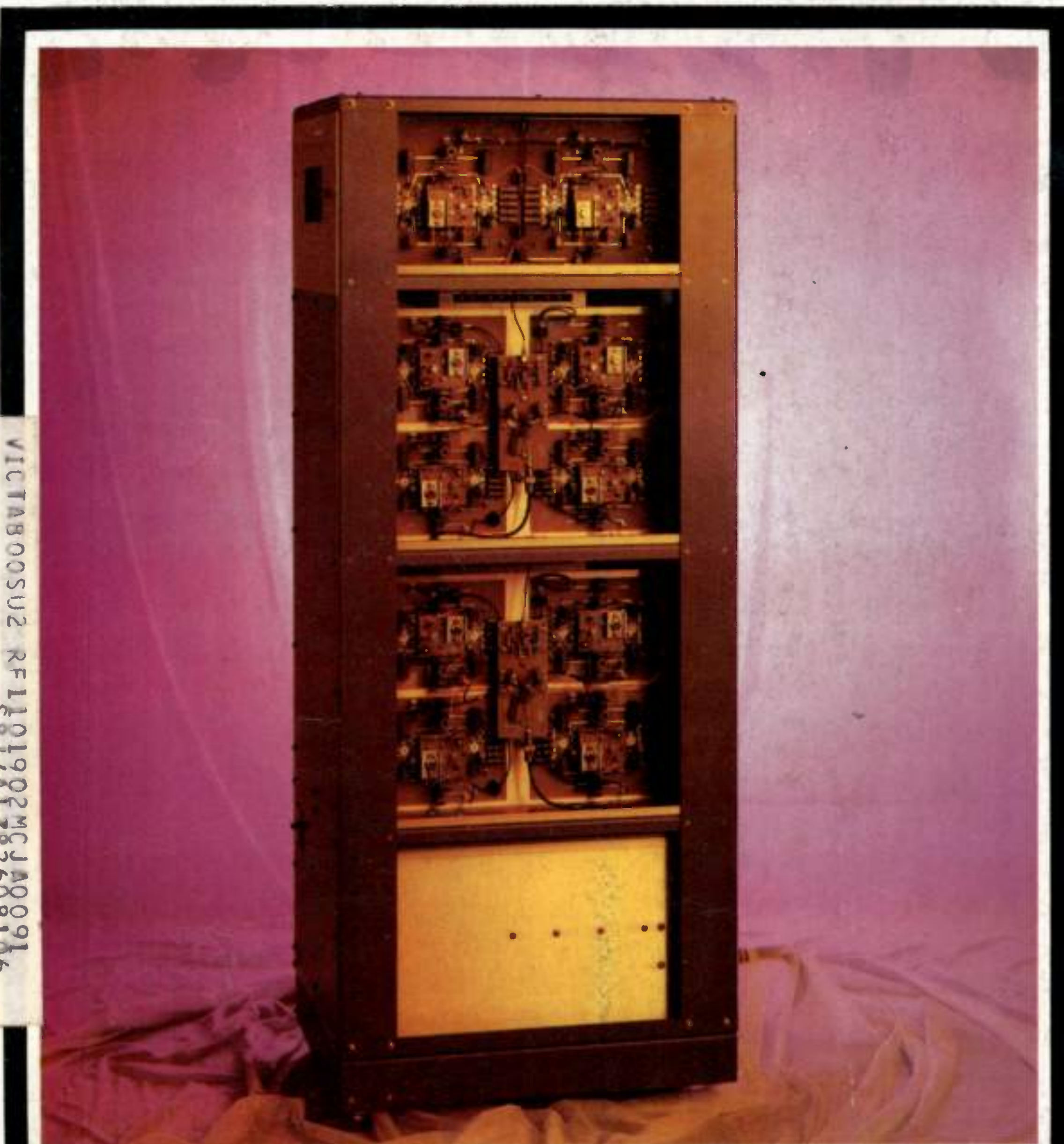
July/August 1981

A Cardiff Publication

rf design



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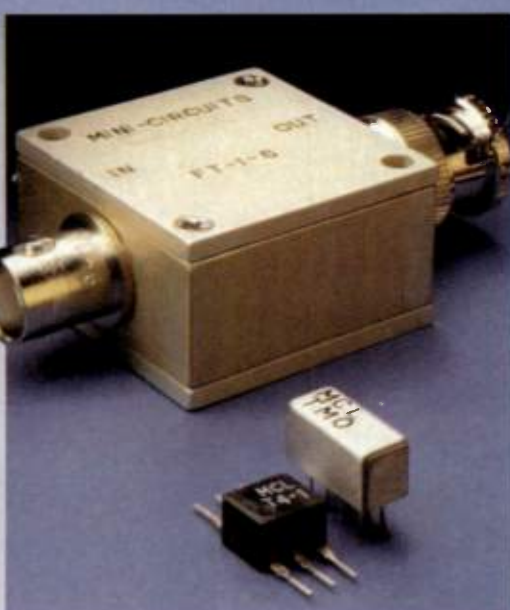

Select from the economical, microminiature T-series (plastic case) or TMO series (hermetically-sealed metal case) covering 10 KHz to 800 MHz. These models operate from 12.5 to 800 ohms with insertion loss typically less than 0.5 dB.

