RF design

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The April

Catalog/Data Sheet Section

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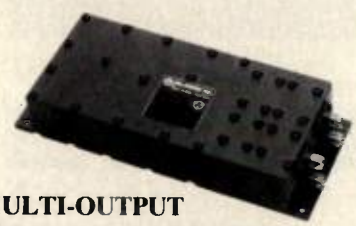

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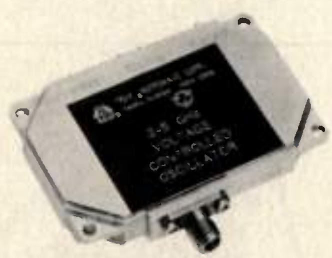

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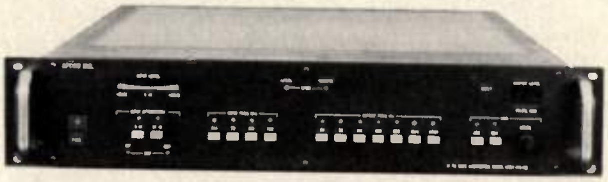

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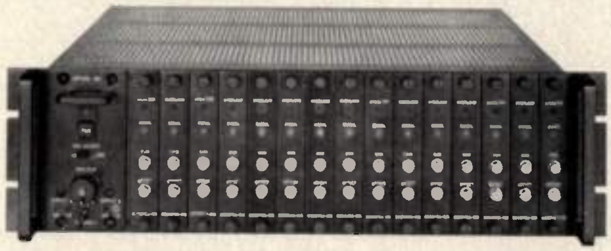

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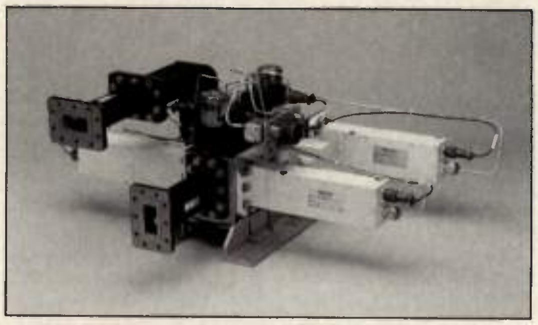

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INFO/CARD 54

Communications Range and Reliability of Part 15 Devices

By Bernard Kasmir
ADEMCO

Low power communication devices authorized under FCC Part 15 have signficant perfomance limitations. Some of these limitations are a result of regulations, and others are part of the mechanical and electrical design, cost, and complexity requirements of the manufacturer. The key performance concerns from the user's perspective are range and reliability. Part 15 device designers have the task of dealing with the limitations while maximizing range and reliability.

When we speak of radio range, we talk in terms of how far we can have communications. The definition becomes ambiguous unless we define initial conditions. There are, for example, best case, worst case, and typical performance; or the more generalized probability of communications under defined conditions.

When using Part 15 low power communication devices for the protection of life and property, reliable range, however it is defined, becomes a significant specification. The typical transmitter is a small PC board housed in some type of plastic case. Usually, the transmitter does not have an external antenna. Indeed, the antenna can often not be completely defined since it consists of a printed loop plus radiation from all components including the battery, and sometimes any external sensor wires.

The FCC specifies "best case" signal level, which is measured by finding the maximum level of radiation considering all possible orientation positions. This signal is expressed in field intensity in microvolts per meter at some specified distance (usually 3 or 10 meters). The FCC is not concerned what type of antenna is used (if any) as long as the best case signal level is maintained.

Radio receivers are usually measured for sensitivity in a 50 ohm system. However, the ability of a receiver to pick up a signal is also determined by the receiver antenna. The true measure of the receiver sensitivity is the ability to intercept a given field intensity. The actual receiver sensitivity can then be expressed in microvolts/meter. The antenna factor of the receiving antenna expresses the ability of the antenna to generate a voltage into 50 ohms for a particular field intensity. In other words:

AF = E/V

where, E is the field intensity in uV/m and V is voltage into 50 ohms. The lower the antenna factor the more voltage is delivered to the receiver for a particular field intensity.

Since the transmitter is specified in microvolts per meter (at some defined distance) and the receiver sensitivity can be defined in terms of field intensity in uV/m, a theoretical best case range can be defined by the use of linear interpolation. Thus, the field intensity (E2) at distance of S2 would be:

$E2 = E1 \times (S1/S2)$ (1)

In this case, S1 would be the distance where the transmitter field intensity E1 is specified, usually 3 meters. This shown in Figure 1 as the "1/X" curve where signal level (voltage) is halved as the distance is doubled.

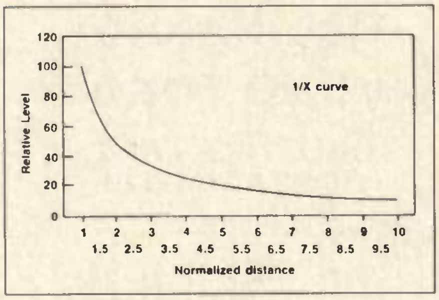

Figure 1. Signal level vs. distance (1/X curve).

