offers significant performance advantages over conventional filters.

The phase-locked loop is not a new technique. It has been known and used for over 30 years. Until recently, because of its cost and complexity in discrete (nonintegrated) form, the PLL had remained as a special design technique, mostly limited to instrumentation applications. The availability of a monolithic PLL as a self-contained system package has rapidly converted it from a highly specialized design technique to a low-cost, general-purpose building block. The economic advantages of monolithic integration has opened up a wide range of new applications for the phase-locked loop where it may not have been suitable before because of cost and complexity considerations. In many ways, this is similar to the case of the operational amplifier, which until less than a decade ago was a fairly expensive building block used mostly in analog computers. Yet today, with the advent of monolithic IC technology, it has become a basic building block in many system designs. The monolithic PLL offers a similar potential, particularly in the communications field.

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Noise in laser recording

A new, and very promising, form of recording has evolved through utilization of a modulated laser beam, but the fundamental parameter that will determine its worth and utility is signal-to-noise ratio

R. F. Kenville RCA Defense Electronic Products

Noise in laser recording is treated from the viewpoint of measurement techniques, equipment design, film choice, and laser power. Signal-to-noise criterions that have been found useful in specifying wide-band magnetic recorders are presented to aid in establishing noise performance parameters for laser recorders. It is shown that such equipment can be built so that it will have an insignificant effect on total noise performance and that, although film granularity is the ultimate limit on noise, improvements in SNR are a function of increased laser power. The design of a 100-MHz analog laser signal recorder is used as the basis for discussion; however, the theory and techniques apply to all forms of laser recording.

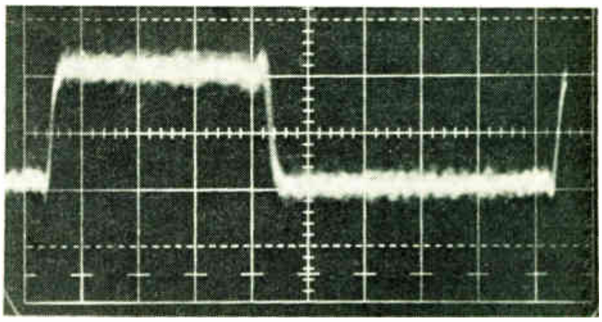
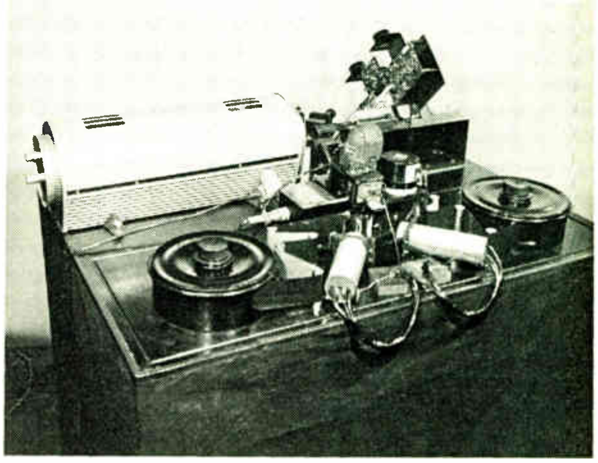

1 μs/div

FIGURE 1. Oscilloscope trace of output pulse from a laser recorder/reproducer.

The use of a modulated laser beam to record wide-band signals on photographic film offers an opportunity to extend bandwidth and packing density over conventional recording techniques. Continuous signals of bandwidths exceeding 100 MHz can be stored through the combination of electronic, mechanical, optical, photographic, and laser technologies called laser recording. The fundamental parameter that will determine the worth and utility of this new form of recording is signal-to-noise ratio.

A laser recording system capable of recording 100-MHz video signals serves as an example for SNR considerations. Since noise can be cataloged and specified in a number of ways, definitions and measurements of SNR are presented in a form applicable to most wide-band recording requirements. Figure 1 shows the oscilloscope trace of a square-wave signal that was recorded and reproduced on the laser signal recorder shown in Fig. 2. The random waveshape at the top and bottom of the signal is the noise to be discussed here.

FIGURE 2. Laser signal recorder.



Laser recorder description

A system* for recording and reproducing signals of 100-MHz bandwidth using laser scanning techniques is shown conceptually in Fig. 3. After the light beam from an argon laser is intensity-modulated by the record signal, the beam is expanded to fill the imaging lens and then focused into a diffraction-limited spot. A polygonal rotating mirror located within the focus of the imaging lens deflects the modulated spot in an arc at a rate of 127 000 cm/s (50 000 in/s). Silver-halide film 70 mm wide is transported past the scanning laser beam and is exposed by it along transverse tracks similar to the format used in television tape recording. In addition, a control track is recorded on the edge of the film to serve later as a form of electronic sprocketing during playback. After the film is developed, it is transported past the scanning beam, which once again becomes intensity-modulated, this time by the transmittance variations in the film. The light is collected by optics and is photodetected, yielding a reproduction of the original signal. Using the control-track information, a capstan servo incrementally aligns the tracks of the film with the

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* Since October 1967, RCA has been under contract with Rome Air Development Center and the Advanced Research Projects Agency to develop a laser recorder.

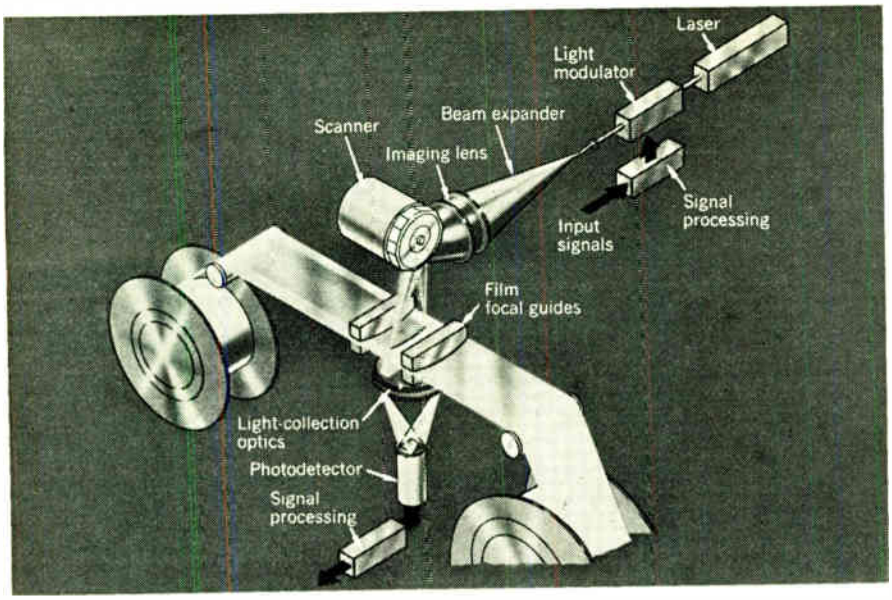


FIGURE 3. Concept of laser signal recorder. The laser beam, which has been intensity-modulated by the input signal and focused to a diffraction-limited spot, is scanned at a high rate across slowly moving silver-halide film to make a recording. Playback is accomplished by transporting the developed transparency past the scanning laser beam detecting the light modulated by the film.

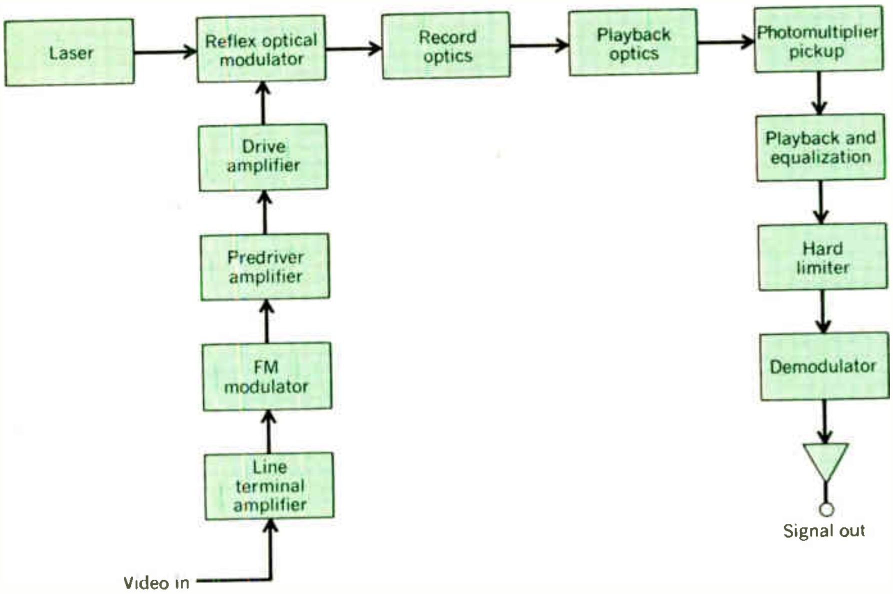


FIGURE 4. Block diagram of FM recorder.

scanning beam to provide continuous playback.

To overcome problems of film nonlinearity, amplitude variations, mechanical tolerances, and multioctave response, the incoming video signal is processed with a frequency modulation system. Signals are stored by the relative position of transmittance variations of the film rather than by the relative level of transmittance. A block diagram of the signal-processing system is shown in Fig. 4. Upon playback, the recovered film signal is demodulated to reconstruct the original wide-band analog information. The waveform in Fig. 1 shows the demodulated playback output for an input square wave.

Figure 5 shows magnified scan lines as recorded on film. The photograph at the left shows several tracks and illustrates the FM record pattern of a square-wave signal. The photograph at the right, a close-up of the same recording, shows the grain of the record film in relation to the size of the recorded track. The photographs also show the format of the tracks, where the track spacing is 36 μm and the track width is 12 μm. (The record spot is 5 by 12 μm.) The discussions that follow consider laser recording noise on a direct basis. Conversion to FM-processed SNR can be accomplished by using standard FM theory.[1]

Signal-to-noise definition and measurement

Noise is a random phenomenon with a Gaussian-like distribution, which, when viewed on a peak basis, statistically has levels greater than those of the signal; hence it is measured on an rms basis. The signal frequently is referenced on a peak-to-peak basis. This section defines the subtleties of the SNR properties normally found in a recording system.[2]

Narrow-band noise. Narrow-band noise (noise spectral density), a useful criterion in evaluating the performance of a wide-band analog recording system, is measured by scanning the passband of the reproduced output with a selective voltmeter, which is a receiver whose detected

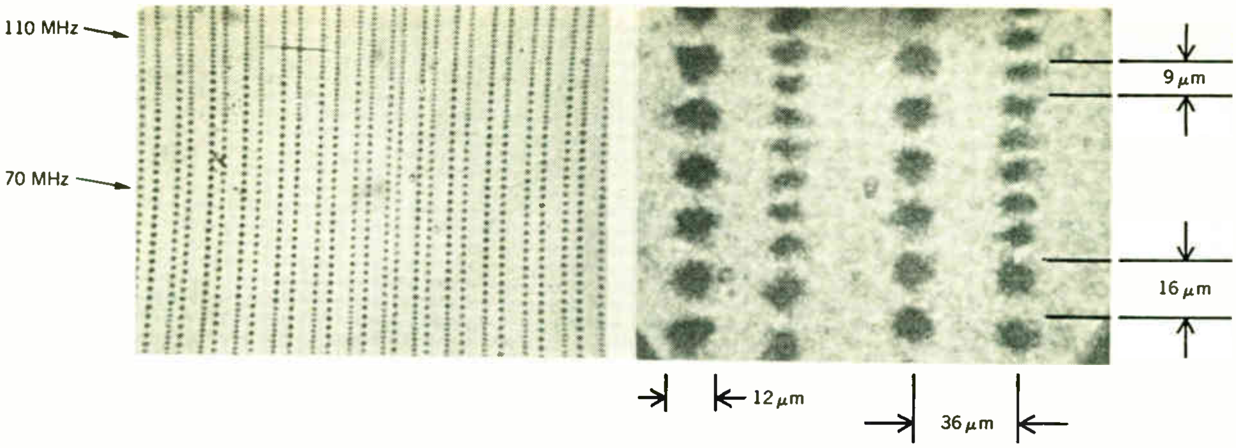


FIGURE 5. **Laser recorder signal tracks.**

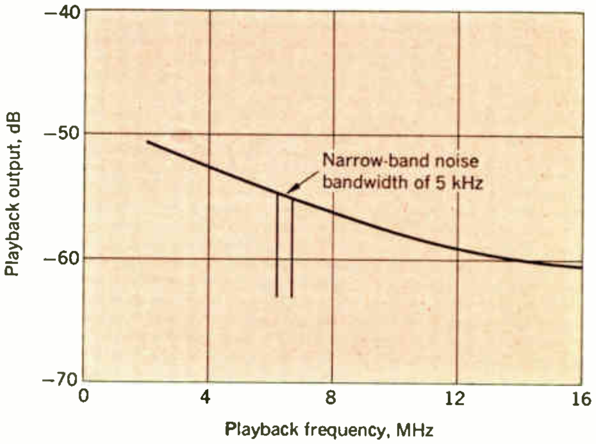
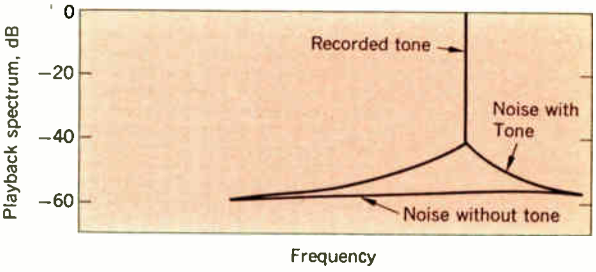


FIGURE 6. **Plot of narrow-band noise vs. frequency for a ten-to-one time-expanded playback of laser recorder.**

FIGURE 7. **Modulation noise with signal as compared with noise without signal.**



output is calibrated. The plot of the narrow-band noise for the direct-record readout at ten-to-one time expansion of the laser record system is shown in Fig. 6. The slot bandwidth is 5 kHz, but any convenient bandwidth may be used to take the data. However, data taken at different bandwidths should be compared on the basis of root-mean-square addition. Most analog recording applications require the condition of uniform narrow-band noise commonly referred to as white noise.

Wide-band noise. Wide-band noise is the playback output noise as measured with a wide-band rms voltmeter. Wide-band noise is, of course, composed of the rms addition of the narrow-band noise. For instance, a 10-MHz bandwidth having uniform narrow-band noise of 60-dB rms signal to rms noise in a 5-kHz slot has wide-band noise 36 dB below the peak-to-peak signal. The pulse in Fig. 1 exhibits peak-to-peak signal-to-rms noise of about 24 dB. When viewed on an oscilloscope, wide-band noise appears on a peak-to-peak basis to be about 14 dB higher than its measured rms level.

Modulation noise. Noise is usually measured in the absence of signal. In many recording systems the noise level depends on the instantaneous level of signal. The term "modulation noise" describes this phenomenon. Modulation noise is common to both photographic film and magnetic tape, since each of these mediums has a grain structure, and variations from grain to grain are detectable only when film is exposed or when tape is magnetized.

Modulation noise can be evaluated by measuring narrow-band noise in the presence of a recorded signal. The plots in Fig. 7 illustrate the phenomenon of modulation noise. The lower curve represents the narrow-band noise without a signal; the upper curve shows the noise spectrum with a recorded tone.

Modulation noise does not necessarily limit dynamic range if the recording system is designed so that the noise decreases as the signal decreases. For instance, in pulse-recording applications it is best to place the baseline of the pulses in the low-noise region of the medium and to allow the top of the pulses to extend into regions of higher noise.

Distortion. Distortion, a measure of the deviation from constant slope of the record/reproduce transfer characteristic, limits the peak level of signals. Many systems specify distortion on a percentage basis. Since this practice generally has not proved to be very useful for wide-band analog recording systems, a specification based on distortion component level below reference level is now generally used in wide-band recording.

Distortion in an analog recording system can be measured readily by the use of a selective voltmeter, as in the measurement of narrow-band noise. Any number of test conditions can be investigated with the selective voltmeter, but the most common technique is the two-tone test. In this test two sinusoidal signals of equal amplitude are recorded at a level totaling the peak-to-peak reference level of the system. Figure 8 is a spectral plot of the test signal. Upon playback, the levels of the worst distortion components below reference level are measured. Figure 8 shows the spectral location of the second-order distortion products.

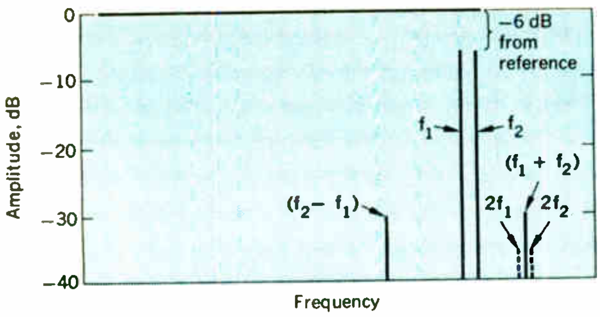


FIGURE 8. **Spectral plot of two-tone test for distortion, showing second-order distortion products.**

Noise-power-ratio test. As implied in the foregoing, distortion and linearity depend upon record level. For instance, the relative level of second-order distortion decreases in a linear fashion with the input level. Third-order distortion decreases in a square-law fashion with the input level. Figure 9 presents curves for distortion versus input level.

On the other hand, SNR increases as the signal increases. Figure 10 shows the relationship between record level and relative playback noise level. It also includes the curve for composite distortion. At the level where the two curves intersect, the effects of noise and distortion are equal. The noise-power-ratio (NPR) test is used to determine the optimum record level and to compare the overall performance capabilities of the record/reproduce system.

The test is conducted by recording white noise. During the recording, a bandstop filter is inserted that eliminates signals from part of the spectrum, as shown in Fig. 11. Upon playback the components introduced into the bandstop region through the distortion and noise of the record/reproduce cycle are measured.

The NPR test will even take into consideration spurious signals and the modulation-noise properties of the system. Modulation noise has the effect of bridging the trough between the curves in Fig. 10, as indicated by the dotted line. However, since the test is an abstract one, it does not directly indicate whether a given system will serve a given recording application. Even so, it does provide a means of comparing the total worth of different recording systems.

Sources of noise

The laser playback process, illustrated in Fig. 12, indicates the following potential causes of noise:

1. Laser—incidental modulation of the light intensity.
2. Photodetector—photon-to-electron conversion plus dark-current noise.
3. Preamplifier—thermal noise of source impedance and noise figure of preamplifier.
4. Film—statistical variations in emulsion properties.

The goal of the recorder designer is to reduce the level of noise generated by the hardware and ultimately to have the film limit SNR. In practice, this goal is achieved in laser-film recording; however, a brief discussion of the noise properties of the laser, photodetector, and preamplifier is given in the following to familiarize the reader with the technique capabilities in these areas.

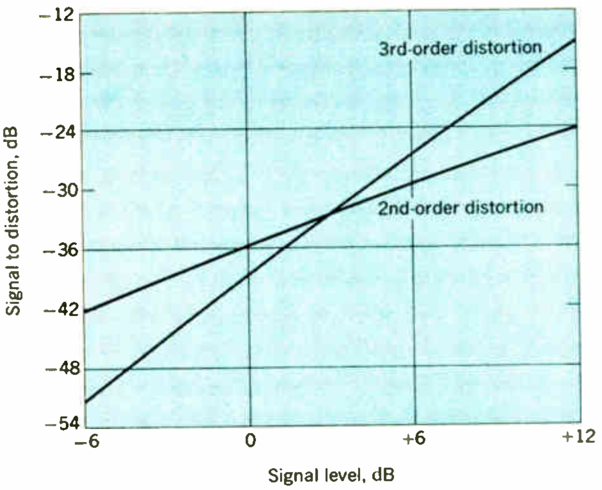


FIGURE 9. **Second- and third-order distortion as a function of record signal level.**

FIGURE 10. **Plot of noise-distortion optimization versus the record signal level.**

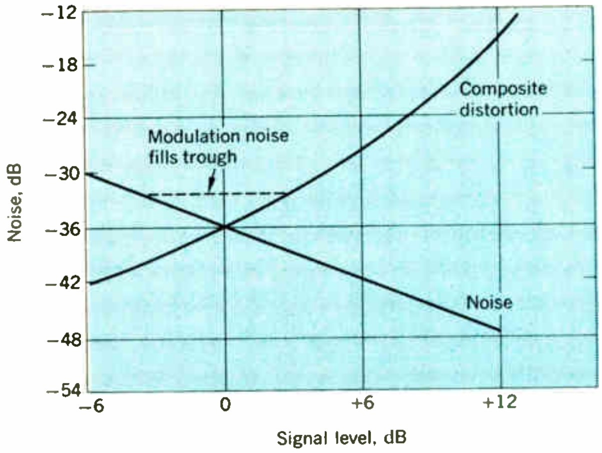


FIGURE 11. **Spectrum of noise-power-ratio test signal with bandstop filter.**

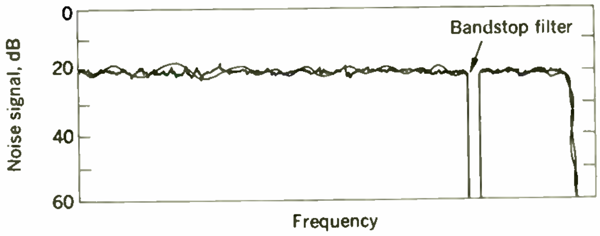
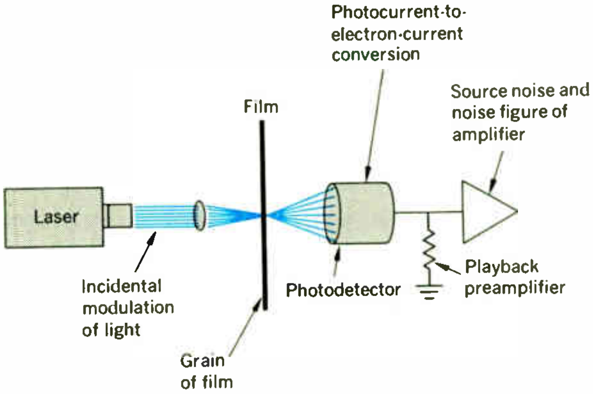


FIGURE 12. **Potential noise sources.**



Laser noise. The noise associated with a laser beam in general is made up of both random and nonrandom (hard) components. When the laser beam used in the recorder shown in Fig. 2 is detected by an ITT FW114-A ultrawide-band planar photodiode, the spectrum shown in Fig. 13 is observed. Random noise, similar to that found in vacuum tubes and semiconductors, is encountered at low levels. Usually, this noise is not excessive, except at low frequencies where low-frequency noise increases to a substantial level. For many applications, the hard components, which result from mode beats within the laser cavity, reach an objectionably high level if left untreated. Implementation of an etalon, used as a secondary optical cavity, eliminates mode-beat problems.[3]

The argon laser (in this case an RCA LD2100) used as the radiation source for the laser recorder in Fig. 2 has a resonant cavity between mirror faces of approximately 60 cm. A prism set in the resonant cavity allows the selection of any one of six wavelengths between 4579 Å and 5145 Å. The dominant line is at 4880 Å and contains more than 50 mW.

Ideally, the 4880-Å line of argon-ion gas should be extremely narrow. In an argon laser, the line width is broadened by the Doppler effect due to the rapid motion (high energy) of the light-emitting atoms in the hot plasma. The result is a Gaussian distribution of excitation energy over a band of wavelengths whose center is 4880 Å.

Enclosing the ionized gas between two mirrors creates an optically resonant cavity, thus allowing the stimulated emission (laser action) of discrete frequencies at cavity resonances with the Doppler line width. Separation of these emitted frequencies of light is determined by the cavity length L; thus the mode-beat frequencies are specified as integral multiples of $F = c/2L$.