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Imped. Ratio	1	1	1.5	2.5	4	9	9	16	16
Freq. (MHz)	15-400	8-300	1-300	01-100	02-200	15-200	2-90	3-120	7-85
T Model (10-49)	$2.95	$4.95	$3.95	$3.95	$3.95	$3.45	$5.45	$3.95	$5.95
TMO model (10-49)	$4.95		$6.75	$6.45	$6.45	$6.45		$6.45	

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Model No.	T1-1T TMO1-1T	T2-1T TMO2-1T	T2.5-6T TMO2.5-6T	T3-1T TMO3-1T	T4-1 TMO4-1	T4-1H	T5-1T TMO5-1T	T13-1T TMO13-1T
Imped. Ratio	1	2	2.5	3	4	4	5	13
Freq. (MHz)	05-200	07-200	01-100	05-250	2-350	8-350	3-300	3-120
T Model (10-49)	$3.95	$4.25	$4.25	$3.95	$2.95	$4.95	$4.25	$4.25
TMO model (10-49)	$6.45	$6.75	$6.75	$6.45	$4.95		$6.75	$6.75

UNBALANCED PRIMARY & SECONDARY

Model No.	T2-1 TMO2-1	T3-1 TMO3-1	T4-2 TMO4-2	T8-1 TMO8-1	T14-1 TMO14-1
Imped. Ratio	2	3	4	8	14
Freq. (MHz)	025-600	5-800	2-600	15-250	2-150
T model (10-49)	$3.45	$4.25	$3.45	$3.45	$4.25
TMO Model (10-49)	$5.95	$6.95	$5.95	$5.95	$6.75

FT FTB

Model No.	FT1.5-1	FTB1-1	FTB1-6	FTB1-1-75
Imped. Ratio	1.5	1	1	1
Freq. (MHz)	1-400	2-500	01-200	5-500
(1-4)	$29.95	$29.95	$29.95	$29.95

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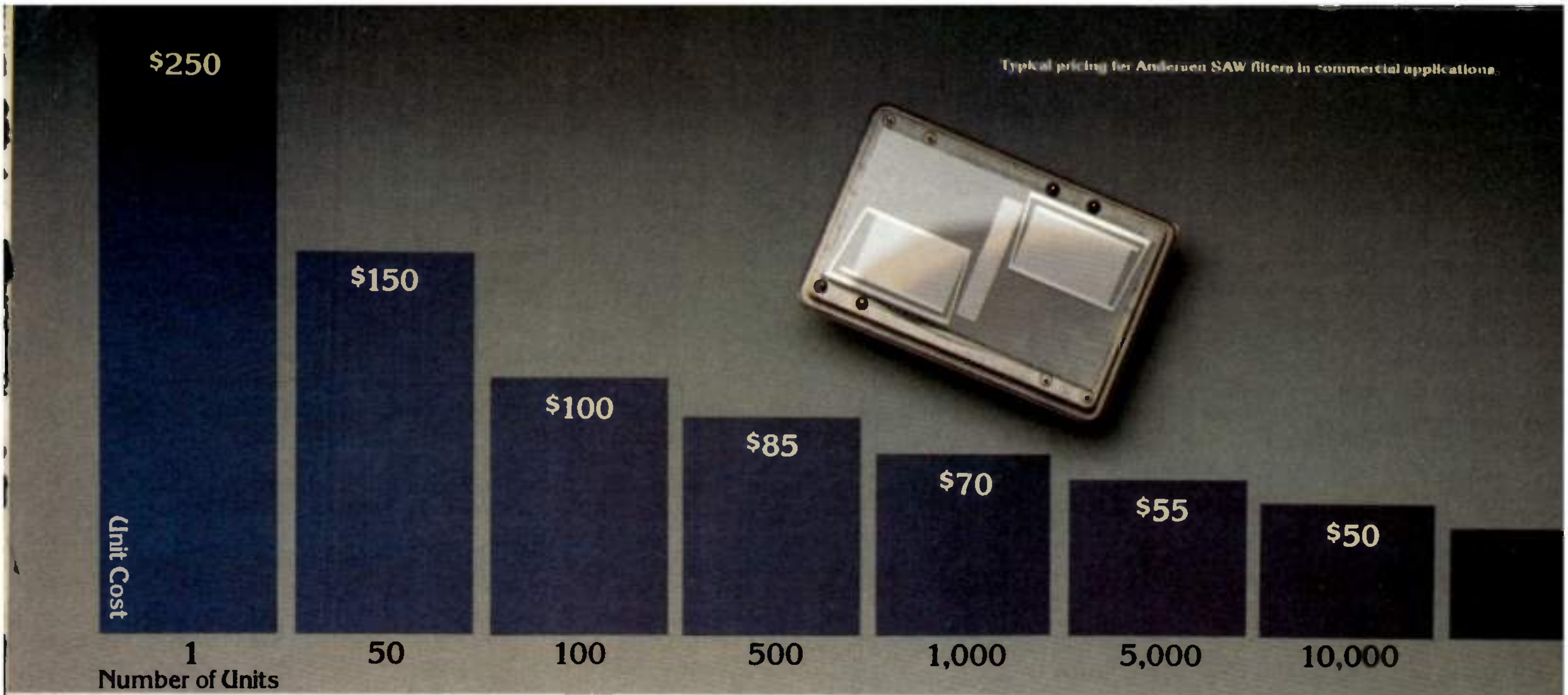
INFO/CARD 1