The transmitter signal level as specified by the FCC is dependent upon frequency and duty cycle. Part 15.202 lists signal levels as follows: from 260 to 470 MHz, 1500-5000 uV/m. The field intensity at any operating frequency can be found by linear interpolation of this range. The field intensity E (in uV/m at 3 meters) at frequency F is equal to:

E = 16.67 (F-260) + 1500

Duty cycle

This is determined by the average-to-peak value of signal level. The FCC

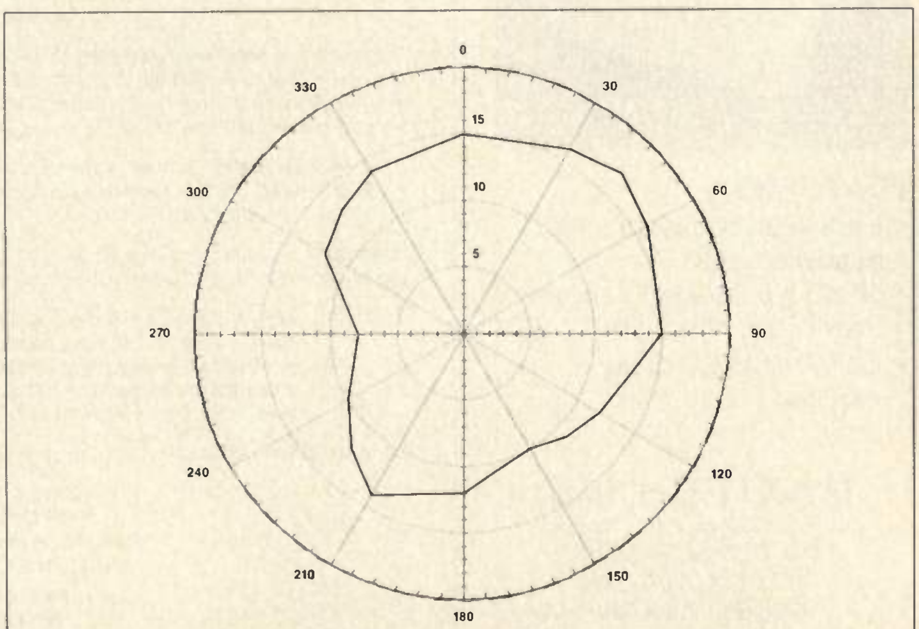

Figure 2. Signal level vs. orientation for a Part 15 transmitter.

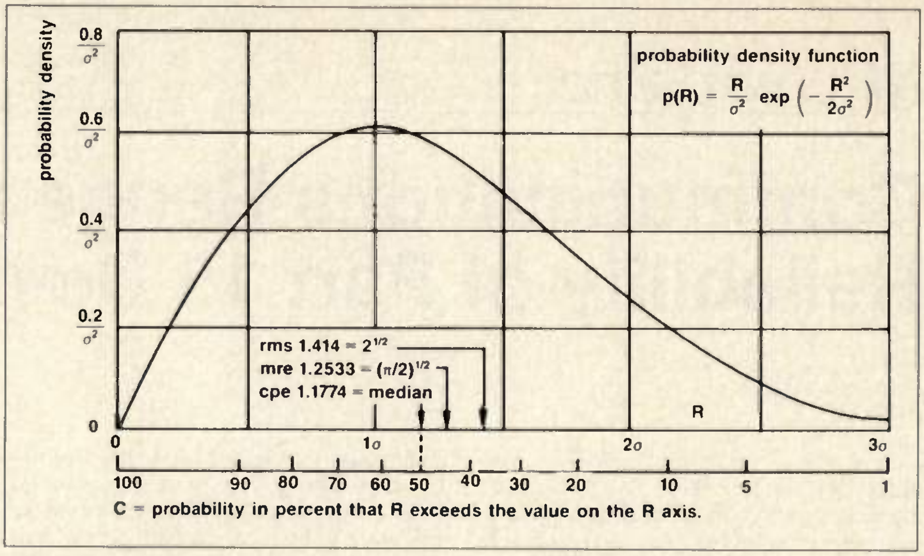

Figure 3. Rayleigh probability distribution model.

allows up to 20 dB (10x) peak signal when pulse modulated and averaged over 100 ms. In the following example we shall calculate range based upon some initial values. The values chosen do not represent any specific product, but rather typical values used in this tutorial.

For example, at 300 MHz, the average signal level allowed would be 2166 uV/m. Assume a duty cycle of .2 (the signal is on 20% of the time averaged over 100 ms). This yields a duty cycle advantage of 5, bringing the peak signal level to 10,830 uV/m at 3 meters.

Also, for this example, assume a receiver with a 30 dB antenna factor (numerically 31.6) and a receiver with a 2 microvolt sensitivity into 50 ohms. The receiver sensitivity in terms of microvolts per meter would be;

$31.6 \times 2 = 63.2$ uV/m

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The "best case" range would then be, from (1):

10,820 × 3 / 63.2 = 513 meters, or 1650 feet.

A typical distance between the control panel and a remote transmitter in a house may be 50 feet. It certainly appears as if we have plenty of signal to spare. In this example, we have about 30 dB signal margin, that is, a best case signal level 30 dB about threshold (20 Log 1650/50).

However, there is a big difference between theoretical best case and actual operation. As we shall see, there are many hazards that tend to reduce the range of a system.

Orientation

Figure 2 shows relative radiated field intensity of a transmitter as a function of orientation or position. The ideal antenna pattern would be omnidirectional (equal signal level in all directions), but this is not practical to achieve. As a result, the field intensity at the receiver is dependent upon the position of the transmitter. This field pattern is also influenced by building materials where the transmitter is mounted.

As with the transmitter, the receiver antenna can also exhibit directional behavior, with varying sensitivity depending on its orientation. In cases where the receiver location is fixed and the transmitter does not move over a wide azimuth relative to the receiver, fairly constant performance can be expected. If neither transmitter nor receiver are fixed, the variation can exceed that shown in Figure 2.

Attenuation

Building materials are translucent (not transparent) to radio waves at the frequencies typically used by Part 15 devices. A signal from a remote transmitter to a receiver will experience some attenuation greater than free space path loss. Studies have been conducted, but some controversy remains on levels of attenuation.

Interference

More and more electronic devices are used in homes and businesses. Devices such as digital clocks, computers, VCRs and even other receivers can cause interference. Although emissions are regulated, the close proximity of an electronic device to a radio receiver can cause significant interference. This results in reduced receiver sensitivity as well as reduced range.

We usually think in terms of deterministic models where performance can be precisely calculated. In the field of communications, there is some randomness of events (apparent randomness due the complexity of the signal propagation) and we cannot obtain a precise calculation. Rather, it is more useful to work with the probability of an event over a defined interval. The signal attenuation caused by the three factors discussed above cannot be given precisely. However, unless we obtain a probabilistic model, we can only operate using intelligent guesses.