In the laser under consideration here, several light frequencies of approximately 5 GHz, or 0.04 Å, separated by F = 234 MHz, are emitted within the Doppler line width, as shown in Fig. 14. The wavelength of each of these longitudinal modes is within this narrow band at 4880 Å. The exact wavelength and energy content vary with time. The result is a varying 10 to 90 percent intensity modulation of the laser beam at the 234-MHz beat frequency and decreasing modulation percentages at the higher-order harmonics of 234 MHz. When the laser beam is intensity-modulated by a 120-MHz signal, a "laser noise" beat frequency appears at 114 MHz at 15 dB below carrier. The energy content of this hard-noise component varies rapidly as thermal conditions change within the laser.

The laser mode beat and the spurious components associated with it can be suppressed by adding a second cavity to the laser to achieve mode selection. One method of mode selection in a laser is to replace the output mirror with an etalon, a glass cylinder whose ends are flat and parallel. The input face is a partially reflecting mirror (about 30 percent) and the output face is highly reflecting (above 95 percent). The laser with etalon acts as a two-stage filter for the 4880-Å line. The length of the etalon is such that the separation between light frequencies for which it is a resonant cavity is much greater than the Doppler spread of light-emission frequency. The etalon will have one sharp resonance in the same Doppler spread in which the laser has several resonances. Tuning the laser cavity mirrors causes the laser and the etalon resonant wavelengths to coincide. The result is a single very narrow line at 4880 Å.

In the laser record system, the etalon reduces the hard-noise components to an insignificant level, and the choice of FM signal processing avoids the low-frequency noise. The foregoing conditions result in a situation in which laser noise is not a limiting factor in laser recording.

Photodetector noise. To develop a high signal level relative to the noise generated by a 50-ohm terminating resistor, a multiplier phototube is used as the photodetector for the laser recorder.

The noise in signal current at the terminating load of a multiplier phototube is[4]

$$i_{NS} = \mu\left[2eI_K\Delta f\left(1+\frac{B}{(m-1)}\right)\right]^{1/2} \qquad (1)$$

where μ = tube current gain; I_K = average photocathode current; m = secondary emission ratio per dynode (typical value, 4); Δf = bandwidth; B = experi-

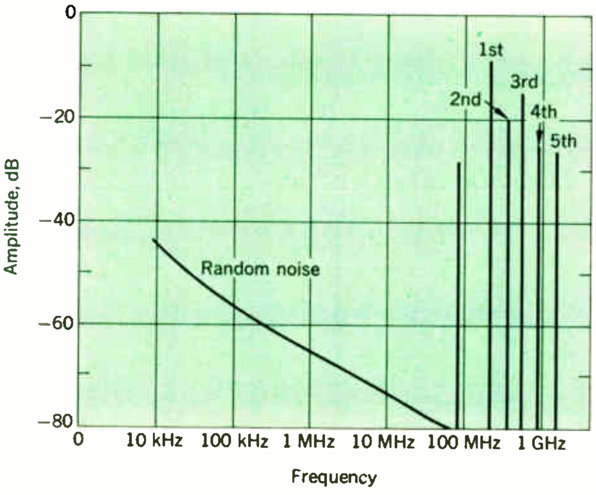


FIGURE 13. **Spectrum of laser-produced noise as observed at 5-kHz slot bandwidth.**

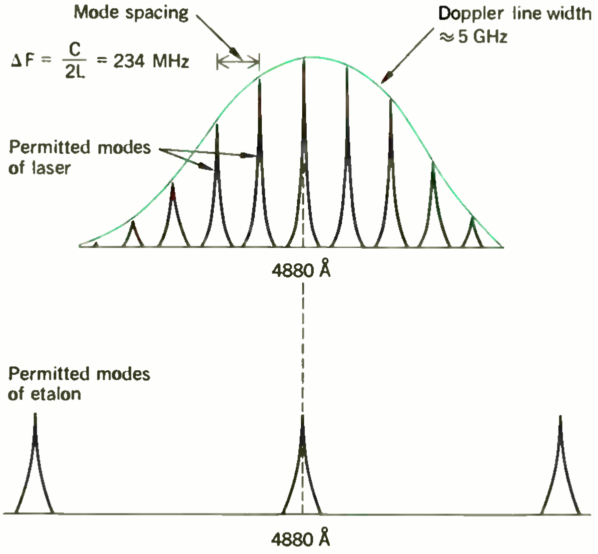


FIGURE 14. **Permitted frequencies in laser and etalon.**

mental factor related to multiplier phototube (typical value, 1.5); and e = charge of an electron.

The peak-to-peak multiplier phototube output current is

$$i_{\text{sig}} = \mu 2 M I_K \tag{2}$$

where M is the transmittance modulation of the signal on the film.

If Eqs. (1) and (2) are combined, the ratio of multiplier phototube peak-to-peak signal to the rms noise in the signal is

$$\text{SNR} = 2M\left[\frac{I_K}{3e\Delta f}\right]^{1/2} \tag{3}$$

where the factor $1 + [B/(m - 1)]$ has been replaced by its approximate value of 1.5. It is evident from Eq. (3) that the SNR will be improved for a given system bandwidth if film modulation M or photocathode current I_K is increased (I_K is proportional to incident light power). Design tradeoff studies and experimental data taken on the laser recorder/reproducer indicate that 0.33 is the value of M that can be achieved in practice.

The multiplier phototube (modified ITT type F4008) in the system has the following salient features: (1) large-area photocathode; (2) S-11 photocathode; (3) high average photocathode and anode current levels; (4) moderate gain (typical value, 100); (5) 150-MHz bandwidth; (6) 8 mV peak-to-peak across 50-ohm output; and (7) 43-dB SNR at 100-μW incident light power. The S-11 photocathode radiant sensitivity at the 4880-Å argon wavelength is typically 29 mA/W. An average photocathode current of 2.9 μA (0.29 μA/cm² for the 10-cm² photocathode) is generated if the average incident light power is uniformly distributed and is regulated to 100 μW. Tube current gain of 86 is required to achieve an average anode current of 0.25 mA for an average photocathode current of 2.9 μA.

Using Eq. (3), the multiplier phototube output SNR (peak to peak rms) is approximately 133 (43.5 dB) for a 150-MHz bandwidth. The equivalent SNR for a 5-kHz slot is 87 dB. Further, the signal level at the multiplier phototube is sufficiently high (165 mA peak to peak into 50 ohms) that the noise introduced by the playback electronics, primarily the preamplifier, is negligible.

Preamplifier noise. The preamplifier noise includes both the source impedance and noise figure of the preamplifier. Thermal noise current generated within the resistive load terminating the multiplier phototube is[5]

$$i_{NR} = 2\left[\frac{kT\Delta f}{R}\right]^{1/2} \tag{4}$$

where k = 1.38 × 10^{-23} J/°K (Boltzmann's constant); T = 300°K; Δf = 150 MHz; R = 50 ohms. Therefore,

$$i_{NR} = 2.22 \times 10^{-7}\ \text{A}$$

$$\text{SNR} = \frac{0.165 \times 10^{-3}}{2.22 \times 10^{-7}} = 750\ \text{(at 57.5 dB)}$$

A noise figure of 6 dB, where noise figure is defined as[6] SNR (input)/SNR (output), is realizable for the FM electronics that process the low-level multiplier phototube output signal. Essentially, the noise current of the source resistance will be doubled to $i_{NR}' = 4.44 \times 10^{-7}$ ampere. The effective SNR of the preamplifier with its source resistance is therefore 51.5 dB over a 150-MHz bandwidth. The equivalent SNR over a 5-kHz slot is 96 dB. This electronic noise component becomes masked by the noise generated within the input photodetector.

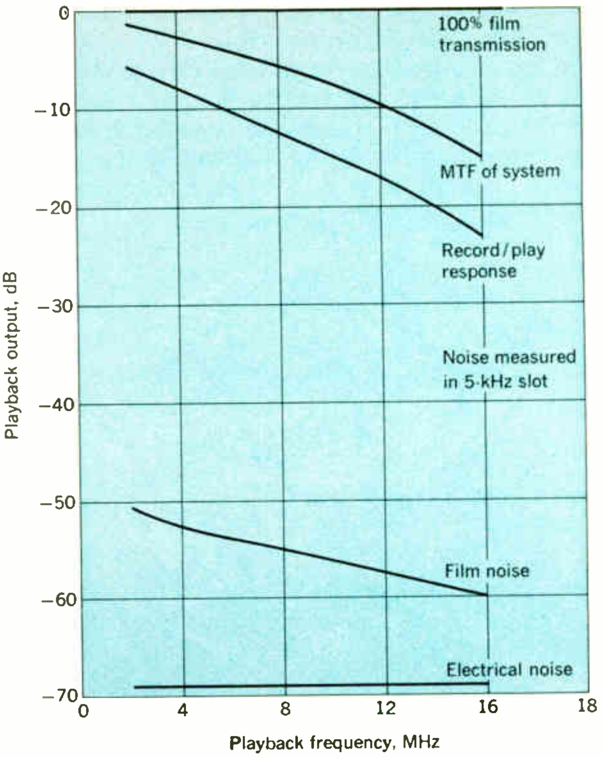


FIGURE 15. Signal measurements of laser recorder at ten-to-one time expansion.

Film noise. As implied earlier, film noise limits the dynamic range of a properly designed laser recorder; system testing of the recorder in Fig. 2 shows that this condition exists. The resultant curves (Fig. 15) combine the frequency response of the system, on a direct record basis, with the noise of the system as measured in 5-kHz increments across the spectrum of the signal. For these tests the recordings were made at the 100-MHz bandwidth rate and played back at a ten-to-one time-base expansion that corresponds to 10 MHz. The 0-dB level represents the full transmittance swing on the film corresponding to 100 percent modulation. The electronic noise (laser, multiplier phototube, and preamplifier) is −70 dB. Film granularity determines the noise limit in the system and is −58 dB at 12-MHz playback. At 12-MHz playback, the signal is −18 dB, giving a 40-dB narrowband SNR for the playback signal. When this SNR is converted to an equivalent 150-MHz record bandwidth, a wide-band SNR of 14 dB peak to peak/rms results. The film used for these tests was Kodak Micro-File AHU type 5459, but it is appropriate to examine several types of film to determine the best one for laser recording.

Film SNR depends upon the transmittance modulation of the signal on film versus the statistical properties of the emulsion constituents (grains). The rms granularity of film, σ_D, is analogous to white noise, limited to a finite spectrum by the aperture effect of the grains. The deviation ΔD from the mean density $\overline{D}$ varies inversely with the sampling area $\overline{a}$ over which the value of mean density is measured. The deviation also varies directly with the grain density level D at which it is measured.[7]

Available information on photographic film that relates to recorder SNR is given in terms of σ_D, which is a measure of variations in density in a film record made at a specific density and which is not found directly in actual experimental measurement. In practice, an rms variation in transmittance σ_T is measured at a specific average film transmittance $\bar{T}$. The published value of σ_D is obtained by calculating

$$\sigma_D = 0.4343 \frac{\sigma_T}{\bar{T}} \qquad (5)$$

This value of σ_D is given for a specific density value $\bar{D}$, which is related to $\bar{T}$ by the familiar relationship,

$$\bar{D} = \log_{10} \frac{1}{\bar{T}} \qquad (6)$$

Experimental data show that

$$\sigma_D \propto \bar{D}^x$$

where the value of x lies between 0.4 and 0.5, depending upon the film being considered. This relationship is illustrated in Fig. 16 for Micro-File AHU film.

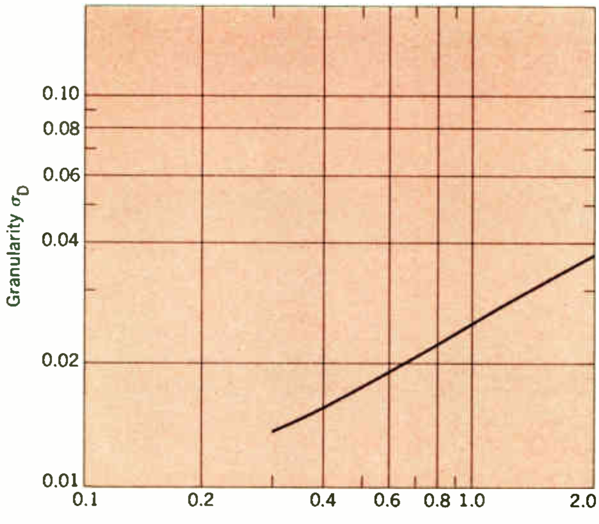
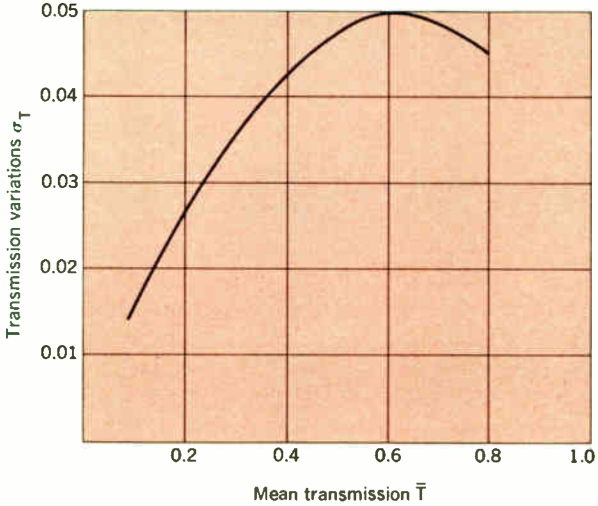


FIGURE 16. Relationship of σ_D to net optical density for Kodak Micro-File AHU type 5459.

FIGURE 17. Relationship of σ_T to mean transmission $\bar{T}$.



The value of σ_D is also a function of the measuring aperture used to determine this value. In general, the relationship between the measuring aperture and the value of σ_D is

$$\sigma_D \propto \frac{1}{d}$$

where d is the diameter of a circular measuring aperture. Since the curve in Fig. 16 is for a 12-μm circular aperture, the information must be converted to its equivalent value at a specific measurement aperture when specific evaluation of system performance is required.

The rms granularity σ_D for different film types is found in the manufacturers' specification sheets. For Micro-File AHU type 5459 (given in Kodak pamphlet no. P151), the rms granularity is 8.2 at a net density of 1.0. A note with the specification reads: "This value represents 1000 times the standard deviation in density produced by the granular structure of the material when a uniformly exposed and developed sample is scanned by a densitometer having an optical system aperture of $f/2.0$ and a circular scanning aperture 48 μm in diameter."

Although films can be compared on the basis of σ_D, the actual calculation of system SNR requires conversion of these values to σ_T, variations in film transmittance. Since in a laser beam recorder/reproducer the film transmittance modulates the reproducing laser beam, random variations in film transmission act as the basic noise source.

The signal-to-noise ratio is given by

$$\text{SNR} = \frac{M}{\sigma_T} \qquad (7)$$

where M is the transmittance modulation of the signal.

Using (5) and (7),

$$\text{SNR} = \frac{0.4343M}{\bar{T}\sigma_D} \qquad (8)$$

Note that σ_T is not constant with film transmittance, as can be seen in Fig. 17.

Table I lists σ_D for some candidate films, together with SNR. Since grain noise depends upon the number of grains under the examining aperture, high-resolution films that contain larger quantities of smaller grains exhibit less noise, as in the case of 649 emulsion. In addition, SNR is proportional to the square root of the aperture of the record/reproduce spot and can be increased by utilizing a greater spot-recording area.

Ultrahigh-resolution films and wider recording tracks require additional laser power. Table II lists SNR for

I. Comparison of film σ_D and exposure power*

Film Type	σ_D	SNR,† dB	Laser Power at Film,‡ μW
649 emulsion	0.008	31	3000
6451	0.016	25.5	300
5459	0.032	19.5	12
SO-243	0.045	16.5	14
3404	0.064	13.5	14

* 5-μm aperture and mean density = 0.3.
† Calculated for $\bar{T}$ = 0.5 and M = 0.33.
‡ Scan velocity = 127 000 cm/s.

II. SNR and laser power for candidate films at various record spot apertures

Film Type	Circular Aperture, 5-μm-diameter SNR, dB	Circular Aperture, 5-μm-diameter Laser Power, mW	Elongated Aperture, 5×20 μm SNR, dB	Elongated Aperture, 5×20 μm Laser Power, mW	Elongated Aperture, 5×80 μm SNR, dB	Elongated Aperture, 5×80 μm Laser Power, mW
649	31	300	37.0	4 800	43	77 000
6451	25.5	30	31.5	480	37.5	7 700
5459	19.5	1.2	25.5	19	31.5	310
SO-243	16.5	1.4	22.5	22	28.5	350
3404	13.5	1.4	19.5	22	25.5	350

candidate films for various record spot dimensions as well as the total laser power required for each condition. (Laser power is an estimate based on one percent light efficiency from laser output to the film. The laser power is for single-mode, single-line operation.) For example, although 649 emulsion will improve SNR 12 dB over Micro-File AHU type 5459, the sensitivity of the 649 emulsion is 1000 erg/cm^2 compared with 4 erg/cm^2 for type 5459. An increase in laser power of 250 to 1 is required to improve SNR by 12 dB through use of the fine-grain film.

If the recording spot is elongated 16 to 1 (5 μm by 80 μm), a 12-dB increase in SNR results at the expense of a 16-to-1 increase in light power at the film. Achieving the wider track by aperturing at the image lens increases the light power required at the imaging lens by another 16-to-1 factor, which ultimately requires a total increase in laser power of 256 to 1. As can be seen from Table II, 43-dB wide-band SNR is achievable but a 77-watt laser is required to make a recording.

Signal-to-noise ratio also can be improved by the usual bandwidth tradeoff (i.e., SNR is inversely proportional to the square root of the spatial bandwidth) and by achieving a greater transmittance modulation of the film. These two factors again require added laser power, but at a rate less than that required by utilization of more film grains. However, lowering the spatial bandwidth directly increases the scanner speed for a given record frequency. Achieving more transmittance modulation involves both an improved modulator capability and higher values for the modulation transfer function of the optics and film.

Image recording

The discussion of noise associated with laser signal recording also applies to laser imaging and facsimile systems. Stated simply, basic properties of noise hold for imagery except that the effect of grain noise is no worse than the effects of conventional photography. When a print is made from an original transparency, grain noise of the two emulsions adds on an rms basis. Thus, even though a facsimile signal scanned from an original image is low in SNR, the reconstructed image is degraded by only 3 dB, as would be a similar copy generated by standard photographic techniques. Equipment noise (laser, photodetector, and preamplifier) must be held to a minimum, of course, to prevent additional degradation of the image.

Conclusions

Noise in laser-film recording has been discussed, including several definitions and measurement techniques for evaluating noise properties. Equipment-generated noise can be reduced through careful design. Photodetector noise is potentially the worst equipment-noise contributor, but it can be effectively controlled by an appropriate multiplier phototube and sufficient laser power. The SNR of the photodetector is proportional to the square root of the incident light power on the photocathode. In a well-designed laser recorder the granularity of the film will limit the SNR. Again, SNR can be improved through the use of higher laser power, which permits use of fine-grain emulsions, wider record tracks, and high signal strengths. A wide-band SNR in excess of 30 dB is practical with present lasers and films; and improvements in laser power, optical efficiency, and film sensitivity offer a potential for considerably better performance. The techniques discussed here have been used in a 100-MHz signal recorder, but they can be extended to other phases of laser signal and image recording.

The writer expresses appreciation to his associates in the Electro-Optic Techniques Laboratory of RCA Advanced Technology Laboratories for consultation during preparation of this article.

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Compensating for propagation errors in electromagnetic measuring systems

Techniques have been developed that permit compensation for errors resulting from propagation effects in the atmosphere, thereby greatly increasing the effective accuracy of measuring systems

Sidney Bertram **The Bunker-Ramo Corporation**

The uncompensated effect of atmospheric refraction can compromise the accuracy of such equipment as tracking radars, photogrammetric cameras, and laser range finders. The refraction is caused by a lowering of the velocity of propagation of electromagnetic waves; this change in velocity has a direct effect on the measurement of distances using propagation time, whereas the deviation of the vertical angle is proportional to the rate of change of velocity with altitude. The errors can be reduced if the pertinent atmospheric parameters (such as temperature, atmospheric pressure, and water-vapor pressure) are known, or can be approximated, for the propagation path. This article derives straightforward techniques for the calculation of these errors and illustrates the procedures by exploring some typical optical refraction problems.

There are a number of familiar examples of equipments whose accuracies can be compromised if the effects of the slowing of electromagnetic signals in the atmosphere are not adequately compensated. The bending of the rays is an important consideration in telescopes, photogrammetric cameras, and tracking radars, and the increase in propagation time is important in today's increasingly accurate distance-measuring devices.

The literature includes descriptions of a number of techniques for handling the refraction problem. Horton[1] presented a variety of methods for handling refraction of underwater sound and showed that a ray traveling in a medium of constant-velocity gradient moves on a circular arc.

A number of authors[2,3] have described ray-tracing methods in relationship to atmospheric refraction. A closed solution has been obtained for the case of an exponentially varying atmosphere.[4] A solution for the angular error by perturbation techniques was given in 1966,[5] and a simplified calculation procedure making it practical to use a slide rule for determining the angular error for specific problems was described in a short note in 1969.[6]

One purpose of the present article is to provide a unified treatment of the application of perturbation techniques to the calculation of refraction errors, including the application of the method to the calculation of the refraction of nearly horizontal rays and to the correction of precision range measurements. The refraction errors for some typical optical problems are computed to illustrate the simplified procedures. The techniques are equally useful in radio-frequency propagation problems, once the pertinent atmospheric data have been acquired.

Velocity of propagation

The velocity of propagation of electromagnetic signals in a vacuum is the absolute constant c, whose accepted value is 299 792.5 km/s. Propagation through the atmosphere of the earth is significantly slower, decreasing with an increase in the atmospheric pressure. As a result of this, and in accordance with Fermat's assertion that propagation takes place over a path that minimizes the transit time, electromagnetic "rays" follow paths that curve downward so as to place the path above the shortest distance path and, therefore, in a region of higher velocity with a decrease in the time.

The propagation velocity v through a gas depends upon the index of refraction n, in accordance with the relationship

$$v = \frac{c}{n} \tag{1}$$

where, for RF waves traveling through the atmosphere of the earth, n_r is given by the empirical formula[7]

$$n_r = 1 + \frac{77.6 \times 10^{-6}}{T}\left(P + \frac{4810e}{T}\right) \tag{2}$$

where P = barometric pressure, millibars; e = water vapor pressure, millibars; and T = temperature, kelvins.

For signals in the visible region, the water vapor is not a factor, so only the first term within the parentheses in Eq. (2) is required for the representation of n_v. For such signals, one can write

$$n_v = 1 + 226\rho \times 10^{-6} \tag{3}$$

where ρ is the density of the atmosphere, in kilograms per cubic meter.*

The variability of the atmosphere makes it impossible

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* The index of refraction is dependent to some degree upon the wavelength[7,8]; over the separated RF and visible regions, the variation is adequately expressed by including the water-vapor term in n_r and excluding it from n_v.

to characterize the general velocity function, except on an average basis. The various detailed results reported here are based on Eq. (3), with the density distribution corresponding to that described in a report by Mizner and others[9]; they are therefore applicable specifically to light rays as used in optical telescopes, photogrammetry, and laser ranging—but only for the model atmosphere. Results for two extreme atmospheric situations are included to show the probable range of variation of refraction phenomenons for the optical case. At sea level in the standard atmosphere, the function $(n - 1)$ is about 15 percent higher for RF waves than it is for visible light; a short table for both is given by Reed and Russell.[10]

Refraction problem

Figure 1 shows a ray traversing a number of parallel layers having indexes of refraction $n_1, n_2, n_3 \cdots$. If we assume that the ray makes an angle θ_1 with respect to the normal to the planes in the first layer, the corresponding angles in the successive mediums are, in accordance with Snell's law, related by the expression

$$n_1 \sin\theta_1 = n_2 \sin\theta_2 = n_3 \sin\theta_3 = \cdots$$

Using $v = c/n$, this can be expressed in the form

$$\frac{\sin\theta}{v} = A \tag{4}$$

where θ is the vertical angle at a point where the velocity of propagation is v, and A is a constant for a given ray. By considering the successive layers to be of infinitesimal thickness it is readily seen that Eq. (4) is valid for the situation where v varies continuously.