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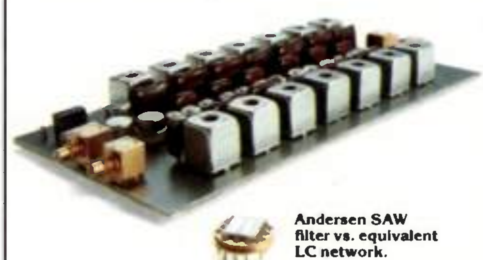

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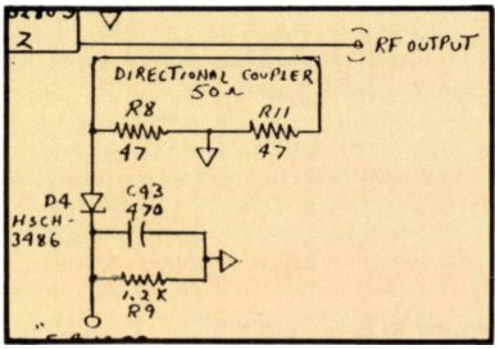
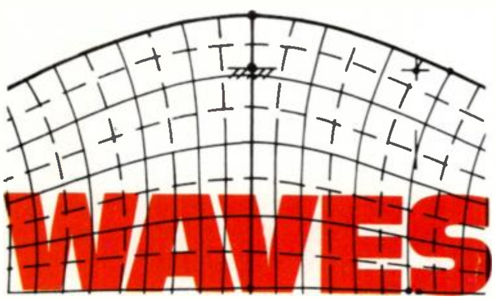
INFO/CARD 2

rf design

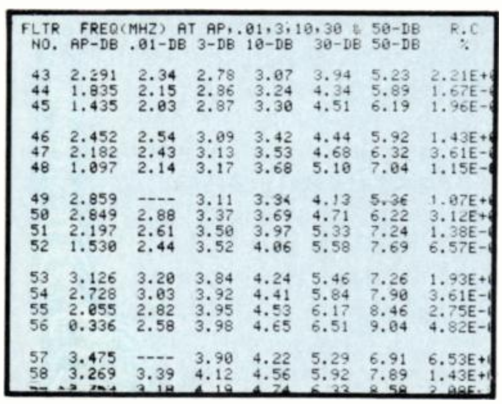
July/August 1981



RF Amp Output Section



SAW Devices



FLTR NO.	FREQ(MHZ) AT AP,.01,3,10,30 & 50-DB AP-DB	.01-DB	3-DB	10-DB	30-DB	50-DB	R.C %
43	2.291	2.34	2.78	3.07	3.94	5.23	2.21E+
44	1.835	2.15	2.86	3.24	4.34	5.89	1.67E-
45	1.435	2.03	2.87	3.30	4.51	6.19	1.96E-
46	2.452	2.54	3.09	3.42	4.44	5.92	1.43E+
47	2.182	2.43	3.13	3.53	4.68	6.32	3.61E-
48	1.097	2.14	3.17	3.68	5.10	7.04	1.15E-
49	2.859	----	3.11	3.34	4.13	5.36	1.07E+
50	2.849	2.88	3.37	3.69	4.71	6.22	3.12E+
51	2.197	2.61	3.50	3.97	5.33	7.24	1.38E-
52	1.530	2.44	3.52	4.06	5.58	7.69	6.57E-
53	3.126	3.20	3.84	4.24	5.46	7.26	1.93E+
54	2.728	3.03	3.92	4.41	5.84	7.90	3.61E-
55	2.055	2.82	3.95	4.53	6.17	8.46	2.75E-
56	0.336	2.58	3.98	4.65	6.51	9.04	4.82E-
57	3.475	----	3.90	4.22	5.29	6.91	6.53E+
58	3.269	3.39	4.12	4.56	5.92	7.89	1.43E+

LPF Table of Values

July/August 1981. Volume 4, No. 4. **r.f. design** (ISSN 0163-321X) is published bi-monthly by Cardiff Publishing Company, a subsidiary of Cardiff Communications, Inc., 3900 S. Wadsworth Blvd., Denver, Colo. 80235 (303) 988-4670. Controlled circulation postage paid at Denver, Colorado. Please address subscription correspondence and Postmaster, please send PS form 3579 to P.O. Box 1077, Skokie, Ill. 60077. Subscriptions: Domestic $10, Canada & Mexico $15 per year; foreign $20 per year. Please make payment in U.S. funds only. Single copies available at $3 each.

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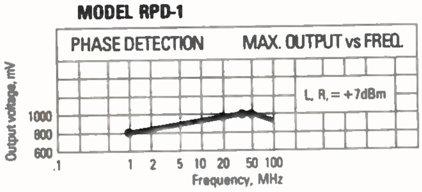
MODEL RPD-1



RPD-1 SPECIFICATIONS

FREQUENCY RANGE:	
L and R ports	1-100 MHz
Output ports	DC-50 MHz
SCALE FACTOR	8 mV/Degree
IMPEDANCE	
L and R ports	50 ohms
I port	500 ohms
L and R SIGNAL LEVELS	+7 dBm
ISOLATION, L-R	40 dB min
MAXIMUM DC OUTPUT, mV	1000 mV typ 750 mV min
DC OUTPUT POLARITY (L and R in-phase)	Negative
DC OUTPUT OFFSET VOLTAGE	0.2 mV typ 1 mV max
FIGURE-OF-MERIT, M	143 Typical