Saving the best for last, the most significant specific phenomena that produces variability in received signal level is multiple reflections. The received signal level is a vector of all direct and reflected signals. This has been mathematically modeled and is represented by the Rayleigh probability distribution function. This model is comprised of an infinite number of small reflectors. Figure 3 shows the shape of this curve, which is called a log-normal curve. The mean represents zero signal attenuation. There is a 50 percent probability that the signal is enhanced anywhere up to 6 dB (left side of the curve). If reflections are all completely in phase the signal can be enhanced up a maximum of 6 dB. However, complete cancellation can theoretically result in infinite signal attenuation.

Table 1 summarizes the probability of additional attenuation caused by multiple reflections.

What does this mean? First, there is a 50 percent probability that the additional attenuation caused by multiple reflections will be 0 dB or less. There is a 90 percent probability that the additional attenuation caused by multiple reflections will be 10 dB or less. Similarly, the chart covers 20, 30 and 40 dB attenuation.

Attenuation	Probability
0 DB	50%
10 DB	90%
20 DB	99%
30 DB	99.9%
40 DB	99.99%

Table 1.

Take any one of the above chart, 20 dB for example. Since there is a 99 percent probability that the additional attenuation is 20 dB or less, then, if we have a 20 dB margin, there is a 99 percent probability that the signal will get through. By knowing the signal margin of a system, we can predict the reliability.

Now, back to our example; since we do not have a good model for items such as attenuation, orientation and interference, let us assign a "best guess" of 10 dB for these above items. This reduces our previously determined signal margin from 30 dB to 20 dB. According to the above table, we have a 20 dB signal margin and a 99 percent reliable system.

Conclusion

To assure maximum systems reliability, more than basic transmitter and receiver performance is required. It is important to use good installation practices to minimize the added attenuation due to the factors covered in this article. There are many techniques used to assure maximum received signal. Advantageous orientation, a clear signal path, no interference sources should be combined with predictable transmitter and receiver performance, including duty cycle and peak power, receiver sensitivity, and antenna factor. Subsequent articles will address these techniques. ***RF***

About the Author

Bernard Kasmir is Senior RF Design Engineer at Alarm Devices Manufacturing Company (ADEMCO), 165 Eileen Way, Syosset, NY 11791. He holds BSEE and MSEE degrees and a P.E. license. Mr. Kasmir has worked with Part 15 devices for more than ten years.

Parasitic Positive Feedback Frequency Acquisition in a PLL

By Jonathon Y.C. Cheah
Hughes Network Systems

When the use of a double balanced mixer as a phase comparator is necessary, frequency acquisition is an inherent problem. Double balanced mixers are commonly used in PLLs with a very high detector frequency or circuits such as analog Costas' loops where digital frequency/phase comparators are not suitable.

The parasitic positive feedback frequency acquisition scheme is not well known (1). A probable reason is that the direct application of this concept has the problems of achieving reliable sweep oscillation and controlling the sweep waveform. Therefore, its use is limited (2). A practical implementation that overcomes these problems is described here. This implementation fully demonstrates the automatic scan and stop action quality of this concept as well as its simplicity. As the feedback circuit is independent of the PLL's parameters other than the sweep rate, this circuit can, in many cases, be a "boiler plate" add-on to where it is needed. In areas such as Costas' loop carrier recovery circuits, where complex frequency scanning is needed to achieve frequency acquisition, this circuit can be a good candidate.

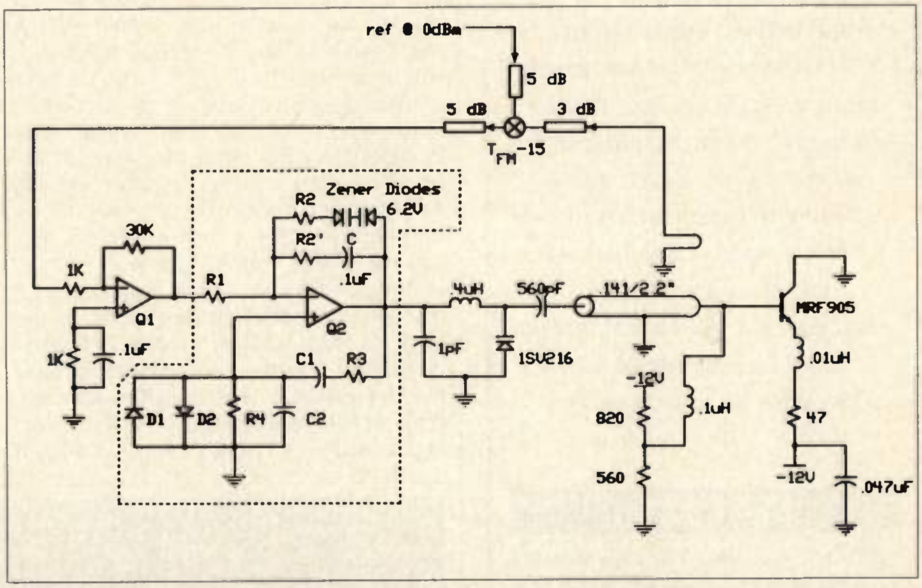

Figure 1. The box indicated by dotted lines shows the 2nd order loop filter with added positive feedback circuitry. The component designators are as used in the test and their values are provided.