It is useful to rewrite Eq. (4) in terms of the tangent so as to obtain the differential equation for the path; thus,

$$\tan\theta = \frac{dx}{dz} = \frac{Av}{\sqrt{1 - A^2v^2}} \tag{5}$$

where x is measured along the horizontal projection of the ray and z is measured along the vertical. The radius of curvature, r, of the ray is required in the sequel. From Eq. (5),

$$\frac{d^2x}{dz^2} = \frac{A(dv/dz)}{[1 - A^2v^2]^{3/2}} \tag{6}$$

and, using the following relationship for the radius of curvature

$$\frac{1}{r} = \frac{d^2x/dz^2}{[1 + (dx/dz)^2]^{3/2}}$$

we obtain

$$\frac{1}{r} = A\frac{dv}{dz} = \frac{\sin\theta}{v}\frac{dv}{dz} \tag{7}$$

Ray path

The curvature of the ray expressed by Eq. (7) provides a useful starting point for a calculation of the ray path. Since the deviation of the path from a straight line is extremely small, it is expedient to use a perturbation technique for its calculation. For this purpose a rotated coordinate system is useful. As shown in Fig. 2, this has one axis (u) tangent to the ray path at the ray origin and the second (w) providing a measure of the deviation from the nominal straight line path (that is, from the u-axis).

Since the velocity is very nearly constant, the ray path is quite straight; thus, $x \approx u \sin\theta$, $z \approx u \cos\theta$, and $dw/du \ll 1$. Using these approximations in the equation for the curvature yields

$$\frac{d^2w}{du^2} = A\frac{dv}{dz}$$

or

$$\frac{d^2w}{du^2} = \frac{\sin\theta}{v}\frac{dv}{dz} = \frac{\tan\theta}{v}\frac{dv}{du} \tag{8}$$

This is readily integrated if θ and v are considered to be essentially constant at their initial values. Thus,

$$\frac{dw}{du} = \left(\frac{v}{v_0} - 1\right)\tan\theta$$

and

$$w = \tan\theta \int_0^u \left(\frac{v}{v_0} - 1\right) du$$

Writing

$$\frac{c}{v} = 1 + N \tag{9}$$

where N is very small, we obtain

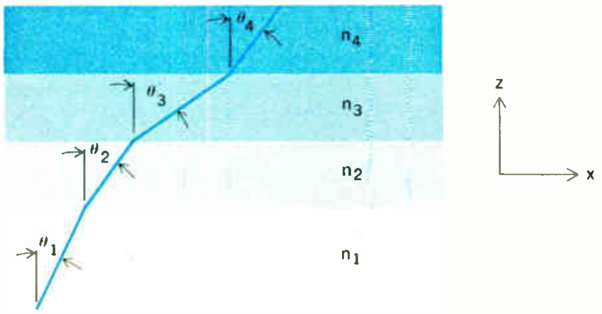
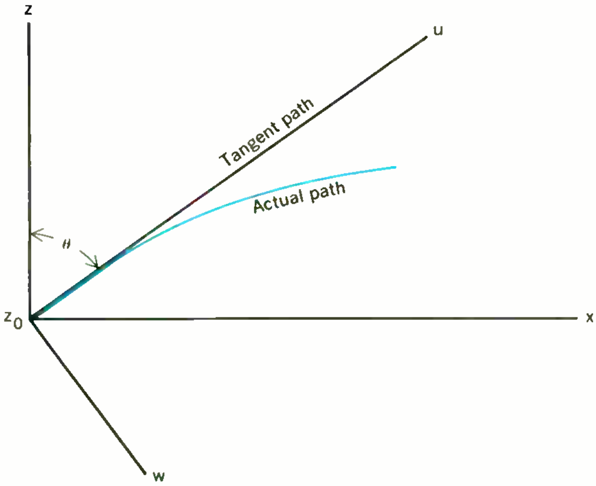


FIGURE 1. Illustration of Snell's law of refraction. ($n_1 \sin\theta_1 = n_2 \sin\theta_2 = n_3 \sin\theta_3 = \cdots$)

FIGURE 2. Rotation of coordinates to agree with tangent at observing position.



I. Inputs for calculation of optical refraction (ARDC model atmosphere, 1959)

Height z Above Sea Level, km	Density ρ, kg/m³	$N \times 10^6$	$I = 10^6 \int_0^z N\,dz$
0	1.2250	276.8	0
1	1.1117	251.2	264.0
2	1.0066	227.5	503.4
3	0.9093	205.5	719.9
4	0.8193	185.2	915.2
5	0.7364	166.4	1091.0
6	0.6601	149.2	1248.8
7	0.5900	133.3	1390.1
8	0.5258	118.8	1516.1
9	0.4671	105.6	1628.3
10	0.4135	93.5	1727.8
11	0.3648	82.4	1815.8
12	0.3119	70.5	1892.3
13	0.2666	60.2	1957.6
14	0.2278	51.5	2013.5
15	0.1947	44.0	2061.2
16	0.1665	37.6	2102.0
17	0.1423	32.2	2136.9
18	0.1216	27.5	2166.7
19	0.1040	23.5	2192.2
20	0.0889	20.1	2214.0
21	0.0760	17.2	2232.7
22	0.0650	14.7	2248.6
23	0.0556	12.6	2262.2
24	0.0475	10.7	2273.9
25	0.0406	9.2	2283.8
26	0.0344	7.8	2292.3
27	0.0291	6.6	2299.5
28	0.0247	5.6	2305.6
29	0.0210	4.7	2310.7
30	0.0179	4.0	2315.1
31	0.0152	3.4	2318.9
32	0.0130	2.7	2321.9
33	0.0112	2.5	2324.5
34	0.0096	2.2	2326.9
35	0.0083	1.9	2328.9
36	0.0071	1.6	2330.6
37	0.0061	1.4	2332.1
38	0.0053	1.2	2333.4
39	0.0046	1.0	2334.5
40	0.0040	0.9	2335.5

$$\frac{v}{v_0} - 1 = \frac{1 + N_0}{1 + N} - 1 \approx N_0 - N$$

and, therefore,

$$w = \tan\theta \int_0^u (N_0 - N)\,du \tag{10}$$

The angular deviation of the ray from the straight line joining the observer's position to the object is of interest. The deviation is

$$\Delta\theta = \frac{w}{u} = \tan\theta\left[N_0 - \frac{1}{u}\int_0^u N\,du\right]$$

or

$$\Delta\theta = \tan\theta\left[N_0 - \frac{1}{z - z_0}\int_{z_0}^z N\,dz\right] \tag{11}$$

It is useful to rewrite this in the form

$$\Delta\theta = K\tan\theta \tag{12}$$

where

$$K = N(z_0) - \frac{I(z_0) - I(z_1)}{z_0 - z_1} \tag{13}$$

in which

$$I(z) = \int_0^z N(z)\,dz \tag{14}$$

The integral $I(z)$ is readily evaluated numerically if the atmospheric parameters are known over the altitude range of interest. Table I shows a representative calculation for the ARDC (Air Research and Development Command) standard atmosphere with $N(z) = 226 \times 10^{-6}\,\rho$. The table permits a straightforward calculation for $\Delta\theta$, using Eq. (13), for altitude pairs (not equal) from sea level to 40 km; since N as listed is in millionths, I is in microradians at 45°.

Nearly horizontal rays

The dependence of the propagation velocity on altitude makes the altitude an expedient choice for the independent variable in refraction calculations. However, it leads to a problem in the consideration of rays that are nearly horizontal, where a small altitude increment is accompanied by a large horizontal change; the problem is exemplified by the tan θ term in Eq. (12).

II. Variation of deviation angle with observation angle for fixed path length of 50 km, z_0 = 10 (ARDC model atmosphere, 1959)

z_1, km	$z_0 - z_1$	$\frac{z_0 - z_1}{R}$	θ	I_1	$I_0 - I_1$	$\frac{I(z_0) - I(z_1)}{z_0 - z_1}$	K	$\Delta\theta$, µrad
0	10	0.20	11.53	0	1727.8	172.8	79.3	388
1	9	0.18	10.37	264.0	1463.8	162.6	69.1	378
2	8	0.16	9.20	503.4	1224.4	153.0	59.5	367
3	7	0.14	8.05	719.9	1008.9	144.1	50.6	357
4	6	0.12	6.90	915.2	812.6	135.4	41.9	346
5	5	0.10	5.74	1091.0	636.8	127.3	33.8	336
6	4	0.08	4.59	1248.8	479.0	119.7	26.2	327
7	3	0.06	3.44	1390.1	337.7	112.6	19.1	318
8	2	0.04	2.30	1516.1	211.7	105.8	12.3	307
9	1	0.02	1.15	1628.3	99.5	99.5	6.0	300
10	0	0	0	1727.8	0	—	—	299

The radius of curvature, given by Eq. (7), suggests that the refraction of a nearly horizontal ray over a path of given length should be fairly independent of the ray angle. Table II, prepared using Eqs. (12) and (13) with the aid of Table I, exhibits the variation of the deviation angle as a function of the source elevation for observations made at an elevation of 10 km (33 000 ft) and a fixed path length (slant range) of 50 km. It will be noted that the deviation for the sea-level origin (viewing elevation angle of $-11.5°$) is 388 μrad with the deviation angle decreasing with increasing source altitude to about 300 μrad for the horizontal ray (the value for $\theta = 0°$ was obtained using the method of the next paragraph).

For moderate horizontal distances and nearly horizontal rays, the altitude variation is small. It is then useful to consider the velocity function to be expanded in a Taylor series; that is,

$$\frac{dv}{dz} = v_0' + v_0''(z - z_0) + \frac{v_0'''}{2}(z - z_0)^2 + \cdots \qquad (15)$$

where $v_0' = dv/dz$, $v_0'' = d^2v/dz^2$, etc., at the observation point. Then, since $z \approx z_0 + u\cos\theta$, Eq. (8) becomes

$$\frac{d^2w}{du^2} = \frac{\sin\theta}{v}\left[v_0' + v_0''u\cos\theta + \frac{v_0'''u^2}{2}\cos^2\theta + \cdots\right] \qquad (16)$$

and, therefore,

$$w = \frac{\sin\theta}{v}\left[\frac{v_0'u^2}{2!} + \frac{v_0''u^3}{3!}\cos\theta + \frac{v_0'''u^4}{4!}\cos^2\theta + \cdots\right] \qquad (17)$$

For very large values of u it would be necessary to correct for the fact that the deviation w corresponds to a z displacement, $z = -w \sin\theta$, with a corresponding change in the velocity. This can be accommodated by stopping the foregoing solution and starting a new one before the error is too great, or by introducing $-w\sin\theta$ as a z change before the integration process; the latter procedure is followed in the analysis of the effect of the earth's curvature in the next section. With w calculated, then $\Delta\theta = w/u$.

Effect of the curvature of the earth on refraction

The curvature of the earth affects the refraction calculation in the following ways:

1. The reference altitude seems to fall away at a rate $x^2/2R$, where R is the radius of the earth.
2. The angle θ is variable, decreasing at the rate $1/R$ with x.

Since x/R is small,

$$\tan\left(\theta - \frac{x}{R}\right) \approx \tan\theta - \frac{x}{R}\sec^2\theta$$

Thus, using x as the independent variable,

$$\Delta\theta = \frac{1}{x}\int_0^x \left[N_0 - N\left(z_0 + x\cot\theta + \frac{x^2}{2R}\right)\right] \times \left(\tan\theta - \frac{x}{R}\sec^2\theta\right)dx \qquad (18)$$

The maximum range for which adequate results can be obtained without the correction for the curvature of the earth is relatively independent of the desired accuracy because of the rate at which the correction changes. The results for a horizontal ray near sea level in the standard atmosphere can be used as a guide.

From Eq. (8), noting that $\theta = 90°$, we obtain

$$\frac{d^2z}{dx^2} = -\frac{v_0'}{v_0} \qquad (19)$$

and, therefore,

$$z - z_0 = -\frac{v_0'}{2v_0}x^2 \qquad (20)$$

is a first approximation to the ray path. A second approximation to the ray path is obtained by replacing v_0' in Eq. (19) by the first two terms of the Taylor expansion, Eq. (15), using the first-approximation deviation and the earth's curvature correction, $x^2/2R$; thus,

$$\frac{d^2z}{dx^2} = -\frac{v_0'}{v_0} - \frac{v_0''}{v}(z - z_0)$$

$$= -\frac{v_0'}{v_0} - \frac{v_0''}{v}\left(-\frac{v_0'}{2v_0}x^2 + \frac{1}{2R}x^2\right)$$

Integrating to find z and dividing by θ to obtain the deviation angle now yields

$$\Delta\theta = \frac{z}{x} = -\frac{v_0'}{2v_0}x + \left(\frac{v_0''}{v_0}\right)\left(\frac{v_0'}{v_0}\right)\frac{x^3}{4!} - \left(\frac{v_0''}{v_0}\right)\frac{1}{R}\frac{x^3}{4!} \qquad (21)$$

The significant parameters in Eq. (21) are

$$\frac{1}{v}\frac{dv}{dz} = -\frac{1}{1+N}\frac{dN}{dz} \approx -\frac{dN}{dz}$$

and

$$\frac{1}{v}\frac{d^2v}{dz^2} \approx -\frac{d^2N}{dz^2}$$

For the ARDC model the lowest 10 km can be adequately represented in the visible range by

$$N = 277 - 25z - 0.66z^2 \qquad (22)$$

(where z is in kilometers), so

$$\frac{v'}{v} = 25 \times 10^{-9} \text{ units/meter}$$

and

$$\frac{v''}{v} = -1.32 \times 10^{-12} \text{ units/meter}^2$$

Using $R = 6.4 \times 10^6$ meters as the radius of the earth,

III. Refraction, in microradians, of horizontal ray observed at sea level

Distance, meters	First Iteration Refraction	Second Iteration Refraction	Earth's Curvature Correction
1 000	−12.5	$+1.37 \times 10^{-6}$	-8.6×10^{-6}
10 000	−125	$+1.37 \times 10^{-3}$	-8.6×10^{-3}
100 000	−1250	+1.37	−8.6

the deviation angle becomes

$$\Delta\theta = -12.5 \times 10^{-9}x + 1.37 \times 10^{-21}x^3 - 8.60 \times 10^{21}x^3$$

The three terms of the deviation angle are shown separately in Table III for several distances. It will be noted that the second-order correction term and the correction due to the curvature of the earth are unimportant to about 100 km, but increase rapidly for longer distances.

In tracking radar applications the vertical deviation, rather than the angle, is important. The first iteration refraction given by Eq. (20) is small compared with the earth's curvature correction, $x^2/2R$; the combined effects are often approximated by considering that the radius of the earth has been increased from R to $^4/_3R$. An equivalent correction for optical instruments operating in the standard atmosphere would be obtained by replacing R by $^6/_5R$.

Distance measurements

The transit time of an electromagnetic signal in passing between two points provides a measure of the distance between the two points. More precisely, it provides the value of the integral

$$\int \frac{ds}{v}$$

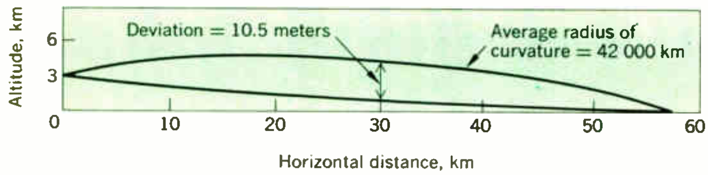
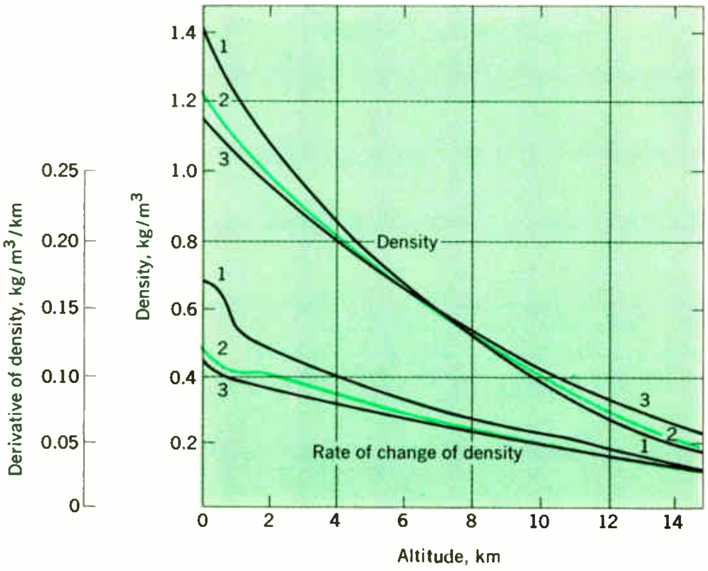


FIGURE 3. Distance calculation problem for stations at 3 km and sea level, 58.3 km apart. (Curvature of ray is greatly exaggerated.)

FIGURE 4. Density and derivative of density for (1) arctic (January, 75°N, T = 249 K), (2) ARDC standard atmosphere, and (3) tropical (15°N, T = 300 K).



over the actual path taken by the ray. This section is devoted to the problem of converting such a measured time to an accurate measure of the distance between the points.

The AGA Laser-Geodimeter, with an advertised accuracy of 6 mm + 1 part per million to distances up to 60 km, provides a useful context for the discussion. Quite obviously the potential accuracy of such equipment cannot be realized without an appropriate compensation for atmospheric effects.

Suppose the distance between two stations—one at an elevation of 3 km above sea level and the second at sea level (as shown in Fig. 3)—is to be calculated from a measured one-way propagation time of $1.945\,276 \times 10^{-4}$ second. Using $N = 239 \times 10^{-6}$, interpolated from Table I, as the refraction index correction term at the average altitude of 1.5 km, the approximate range is seen to be

$$\text{Range} \approx \frac{(299\,792.5) \times (1.945\,276) \times 10^{-4}}{1.000\,239}$$

$$= 58.303\,98 \text{ km}$$

The denominator factor (i.e., 1.000 239) accounts for a correction of 13.93 meters; the equipment uncertainty is 0.06 meter.

Equation (7) can be used to estimate the effective curvature of the ray. In the interval $0 < z < 3$ km, the average

$$\frac{1}{c}\frac{dv}{dz} \approx -\frac{dN}{dz}$$

is

$$\frac{1}{c}\frac{dv}{dz} = -\frac{(276.8 - 205.5)}{3} = -23.8 \text{ microunits/km}$$

Since the ray is nearly horizontal, $\sin\theta \approx 1$; the radius of curvature is then

$$r \approx \frac{c}{dv/dz} = \frac{10^6}{23.8} = 42\,000 \text{ km}$$

The angle subtended by a chord 58.3 km long for a circle of this radius is $58.3/42\,000 = 1.39 \times 10^{-3}$ radian, and the maximum deviation from the straight line is

$$d \approx 42\,000\left[1 - \cos\left(\frac{1.39 \times 10^{-3}}{2}\right)\right] \approx 1.05 \times 10^{-2} \text{ km}$$

or about 10 meters. This is insignificant both with respect to the change in N and with respect to the change in the length of the line. Hence, adequate results can be obtained by assuming that the line is straight.

The actual range calculation is performed as follows. If t is the measured time, then

IV. Determination of distance correction

u	$\frac{2.74u}{58.3}$	$\frac{x^2}{2R}$	z	N	$\overline{N}$ du
0	0	0	0	277	
10	0.47	0.01	0.48	264	2 705
20	0.94	0.03	0.97	252	2 580
30	1.41	0.07	1.48	239	2 455
40	1.88	0.12	2.00	227	2 330
50	2.35	0.20	2.55	215	2 210
58.3	2.74	0.26	3.00	205	1 743
					$\Sigma\overline{N}$ du = 14 023

$$\text{Range} = c\int_0^t \frac{dt}{1+N} \approx ct - \int_0^u N\,du \qquad (23)$$

where u is the nominal range. (For $N < 300 \times 10^{-6}$, the approximation $1/(1+N) \approx 1 - N$ is good to one part in 10^7. The use of the nominal range as the limit of integration does not introduce a significant error because the integral provides a very small correction.) The integration is shown carried out numerically in Table IV. At the maximum range (58.3 km), the earth's curvature correction is $58.3^2/[2 \times 6.4 \times 10^6] = 0.26$ km. This quantity has been subtracted from the 3-km elevation of the upper station to obtain the altitude with respect to a tangent plane at the sea-level station. The intermediate altitudes with respect to the tangent plane are taken to be proportional to the distance. The earth's curvature correction is then added to obtain the height above sea level for use in determining the N values. The range, corrected in accordance with Eq. (23), is therefore

$$U = (299\,792.5) \times (1.945\,276 \times 10^{-4}) - (14\,023 \times 10^{-6}) = 58.303\,90 \text{ km}$$

This is slightly less than the value estimated earlier from the average altitude. The difference is largely attributable to the fact that the altitude used should have been somewhat lower because of the effect of the earth's curvature.

If both stations were at sea level, the radius of curvature would be at its minimum value of about 40 000 km; this would yield a vertical offset of about 30 meters for a 100-km path and a consequent change in N of about 0.7. This is a change of less than 1 part in a million (and that change over only a part of the path). The offset increases as the square of the range, and would be a significant factor at ranges over 100 km if comparable accuracies were desired. It seems unlikely, however, that the atmospheric conditions for such long paths would be known in sufficient detail to permit meaningful calculations to such high accuracy.

Refraction for other atmospheres

It is possible to obtain some insight into the variability of the refraction problem from published data for some extreme atmospheric conditions.[11] Figure 4 shows the density variation for the ARDC model atmosphere (curve 2), along with corresponding curves for the density in the arctic (curve 1; January, 75°N, $T = 249$ K) and for the tropics (curve 3; 15°N, $T = 300$ K). There is seen to be a total variation of about 20 percent between the two extremes at sea level, with a smaller variation at the higher altitudes; this variation would reflect directly into the range-correction calculations. The curves indicate that precise knowledge of the atmospheric conditions is not very important for range measurements.

The change in the deviation angle is more severe, since it depends upon the rate of change of density. As shown in Fig. 4, this angle changes nearly 50 percent between the tropics and the arctic winter for elevations near sea level. At the higher altitudes the local effect is much less severe.

Conclusion

The correction of angle and distance measurements for the effects of refraction has been shown to be a straightforward procedure once the index of refraction function is known for the area of interest. For visible rays the technique requires a knowledge of the variation of the pressure and temperature that together determine the density of the atmosphere. For radio waves it is also necessary to know the pressure of the water vapor. The accuracy of the refraction calculations is limited by the accuracy with which the index of refraction function over the propagation path is known.

The solution assumes that the horizontal gradient is negligible. If this is not true for a given situation, the problem could be handled by breaking the ray into a number of horizontal segments where the variation is negligible on each segment.

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Sidney Bertram (F) received the A.A. degree from Los Angeles City College, the B.S. degree (with honor) in engineering from the California Institute of Technology, and the M.S. degree in electrical engineering and the Ph.D. degree in physics from Ohio State University. He was an instructor at the Radio Institute of California (1935–36), a research engineer for International Geophysics Co. (1938–39), and a research associate at Ohio State University's Antenna Laboratory (1940–41). During World War II his circuit development work at the University of California Division of War Research was a key factor in the success of a PPI sonar system. In 1945 he became involved in the development of a missile guidance system at Boeing Aircraft Co. He returned in 1946 to Ohio State University as an assistant professor in electrical engineering. He joined the Rand Corp. in 1951 and in 1953 transferred to the System Development Corp., where as head of the Engineering Department he was in charge of efforts in implementing the Air Defense Command's System Training Program. He is now senior staff scientist at the Bunker-Ramo Corp., Westlake Village, Calif., where his recent work has led to the practical realization of equipment to measure altitudes automatically from stereo aerial photographs. He is the author of many papers and holds a number of patents.