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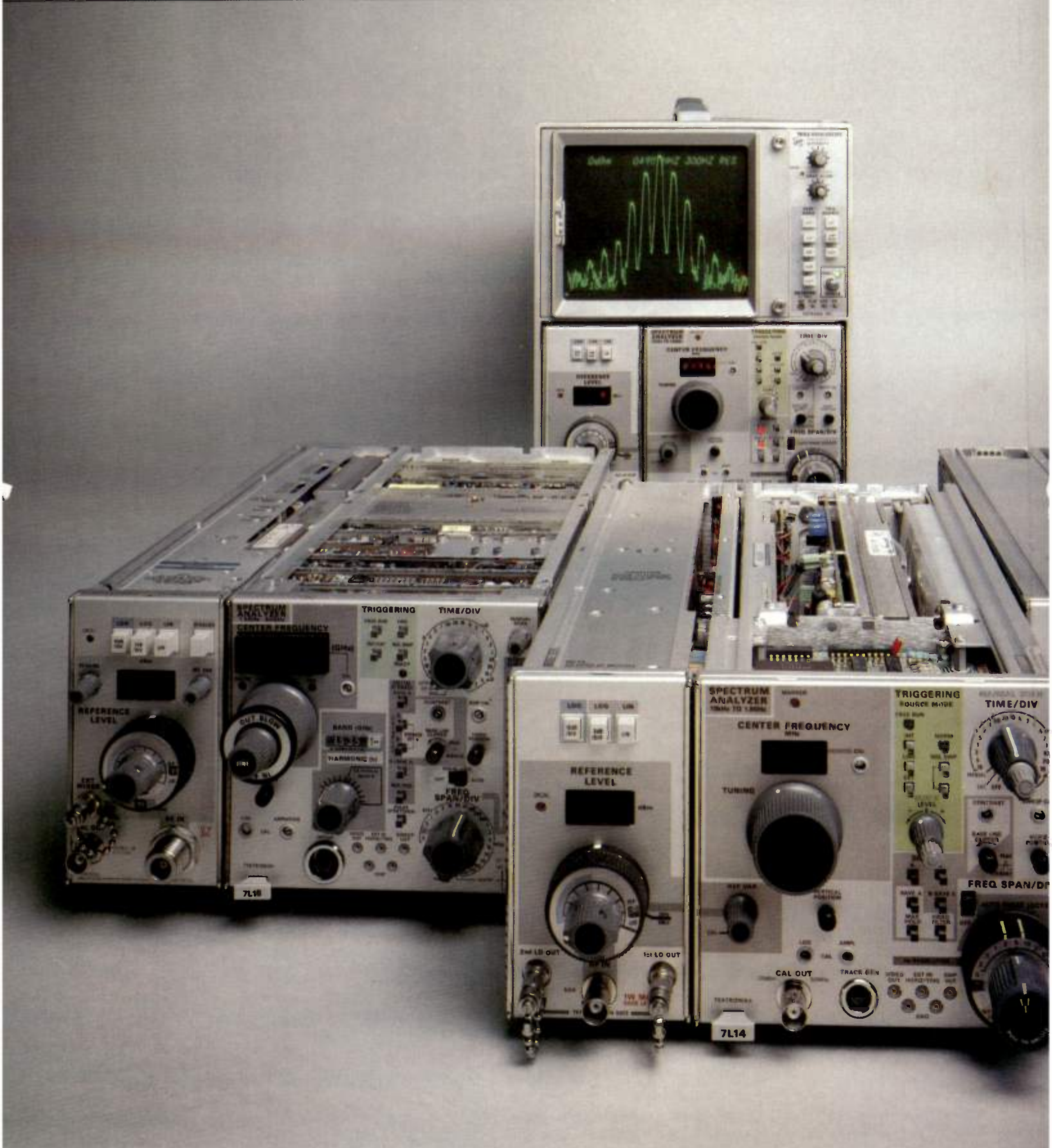

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A 'Slight' Change

Editorial calendars for the year are usually generated well in advance with the intent in mind of alerting the readership and advertisers of upcoming articles and events.

r.f. design has experienced rapid growth over this past year and is still in the process of changing and experimenting. This editor, believing that he feels the pulse — oops, pardon — complex waveform of the readership, continues to try to provide articles that are pertinent and useful. Consequently the lead-off article for this issue is a 1000 watt VHF linear amplifier. It is a detailed design presentation which includes a technique entitled "Envelope Corrected Feedback." It has supplanted, at least temporarily, an article on "Data Transmission."

Jeff Schoenwald of Rockville Space Center has provided *r.f. design* with a second manuscript on SAW devices covering the Interdigital Transducer. It delves into the equivalent circuit of a SAW transponder, its frequency response, delay-line application and parasitics. It too has supplanted a a tentatively-scheduled article on pin diodes.

It is my belief that filters and their subsequent design represent a large percentage of an RF engineer's time. Consequently this is the second issue this year to have a filter article. The tabulated format should help even the beginner in choosing the closest design to meet his requirements without the need of lengthy calculations or computer time.

Lastly, Messrs. Blumkin and Schabauer's article on Ceramic and Porcelain Multilayer Capacitors concludes in this issue.

The other "slight" change involves a buyer's guide that was promo'd for the last issue of this year and will be a little delayed. To the hundreds who have so enthusiastically replied please accept my warm thanks. Literally all your cards have been read and comments noted. Gentlemen, I know how busy you are and what a pain in the butt it is to stop what you're doing and fill out "another" card. But please realize that there are about 49,000 buyer's guide cards still floating around out there that want to come home to me. Please flood these offices with them. (Fill them out first please.)