The frequency acquisition design can be described as a Wein-Bridge oscillator superimposed on a 2nd order loop filter. When the loop is unlocked, the positive feedback oscillator is activated to provide frequency sweeping action. When the PLL attains the lock condition, the required negative gain feedback to sustain a continual oscillation condition is

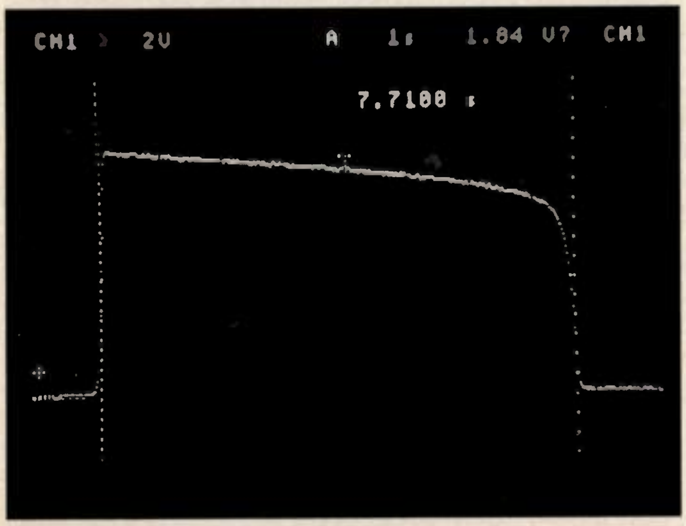

Figure 2. The sweep voltage at the output of the 2nd op-amp when the PLL is unlocked.

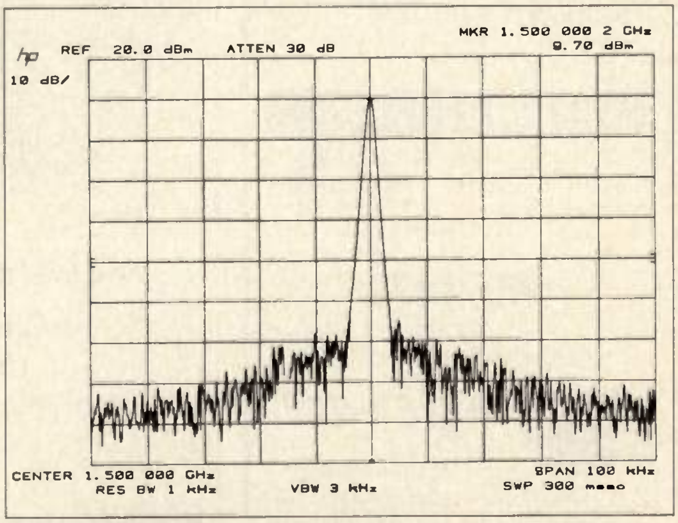

Figure 3. The frequency spectrum of the VCO when the loop is locked.

discontinued, while the infinite DC gain required by the phase lock loop is maintained.

Consider the transfer function of an active 2nd order PLL:

$$F(s) = \frac{s\tau_2 + 1}{s\tau_1} \quad \text{where} \quad \begin{cases} \tau_1 = R_1C \\ \tau_2 = R_2C \end{cases} \qquad (1)$$

as shown in Figure 1.

When the frequency offset is large, the low pass loop filter effectively reduces the gain of the op-amp. For microwave VCOs, the offset frequency will most likely be beyond the 2nd pole of the op-amp's frequency response. However, at DC, the amplification gain is essentially the open loop gain of the op-amp. The output voltage of the amplifier will therefore reach the positive or negative rails of its maximum DC voltage swing. Referring to Figure 1, C is designed to be a high impedance at the sweep oscillation frequency. If the back to back Zener diodes have the combined Zener and forward diode biasing voltages less than the rail voltages of the op-amp and series resistor, R_2, provides a negative feedback gain of 2 or higher, the Wein-Bridge oscillator gain condition of $G(j\omega) \geq 1$ is satisfied where:

$G(j\omega) =$

$$\frac{1 + \frac{R_2}{R_1}}{1 + \frac{R_3}{R_4} + \frac{C_2}{C_1} + j\left(\omega R_3 C_2 - \frac{1}{\omega R_4 C_1}\right)} \qquad (2)$$

The purpose of using the Zener diodes is two-fold. First, they are in place to obtain a low frequency negative feedback path for the oscillation condition to be met.

Second, when the amplifier output voltage falls below the Zener voltage, they provide a virtual open circuit. In this way, the PLL function is isolated from the sweep oscillator.

The shunt small-signal diodes D_1 and D_2 provide a means of ensuring that the output voltage swing is clamped by the Zener diodes at the correct level. When an excessively large value of R_2 is used, the effectiveness of the clamping function will be lost.

For the PLL to acquire lock under a frequency sweep condition (1) with bet-

Zener Diode Voltage	Δ V^{-1}
3.3 V	160 ± 20 msec/V
3.6 V	160 ± 10 msec/V
3.9 V	160 ± 20 msec/V
4.3 V	120 ± 30 msec/V
4.7 V	110 ± 20 msec/V
5.1 V	75 ± 20 msec/V
6.2 V	40 ± 10 msec/V

Table 1. ΔV was measured at 1 to 0 Volt crossing.

ter than 90 percent successful lock probability, the rate of change of the output VCO frequency ω_{VCO} is an approximate function of the loop bandwidth ω_n.

$$\frac{d\omega_{vco}}{dt} \leqslant 0.5\omega_n^2 \tag{3}$$

By definition, the rate of change of the output VCO frequency can be written as,

$$\frac{d\omega_{vco}}{dt} = \Delta V K_{vco}$$

$$\Delta V = 2f_{so}V_{pk} \tag{4}$$

$$= \frac{\omega_{so}}{\pi} V_{pk}$$

where,

V_{pk} is the maximum voltage swing potential of the sweep oscillator

$\omega_{so} = 2\pi f_{so}$ is the angular frequency of the sweep oscillator

K_{VCO} is the gain constant of the output VCO.

Thus the maximum ΔV, and therefore the frequency of the sweep oscillator can be calculated.

It is also clear that the lock condition should occur at a point where the output voltage V_l of the loop filter is less than $|V_z|$, where V_z is the Zener voltage of the back to back Zener diodes. Under the lock condition, the negative feedback condition required by the Wein-Bridge oscillator is almost non-existent due to the fact that very little or no current flows through R_1. In addition, the open-circuit action of the Zener diodes prevent any further positive feedback tendency.