Biomechanical studies in aerospace physiology

How well will an astronaut's cardiovascular system supply blood to his brain and muscles after a month of weightlessness? What causes space sickness? Interdisciplinary research teams are using engineering tools and techniques to answer such questions.

Max Anliker Stanford University
John Billingham Ames Research Center

The field of aerospace medicine has long called for an engineering approach to modeling human systems. The behavior of the cardiovascular and vestibular systems in an aerospace environment is an area where the collaborative research of physicians, engineers, and physical scientists has been rewarding.

When men are subjected to unusual conditions such as high accelerations and weightlessness, we would like to be able to predict how well they function and to what extent their health, safety, and performance are influenced. Reliable predictions, however, can only be made if we have a thorough quantitative understanding of the physiological functions affected or if we have extensive empirical data from which we can deduce patterns of possible responses to this kind of environment. But since we have only a very limited knowledge of the effects of weightlessness, radiation, and confinement on man, we have to anticipate the potential hazards, develop methods and techniques of assessing the corresponding response of astronauts, and devise preventive measures.

Motivated by these facts, a number of interdisciplinary research teams have been formed at various universities and NASA centers. Most of these teams consist of biomedical scientists and engineers who are directing their attention to problems of a general biomedical nature[1,2] or to problems within specific areas such as respiration,[3] circulation,[1] and vestibular function.[5,6] One of the common features of many of the efforts is that they are aimed at two major goals: to establish a better quantitative understanding of certain physiological phenomenons and their control mechanisms, and to develop noninvasive methods and the necessary instrumentation to measure essential physiological parameters and quantities.

A thorough quantitative understanding of human physiology would permit us to design new diagnostic tools for clinical medicine and to utilize better naturally occurring physiological phenomenons as indicators of health or disease. Within the field of cardiovascular dynamics, for example, we still have numerous theories for the genesis of heart sounds, the spatial variations in the arterial flow and pressure pulses, and the relationship between the electrical and mechanical activities of the heart. In addition, our comprehension of the regulatory mechanisms of circulation and physiological functions in general is still somewhat vague[7,8] since we are lacking reliable methods of distinguishing between active and passive responses of the system and also techniques for accurate noninvasive measurements of changes in the controlled and controlling parameters. With regard to vestibular functions we even have to question our qualitative understanding of the mechanisms responsible for the sensation of changes in linear and angular motions.

Circulation

The cardiovascular system is significantly affected by the stresses of the aerospace environment, and particularly by the alterations in force fields encountered during weightlessness and the high accelerations associated with the maneuvers of aerospace vehicles. We are still uncertain how well the cardiovascular system will retain its ability to supply adequate blood flow to the brain and muscles after a prolonged exposure to weightlessness, particularly during reentry and on reassumption of an upright posture in the normal gravitational field.

As opportunities for carrying out experiments in orbiting vehicles slowly increase, it is important to develop experimental methods that will give us the best possible information on what is happening to the human cardiovascular system. It would be much easier to acquire information from chronically instrumented animals. However, this would still leave us with two major problems. First, extrapolations have to be made from the animals to man. Second, restrained animals subjected to a strange environment and to surgical trauma do not necessarily give the same data as the intact animal in his normal habitat. Clearly, it is better to conduct studies on astronauts whenever possible. Yet extensive experimental procedures such as cardiac catheterization cannot be contemplated at this stage since the conditions in a space vehicle are simply not equivalent to those of the appropriately equipped clinical research laboratory. Moreover, it is very important to avoid alterations in the physiological reactions that may be induced by traumatic procedures. The conclusion is obvious. We must develop experimental procedures that allow us to acquire accurate data on the cardiovascular system parameters without significantly disturbing the subject. For example, we have to find ways of measuring, without penetrating the skin, the local *distensibility* of blood vessels and their geometry, which together play a particularly relevant role in maintaining and regulating blood flow. Since there is generally no direct method of determining the local distensibility of the blood vessels and other mechanical parameters without skin penetration, we resort to indirect approaches. We can, for example, make use of a relationship between the distensibility of a vessel, its geometry, and its wave transmission properties.[9] Such a relationship can be derived by using methods of aeroelasticity; the wave transmission characteristics of certain vessels and their geometry can be determined with the help of ultrasound devices that are currently available[10,11] or being developed. The interpretation of the data obtained in this manner is, however, only as accurate as the mathematical model that describes the mechanical behavior of the vessels. Studies are therefore in progress to develop a sufficiently accurate mathematical model and the instruments that would permit us to evaluate the changes in the mechanical properties of arteries and veins as a result of aging, disease, or exposure to weightlessness.

Without exception, the models postulated to date for the cardiovascular system appear to oversimplify the actual situation, and the data supporting them are at best fragmentary. Many of these models may well be suitable for order of magnitude type studies of the gross behavior of the system. However, they probably will have to be refined if we want to use them to quantify small changes in the cardiovascular parameters from indirect measurements. To clarify this point a team of engineers and biomedical scientists from Stanford University and NASA's Ames Research Center have begun an evaluation of the limitations of various models by comparing theoretical predictions with the results of appropriately designed experiments on anesthetized dogs.

In several recent theoretical studies[12,13,14,15] veins and arteries have been characterized as elastic or viscoelastic circular cylindrical shells with a prescribed axial stretch and transmural pressure. Perturbations or waves were defined in terms of three displacement components of a generic point of the middle surface in the vessel wall. By adhering strictly to linear theory, one finds that there are basically three types of waves that can propagate in blood vessels. These waves are usually distinguished by the dominant component of the wall displacement at high frequencies and are accordingly referred to as *radial, circumferential*, and *axial* waves. Since the radial waves are associated with strong transmural pressure fluctuations, they are frequently also called pressure waves. Likewise, the circumferential waves are usually also referred to as torsion waves to characterize the mode of deformation. Except at very low frequencies, the speed c of pressure waves is approximately given by the so-called Moëns-Korteweg equation

$$c^2 = \frac{E}{2\rho_f}\left(\frac{h}{a}\right)$$

where E is the effective Young's modulus for the vessel wall material in the circumferential direction, ρ_f is the density of the blood, and h/a is the wall thickness-to-radius ratio. For isotropic wall properties, the theory predicts that the pressure waves are the slowest, while for a typical artery like the carotid, the torsion waves should be about two and a half times faster, and the axial waves

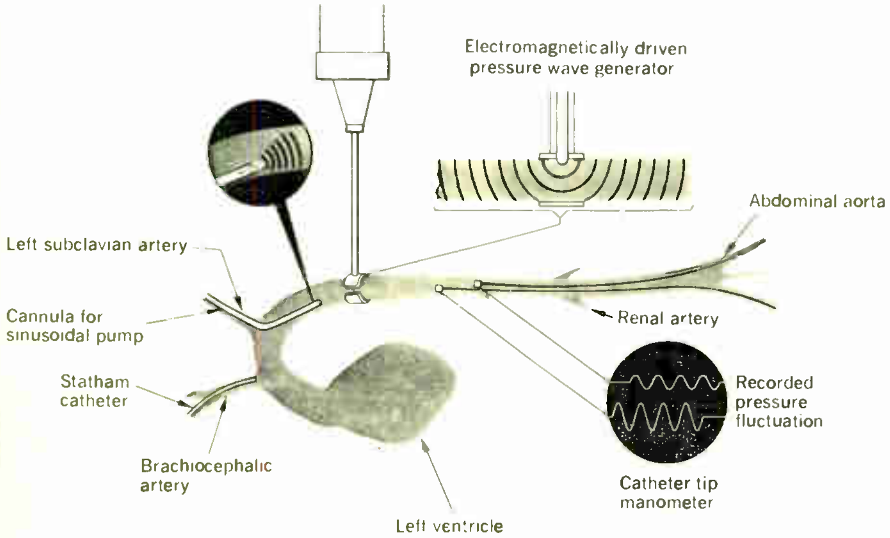


FIGURE 1. Experimental setup for generating small sinusoidal pressure waves in the canine aorta by a sinusoidal pump or an electromagnetically driven pressure wave generator. The transient signals are sensed by wind tunnel pressure gauges that were adapted for use as catheter-tip manometers in blood vessels.

about five times faster. Also, all three types of waves show either no or only mild dispersion for frequencies up to 200 Hz.

In the past all experimental studies of wave transmission in blood vessels have been restricted to large-amplitude pressure waves that do not allow a validation of theoretical predictions based on a linear analysis. A series of experiments[16,17,18,19] have therefore been initiated in which small sinusoidal waves of all three types at frequencies between 20 and 200 Hz have been induced in blood vessels of anesthetized dogs. Figures 1 and 2 illustrate the experimental arrangements used for the aorta and the inferior vena cava, respectively. The physiological experiments conducted were unique insofar as they utilized instrumentation originally developed for strictly engineering purposes. For example, the pressure transducers employed were designed for wind tunnel experiments and as such allowed for the resolution of pressure fluctuations of less than 1 mm H_2O at frequencies far beyond the normal physiological range of 0–1000 Hz. To avoid reflection interference, which would complicate the determination of wave dispersion and attenuation, the experimenters in this group have induced signals that were either of such a short duration that they could be recorded before their reflections arrived at the transducer sites, or of such small amplitudes that the reflections were damped below recognition before they reached the transducer locations. Since blood vessels are apparently only mildly dispersive in the frequency range considered, it was possible to circumvent laborious Fourier transform computations and consider the signal speed of the finite wave trains as an approximation for the phase velocity corresponding to the frequency of the sine waves in these trains.

Figure 3 illustrates representative tracings of a recording of the natural pulse wave in the thoracic aorta of an anesthetized dog with artificially superimposed trains of sinusoidal pressure waves. From the absence of any marked distortions of these waves during transmission we can infer that the thoracic aorta is indeed only mildly dispersive with respect to radial waves as predicted by the theory and that there is no noticeable evidence of reflection interference.

The results from the aortic studies have shown that pressure waves with frequencies between 20 and 200 Hz are essentially nondispersive but also are strongly attenuated, primarily due to the viscoelastic nature of the vessel wall. It appears that for all frequencies the amplitude ratio A/A_0 decreases in the same exponential fashion with distance measured in terms of wavelengths:

$$\frac{A}{A_0} = e^{-k\Delta x/\lambda}$$

where k is the logarithmic decrement, λ the wavelength, and where A_0 and A are defined as shown in Fig. 3. This attenuation pattern identifies to a considerable extent the viscoelastic nature of the vessel wall. The experiments on the aorta have also verified the experimental prediction that in the presence of a mean flow the pressure signals are being convected by the mean flow velocity.[20] In addition, the results demonstrated that the wave speed increases

FIGURE 2. Experimental setup for the study of the transmission characteristics of small, artificially induced pressure waves in the abdominal vena cavae (large veins that feed blood back to the heart) of anesthetized dogs. The volumetric displacements of a vibrating piston in the right common iliac vein generate sinusoidal pressure signals. A catheter-tip balloon, positioned between the heart and the hepatic veins, assesses the changes in the mechanical behavior of the vessel as a function of pressure. Inflation of the balloon produces increases in venous pressure of about 100 mm H_2O within 10 to 20 seconds. The internal diameter of the vena cava is monitored with a modified Pieper diameter gauge. Figures 5 and 6 illustrate the effects of these pressure waves.

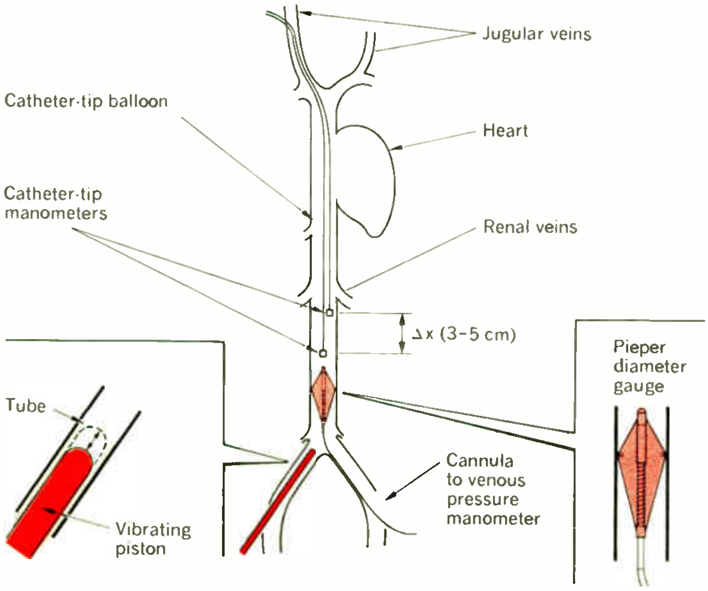
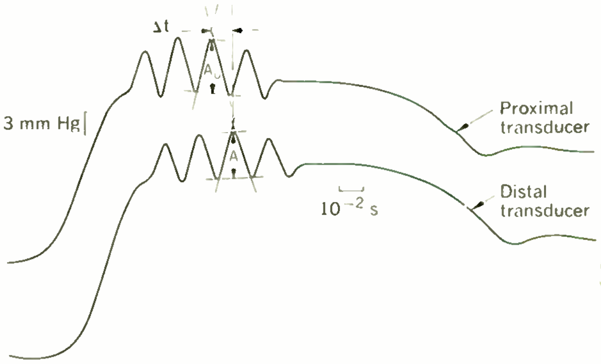


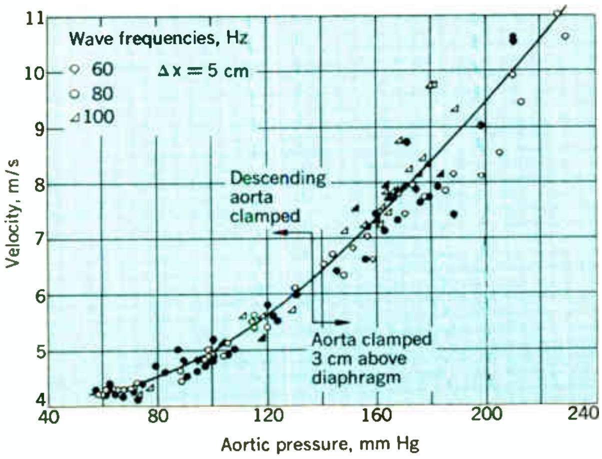
FIGURE 3. A recording of a natural pulse wave propagating downstream in the canine thoracic aorta. Transient signals in the form of finite trains of sinusoidal waves have been superimposed on the natural pressure pulse at various instances of the cardiac cycle. The perturbations retain their sinusoidal waveform during transmission over the distance Δx (4 cm) between the proximal and distal transducers, but they are slightly damped. No forerunners can be discerned. In this Visicorder recording of the train of four sine waves induced at systole (the rhythmic contraction of the heart), A_0 is the amplitude of wave measured by proximal transducer, and A is the amplitude measured by distal transducer. The speed of sinusoidal perturbations at 70 Hz is determined by measuring the time Δt that it takes a characteristic signal point to travel the distance Δx. Here, the characteristic point is the intersection of tangents in two successive inflection points. Although the aortic pressure does not vary much between the beginning and end of the perturbation, a detailed inspection reveals a systematic change of the signal speed with selection of the inflection points. This speed variation is due to the rapid changes in mean flow that occur during systole.



sharply with transmural pressure (Fig. 4), which is contrary to what we would expect in an elastic fluid-filled tube whose wall exhibits the mechanical properties of conventional materials. Any rise in transmural pressure should cause an increase in the diameter and a thinning of the wall, which together would cause a decrease in the speed of pressure waves according to the Moëns-Korteweg equation if the effective Young's modulus of the wall material remains substantially unchanged. We conclude therefore that the increase in the circumferential wall stress produced by the rise in transmural pressure must lead to a dramatic increase in the effective Young's modulus. This means that large-amplitude pressure signals such as those generated by the heart will undergo a steepening of the wavefront and other shape changes as a result of the pressure dependence of the speed. As a matter of fact, when the aortic valve is incompetent (that is, when it no longer provides a good seal between the aorta and the left ventricle and thus allows for extension regurgitation of blood back into the ventricle), we may have such a strong pressure pulse that a shock wave can form within a relatively short distance from the heart. Such shock waves have been predicted in theoretical studies using the method of characteristics and data acquired in dog experiments,[21] and it can be argued with reasonable certainty that these shock waves are responsible for the so-called pistol shot phenomenon that is one of the diagnostic indicators for aortic valve incompetence.[22]

Findings of studies on the carotid artery, in which not only pressure waves but also torsion and axial waves were generated, show that both the torsion and axial waves are not being propagated with the predicted speeds.[23] This discrepancy between theory and experiment suggests that the wall of the carotid is not isotropically elastic and has an effective Young's modulus that is lower for the axial direction than it is for the circumferential direction.

FIGURE 4. Wave-speed variation for a segment of the thoracic aorta in an anesthetized dog. Occluding the aorta above and below the segment of interest caused the pressure to vary beyond its normal range. The solid points represent waves recorded as they propagate in a downstream direction (away from the heart), and the open points were obtained from waves traveling in the upstream direction. The sharp rise of wave speed with pressure suggests that the peak of a large-amplitude pressure pulse travels at a higher speed than does the foot of the pulse. Accordingly, such pulses change their shape with propagation.



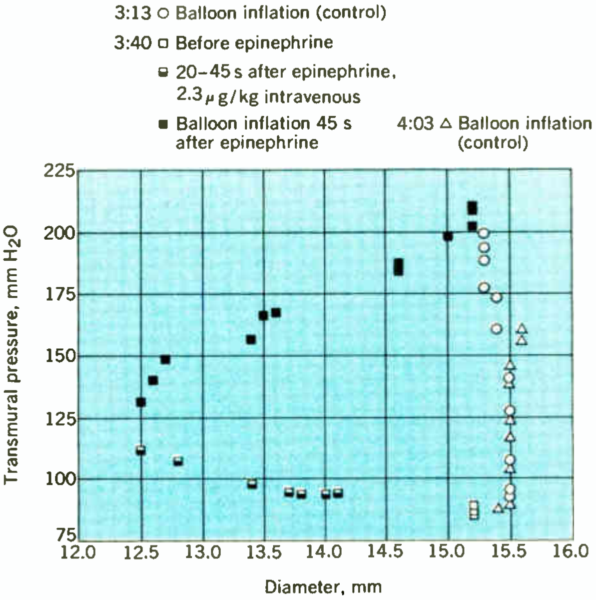
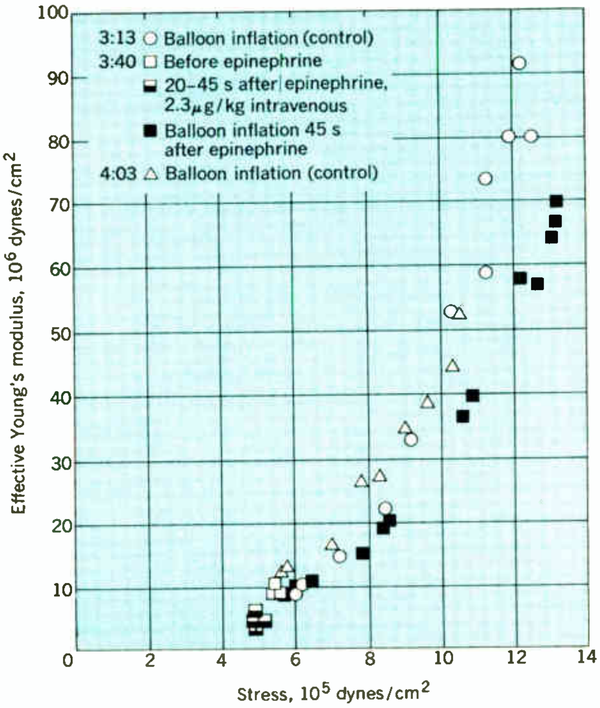


FIGURE 5. Effects of pressure and epinephrine on the diameter of the vena cava. Control balloon inflations (Fig. 2) before the injection of epinephrine show that the vessel is "relaxed," but also stiff, since it does not change its diameter with pressure. In response to the epinephrine, the vessel constricts markedly, while the pressure increases about 50 percent. A subsequent balloon inflation shows that with rising pressure the diameter of the "stimulated" vessel increases again to its control value, which indicates that the constricted vessel is more distensible than the relaxed vessel.

FIGURE 6. Effects of pressure and epinephrine on the effective Young's modulus of the vena cava. Measurements of the diameter of the vena cava and the wave speed, together with the transmural pressure, can be used to determine the instantaneous effective Young's modulus as a function of the circumferential wall stress. Young's modulus is given by the Moens-Korteweg equation (see text).



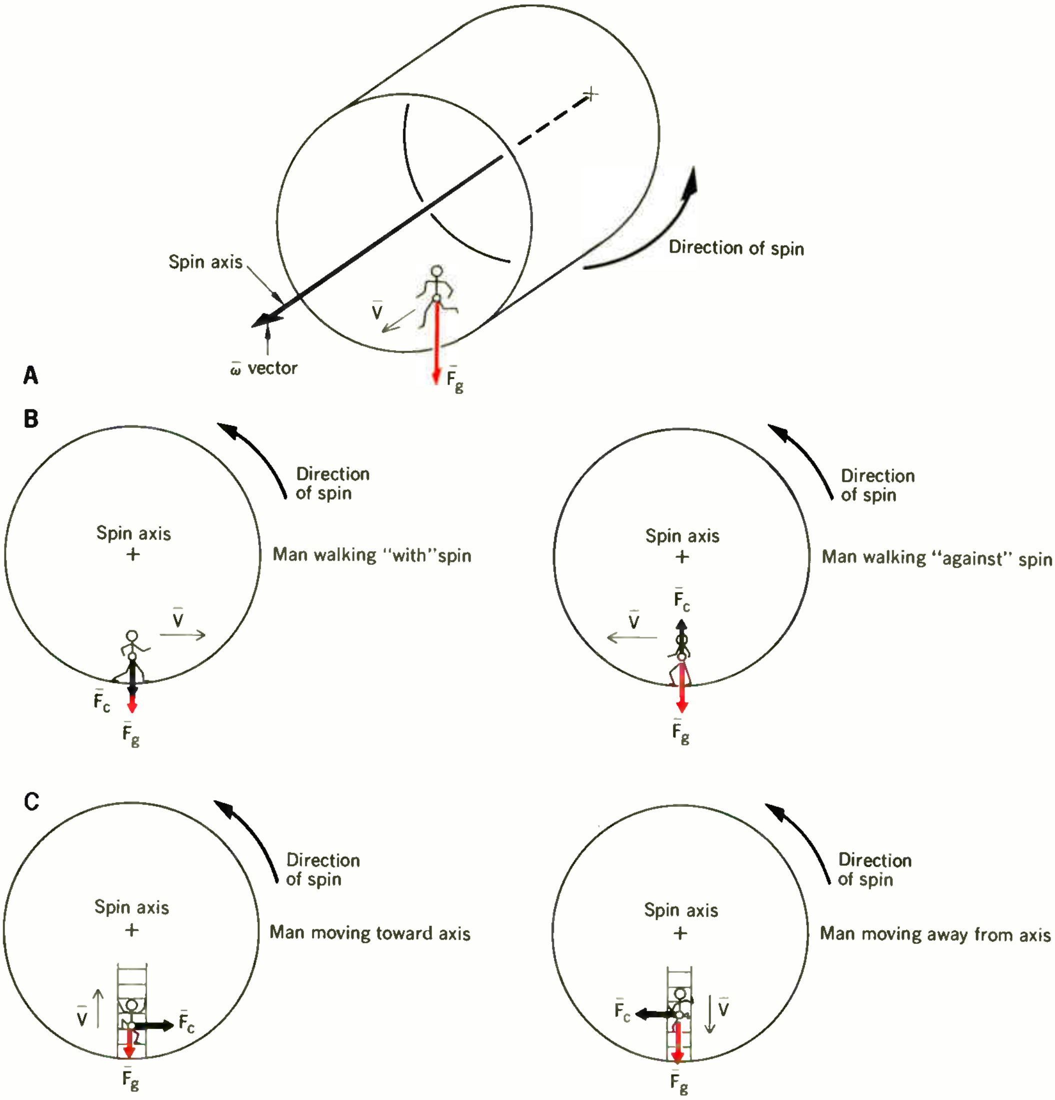


FIGURE 7. A—During rotation of the spacecraft, movement of the man in a direction parallel with the axis of rotation produces no Coriolis force and therefore no disorientation. B—Movement of the man in or against the direction of spin produces a Coriolis force that increases or decreases his weight. C—Movement of the man toward the axis of rotation produces a Coriolis force that is felt as a lateral force.