Lastly, about the subscription card. There's been a "wee" bit of confusion about it. If you have received *r.f. design* in the past and wish to continue receiving it just check the "yes" box and send it in. Period! For all *new* subscribers the cost is $10/year and we appreciate your help.

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Thanks,

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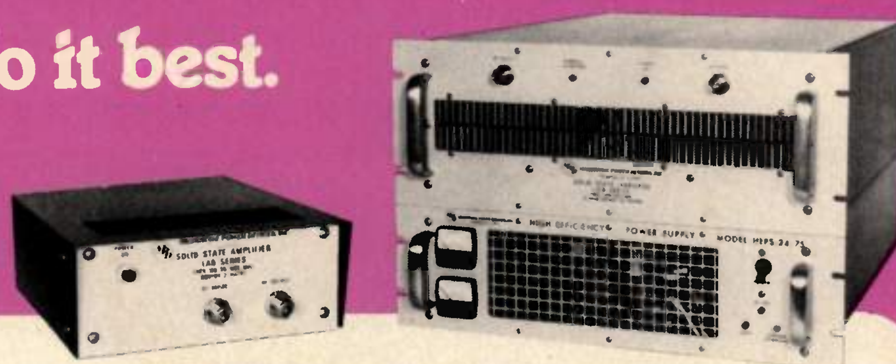

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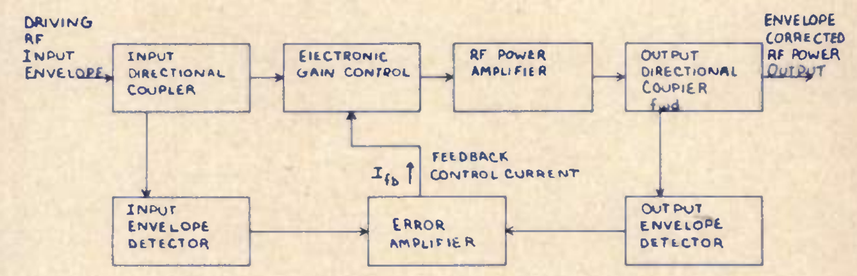

Figure 1. Basic Envelope Correction Feedback (ECFB) System.

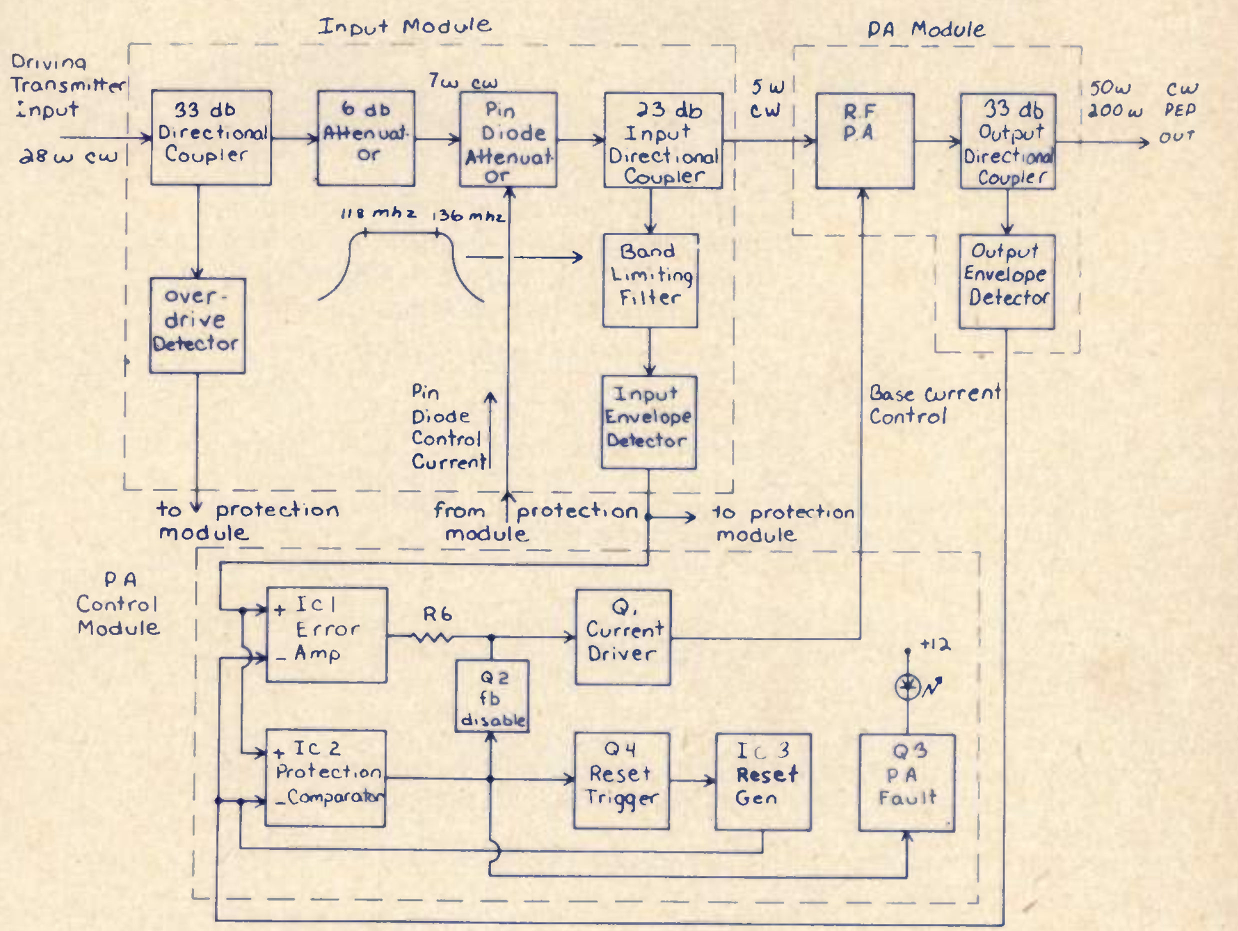

Figure 2. Detailed ECFB System.