There is a small penalty suffered in the PLL loop dynamics. Since it is inconvenient to limit $R_4 = R_1 // R_2'$, the op-amp output bias current will create a small amount of DC offset. The selection of a low input bias current op-amp will mitigate this problem. In most cases, this problem is not important when compared with other sources of DC offset phenomena such as that of a mixer.

It is essential to ensure that the input of R_1 does not have a significant amount of fixed DC offset. This offset would prevent sweep oscillation from taking place.

A test circuit was designed to illustrate the usefulness of this method. A 1.5 GHz VCO was designed to provide ± 55 MHz deviation for the control input voltage of 0 to 10 V. The relatively wide tuning range of this VCO is intentional, so that the frequency acquisition action can be more clearly demonstrated. A second order loop is constructed as shown in Figure 1 where $\omega_n = 2\pi \times 10{,}000$ radians/sec and $\zeta = 2$. Thus the maximum rate of voltage change required at the DC crossing is $\Delta V < 28.6$ V/sec.

A heuristic estimate of the V_{pk} to V_z of about 25 is used, as V_{pk} is not always immediately obvious. With this initial value, all the circuit components can now be determined.

Table 1 shows the measured maximum ΔV with respect to the choice of Zener diodes. Obviously, the smaller the Zener voltage, the smaller the VCO tuning range is allowed. Figure 2 shows the sweep voltage waveform and Figure 3 shows the quality of the lock output spectrum of the VCO.

The requirements on all the circuit components are not critical and are simple to implement. In fact, it is convenient to adopt a slow sweep rate configuration, such as the one shown in the example, as general framework for PLLs of this kind which can tolerate equal or faster ΔV. ***RF***

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1. F.M. Gardner, *Phase Lock Techniques*, John Wiley and Sons, 1979, pp. 79-84.
2. J.Y.C. Cheah, "Low Cost DDS Applications in Ku-band Frequency Synthesis," *Proc. 2nd ISRAMT*, Sept. 1989, pp. 498-501.

About the Author

Dr. Jonathon Y.C. Cheah is a Senior Principal Engineer with Hughes Network Systems-West. He is currently involved in RF LAN and DBS areas. He may be reached at 10790 Roselle St., San Diego, CA 92121.

Biasing Solid State Amplifiers to Linear Operation

By Helge Granberg
Motorola, Inc., SPS

All solid state devices intended for linear operation must have a certain amount of "forward bias" (idle current) in order to place their operating points in the linear region of the transfer curve. Bipolar devices require a constant voltage source, whereas MOSFETs can be biased with simple resistor divider networks. Both will get more complex however, if temperature stability is required. Examples of applications requiring amplifier linearity include all amplitude modulated systems for communications and broadcast, nuclear magnetic resonance, magnetic resonance imaging, digital cellular telephone and signal sources for instrumentation. Circuit examples are presented from the most simple ones to sophisticated closed loop systems.

Since the base current of a bipolar transistor is equal to I_c (peak)/h_{FE}, the base bias supply must be able to supply this current without considerable excursions in the base-emitter voltage between the no-signal and the maximum signal conditions. This requires a constant voltage source, as variations of a few millivolts represent a large portion of the nominal 0.63 - 0.67 volt typical value. Depending on the specification of a specific application, various degrees of requirements are set for the base bias voltage source. In some instances a large value capacitor can be seen connected across the bias voltage supply to further reduce its AC impedance. However, this makes the impedance dependent on the frequency of modulation and is a good and practical solution only in applications where the modulating frequency is in the medium to high audio frequency range. One of the simplest biasing circuits for bipolar transistors (4,5,6), is shown in Figure 1a. It uses a clamping diode to provide a low impedance voltage source. The diode forward current must be greater than the peak base current of the transistor. This current is adjusted with R2 and the resistance of RFC1 and R1 combination is used to reduce the actual base voltage to a slightly lower value than the forward voltage of D1. The diode can be mechanically connected to the heat sink or the transistor housing to perform a temperature compensating function to Q1. This technique works adequately, although for perfect temperature tracking, Q1 and D1 should have similar DC parameters. One disadvantage with the circuit shown in Figure 1a is its inefficiency especially in biasing high power devices since (V_{cc} -V_b)I_b(max) will always be dissipated in the dropping resistors.

This deficiency of the circuit shown in Figure 1a can be overcome by amplifying the clamping diode current with an emitter follower (1,2) as shown in Figure 1b. Two series diodes (D1-D2) are required since one has to compensate for the V_{BE}(f) drop in Q1. In this case, low current signal diodes can be used and their forward current is equal to I(bias)/h_{FE}(Q1). For best results, Q1 should have a linear h_{FE} up to the peak bias current required and in higher power systems it must be cooled by some means. Ideally, Q1 and one of the series diodes should remain at ambient temperature, whereas the other diode (D1 or D2) can be used for temperature compensation of the RF device. An effective and fast responsive system is obtained if the diode (having long leads) is located near the RF transistor. The leads can be suitably formed allowing the body of the diode to be pressed against the ceramic lid of the RF transistor and fastened in place with thermal conductive epoxy. R1 is used to set the bias idle current and R2 limits

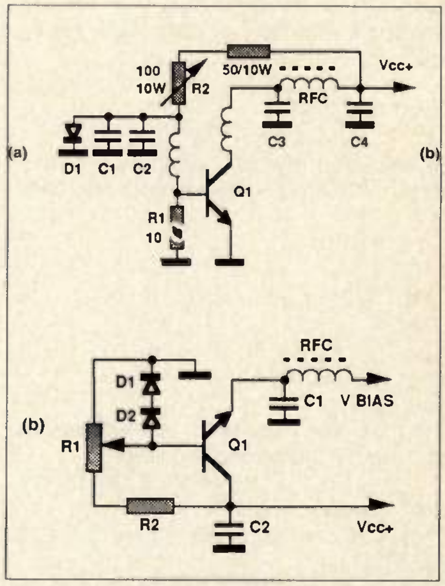

Figure 1. Two simple bias sources for bipolar transistors. (a) Uses all passive components, but is inefficient. This drawback is not present in (b), although it is slightly more complex.

its range of adjustment. The value of R2 depends on the supply voltage employed. The function of C1 and the RFC is simple to prevent the RF signal from getting into Q1.