Perhaps most intriguing to engineers and physical scientists are the observations made in the experimental analysis of the distensibility of the vena cava and its response to changes in transmural pressure, nerve stimulation, and hormone (epinephrine) injections into the vessel.[19] In general, the speed of the pressure waves increases with the transmural pressure. This means that for the vena cava, just as in the case of the aorta and carotid artery, the effective Young's modulus for the circumferential direction increases with rising transmural pressure or wall stress. Also, simultaneous measurement of the internal diameter and the wave transmission characteristics of the vessel demonstrates that the stimulation of the smooth muscle in the vena cava wall causes the vessel to constrict and generally lowers the effective Young's modulus for a given wall stress. The stimulation elicits an active response of the vessel that can manifest itself by a paradoxical situation where a decreasing diameter exists in the presence of an increasing transmural pressure. Figure 5 illustrates such a situation and shows that the stimulated vessel can be expected to be more distensible than the "relaxed" vessel. Since we also know the diameter, the wall thickness, and the speed of pressure waves at any instant, we can determine the active changes of the effective Young's modulus E with stimulation. Active changes in E observed in the experiment are shown in Fig. 6. If we wish to delineate and analyze the control systems that are responsible for maintaining and regulating circulation or any other physiological processes, it is essential that we quantify the changes of the controlling parameters and the controlled variables.

Vestibular functions

Everyone is familiar with the effects of spinning around for a minute and then trying unsuccessfully to walk. This is an example of exposure of the body to a prolonged and steady angular velocity for which it was not designed. In the aerospace environment, many types of disorientation occur in situations where the imposed force characteristics

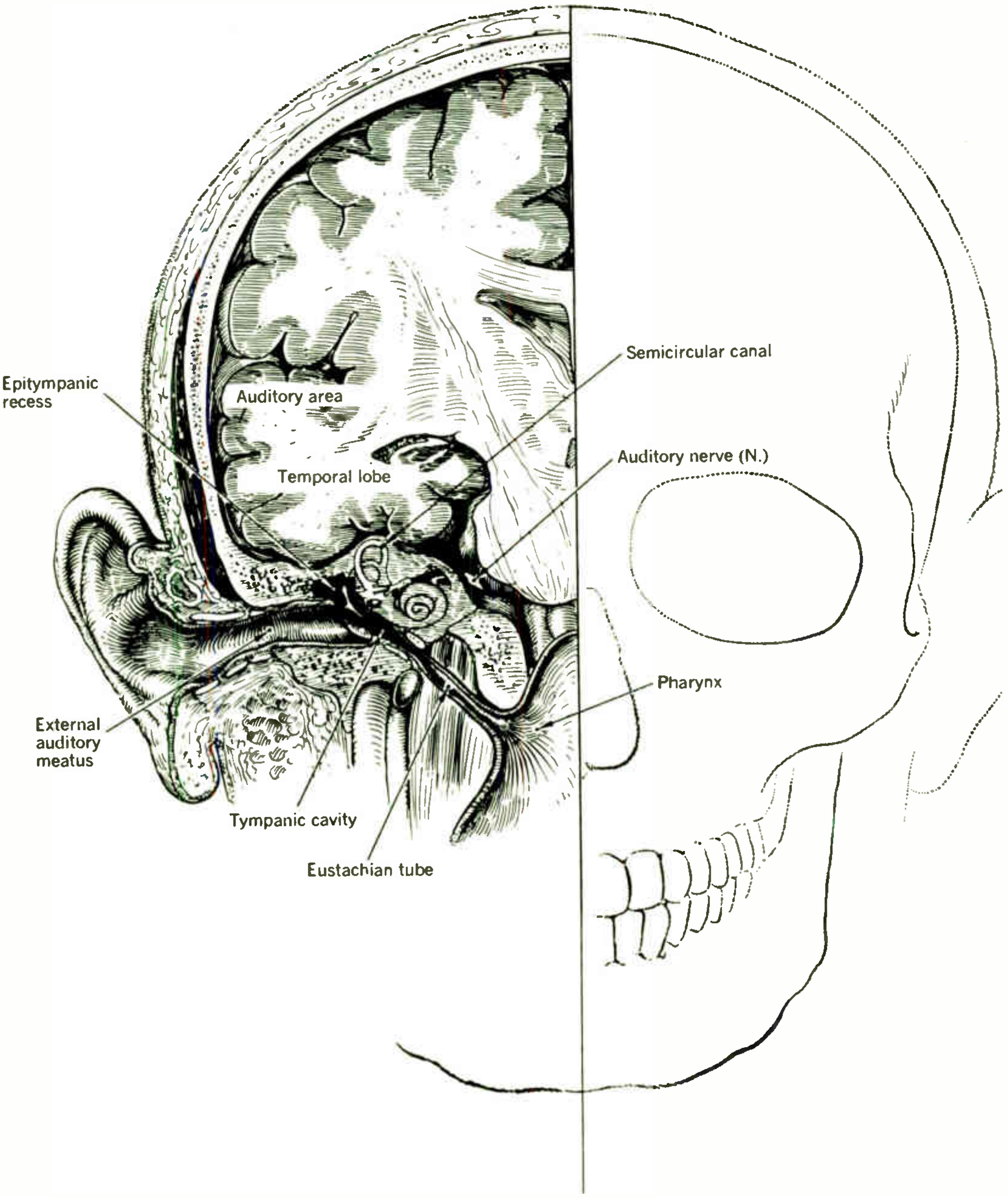


FIGURE 8. An anatomical illustration of the position of the vestibular apparatus within the temporal bone. (Courtesy Dr. B. J. Melloni, Georgetown University and Abbott Laboratories)

are also outside the response capabilities of the vestibular system. For example, if the head is rotated while the body is being spun in a different axis during an aircraft maneuver such as a roll, Coriolis forces generate sensations of movement of the body in relation to the airplane, although these movements are not occurring. The pilot tries to correct for the imagined force and may get his aircraft into difficulties. Exactly the same phenomenon can occur in a spacecraft, as during the emergency situation in Gemini 8, where rapid rotation was produced by thrusters that could not be turned off. Some illustrations of the effects of such Coriolis forces ($\overline{F}_c$) on the body as a whole are shown in Fig. 7. The abnormal body movements, dizziness, disorientation, and motion sickness that can result if the forces are large enough are clearly a hazard to the crew members affected.

Another type of disturbance is seen in weightlessness. This is the phenomenon of *space sickness* first experienced by the Russian cosmonauts, and more recently by some of the Apollo astronauts. It is possible that this is the result of a sensory conflict of the type already described for situations involving body movement, but now occurring sometimes without movement at all. One theory that has been advanced is that two major sensory systems, the otolith, and the body gravity and posture sensors, both indicate that the body is in free fall, while two other important systems, vision and touch/pressure, often indicate a normal environment. However, this remains to be confirmed.

Information on the accelerations that the head is subjected to emanates from two sensitive transducer systems —the otolith end organs and the semicircular canals. They are located within small cavities of the temporal bone in proximity to the receptor system that is responsible for hearing (Fig. 8). The otolith, shown diagrammatically in

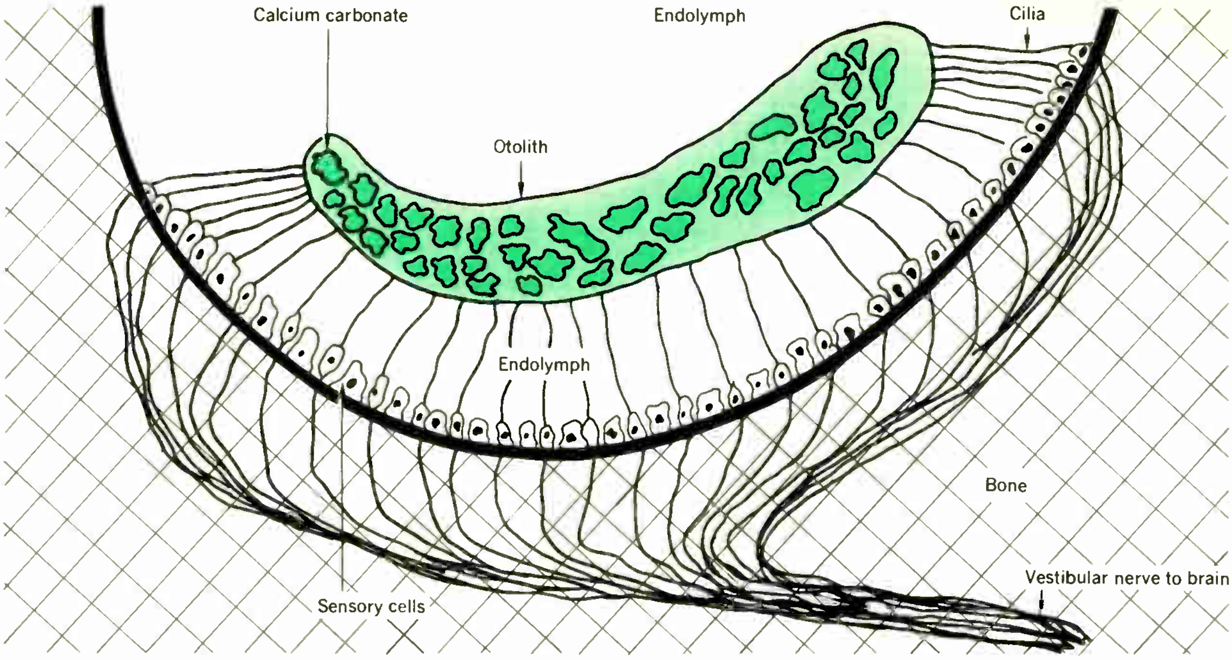
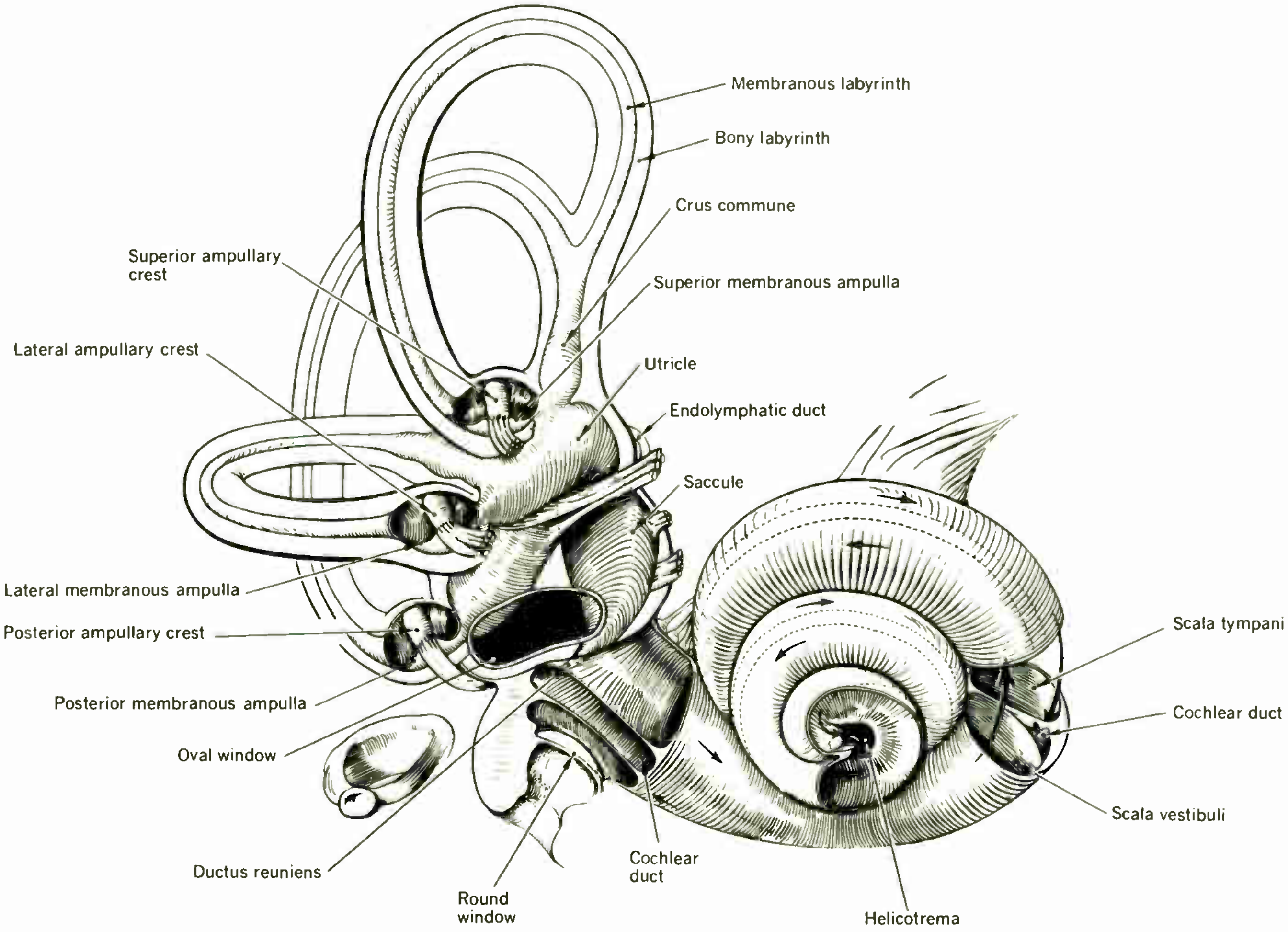


FIGURE 9. The otolith has a specific gravity about twice that of the fluid in which it is immersed. Otoliths—one in the saccule and one in the utricule (Fig. 10)—respond by moving when the direction and/or magnitude of inertial or gravitational forces is changed. The cilia are deflected and the sensory cells change their firing rates.

FIGURE 10. Medical illustration of the inner ear. (Courtesy Dr. B. J. Melloni, Georgetown University and Abbott Laboratories)



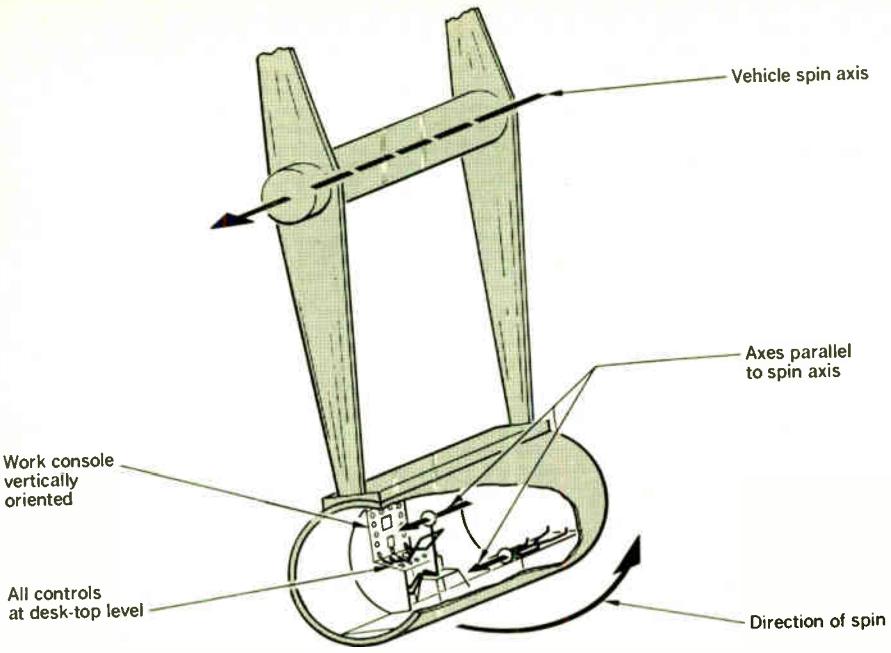


FIGURE 11. Desirable design characteristics of a rotating spacecraft; the radius of rotation should be as large as possible, which means that angular velocity should be as low as possible; head rotation should normally be about an axis parallel to that of station rotation; the long axis of the cabin should be parallel to the axis of rotation; and couches should be parallel to the axis of rotation.

Fig. 9, senses linear accelerations and body forces (induced, for example, by gravitation) in terms of both magnitude and direction. Since the otoliths have a density about twice that of the fluid they are immersed in, any change in magnitude or direction of imposed accelerations induces a displacement of the otoliths and thereby alters the discharge rate of the sensory nerve endings they are attached to. The semicircular canals (Fig. 10) respond to angular motions in the following fashion. A cupula, situated in the fluid of each of the three rings (tori) making up the canal system on each side of the head, is displaced by inertia forces acting on the fluid because of the angular acceleration of the skull. Sensory nerve endings that are attached to the cupulae again transmit information on the rate and degree of the displacement to the brain via the vestibular nerve.

Within the brain the information from otoliths and semicircular canals is processed to give subjective sensations of movement, and also to produce reflex responses of body musculature so that appropriate body reactions may be induced in response to the forces imposed on the head. For example, as the head is rotated to the left, the angular accelerations sensed by the semicircular canals are integrated in the brain and sent as velocity signals to the eye muscles, causing eye movements to the right. These movements follow closely the changing head velocity so that the eyes remain essentially stationary in relation to the environment. The image of the scene at which the subject is looking remains fixed on the retina, and clarity of vision is maintained. A similar fixation of the direction of view is achieved by the movie camera mounted on a swaying vehicle and controlled by an inertial guidance system using angular accelerometers.

The information on head movement received by the brain in this way is also combined with signals from sensory systems in the rest of the body that are responding to neck and limb displacement and movement, touch, pressure, sound, and the relationship of the body to the environment as detected by the visual system.

Under normal circumstances all the systems are finely tuned to respond to the usual sorts of forces imposed on the body. The integration of the signals produces continuously appropriate reflex responses of body musculature and the subjective sensations of body position and movement. However, if one or more of the systems is damaged by disease or trauma or is subjected to abnormal force fields, the delicate integrating system is upset by what appears to be conflicting information arriving in the brain. There are three possible consequences that may be present separately or together—inappropriate muscular activity, disorientation, and motion sickness. All are commonly seen in various disorders of the sensory nervous system. In the case of the vestibular system, infection or trauma may cause abnormally increased or reduced signals that the body interprets as forces acting on the head. The body musculature attempts to correct for these and responds with inappropriate movements. It may be impossible to stand or walk, and there is a profound dizziness. These effects can be observed in a normal subject by simple clinical tests, such as irrigating the ear with cold water (caloric stimulation). Although the subject is perfectly still, convection currents induced by the cold water in the fluid of the semicircular canals signal apparent rapid head rotation. This sensation of movement conflicts with visual and proprioceptive cues and elicits dizziness, disorientation, and, if severe, motion sickness. Similar situations may arise when the body is subjected to unusual force fields.

Figure 11 illustrates how the problem can be alleviated by an appropriate vehicle design in the case of space flight. If the principal body movements are restricted to those that include only rotations of the head about an axis parallel to the axis of vehicle rotation and translatory motion in the direction of this axis, the conflicting cues can be kept minimal.

A number of important engineering studies of the vestibular apparatus conducted at M.I.T. and elsewhere have recently been summarized in a survey article by L. R. Young.[24] The small dimensions of the semicircular canals and otolith, together with their inaccessibility, make it rather difficult to measure or record directly the variables that manifest the functions of these organs. In the absence of miniaturized instrumentation and adequate techniques for measuring the response of the end organ proper, it was necessary to resort to ingenious systems approaches that allow us to infer some of the details of the sensory mechanisms. While these systems approaches have also been supported to some extent by theoretical and experimental model studies of the semicircular canals,[25,26,27,28] we do not yet have a firm basis for a full understanding of the sensation of orientation and motion.

We have focused our attention here exclusively on the cardiovascular and vestibular systems and some of their problems in relation to the aerospace environment. Similar types of problems exist for other physiological systems and many of these are equally amenable to analysis based on methods customarily used only in engineering and in the physical sciences. For example, studies of environmental effects on respiration, bone physiology, endocrinology, and neural functions should be as rewarding for engineers and physical scientists as the areas described.

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IEEE Recommended Practice: Rules for the Use of Units of the International System of Units

Adopted by the IEEE Standards Committee on December 3, 1970

The IEEE Standards Committee has adopted ISO Recommendation R1000, thus continuing its practice of providing authoritative guidance on the application of the metric system

As a service to its members and in support of the principle of international standardization, the IEEE Standards Committee reviews international standards recommendations, particularly those of the International Organization for Standardization (ISO) and the International Electrotechnical Commission (IEC), in its area of interest. This document is identical with ISO Recommendation R1000 except that in accordance with other IEEE recommendations the spellings "kilogram," "meter," "liter," and "deka-" have been used in place of "kilogramme," "metre," "litre," and "deca-." ISO R1000 was drawn up by Technical Committee ISO/TC 12 (Quantities, Units, Symbols, Conversion Factors, and Conversion Tables) and was first published in 1969. It was reviewed and recommended for adoption by the IEEE Standards Coordinating Committee on Quantities and Units and was adopted by the IEEE Standards Committee on December 3, 1970.

Comments concerning this document may be sent to the Secretary, IEEE Standards Committee, 345 East 47 Street, New York, N.Y. 10017.

1. SCOPE

This Recommended Practice gives rules for the use of units of the International System of Units and for forming and selecting decimal multiples and submultiples of the SI units for application in the various fields of technology.

2. GENERAL

2.1 The name Système International d'Unités (International System of Units), with the abbreviation SI, was adopted by the 11th Conférence Générale des Poids et Mesures in 1960.

The coherent units are designated "SI units."

2.2 The International System of Units is based on the following six base units

meter (m)	ampere (A)
kilogram (kg)	kelvin (K)
second (s)	candela (cd)

as units for the base quantities: length, mass, time, electric current, thermodynamic temperature, and luminous intensity.

2.3 The SI units for plane angle and solid angle, the radian (rad) and the steradian (sr) respectively, are called supplementary units in the International System of Units.

2.4 The expressions for the derived SI units are stated in terms of base units; for example, the SI unit for velocity is meter per second (m/s).

In the United States, the first week of March is each year designated as "Weights and Measures Week." It is very fitting, therefore, that this issue of *Spectrum* should carry the recently adopted Recommended Practice, which appears on this page and the next. In a way this establishes a tradition, for it was exactly five years ago, in March 1966, that *Spectrum* published the IEEE Recommended Practice for Units in Published Scientific and Technical Work (IEEE No. 268), a document that has had wide influence both inside and outside the IEEE.

The five years since then have seen ever-increasing use of the metric system. The United Kingdom and Canada have announced their intention of "going metric" and the United States is studying the problem. The question seems no longer to be "whether" but rather "when" and "how."

By adopting ISO R1000 as an IEEE Recommended Practice, the Standards Committee continues its service of providing authoritative information on the metric system. Observant readers will note, for example, that the base unit for thermodynamic temperature is now called the kelvin, instead of "degree Kelvin." It is now properly given the symbol K, without the symbol °. This is a result of a decision taken by the General Conference of Weights and Measures in 1967. Authorized English-language definitions of the six base units are given in the February 1969 issue of *Spectrum*, page 8.