Linear Amplifier

Extended range requirements in the VHF communications band have made high power solid-state transmitters a product line necessity rather than a laboratory curiosity. The most difficult part of such a development is to achieve a product design which is consistent in production and is field maintainable by the average technician. It is the purpose of this article to present one such design.

Richard W. Brounley, P.E.
RF Consulting Engineer
1414 Madison St.
Hollywood, FL 33019

Basic Design Considerations and Goals

The transmitter is designed for the Aeronautical Communications air to ground service which operates in the 118 MHz to 136 MHz band. The transmitter as part of a ground station with AM service must deliver 250 watts of carrier. Because of the many existing AM transmitters of power levels between 5 and 50 watts, a linear power amplifier design approach was undertaken. Customers could then have the option of adding the linear to their existing transmitters, resulting in substantial savings over purchasing a complete unit. In addition, the self-contained driving transmitter can be used alone by bypassing the linear amplifier in case of failure or for conditions where less coverage is required.

Basic design criteria discussed below are considered necessary to meet the requirements of the desired market:

Output Power — 300 Watt Carrier with 90 percent modulation capability, (1100 watts PEP) in order to deliver

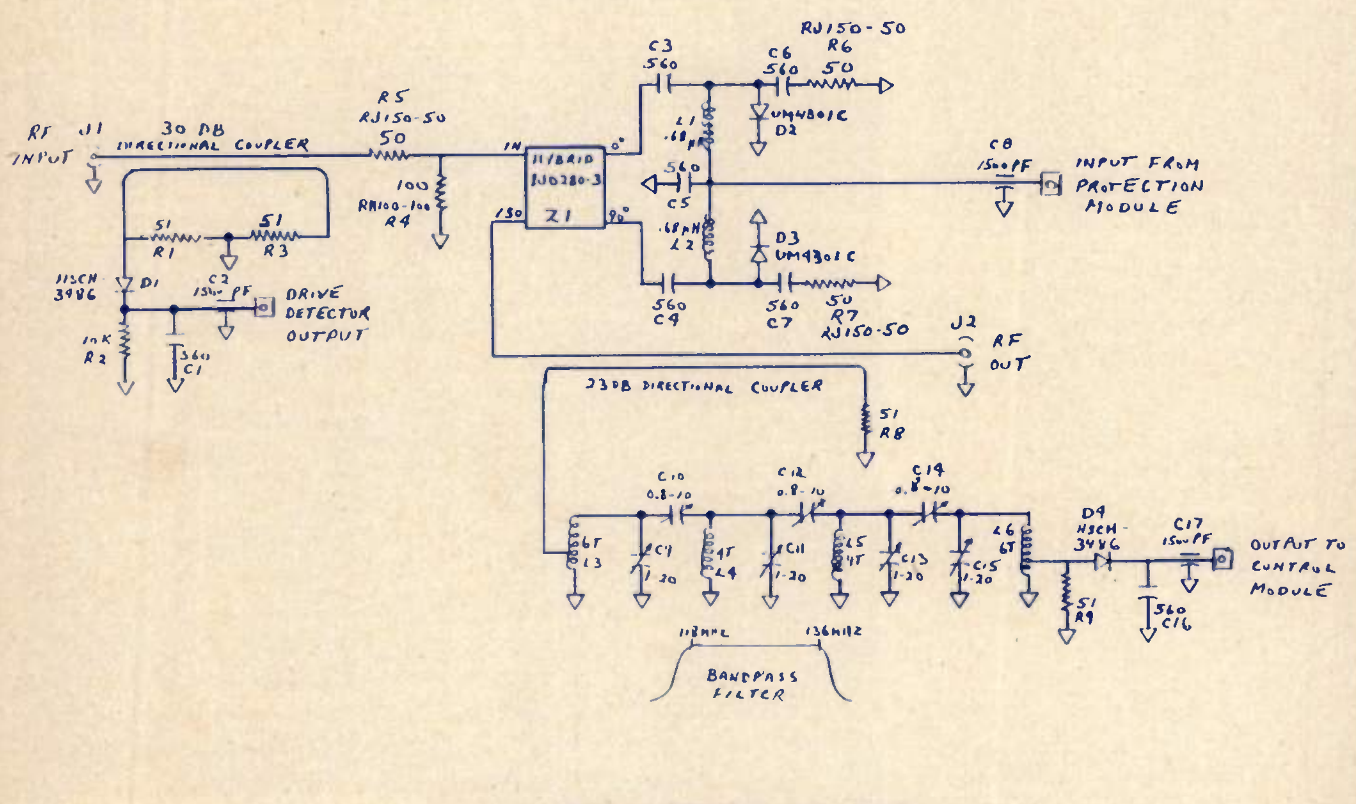

Figure 3. RF Input Module Schematic.

250 watts minimum under all conditions and production tolerances.