Another fairly simple bipolar bias source (3) is shown in Figure 2. Its output voltage equals the base-emitter junction drop of Q1 plus the drop across R3. R1 must be selected to provide sufficient base drive current for Q2, set by its h_{FE}. Normally this current is in the range of a few milliamperes, and Q1 can be any small signal transistor in a package configuration that can be easily mounted to the heat sink or RF transistor housing for temperature compensation. The only requirement is that its V_{BE}(f) at that

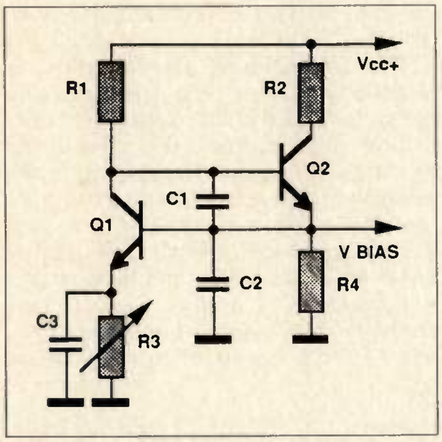

Figure 2. This circuit features the lowest source impedance of the less complex ones. Therefore it is recommended for high power device biasing and for demand applications.

current must be lower than that of the RF transistor at its bias current level. The maximum current capability depends on Q2 and R2. The power dissipation of Q2 can be up to a few watts and in most cases should be heat sunk, but must be electrically isolated from ground. The value of R2 can be calculated as: $(V_{CE}\text{-}V_{CE}(sat))/I_b$. C1 through C3 are a precaution to suppress high frequency oscillations, but may not be necessary depending on the transistors used and the physical circuit layout. Output source impedances for this circuit, when used in conjunction with a 300 W amplifier, have been calculated as low as 200-300 mOhms.

More sophisticated bias sources can include an integrated circuit voltage regulator (2). In most cases a pass transistor is required for current boost and to lower the source impedance. There are high current regulators available today, such as the LM317, LM337, etc., but the author has not investigated whether they are suitable for applications such as this. The circuit in Figure 3 uses a 723 regulator, which is available from several manufacturers with a variety of prefixes. It has been used for bipolar bias sources since the early 70s and more recently for MOSFET biasing as well. The 723 is specified for a minimum V_{out} of 2 volts, but with certain circuit modifications can be lowered to less than 0.5 V. The main advantages of this type bias source are: 1) It provides the lowest source impedance at a relatively low cost. 2) The bias voltage remains independent of variations in the power supply voltage. 3) Temperature compensation is easy to implement. In Figure 3 D1 performs this function and should be in thermal contact with the heat source. The same technique discussed with the circuit shown in Figure 1b can also be adapted here. Depending on the current requirement and the pass transistor used, Q1 may have to be cooled. It has a positive temperature coefficient to the bias voltage, but is negligible compared to the negative coefficient of D1. This permits Q1 to be attached to the main heat sink. R1 and D2 are only necessary if the RF amplifier is operated at a supply voltage higher than 40V, which is the maximum rating for the regulator.

Biasing of MOSFETs

Since MOSFETs have gate threshold voltages up to 5-6 volts, they require some base bias voltage in most applications. They can be operated in class C, (zero gate bias) but at a cost of low power gain. In such cases the input

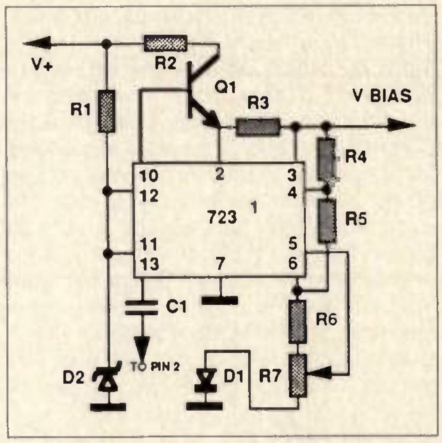

Figure 3. An integrated circuit bias source. The idle current remains constant regardless of the supply voltage. The source impedance mainly depends on the h_{FE} of Q1.

voltage swing must have an amplitude sufficient to overcome the gate voltage from zero to over the threshold level. The drain efficiency is usually higher than in other classes of operation. Especially if overdriven, the class of operation can approach class D. Zero bias is often used in amplifiers intended for FM or CW; and efficiencies in excess of 80 percent are not uncommon. In class B the gate bias voltage is set just below the threshold resulting in zero drain idle current flow. The power gain is higher than in class C, but the drain efficiency is 10-15 percent lower. Class B is suitable only for the FM and CW modes. Between these two classes of operation, one must decide whether the system has power gain to spare and how important is efficiency. At higher frequencies, such as UHF, a good compromise may be class B or even class AB. In class AB the gate bias voltage is somewhat higher than the device threshold, resulting in drain idle current flow. The idle current required to place the device into the linear mode of operation is usually given in a data sheet. In this respect MOSFETs are much more sensitive to the level of idle current than are bipolar transistors. They also require somewhat higher current levels compared to bipolars of comparable electrical size.

The temperature compensation of MOSFETs can be most readily accomplished with networks consisting of thermistors and resistors. The ratio of the two must be adjusted according to the thermistor characteristics and the g_{fs} of the FET. The changes in the gate threshold voltage are inversely proportional to temperature and amounts to approximately 1 mV/degree C. These changes have a larger effect on the I_{DQ} of a FET with high g_{fs} than one with low g_{fs}.