Bruce B. Barrow, Chairman
IEEE Standards Committee

For some of the derived SI units special names and symbols exist; those approved by the Conférence Générale des Poids et Mesures are listed below:

Quantity	Name of SI Unit	Symbol	Expressed in Terms of Basic or Derived Unit
frequency	hertz	Hz	1 Hz = 1 s^{-1}
force	newton	N	1 N = 1 kg·m/s^2
work, energy, quantity of heat	joule	J	1 J = 1 N·m
power	watt	W	1 W = 1 J/s
quantity of electricity	coulomb	C	1 C = 1 A·s
electric potential, potential difference, tension, electromotive force	volt	V	1 V = 1 W/A
electric capacitance	farad	F	1 F = 1 A·s/V
electric resistance	ohm	Ω	1 Ω = 1 V/A
flux of magnetic induction, magnetic flux	weber	Wb	1 Wb = 1 V·s
magnetic flux density, magnetic induction	tesla	T	1 T = 1 Wb/m^2
inductance	henry	H	1 H = 1 V·s/A
luminous flux	lumen	lm	1 lm = 1 d·sr
illumination	lux	lx	1 lx = 1 lm/m^2

It may sometimes be advantageous to express derived units in terms of other derived units having special names; for example, the SI unit of electric dipole moment (A·s·m) is usually expressed as C·m.

2.5 Decimal multiples and submultiples of the SI units are formed by means of the prefixes given below:

Factor by Which the Unit Is Multiplied	Prefix	Symbol
10^{12}	tera	T
10^{9}	giga	G
10^{6}	mega	M
10^{3}	kilo	k
10^{2}	hecto	h
10	deka	da
10^{-1}	deci	d
10^{-2}	centi	c
10^{-3}	milli	m
10^{-6}	micro	μ
10^{-9}	nano	n
10^{-12}	pico	p
10^{-15}	femto	f
10^{-18}	atto	a

The symbol of a prefix is considered to be combined with the symbol to which it is directly attached, forming with it a new unit symbol which can be raised to a positive or negative power and which can be combined with other unit symbols to form symbols for compound units.

Examples:

$$1\ \mathrm{cm}^3 = (10^{-2}\ \mathrm{m})^3 = 10^{-6}\ \mathrm{m}^3$$
$$1\ \mu\mathrm{s}^{-1} = (10^{-6}\ \mathrm{s})^{-1} = 10^{6}\ \mathrm{s}^{-1}$$
$$1\ \mathrm{mm}^2/\mathrm{s} = (10^{-3}\ \mathrm{m})^2/\mathrm{s} = 10^{-6}\ \mathrm{m}^2/\mathrm{s}$$

Compound prefixes should not be used; for example, write nm (nanometer) instead of mμm.

3. RULES FOR THE USE OF SI UNITS AND THEIR DECIMAL MULTIPLES AND SUBMULTIPLES

3.1 The SI units are *preferred*, but it will not be practical to limit usage to these; in addition, therefore, their decimal multiples and submultiples, formed by using the prefixes, are required.

In order to avoid errors in calculation it is essential to use coherent units. Therefore, it is strongly recommended that in calculations only SI units themselves be used, and not their decimal multiples and submultiples.

3.2 The use of prefixes representing 10 raised to a power which is a multiple of 3 is especially recommended.

Note:

In certain cases, to ensure convenience in the use of the units, this recommendation cannot be followed.

3.3 It is recommended that only one prefix be used in forming the decimal multiples or submultiples of a derived SI unit, and that this prefix be attached to a unit in the numerator.

Note:

In certain cases convenience in the use requires attachment of a prefix to both the numerator and the denominator at the same time, and sometimes only to the denominator.

4. NUMERICAL VALUES

4.1 When expressing a quantity by a numerical value and certain unit it has been found suitable in most applications to use units resulting in numerical values between 0.1 and 1000.

The units which are decimal multiples and submultiples of the SI units should therefore be chosen to provide values in this range; for example,

Observed or Calculated Values	Can Be Expressed As
12 000 N	12 kN
0.003 94 m	3.94 mm
14 010 N/m^2	14.01 kN/m^2
0.0003 s	0.3 ms

4.2 The rule according to clause 4.1 cannot, however, be consistently applied. In one and the same context the numerical values expressed in a certain unit can extend over a considerable range; this applies especially to tabulated numerical values. In such cases it is often appropriate to use the same unit, even when this means exceeding the preferred value range 0.1 to 1000.

5. LIST OF UNITS

For a number of commonly used quantities, examples of decimal multiples and submultiples of SI units, as well as of some other units which may be used, are given in the Annex to this document. (See IEEE Standard 322).

Linear induction machines

II. Applications

Not only are linear induction motors used in popular high-speed transportation systems, but in materials handling, industrial weaving, impact testing, and belt conveyers as well

Michel Poloujadoff Université de Grenoble

The history of linear induction motors extends as far back as the 19th century. Although these machines have been practically forgotten for the past 30 or 40 years, there appears to be a genuine revival of interest in them. Part I of this article dealt with the fascinating history of these "unrolled" motors and their theory of operation. The present installment, which concludes the article, deals with the unique advantages that such high-performance systems have for modern-day applications.

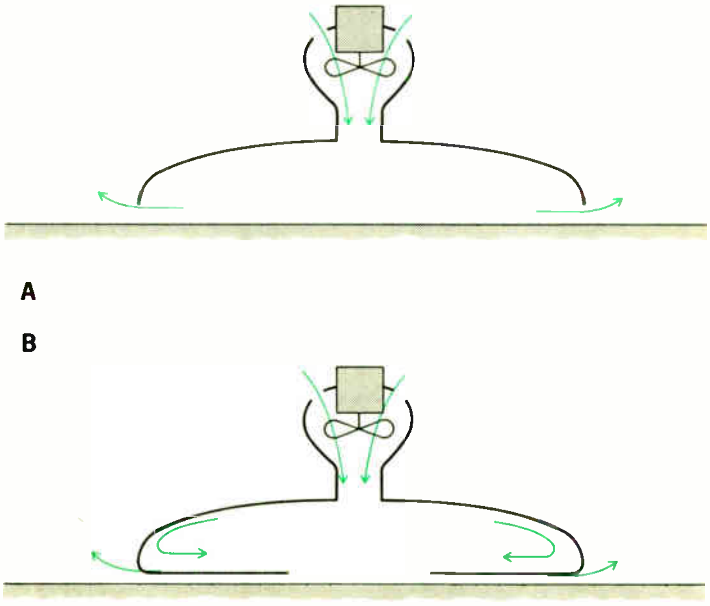
With the great increase in the traffic of people and commodities, the domestic airways of the world have reached a high degree of saturation. It would seem highly desirable, therefore, to develop some type of rapid interurban transit system to help alleviate the problem, with particular emphasis upon railways. Unfortunately, the irregularities of railway tracks produce excessive noise and inordinate wear as the speed of a wheel-supported railway car increases. Such difficulties as this do much to hinder the search for adequate solutions to the transportation problem and, as a result, there has been an accelerated research effort directed toward the most promising solution yet found—the fluid-cushion vehicle.[1-3]

The essential principle underlying such systems is explained by Fig. 18. Within a simple cusp, a blower is installed to increase the interior pressure. When enough pressure is created to lift the cusp and the blower, air linkages are formed that oppose this increase of pressure; therefore, there is an equilibrium value h that must be obtained for successful operation. If h is too small, the inside pressure increases and the cusp is lifted up; if h is too large, linkages become critical and the inside pressure decreases, resulting in the cusp falling down.

It is interesting to note that any fluid can be used in place of air and that, at the 1889 International Fair, Louis Girard presented his version of a water-cushion vehicle. There is also an important advantage to air-cushion vehicles, which is that the total force exerted by a vehicle upon the track through its cushions is divided over a large surface, thereby corresponding to a low pressure (30–50 g/cm^2), as opposed to the extreme pressure of a railway wheel's highly punctuated operation. Therefore, if it is necessary for a track to be built on a superstructure, loading values will be lighter for air-cushion vehicles than for vehicles that are wheel-supported. For similar reasons, air-cushion vehicles are lighter than conventional railway cars and may be built like airplanes (Fig. 19).

These principles were the basis for the initial undertaking of the Hoover Company and the Société de l'Aérotrain (under the direction of Jean Bertin), whose efforts are represented by the model in Fig. 20. As one can see, such cars cannot be driven by direct reaction with the ground since there are no contact surfaces. The first experiment with such apparatus was performed with an aircraft propeller; the horizontal part of the track was 2 meters wide, the vertical part 50 cm high and 10 cm thick, with both portions made of cement. With this system, a top speed of 250 km/h was reached; later, a similar car was operated at 442 km/h, propelled by a turbo engine and two rockets. However, pollution and noise problems eventually led to the use of a linear induction motor, as indicated by Fig. 21. The vertical part of the track had to be both conductive and strong enough to guide the car, while at the same time possessing an electrical resistance high enough to have a convenient thrust–speed curve. As a result, a hollow vertical rail was generally adopted. Actual experience with the Aerotrain was gained on a track 3 km long and a passenger vehicle that accommodated 40 persons.

FIGURE 18. A—Principle of the positive air-cushion vehicle. B—The use of a plenum does not affect this principle, but reduces lifting power.



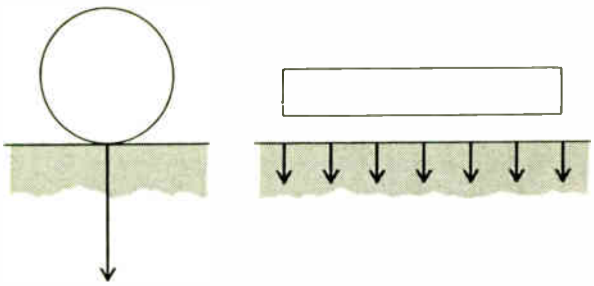


FIGURE 19. The use of wheels exerts a high pressure on a small surface, whereas an air cushion exerts a low pressure over a large surface.

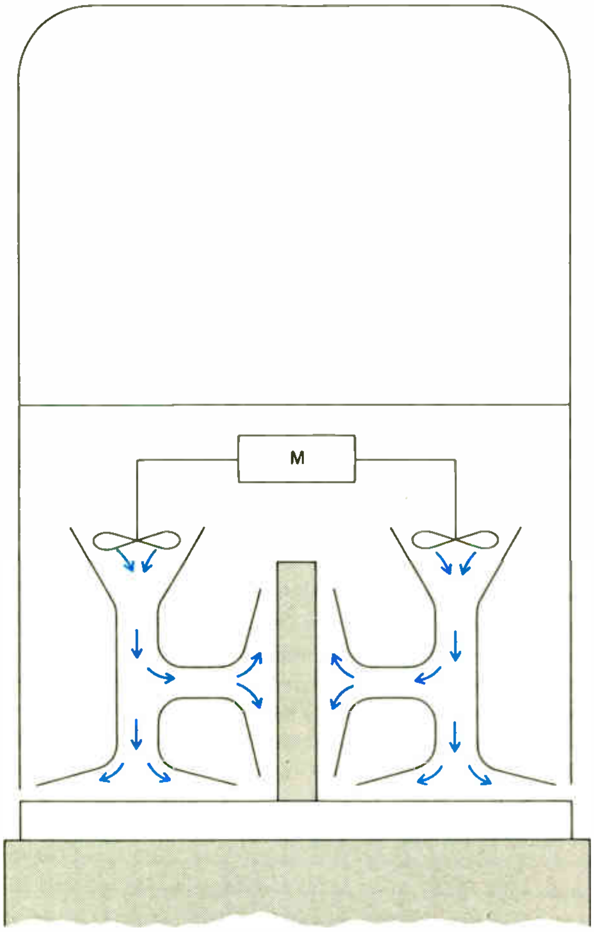


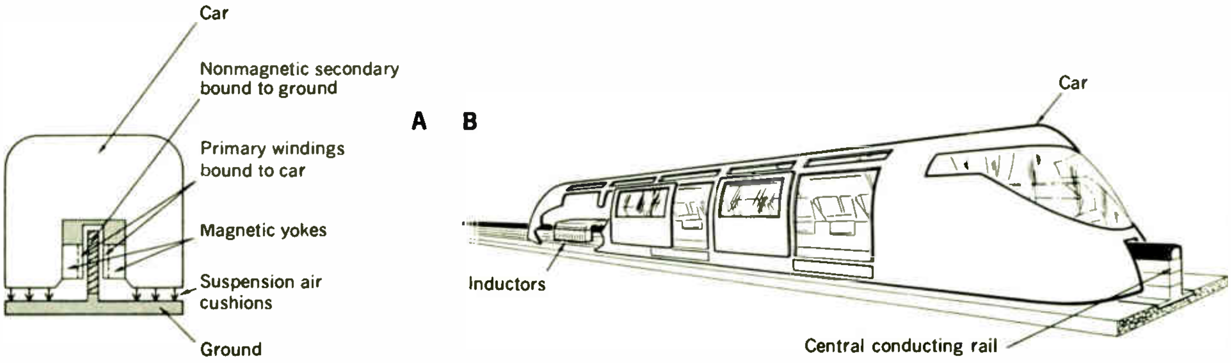
FIGURE 20. A train must be simultaneously suspended above the ground and guided along the track; hence, in this T-shaped system, one set of cushions is used for suspension, the other for guidance.

An interesting ramification of the "positive" air-cushion principle displayed in Fig. 18, not very well known about at the present time, is exhibited in Fig. 22. In this "negative" vehicle, which is based upon a reversal in air direction, the ground is replaced by a "ceiling," and the blower decreases the pressure inside of the cusp instead of increasing it. For every blower speed, there exists a value of h such that the differential pressure equilibrates exactly the weight of the cusp and the blower. If, by accident, the cusp rises, the differential pressure will increase and the cusp will adhere to the ceiling. On the other hand, if h becomes too large, equilibrium will again be lost and the cusp will fall downward. Unhappily, therefore, this means of suspension is extremely unstable.

It is amazing to ponder how the human imagination, which generally is considered brilliant, can be really poor at times. Indeed, all of the hundreds of researchers who considered the configuration of Fig. 22 seem to have cut short their reflections, only to go on to deal with other problems. If they had persisted, they might have observed that, to obtain stability, it is essential that the leakage (and h) decrease when the cusp falls, and increase when the cusp rises. Restating the problem in these terms is almost equivalent to the representation of Fig. 23, where the cusp "envelops" a finite ceiling and equilibrium becomes stable. This idea is attributed to M. Barthalon, under whose direction the experimental car "Urba" was constructed (see Fig. 24). In terms of the positive air-cushion vehicle, it is necessary to provide both suspension and guidance; Fig. 25 presents the complete details. Needless to say, the traction for this project was supplied by a linear induction motor.[4,5]

One of the important problems we must now contend with is how to supply a powerful electric motor mounted aboard such air-cushion vehicles. One possibility is to have a generating unit aboard, a solution that is now being tried by the Garrett Corporation.[6-8] With this method, voltage regulation is no problem; however, the main advantage is that it is possible to vary the frequency as a function of speed in such a way that slip frequency and thrust are constants. Unfortunately, the added weight of a generating unit is a disadvantage, with the great quantity of fuel that must be burned a source of pollution. Of course, one alternative would be to attach three or four brushes on the car, which would glide on feeding rails fixed directly to the track [Fig. 25(C)]. For low speeds, this presents no particular problem; however,

FIGURE 21. Graphic representation of the Aerotrain with its linear induction motor.



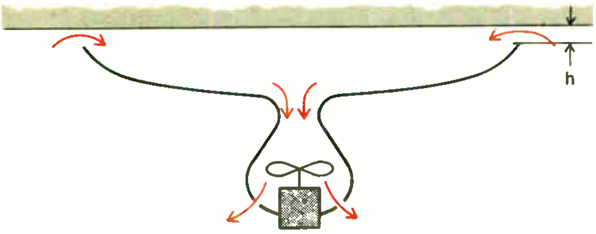
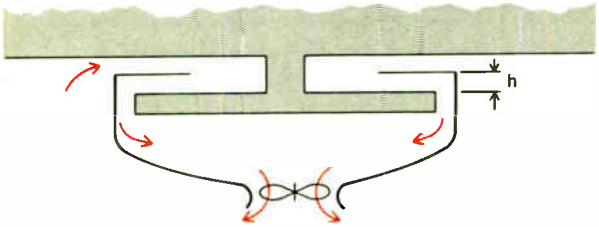


FIGURE 22. Principle of the negative air cushion.

FIGURE 23. The Dynavac or stable negative air cushion. As in Fig. 22, there exists an equilibrium value h, but the suspension is no longer unstable since the leakage decreases as the cusp falls and increases as the cusp lifts.



during high speeds, any deviations of the car from its ideal forward movement would cause very important arcing problems, as well as fast wear and deterioration of both rails and brushes. One elegant solution to this difficulty is shown in Figs. 26 and 27. Captation is made by three brushes attached to a "shuttle," which is tied to the car itself by struts and cables. The shuttle, in turn, is guided by the feeding rails themselves. With this disposition, deviations of both the car and the shuttle from their trajectories are more or less independent. More unconventional ideas can be seen in Figs. 28[2] and 29.[9]

In Fig. 28, two rails lying along the track carry an alternating current I_1 and act as the primary winding of a transformer, with the secondary providing the current I_2 to the car. The magnetic circuit must be open, hence a considerable magnetizing current is required. Furthermore, the system is heavy, which is a detriment for light air-suspended vehicles. A three-phase, instead of single-phase, scheme is also possible.

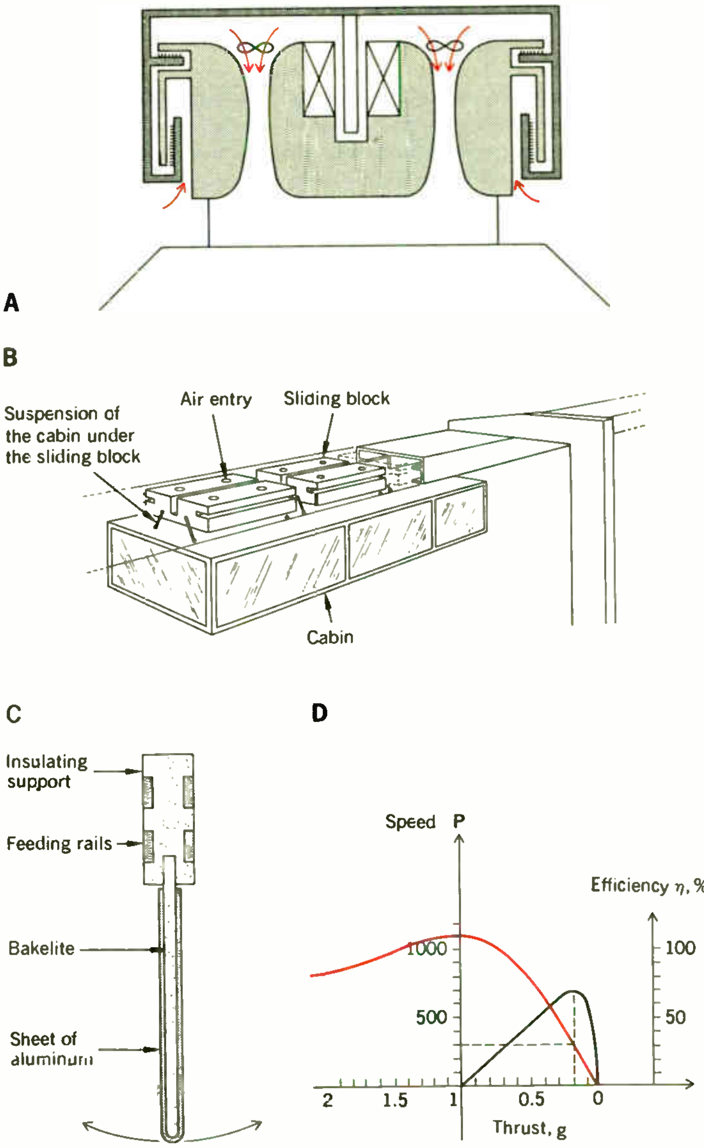
Materials handling

Not only is the linear induction motor ideal for air-cushion vehicles, but it is a unique propelling device with space-saving qualities for bulk coal handling. The idea is described in Fig. 30. Transportation is provided by small cars, with empty ones running under full ones upside down, and discharging of coal taking place during reversion of the car. The great advantage of this system is that the displacement of the cars uses less room than if they were put side by side. Naturally, a conveyer belt would occupy still less room, but it must follow an almost straight path; also, the belt is more sensitive to wear than a train, and is slower, being limited to 10 or 12 km/h. Additional advantage may be ensured if the upper parts of the small cars are suppressed and replaced by a long flexible material constituting a kind of moving gutter.

Such a system cannot work without a convenient propelling device, however. Self-propulsion may be elimi-

FIGURE 24. First prototype of the Urba (six seats). (Courtesy Compagnie d'Energetique Linéaire)

FIGURE 25. A—Cutaway view of the Urba Dynavac, showing vertical and lateral sustentation. B—A perspective view of the system. C—The secondary used on the Urba. D—Thrust-speed curve of the Urba motor (colored curve), which shows good braking characteristics as a result of reversing two phases (black curve is efficiency).



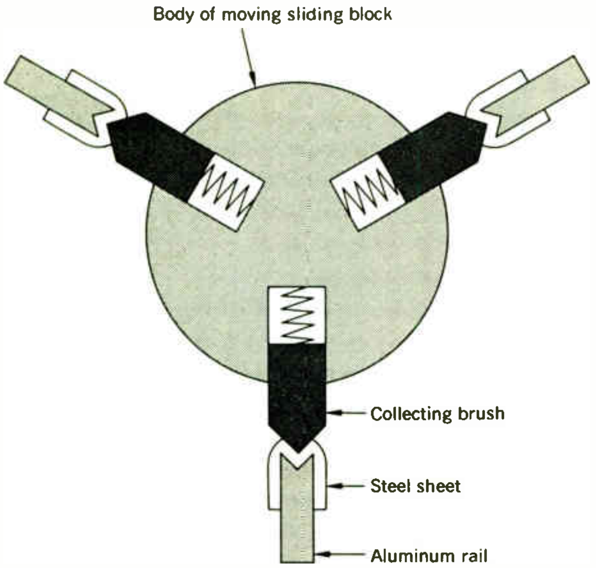


FIGURE 26. One method of supplying electric power. Three aluminum rails are fixed along the track, with wear reduced by a steel sheet. The sliding body, tied to the vehicle, is made mostly of insulating material.

FIGURE 27. View of collector shuttle, showing struts and electrical connections to the cars and insulators that support the rails.