Bandwidth — Broadbanded over the 118 to 136 MHz frequency range without adjustment.

Envelope Distortion — Within 1 percent of the driving transmitter RF envelope.

Gain — 18 dB in order to operate with a minimum of 5 watts of drive. Adjustable to a maximum of 50 watts of drive.

Power Supply Requirements — +28 VDC nominal. Because of the complexity of regulators at this power level, their minimum use was considered important.

Protection — Automatic protection against excessive VSWR, temperature, drive power, and line voltage.

Monitoring — Sufficient monitoring to enable on site adjustments and identification of system failure modes without the use of complex test equipment.

Production Efficiency — The design must result in a system of fabrication and test which, when assembled as a complete transmitter, is free from "tweaking" and other time consuming adjustments. Unless this criteria can be met, the complexity of such a transmitter design will prevent the realization of a successful product.

Envelope Corrected Feedback (ECFB)

Since the linear amplifier is to be driven by existing transmitters, a technique had to be developed which would produce low envelope distortion independent of the driving source. This ruled out commonly used methods whereby the output envelope is corrected by feedback to the low level stages of the transmitter. It was decided to design the Linear Amplifier using Envelope Correction Feedback.

The basic Envelope Correction Feedback (ECFB) scheme is shown in Figure 1. The driving RF Input Envelope is fed to the Input Directional Coupler where a portion of the forward power is sampled. The Input Envelope Detector recovers the RF envelope waveform and connects it to the Error Amplifier.

The output of the Error Amplifier connects to the input of the Electronic Gain Control which modifies the drive power to the RF Power Amplifier in accordance with the feedback control current from the Error Amplifier. The RF Power Amplifier connects to the Output Directional Coupler and then on to the antenna or other external circuitry. The Output Directional Coupler samples a portion of the forward power (which is detected in the Output Envelope Detector). The detected output envelope is then compared with the detected input envelope in the Error Amplifier. The error between the two detected envelopes is amplified and modifies the gain or loss of the Electronic Gain Control Circuit.

The error between the input and output envelope is a function of the gain of the feedback loop. If 20 dB of feedback is used, the error will be reduced by a factor of 10:1. The ECFB scheme is only effective within the bandwidth of the feedback loop and requires the normal gain and phase shift considerations. The linear amplifier feedback loop has sufficient gain so that the output envelope essentially duplicates the input envelope over approximately a 10 kHz bandwidth.

The one problem in this basic scheme is in implementing the Electronic Gain Control function. Pin-diode modulators suffer from envelope distortion if very much control range is required as well as requiring an initial insertion loss to make the envelope correction bidirectional. VMOS power FET's currently offer good gain control characteristics and would be worthy of comparison but were not available when the design was initiated. Experimentation revealed that a significant amount of gain control could be achieved in a class C bipolar amplifier by externally controlling the base current as long as the drive power is kept below the saturation level. This is the technique used in the linear amplifier and provides good correction control up to 200 watts PEP output from each module.

A detailed block diagram of the actual ECFB system is shown in Figure 2. Figures 3, 4, and 5 are schematic diagrams of the modules involved.

The driving transmitter* delivers 28 watts of CW power (100 watts PEP at 90 percent modulation) to the Input Module. Overdrive protection is provided by the Protection Module which receives its signal sample from the 33 dB Directional Coupler and Overdrive Detector. (The Protection Module is discussed later.) A 6 dB power Attenuator is used to reduce the input to the Pin Diode Attenuator to 7 watts CW. The Pin Diode Attenuator adjusts the drive power to the PA Module and also serves as the control element for the Protection Module. Schematically, the attenuator uses a 90° hybrid coupler feeding two pin

*This would be a typical case for a nominal 25 watt driving transmitter.