Unfortunately the situation is complicated by the fact that g_{fs} is also reduced at elevated temperatures, making the drain idle current dependent on two variables. In spite of this, this method of temperature compensation can be designed to operate satisfactorily and is repeatable for production. The thermistor is thermally connected into a convenient location in the heat source in a manner similar to that described for the compensating diodes with bipolar units discussed earlier. An example of a simple MOSFET biasing circuit (2) as described here is shown in Figure 4.

Most MOSFET device data sheets give V_{GS}(th) versus I_D data, but the values are only typical, and in some cases the g_{fs} can vary as much as 100 percent from unit to unit. Thus, in production the devices should have g_{fs} values that are matched on the parameter above to at least 20 percent. Otherwise each amplifier must be individually checked for temperature tracking. Some manufacturers such as Motorola supply RF power FETs with color coded g_{fs} matching.

The circuit in Figure 5 shows a typical MOSFET bias voltage source using the 723 IC regulator (2), which was earlier presented for bipolar transistor biasing. Since a MOSFET draws no gate bias current, except in the form of leakage, the pass transistor (Q1) has been omitted and D1 replaced by R5-R6 combination. The values of other passive components have also been modified to produce a maximum output of 8 volts. The temperature slope is adjusted by the ratio of the series resistor (R5) and the thermistor (R6). In addition to maintaining a constant bias voltage, this circuit also features bias voltage regulation against changes in the power supply voltage.

Figure 6 shows a closed loop system for MOSFET biasing. It provides an automatic and precise temperature compensation to any MOSFET regardless of its electrical size and g_{fs}. No temperature sensing elements need be connected to the heat sink or to the device housing. In fact, FETs with different gate threshold voltages can be changed in the amplifier without affecting the level of the idle current. The gate threshold voltage range is about ± 0.5 V with a single initial setting of the idle current. This means that the gate threshold voltage can vary within these limits over short or long periods of time for whatever reasons. In addition to temperature, other factors affecting V_{gs}(th) might be moisture, atmospheric pressure, etc.

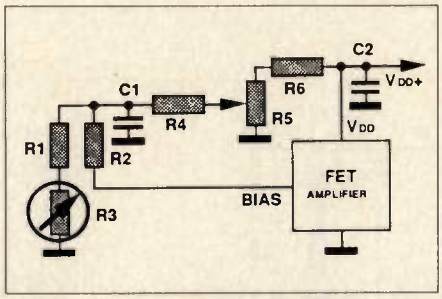

Figure 4. A simple MOSFET bias circuit using a thermistor-resistor network for temperature compensation.

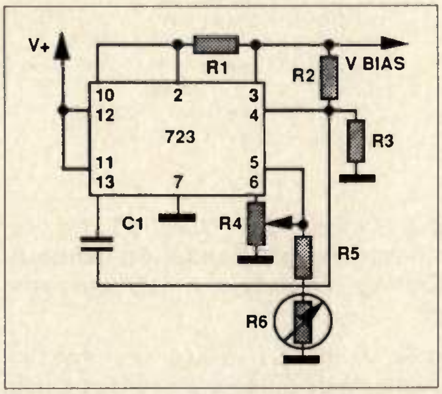

Figure 5. A more sophisticated MOSFET bias system with an integrated circuit voltage regulator. It also employs a thermistor for temperature tracking.

The principle of operation of circuit in Figure 6 is as follows: The idle current of the MOSFET amplifier is initially set to class A, AB or anywhere in between these bias limits by R8, which also provides a stable voltage reference to the negative input of the operational amplifier U1. This results in a current flow through R1 with a consequent voltage generated across it. This voltage is fed to the positive input of U1, which results in the output of U1 following it in polarity, but not in amplitude. Due to the voltage gain in U1, which operates in an open loop mode, its output voltage excursions are much higher than those generated across R1. Thus, if the current through R1 tends to increase for any reason, part of the output voltage of U1 fed to the amplifier gate bias input will adjust to a lower level, holding the current through R1 at its original value. A similar self adjustment will take place in the opposite direction as well.

The values for the resistive voltage divider R5-R6 have been selected for a range of ± 0.5 Volts at the amplifier FET

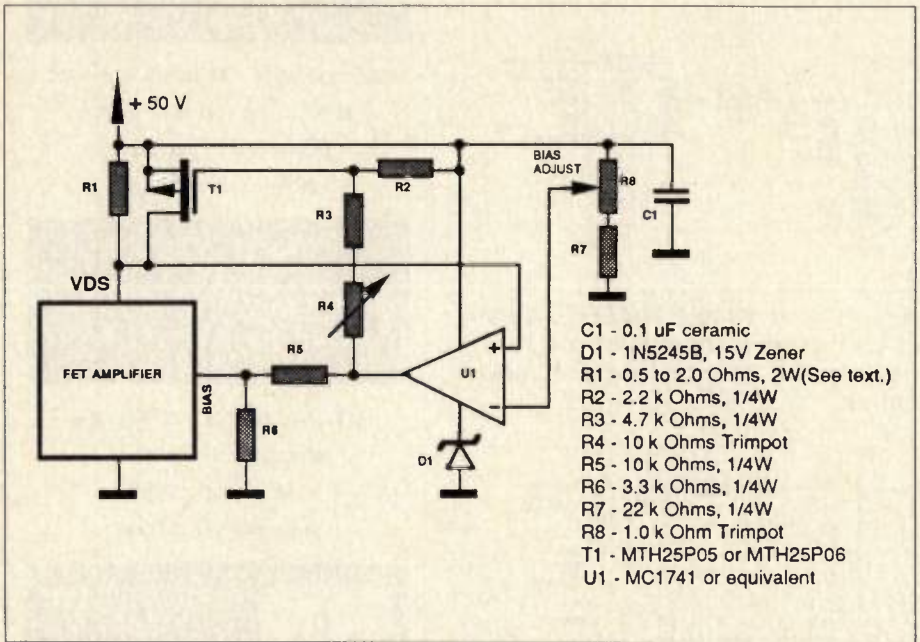

Figure 6. An automatic bias tracking system for MOSFET power amplifiers. It provides automatic temperature compensation without sensors as well as versatility for substituting a variety of electrical sizes of FETs operating at any supply voltage.