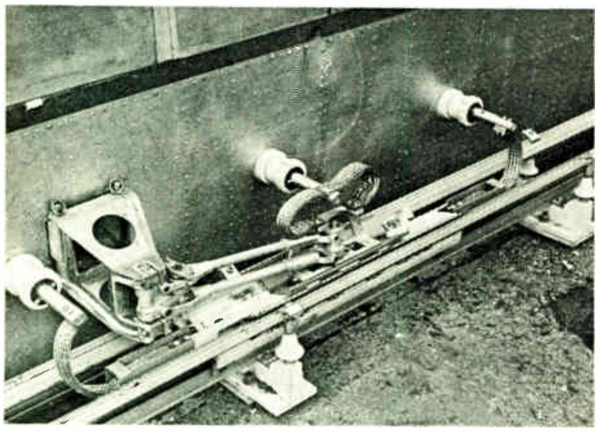
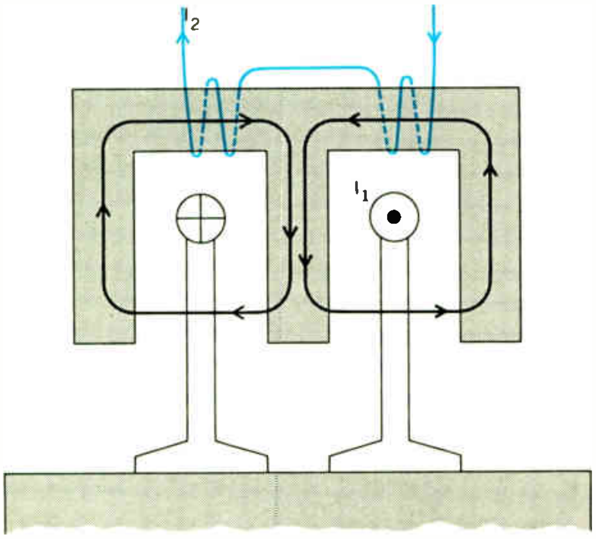


FIGURE 28. This scheme shows how power can be collected magnetically by the vehicle.



nated for the following reasons: lack of room for the motors, difficulty in controlling speed from the track, and danger of arcing in the hazardous mines when using electric propulsion. Propulsion by mechanical impulsions provided by motors disposed along the track has been used in some cases; Fig. 31 illustrates the disposal system used by SECCAM[10] that consisted of tires and electric motors. But the best solution, by far, entails the use of the method outlined by Fig. 32, which follows the scheme described in Fig. 10. The bottoms of the cars carry both a magnetic circuit and a squirrel cage, which interacts with inductors laid down on the track. An industrial model, 50 meters long (Fig. 33), has already been made to check the performance of this system, which can transport thousands of metric tons (tonnes) of material per hour over dozens of kilometers.

At present, such equipment is being installed in a coal mine at Gardanne in southern France. The track has been placed 500 meters underground, with a total length of 2325 meters. The radius of the overturning portion of the track is 2 meters, and four 225-meter long lifts of 150 cars will be driven by 180 motors at a speed of 7 m/s (25 km/h), transporting 460 tonnes per hour. Accurate regulation will be made through thyristor voltage regulators.

Another application of linear motors to materials handling is in the realization of "automatic stores." Before explaining the basis underlining such stores, let us first examine the positioning and switching principles of a specific set of linear motors. In Fig. 34, several primary windings are disposed along a common line. They

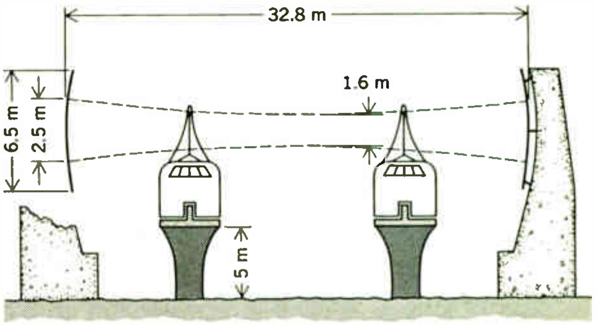
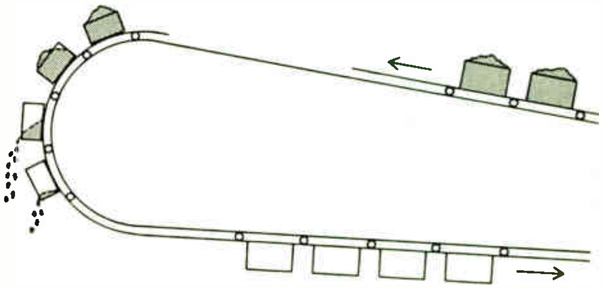


FIGURE 29. In this microwave scheme, a set of cylindrical antennas 30 km long is fed by a 915-MHz, 3.8-MW generator. Energy is received by a semiconducting antenna (Rectenna) (which weights 1.5 tonnes and is 2 meters high by 30 meters long) mounted above the cars, delivering dc power to an inverter, which in turn provides variable-frequency power to the linear motor. Traction power is estimated at 1.6 MW. (From Basisio and Foggia)

FIGURE 30. Example of bulk coal handling by linear motor; loading, not shown here, is done through a funnel.



can be energized to produce a field in either direction along this line, and are grouped by pairs, every pair defining a certain "position." A rolling vehicle, supporting a secondary, will be propelled if it is situated over a position where a pair of primaries cooperate; it will stop, however, wherever two primaries oppose each other.

An extension of this principle takes the form of the direction switching of Fig. 35, made by the crossing of two linear positioning systems. A rolling vehicle *A*, supporting a secondary, can be directed in the *Y* direction if the primaries are correctly energized, and can be stopped at the crossing, then directed left or right, according to how the *X*-directed primaries are energized.

Now we can examine the automatic-store problem. From Fig. 36, we see a set of positions that are regularly situated in a plane and defined by four primary windings. A square secondary, which may roll on rails in either the *X* or *Y* direction, will be maintained in the position that it occupies if the four corresponding primaries are energized so that they produce fields directed toward the center of the position. Reversing any one of the fields results in the secondary being pushed toward one of the neighboring positions. If there were one less as many secondaries as there are positions, it would be possible to move a given secondary from its given position to any other position. For example, it is possible to park 99 automobiles on a surface equal to 100 times the surface occupied by one of them, solely at the cost of the equipment just described, plus a small computer to memorize the location of every customer's car and determine the most efficient way to remove a given car. The same procedure may be applied to a wide variety of storage problems.

It should be noted that a positioning device using linear induction motors based on a different design from that of Fig. 34 has been developed and is represented in Fig. 37.[11] In this case, two primary windings are bound to a mobile vehicle and act on the secondaries that are "S"-spaced on the two sides of the track. If the two motors are set to move the carriage in the same direction, the vehicle goes steadily in this direction; if the two motors move the carriage in opposite directions, it is easily seen that the boundary between the two primary windings will be carried to the nearer of the intermediary positions indicated, which will produce the desired stability.

There is another application of the linear motor to materials handling that is called the "Linerail" (see Fig. 38). In this system, a semiclosed hollow girder is suspended by a superstructure, and within the girder a movable automotive truck consisting mainly of a primary winding acts on a secondary fixed to the upper part of the girder. The truck comprises a hook passing through the aperture of the girder from which is suspended the load, and attains a speed of 3 m/s with a load of 100 kg, corresponding to a thrust of 20 newtons. Such a device can be compared to the chain conveyer, which it resembles, but is faster, less noisy, and allows a variety of speeds along a given track, the trucks speeding up during a return on the main path or slowing down when arriving at a

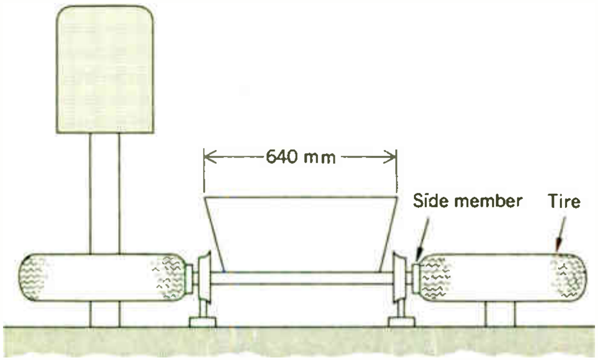


FIGURE 31. The SECCAM system for propelling railway trains without an on-board motor. The side members of the carriage are pushed by pairs of tires driven by electric motors. This system has some of the advantages claimed by the tangential traction of Fig. 10 (it is, in fact, mechanical tangential traction).

FIGURE 32. Cutaway view of the rolling gutter system showing arrangement of the linear motors.

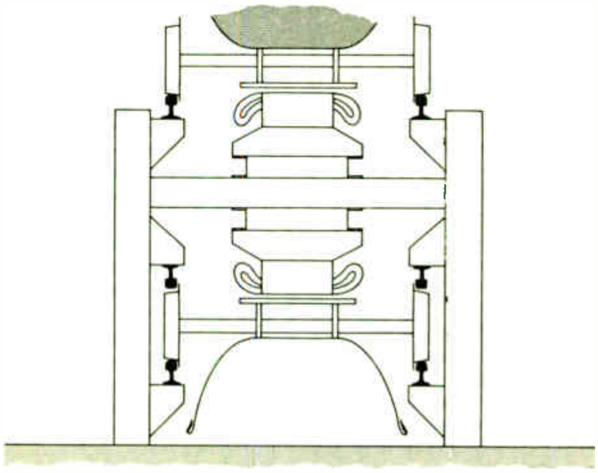
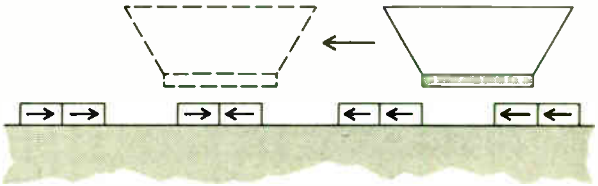


FIGURE 33. Full-scale model of the Gardanne equipment.

FIGURE 34. Method of positioning with linear motors (stationary primaries).



switch. The main application of the Linerail will be the handling of individual loads between a store and a workshop. Temporary storage can be obtained by loaded trucks along a side track, or even on the main track.

Other applications

The application of linear induction motor railways has already been discussed in connection with adhesion problems. However, the thought might arise that conventional rails could be used as the secondary member for a one-sided primary fixed to a locomotive, without any additional cost (see Fig. 39). This idea has been tried,[12] and it has been experimentally possible, with a 75-mm-wide rail (U80 gauge), to obtain a thrust of 110 kgf (1080 newtons) per meter of inductor at constant nominal speed (88 percent of maximum speed). Starting thrust was 190

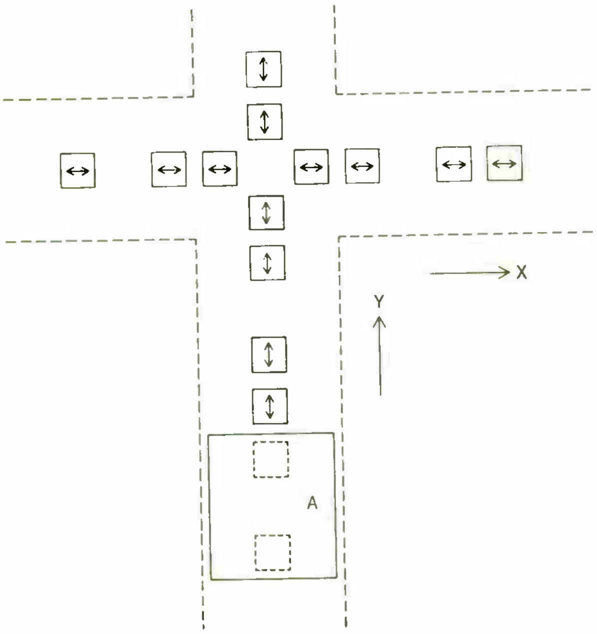
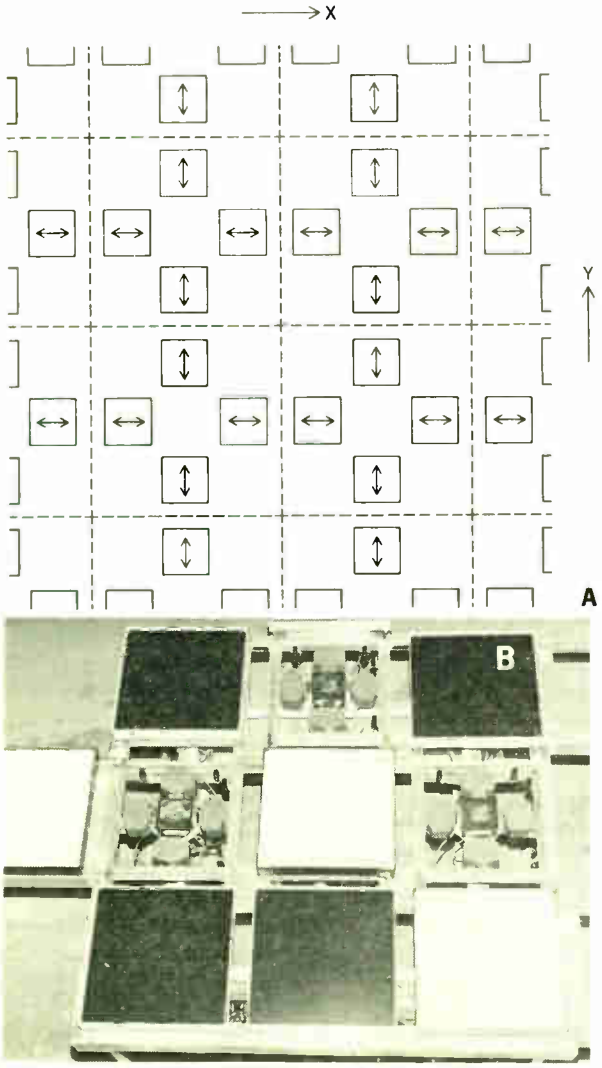


FIGURE 35. Method of positioning and switching with linear motors (stationary primaries).

FIGURE 36. A—The automatic-store method in a two-dimensional system (rectangular positions may be realized in the same manner). B—A six-place model using black and white pallets and stationary primaries.



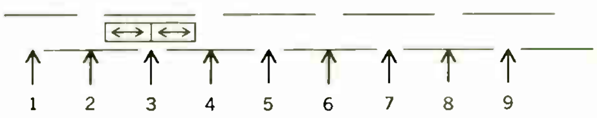
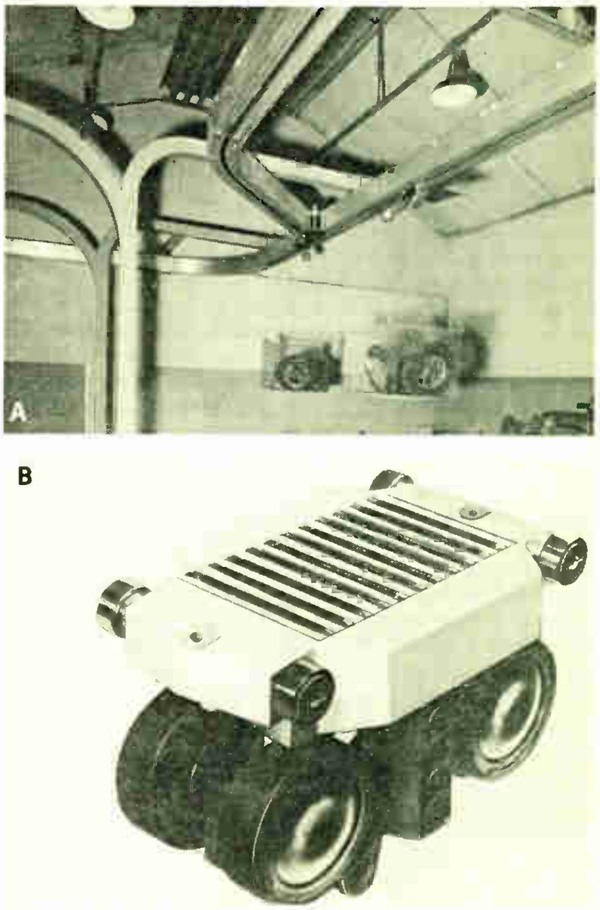


FIGURE 37. A positioning method using moving primaries bound to the carriages and fed by stationary rails.

FIGURE 38. A—General view of the Linerail. B—Close-up of truck that runs within the girder. The top portion contains the primary of the motor, with the upper part of the girder as secondary and the air gap kept constant by four small rollers. Collecting brushes are seen to the right, with four main wheels to support the load on the bottom (it should be noted that the motor is vertically independent of the remainder of the truck).



kgf/m (1860 N/m) for 90 seconds' duration; braking effort, obtained by field reversal, was 250 kgf/m (2450 N/m) for 60 seconds. Still higher braking thrusts could have been achieved by dc feeding of the windings. The actuator, which may be based on the design indicated by Fig. 5, is very well known as it has been widely discussed in the technical literature. Its principal advantages are

1. Flexibility in design, owing to the variety of shapes that can be used.
2. Easy alteration of stroke length by just changing the secondary length.
3. Ease in obtaining extra force by adding extra stator units as required.

Unfortunately, these actuators suffer both from a force/weight ratio that is inferior to those of pneumatic and hydraulic systems, and from cooling difficulties when operating at standstill.

An example of the use of a linear motor as an actuator is given by Fig. 40; here, the motor is of the type indicated by Fig. 1(D) (double-sided short fixed primary). Another application as an actuator has been made in connection with 10- and 35-kV oil circuit breakers. Field trials covering four years suggest that the drive could be applied on power stations having an ac operating supply.[13]

The application of linear induction motors to overhead cranes could just as well have been included in the section on materials handling. The Herbert Morris Society has been successful for several years with such equipment.

In any application of linear induction motors, insuring a high-frequency, long-stroke reciprocating movement has always been a difficult problem. For example, consider the problem of guiding filamentary material such as textile yarn to wrap packages. The solution involves traverse mechanisms that, if mechanically actuated, are quite satisfactory at low speeds. Since both wear and power requirements increase with speed and frequency, the system must be pneumatically actuated, which makes it difficult to control the delivery of power to the traverse element throughout the entire length of its stroke. Problems related to loom shuttle propelling are made still more complicated because of the presence of the warp. Such difficulties are well described in Refs. 14–16.

The ideas associated with such systems are basically as follows. A long inductor is excited to produce a traveling field periodically reversed by the inversion of two phases. Springs are used to store momentarily some kinetic energy; however, if the motor is not synchronous, a switch has to be used to control field reversal, and this is a source of noise and wear. A good idea has been offered by Laithwaite in connection with his general study of oscillating motors: Both parts of the stroke length are energized to produce fields traveling toward the center; therefore, if the shuttle is at the center of the stroke, it is in equilibrium, being equally attracted to both ends. If the shuttle is moving toward one of the ends, a relaxation motion takes place, the moving part being braked as it enters one of the end regions, then stopped and accelerated toward the other end, where the same phenomenon takes place. Ten years ago, using this same principle, Laithwaite and Nix succeeded in operating a secondary of 2 ¾ oz (78 g) at 540 traverses per minute over a length of 35 cm with a power input of 120 watts. The motor was of the type described by Fig. 7, with a short secondary (the secondary yoke moving and short as well).[16-19]

One unexpected application of linear motors has been in the impact testing of vehicle structures. It is now becoming customary to test the safety features of motor vehicles and their components by crashing them into a concrete block. In one such test setup (at the Motor Industries Research Association), a linear motor is mounted on a trolley that moves in a trough beneath the test rig and passes under the concrete block. Made of a moving primary and an aluminum reaction plate, it can accelerate a maximum of 3500 kg to a speed of 48 km/h without exceeding an acceleration of 1.5 *g*. The design includes a special provision for accurate control of the impact velocity to between 48 and 50 km/h.[20,21]

FIGURE 39. Linear motor mounted on a conventional railway, using the track rail as a secondary.

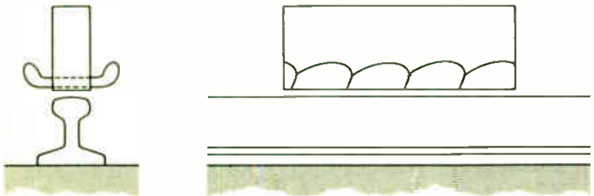
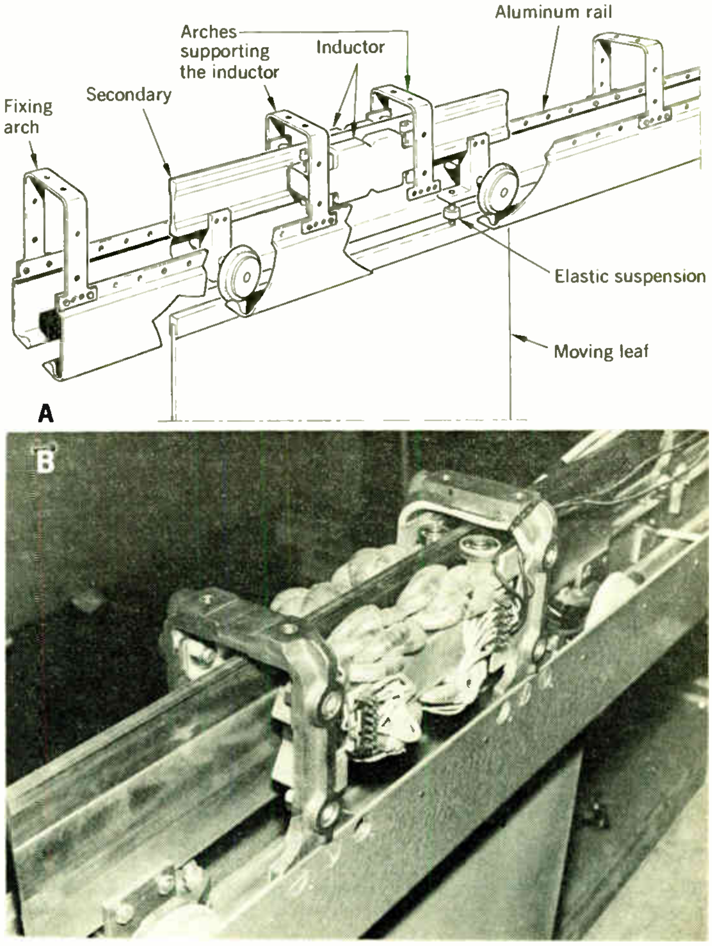


FIGURE 40. A—Example of the linear motor as an actuator for sliding doors. B—An industry model.



It is also possible to conceive moving a belt conveyor by linear motors, with the sole condition that the belt be conducting. This requirement can be met by coating a woven copper belt with rubber. Unhappily, it is difficult to insure good electrical contact between warp and thread. Up to now, all trials have been failures because of this critical difficulty. (See Refs. 22–34 for more information concerning the applications and future of linear induction machines.)

Conclusion

After having been practically forgotten for 30 or 40 years, the linear induction motor is being reevaluated. Since it was very likely originally given up primarily for economic and practical reasons, we may ask if these reasons have really disappeared. It can be said that "fashion" prevails in engineering just as in any other field. If this is true, then the previous fashion was to say "do well at a good price," and emphasize efficiency when, in reality, the trend was to "do better at any price" and emphasize performance. It must be realized that economic factors have drastically changed since the beginning of the century. In that time, energy and power equipment have become cheaper, manpower more expensive, and the need for automatic control more in demand. Therefore, a device that is easy to control and maintain can now compete with devices that give better performance but are not so easily integrated. Whatever philosophy one adopts, there is no doubt that the number of industrial companies expressing interest, and the attendance at meetings dealing with linear motors held in Leipzig (March 4–5, 1970) and Grenoble (April 17, 1970), have shown that many engineers earnestly believe in the future of linear induction motors.

Furthermore, quite a few applications of common rotating machines are now directed at obtaining linear movement (including dc linear machines), a sure sign that the linear motor is an idea whose time has come.

Appendix

Magnetic dissymmetry may be evaluated from a factor k, which is the ratio of the permeance of the lateral fringes to the total permeance of the air gap. If we consider a current i flowing through a given coil and neglect the fringes, magnetic induction is $b = Ki\cos(\nu x - \psi)$, where K takes into account the number of turns, the winding factors, and the length of the air gap, and ψ indicates the position of the coil. If we take the fringes into account, we find

$$b = Ki\left[\cos(\nu x - \psi) - \frac{k}{k+1}\cos\psi\right]$$

from which it is possible to derive the self-inductance of several coils and their mutual inductances. If we consider a set of three coils corresponding to the three values $2\pi/3$, $4\pi/3$ and 2π, the inductance matrix will be

$$\frac{L_0}{k+1}\begin{vmatrix} 3k+2 & -1 & -(3k+1) \\ -1 & 3k+2 & -(3k+1) \\ -(3k+1) & -(3k+1) & 6k+2 \end{vmatrix}$$

E. Remy, the first French engineer to be actively interested in linear induction motors, Y. Pelenc, P. Reyx, J. C. Sabonnadière, and M. Victorri have contributed much to research and development in the field. To all of them, I express my gratitude for their sincere dedication to this common effort.