gate for the full voltage swing at the output of U1. This larger voltage swing is required to provide negative bias to the gate of the P-channel FET Q1, turning it on harder the higher the current is through R1. The r_{DS}(on) of Q1, which can decrease as much as 50 percent with 1V increase in gate voltage, depending on where the operating point is set by R4, is used in parallel with R1 to function as a voltage variable resistor. r_{DS}(on)(Q1) in parallel can be less than 0.2 ohms at 10 amperes. The value of R1 must be selected according to the desired level of idle current and R2 through R4 combination. R4 is adjusted only once (starting with its maximum value) according to the V_{GS}(th) and other parameters of Q1 to the point where the idle current previously set starts to increase. With correct component values, the gate bias voltage can be made to remain constant, or even increase (if desired) from the idle current level to the maximum drain current drawn under RF drive conditions. A bipolar Darlington transistor such as MJH6085 can be also used in place of the P-channel FET (Q1), but its h_{FE} must be 3000-4000 minimum in order for U1 to be able to drive it.

Summary

The circuits presented in this paper are of a basic nature, and may require refinement or modification according to specific applications. Equations to calculate the component values for most circuits described herein are given in the references and have not been duplicated here. No known description exists for the circuit shown in Figure 6. Circuit analysis and calculation of component values will be presented in a forthcoming paper by the author.

Many of these circuits have been in use by the industry in various forms for years, but the designers are constantly looking for simpler and better performing bias sources for solid state amplifiers. One possibility might be an Application Specific Integrated Circuit (ASIC) in a form of SMART power. These integrated circuits are gaining popularity and are manufactured by several companies including Motorola. Most are intended for automotive applications.

An integrated circuit can certainly be designed that satisfies the requirements for bipolar transistor and MOSFET biasing combining all the features presented in this paper. It is not certain though, that the existing market can justify such an IC development. However, new applications requiring linear amplification, such as digital cellular telephones, are being created and older vacuum tube designs are being converted to solid state designs at an ever increasing rate. An estimated increase in solid state linear amplifiers for all applications combined is at least a factor of 4-5 within the past ten years and is expected to double by the end of the decade. ***RF***

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A Stable Future for the Oscillator Market

By Liane G. Pomfret
Associate Editor

It comes as no surprise that the SAW and crystal oscillator business is doing well in the face of economic uncertainty. Their already widespread use, coupled with emerging technologies has provided a large market for manufacturers. SAW and crystal oscillators have not undergone any recent technological breakthroughs. Rather, changes continue to be small and cumulative.

The technology behind SAW and crystal oscillators has not changed much recently, nor is it expected to undergo any major changes in the near future. However, a number of small changes are occurring. One visible area of improvement is that component suppliers are improving their own quality control processes and passing these improvements on to the oscillator manufacturers. "We're getting nicer crystals," notes Liz Ronchetti, Vice President at Wenzel Associates, "the crystal manufacturers are doing better." Frank Perkins, Vice President of Marketing at RF Monolithics commented on the same trend in the MMICs and transistors that his company buys. In both cases, the suppliers, in trying to improve their own product, have helped improve SAW and crystal oscillator products.

Manufacturers are also paying closer attention to their own design processes. The use of computer aided design techniques in the design of circuits has improved reliability and performance and reduced the time and costs associated with developing new products. Problems such as phase noise under vibration, stability, low power, temperature performance, and special frequencies are all problems that are easier to solve because of the versatility of CAD programs. In the high performance area, the advent of microprocessor controlled oscillators has caused a bit of a stir within the industry. The technology first became available about four years ago and it's only been in the last two that it has become popular.

Another push lately has been towards surface mount packaging and miniaturi-

zation. While surface mount has been in use for years within the RF industry, it is just beginning to appear in the oscillator industry. One example of this, is from Wenzel Associates who recently introduced three new products, all in miniature packages. Marty Finklestein, President of MF Electronics observes, "We feel that surface mount is starting to catch on. Not only high volume parts, but more difficult products as well." Often the people who are pushing for surface mount are not the large, commercial end-users but the high-reliability, tight specification military users. Bob Murphy, Director of Aerospace Business at Frequency and Time Systems indicated that their switch to surface mount for some of their products was due to requirements for smaller sizes and lower power for Lightest programs. Since many of the companies in this report manufacture military, space or high reliability products, their attention to specifications and standards is much stricter than a high volume, popcorn part manufacturer. The military appears to be a driving force these days in development of smaller and better oscillators. Their need for small, high tech defense systems has forced manufacturers develop the necessary components on a much smaller scale than previously required. Tom Mihalek, Contract Administrator at Anderson Labs noted that improving quality is always an ongoing process — an idea few people would argue with. As Brian Rose, Vice President of Engineering at Q-Tech observes, "There's a responsibility to the customer."

These days, customers are using oscillators in just about every imaginable technology. With the new wave in communications technology, SAW and crystal oscillators are being used in everything from military spread spectrum to children's toys. Wenzel Associates has noticed that many of their new orders are coming from high performance products using spread spectrum and frequency hopping. Others have found that digital communications is the area with the most activity. Marty Finklestein indicated that standard telecommunications, or telephones are very strong for his company. RF Monolithics on the other hand, has seen their orders coming from areas such as air traffic control, aircraft and microwave radios. The list of products that use SAW and crystal oscillators is long and includes other areas such as radar, computers, spacecraft, satellites, cellular telephone and many more.

Everything points to the fact that the SAW and crystal oscillator business is doing well. Comments range from "we're tickled to death" to "we're pleasantly surprised." There were indications of a slow period at the end of last year, but business picked up again in January and is forging forward. With constant refinements and fine tuning, the oscillator industry is creating better products — from high cost, low volume military all the way down to high volume, low cost commercial. ***RF***

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