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Inside IEEE

A report from IEEE's President

Employment and education were only two of the many topics that received in-depth consideration at the February meeting of the IEEE Executive Committee. At that meeting the committee reviewed an integrated IEEE program specifically designed to serve our members in these economically troubled times and took appropriate steps to start action on two elements in the program. Many members in the United States, and some outside, have urged that the Institute take constructive action for those affected by deteriorating economic conditions. The program at hand is believed to be one step in that direction. It is aimed at assisting a member to make a realistic assessment of his future career possibilities as an electrical engineer and at providing guidance in taking appropriate action.

The return postcard appearing on page 149 in this issue is intended to give us an accurate count of our members who are either unemployed or underemployed. If you are in either category please complete and return the card as soon as possible. The data collected will be used by the Institute for future action programs and a summary will be forwarded to appropriate groups in the U.S. Government and elsewhere for use in the development of programs aimed at reducing unemployment of scientists and engineers.

While this information is being collected from the membership and analyzed, a task force under the chairmanship of Executive Committee member W. O. Fleckenstein will be directing its attention to obtaining a quantitative picture of the economic conditions in the electronics, electrical, and related industries. It will attempt to determine where the profession stands economically, what the principal forces of change are, and what the trends are likely to be in the 5–10-year period ahead. Finally, from studies presently being conducted by groups in the U.S. Government, we expect to have data by about mid-year regarding anticipated demand for engineers in various specialties and geographical locations. The totality of the information collected should allow an individual to make decisions concerning his future career on a considerably better factual basis than now appears to be possible.

We believe, however, that more than factual data can be provided to assist one in mid-career (age 35 to 50) in making decisions about one's future. Unemployment at the peak of one's career can raise major questions of individual career commitment, skill transfer, physical mobility, and related factors. This condition, combined with the highly specialized skill and level of income and standard of living that this group has, has caused increasing concern over the lack of definitive knowledge covering mid-career counseling. Unfortunately, normal job vocational guidance and placement techniques do not offer the models for new approaches that are perceived to be required. Because of this situation the Executive Committee has arranged for a planning and review conference, to be held early in March, concerned with the explicit and implicit problems in mid-career guidance. Those present will be specialists from various professional fields associated with this problem area. It is intended that their recommendations will be translated as appropriate into programs under the purview of the Educational Activities Board aimed at providing direct benefits to the membership of the Institute. The results of the conference will also be made available to interested agencies of the U.S. Government and others concerned with manpower development and utilization.

The activities just described, however, do not respond to the immediate need of many of our members who now face unemployment. To this end, we intend to continue the employment counseling workshops conducted in cooperation with the AIAA and the Department of Labor. These are aimed directly at improving an individual's effectiveness in identifying and obtaining a suitable career position. Series of these workshops have been conducted with considerable success in Baltimore, Boston, Long Island, Los Angeles, San Francisco, and Washington. Literature used in the workshops is available without charge to any member by contacting J. M. Kinn, Director of Educational Services, at IEEE Headquarters. Furthermore, I shall continue to present to various groups of the U.S. Government concerned with reduction of unemployment of scientists and engineers specific suggestions for action that will be based on the communications received from the Institute membership.

Education

A major function of the Institute is the continuous education of the membership—a process using a variety of methods in diverse subject areas. A major effort to meet the needs of our members in this area is a cooperative program among the Regional Activities Board (RAB), the Educational Activities Board (EAB), and the Technical Activities Board (TAB). These operating units of the IEEE have embarked upon unprecedented and ambitious programs of interrelated activities aimed at bringing each IEEE member the maximum possible benefits from his IEEE membership.

A partnership is emerging between TAB and RAB whereby TAB will produce technical programs and material—with the assistance of EAB—and this material will be made available to the individual member through the communications network of RAB. Section officers, who at times may have felt that they were out on a long limb in providing suitable programs, will be furnished with many aids to clarify and simplify their jobs. They will be provided by TAB with speaker rosters, lists of educational materials available, and assistance in running Section educational courses. There is now an increased emphasis on encouragement of inter-Section activities, including cooperation in exchanging speakers and organizing speaker tours in various local geographical areas. All of these programs have a single objective in mind: to serve the individual IEEE member better.

The close liaison between EAB and TAB is resulting in the various Groups setting up committees on education or naming education officers. Eleven Groups have already done so. According to EAB, such committees or officers might do the following: develop tutorial material on particular aspects of our technology; produce bibliographies in specific technical areas; make suggestions for IEEE Press offerings in the applications area; provide inventories of educational materials that the Group has produced in the past; sponsor educational tours abroad; contribute to IEEE's Cassette Colloquia and Dial Access Technical Educational Service;

510 Sixth Ave., New York, N.Y. 10011, 1970; 399 pages, illus., ppbk., $4.95. In its description of valves and aerials, it is apparent that original publication, in 1966, was for the United Kingdom. Certain information, such as television scanning, is not directly usable outside the United Kingdom, but basic concepts are universal.

Modern Electronic Maintenance Principles, D. W. Garland and F. W. Stainer—*Pergamon Press Inc., Elmsford, N.Y. 10523, 1970; 86 pages, illus., ppbk., $2.70.* General rules and principles are expounded without much practical detail possible in a book devoted mainly to logical methods.

RCA Linear Integrated Circuits, Staff of Solid State Division—*RCA Commercial Engineering, 415 S. Fifth St., Harrison, N.J. 07029, 1970; 416 pages, illus., $2.50.* This revised manual can be used as a guide for circuit and systems designers as well as educators and technicians who need both the information and the technical data presented in the application guide.

Writing for Science and Technology, S. Mandel—*Dell Publishing Co., Inc., 750 Third Ave., New York, N.Y. 10017, 1970; 353 pages, illus., ppbk., $3.95.* Subtitled as a practical guide, this volume presents a large number of practical illustrations of the craft.

Computer Science: Basic Language Programming, A. I. Forsythe, T. A. Keenan, E. I. Organick, and W. Stenberg—*John Wiley & Sons, Inc., 605 Third Ave., New York, N.Y. 10016, 1970; 124 pages, illus., ppbk., $3.95.* One in a series of texts developed for introductory course in college (or high school) with an underlying theme of separate identification and teaching of algorithmic concepts.

Amateur Radio Extra-Class License Study Guide, W. Green et al.—*TAB Books, Blue Ridge Summit, Pa. 17214, 1970; 217 pages, illus., ppbk., $4.95, hardbound, $7.95.* Although directed toward amateur radio operators, the book contains material that could be helpful in preparing for the Federal Communications Commission first-class radiotelephone operator license examination.

TV Trouble Diagnosis Made Easy, A. Margolis—*TAB Books, Blue Ridge Summit, Pa. 17214, 1970; 255 pages, illus., ppbk., $4.95, hardbound, $7.95.* Practical advice that could help make it possible for an engineer to repair his own set is the message of this volume.

Introduction to Telephony and Telegraphy, E. H. Jolley—*Hart Publishing Co., Inc., 510 Sixth Ave., New York, N.Y. 10009; 1970; 413 pages, illus., $32.50.* Describing electromechanical switching practices used in the United Kingdom, this book is geared to study for the guild examinations.

New Math Puzzle Book, L. H. Longley-Cook—*Van Nostrand Reinhold Co., 450 W. 33 St., New York, N.Y. 10001, 1970; 176 pages, illus., $4.95.* The puzzles in this book are all good. Since they were selected to be appropriate for the new math, an engineer's children will be able to help him when he gets stuck.

Introduction to Microelectronics, D. Roddy—*The Pergamon Press, Inc., Maxwell House, Fairview Park, Elmsford, N.Y. 10523, 1970; 151 pages, illus., ppbk., $1.35.* A discussion of integrated circuits must begin, as does this book, with some background on semiconductors and individual devices.

Collected Basic Circuits, D. I. P. Stretton and A. W. Hartley—*Howard W. Sams & Co., Inc., 4300 W. 62 St., Indianapolis, Ind. 46268, 1970; 176 pages, illus., ppbk., $4.95.* This collection of tube and semiconductor circuits is accompanied by explanations of the circuit functions, but does not include the values of the circuit components used.

Direct Transistor Substitution Handbook, H. A. Middleton—*Hayden Book Co., Inc., 116 W. 14 St., New York, N.Y. 10011, 1970; 224 pages, illus., ppbk., $2.95.* Containing information on some 130 000 transistor substitutions for nearly 12 000 different transistors, this small computer-derived book is useful to engineers building equipment.

Circuit Consultant's Casebook, T. K. Hemingway—*TAB Books, Blue Ridge Summit, Pa. 17214, 1970; 210 pages, illus., $9.95.* The aim of this book is to bridge the gap between an engineering degree and the harsh realities of designing equipment that works, including the textbook, nonstarting multivibrator.

Tensor Properties of Materials, A. R. Billings—*John Wiley & Sons, Inc., 605 Third Ave., New York, N.Y. 10016, 1970; 171 pages, illus., $9.75.* This study of generalized compliances and conductivities aims to assist the extrapolation of experience from one area into an area of ignorance.

Understanding Solid-State Circuits, N. H. Crowhurst—*TAB Books, Blue Ridge Summit, Pa. 17214, 1970; 189 pages, illus., ppbk., $4.95, hardbound, $7.95.* The author provides practical assistance in understanding solid-state circuits, giving circuit diagrams with component values.

Electronic Designer's Handbook, T. K. Hemingway—*TAB Books, Blue Ridge Summit, Pa. 17214, 1970; 296 pages, illus., $9.95.* A revision of a book on practical transistor-circuit design originally published in 1966, this edition brings the subject up to date.

The Computer Impact, Irene Taviss, ed.—*Prentice-Hall, Inc., Englewood Cliffs, N.J. 07632, 1970; 297 pages, ppbk., $4.50, hardbound, $7.95.* This book is a selection of previously published papers dealing with the relationship between society and technology.

Fourth Generation Computers: User Requirements and Transition, F. Gruenberger, ed.—*Prentice-Hall, Inc., Englewood Cliffs, N.J. 07632, 1970; 177 pages, illus., $8.95.* The proceedings of a symposium held at the University of California, Los Angeles, in 1969, which is addressed to industry, education, and government representatives, is now available between hard covers.

Selected Papers on Frequency Modulation, J. Klapper, ed.—*Dover Publications, Inc., 180 Varick St., New York, N.Y. 10014, 1970; 417 pages, illus., $7.50.* Starting with the original Armstrong paper, first published in *Proceedings of the IRE*, the editor has collected many of the important papers that provide an overview of the subject.

Basic Engineering Sciences and Structural Engineering for Engineer-in-Training Examinations, H. J. Apfelbaum and W. O. Ottesen—*Hayden Book Co., Inc., 116 W. 14 St., New York, N.Y. 10011, 1970; 408 pages, illus., $13.95.* Designed for both study and reference, the problems chosen to illustrate engineering points in this book serve to call up review of broader principles.

An Illustrated Guide to Linear Programming, S. I. Gass—*McGraw-Hill Book Co., Inc., 330 W. 42 St., New York, N.Y. 10036, 1970; 173 pages, illus., $9.95.* An unusual sugar-coated approach to the subject, this book utilizes a number of cartoons to make some of its points.

Intermetallic Semiconducting Films, H. H. Wieder—*Pergamon Press, Inc., Maxwell House, Fairview Park, Elmsford, N.Y. 10523, 1970; 361 pages, illus., $16.00.* Besides covering the properties of films, this treatment discusses devices, applications, and measurement techniques.

Advances in Electronics and Electron Physics, L. Marton, ed.—*Academic Press, Inc., 111 Fifth Ave., New York, N.Y. 10003, 1970; 358 pages, illus., $18.50.* Three of the five contributions to this review are gaseous discharge oriented. Magnetic resonance phenomenons and the use of RF quadrupole fields in mass spectroscopy round out the coverage.

Handbook of Electromagnetic Propagation in Conducting Media, M. B. Kraichman—*U.S. Naval Ordnance Laboratory (order from Superintendent of Documents, Washington, D.C. 20402), 1970; 116 pages, illus., $3.00.* Written as a single source aid for students and researchers, this book summarizes information published on the subject over the last couple of decades.

sponsor national lectureships; and help educate IEEE on matters of special technical interest such as the power needs of the United States and siting of power plants.

Role of the Section Chairman

It is vital to your interests to get to know your local Section Chairman, for it is on the local level that you have an excellent opportunity to express your desires, interests, and complaints in person. Once you know him, get him to know your needs. Your Section Chairman is one of your important spokesmen in IEEE affairs, but he cannot represent your interests if he doesn't know what they are.

If you don't know who your Section Chairman is, you can find out by looking at the listing at the end of this column.

Regional Committee Meetings

One good example of how you can benefit by working through your Section Chairman is tied in with this month's IEEE International Convention. At the Convention a series of Regional Committee meetings will be held. Each Regional Committtee includes all the Section Chairmen in that particular Region. Many important matters will be discussed at the meetings. One topic expected to receive serious consideration is how IEEE should budget its finances for 1972. Here is an opportunity for you to get your thoughts known on how IEEE should allocate its funds among its various activities. Speak to your Section Chairman. Make certain that he or his representative is planning to attend his Regional meeting and make sure that he is aware of your views on priorities in budget allocations and your feelings on any other IEEE matters of importance to you.

Section Chairmen Workshops

Another important event during the IEEE International Convention will be the holding of Section Chairmen Workshops. You may ask how you will benefit from these if you are not a Section officer. The answer is obvious once you are aware of their purpose. The workshops are designed to educate the Section officers in their duties and to teach them how to serve better the members in their local Sections. Section Chairmen and Vice Chairmen attending these workshops will be asked such questions as: Are you serving the interests of the members in your Section? What kind of personal contact do you have with the members in your Section? How do you go about finding out the needs of the members in your Section? How do you implement these needs once they are known to you?

Once again, if you want to obtain maximum benefit from your membership make certain that your Section Chairman or his representative will be attending these workshops.

Communications

The Regional meetings and Section workshops are two elements in our continuing program of improving the two-way communication process in our Institute. The Groups will be conducting a comparable series of leadership training programs during the International Convention. These will be for Group/Society officers and other members of Group/Society Administrative Committees and will consist essentially of an introduction to modern management concepts and their application to Group/Society operations.

Finally, I direct your attention to the Highlight Session at the International Convention on Monday evening, March 22. If you attend the Convention, I hope that you will be present at this session. At that time I expect to report further on activities of the Institute, and the other officers and chairmen of selected boards and committees will be available to respond to questions from our members.

J. H. Mulligan, Jr.
President, IEEE

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Nominations requested for Donald Eckman Award

The Donald P. Eckman Award has been established by the American Automatic Control Council in memoriam to encourage creative contributions to the field of automatic control by young scientists, engineers, and students.

Nominations are now being solicited for the 1971 award, which will be presented at the Joint Automatic Control Conference on August 11–13 at Washington University, St. Louis, Mo. Nominations are due by May 1.

In order to qualify for the award, an individual must be less than 30 years old and have contributed to the field of automatic control in the form of publications, theses, patents, inventions, or any combination of the preceding. The contribution for which the award is sought must represent work done prior to the age of 27 while resident in the United States. Supporting evidence must include a full endorsement by at least one responsible supervisor.

Nominations should be sent to the Donald P. Eckman Award Committee of AACC, c/o Systems Research Center, Case Western Reserve University, University Circle, Cleveland, Ohio 44106.

Professor Donald Eckman, who died in an automobile accident near Verdun, France, in 1962, conceived, founded, and directed the Systems Research Center at Case Institute of Technology, the first of its kind in the world. Professor Eckman, whose career was marked by pioneering contributions to the field of automatic control and systems engineering, was credited with a considerable number of technical inventions, scholarly textbooks, and scientific publications.

Educated at Jackson Junior College and the University of Michigan, Professor Eckman received the Ph.D. degree at Cornell University in 1950. A member of Phi Kappa Phi, Tau Beta Pi, and Sigma Xi, he received the ASME-IRD Award for an Outstanding Contribution to Engineering in 1958 and was honored as an ISA National Lecturer for 1959.

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Alternatives Now Open to IEEE Members Are. . .

Subscription to NSPE services. Available at a fee of $15 a year. Services include eligibility for the Professional Engineer Employment Referral Service (PEERS); eligibility for the NSPE retirement program; participation in the NSPE Salary Survey; subscription to the *Professional Engineer* magazine; a member rate on all NSPE publications; and receipt of any or all of the following on request: Legislative Bulletin, Legislative Action Report, Legislative Opinion Request, one practice section newsletter (except Professional Engineers in Private Practice newsletter).

Services from state society and national organization. Available at a fee of $30—$15 to be service charge to national, $15 to be service charge at state level. In addition to the services already mentioned (alternative 1), an IEEE member would receive the state publication and such other state services and participation as might be available, but would have no vote in state society or national activities. Several states have programs of discussions with employers on desirable employment practices for engineers.

Full membership in NSPE. This is available to IEEE members subject to eligibility provisions of NSPE—fee $35 to $75, depending on the particular state and chapter. NSPE members receive all services of the national, state, and local society and have voting privileges. Full membership would enable participation in all activities of the local, state, and national societies in such matters as improving the status of engineers, public relations, community involvement, ethics, career guidance, committee participation, holding office, voting on policy matters, and many others.

NSPE publications and services

Legislative Bulletin. A four-page publication issued monthly, the bulletin summarizes Federal legislation introduced in the Congress.

Legislative Opinion Request. This publication requests the opinion of those receiving it as to their position on various legislative proposals that may be introduced from time to time. It attempts to summarize the elements of proposed legislation. It may be limited to a particular bill, or to several bills introduced on the same subject.

Legislative Action Report. Action reports are issued as legislation progresses through the various stages of consideration. The report suggests to those receiving it the desirability of contacting their legislators concerning the position of the Society and is the mechanism by which the profession can be most effective.

Practice section newsletters. NSPE has four newsletters; one each for engineers employed in government, in industry, in private practice, and in education. The newsletters primarily report important developments in the area of employment for each group of engineers. Items range from significant statistical information to reporting on studies of value to that type of employment, employment practices, and salary and economic considerations.

Professional Engineer Employment Referral Service (PEERS). This service, for which only NSPE members are eligible, permits an unemployed or soon-to-be unemployed engineer to place his name and a short résumé on file with NSPE. These are summarized and sent without individual identification to more than 1000 employers. If an employer is interested in an anonymous summary, NSPE will provide him with the individual's name and address so that he may contact him for a full résumé and any details concerning employment.

NSPE retirement program. The retirement program is a combined annuity and mutual fund program for which NSPE members are eligible. It provides an investment program with guaranteed annuity through an insurance aspect and a variable annuity return via the mutual fund.

Salary survey. The NSPE conducts a biennial survey of its members' incomes and reports the results in breakdowns by age, field of employment, type of work assigned, and degree of responsibility.

Several programs were undertaken to assist in achieving this high objective.

A slide presentation "Why M&T?" was extensively used during the year at Regional Committee meetings, Section Workshops, and Section Executive Committee meetings to stress the importance of organizing and monitoring effective M&T programs at the Section level.

Regional Directors gave the support necessary to generate M&T activity at the grass-roots level, since the success of the membership goal depended to a great extent upon the degree of local attention given it by the Section Chairman and those serving with him on the Section Executive Committee. In an organization the size of the IEEE, the person-to-person approach is vital not only in recruiting new members, but in identifying and attempting to serve the special needs and interests of current members by encouraging their involvement and interest in the Institute. With the cooperation of the Regional Directors, a concentrated effort was made during the past year, through the Sections, to reinterest former members in continuing their membership. The subsequent substantial reduction of members in dues arrears can be traced to this direct communication with the members, which only the Sections can provide.

During the year members of the IEEE staff operated Member Services Desks at 20 major conferences sponsored or cosponsored by the IEEE; 575 membership applications and 656 applications for enrollment in the IEEE Groups and Societies were deposited. In addition, attention was given to 3011 inquiries from members and prospective members during these conferences.

The results of the 1970 M&T program are gratifying. At the year's end, the number of new elections (Student Members as well as other grades), the reinstatement of former members, and the renewals of membership exceeded the numbers in all of these categories over the prior two years. Even in this period of economic recession, the IEEE membership reached an all-time high of 169 059 at the end of 1970. Forty IEEE Sections, identified in the following, achieved a net increase of 10 percent or more in membership during the year.

REGION 2

Johnstown
Lehigh Valley
Ohio Valley
Susquehanna

REGION 3

Central North Carolina
Chattanooga
Eastern North Carolina
Fort Walton
Hampton Roads
Mobile
Palm Beach
Panama City
Pensacola

REGION 4

Northeastern Wisconsin

REGION 5

Central Texas
Corpus Christi
Kansas City
Ozark
South Plains

REGION 6

Boise
China Lake
Fort Huachuca
San Diego
Spokane
Utah

REGION 7

Canadian Atlantic
Hamilton
Quebec
Southern Alberta

REGION 8

Denmark
Egypt
France
West Germany
Israel
Middle and South Italy
North Italy
Switzerland

REGION 9

Colombia
Puerto Rico and Virgin Islands
Rio de Janeiro

M&T activity is well under way for 1971. The committee has been reorganized to include technical and geographical representation and the involvement of the Student Activities Committee to coordinate M&T activity effectively on an Institute-wide basis. The emphasis on successful M&T programs through the Regions and Sections will continue. Representatives from the six Technical Divisions are developing plans for attracting the membership, and new members, in the activities of the Groups and Societies. A concentrated program to involve graduating Student Members in the affairs of the Institute will have high priority in 1971. The M&T looks forward to continued interest and support from all IEEE organizational units and from individual members in these activities in the ensuing year.

W. L. Sullivan
Chairman, M&T

Divisions 1 and 5 to elect new Directors

In accordance with the provisions of the IEEE Bylaws, Division 1 (Audio and Electroacoustics, Circuit Theory, and Information Theory Groups, and the IEEE Control Systems Society) and Division 5 (IEEE Computer Society) will nominate and elect Directors in 1971.

Each Group Chairman or Society President is to submit nominations by April 30. The IEEE Nomination and Appointment Committee then will prepare and publish a slate with no fewer than two candidates for each position.

Individual members will have until August 15 to submit further nominating petitions. To be valid, such petitions must carry the signatures of at least 100 voting members. The IEEE Board will approve the final ballot, which will be mailed about September 1.

Directors should be of Fellow or Senior Member grade. Nominations should give biographical information, emphasizing prior IEEE experience.

Electric process heating conference plans program

A final technical program has been issued for the 1971 IEEE Conference on Electric Process Heating in Industry (G-IGA, Milwaukee Sect.). The conference will be held at the Pfister Hotel in Milwaukee, Wis., on April 20–21. Papers presented at the conference will cover applications, design, cost savings, new products, and new techniques related to the users of industrial electric equipment.

Session titles follow.

TUESDAY, APRIL 20

Morning Sessions

Solid-State Power Sources I
Melting

Afternoon Sessions

Special Processes
Application Techniques

WEDNESDAY, APRIL 21

Morning Sessions

Mass Heating
Solid-State Power Sources II

Afternoon Sessions

Joining
New Processes

For additional information consult Byron L. Taylor, Induction Process Equipment Corp., 32251 N. Avis Drive, Madison Heights, Mich. 48071.

Surface elastic waves is subject of symposium

On April 6 the Albuquerque Section will sponsor a one-day symposium on surface elastic waves. This symposium, tutorial in nature, will be held at Sandia Laboratories in Albuquerque, N. Mex.

Several aspects of surface-elastic-wave technology will be explored with emphasis on potential device applications. The invited speakers include Dr. W. R. Jones of Hughes Aircraft, Prof. R. M. White of the University of California at Berkeley, Dr. John deKlerk of Westinghouse, Prof. Irving Kaufman of Arizona State University, Dr. D. B. Armstrong of Litton Industries, Prof. C. F. Quate of Stanford University, and Dr. Mark Dakss of General Telephone and Electronics.

For further details, contact Dr. Willis D. Smith, Sandia Laboratories, Division 5153, Albuquerque, N.Mex. 87115.