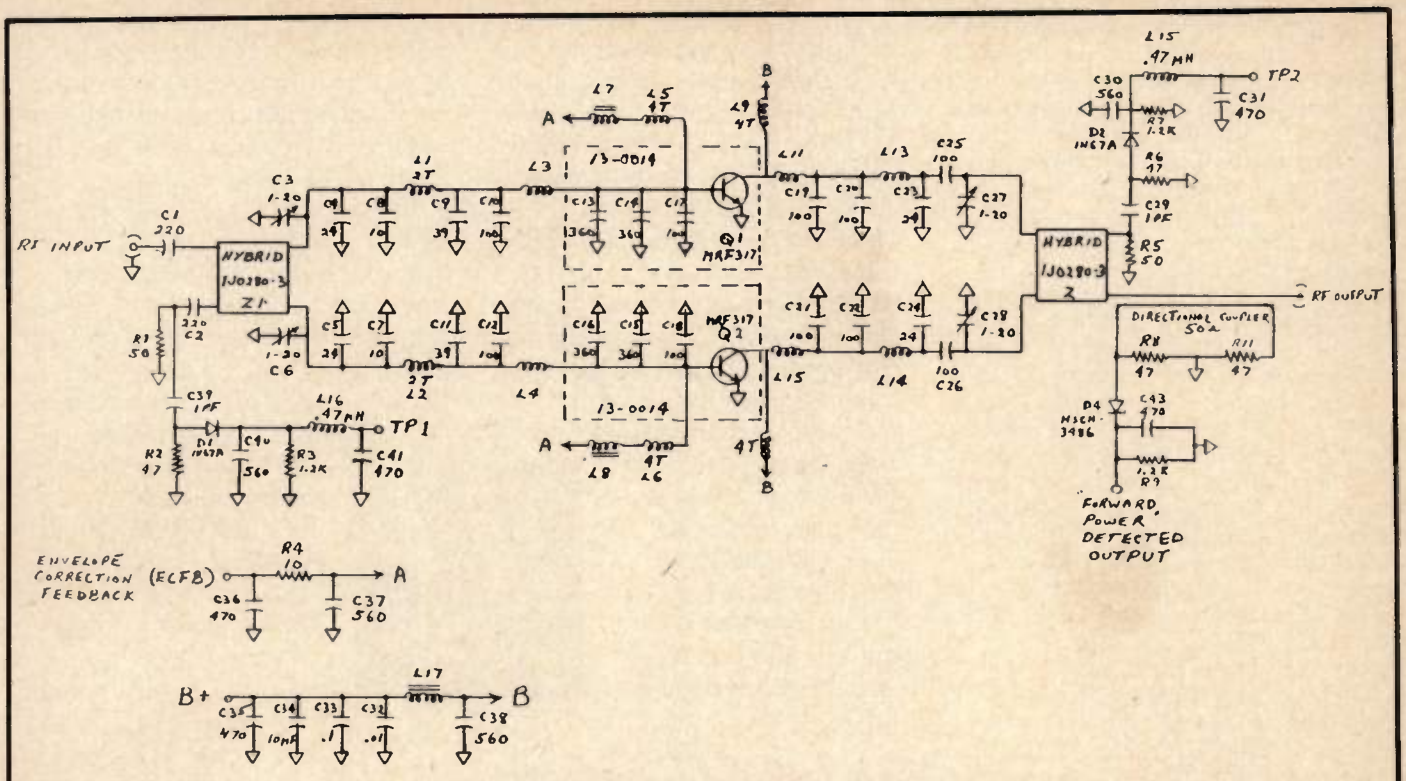

Figure 4. VHF Power Amplifier Module Schematic.

diodes terminated by 50 ohm loads as shown in Figure 3.

The attenuation is maximum when no current is being delivered to the diodes and decreases as the current increases. This type of attenuator presents a constant VSWR to the driving transmitter throughout its range of attenuation since any reflect-

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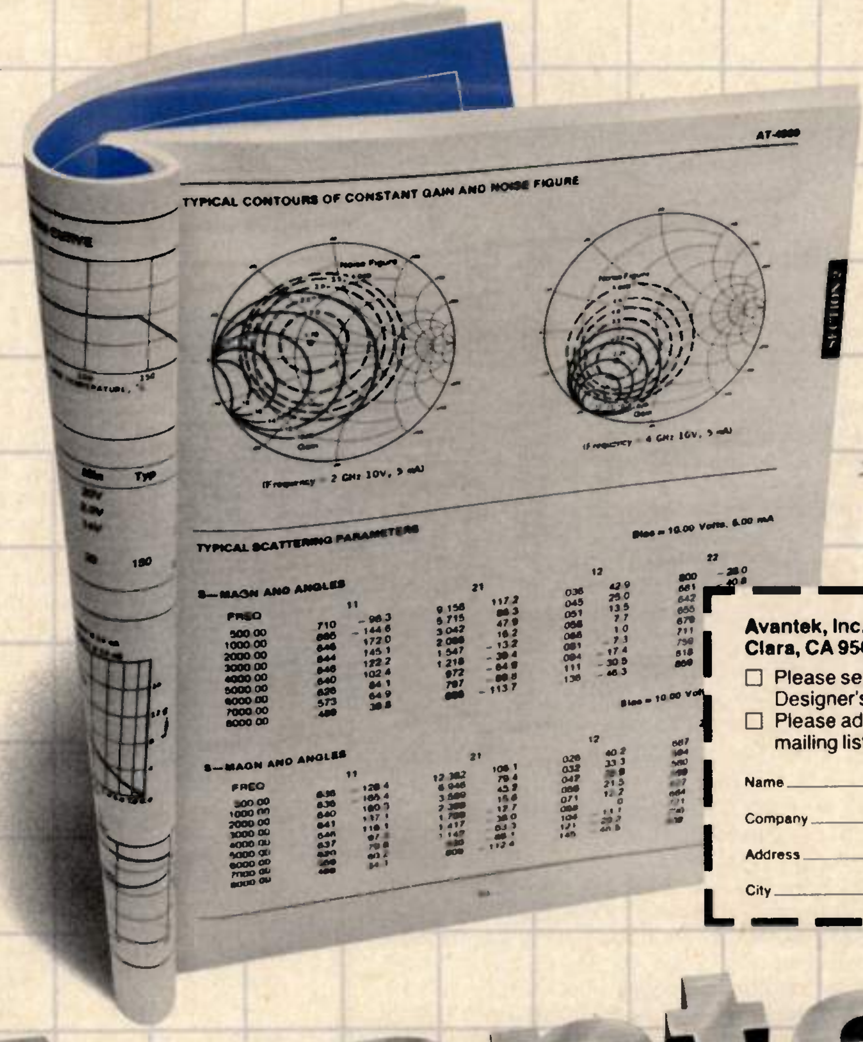

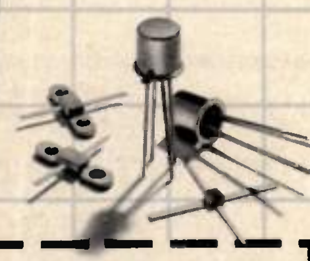

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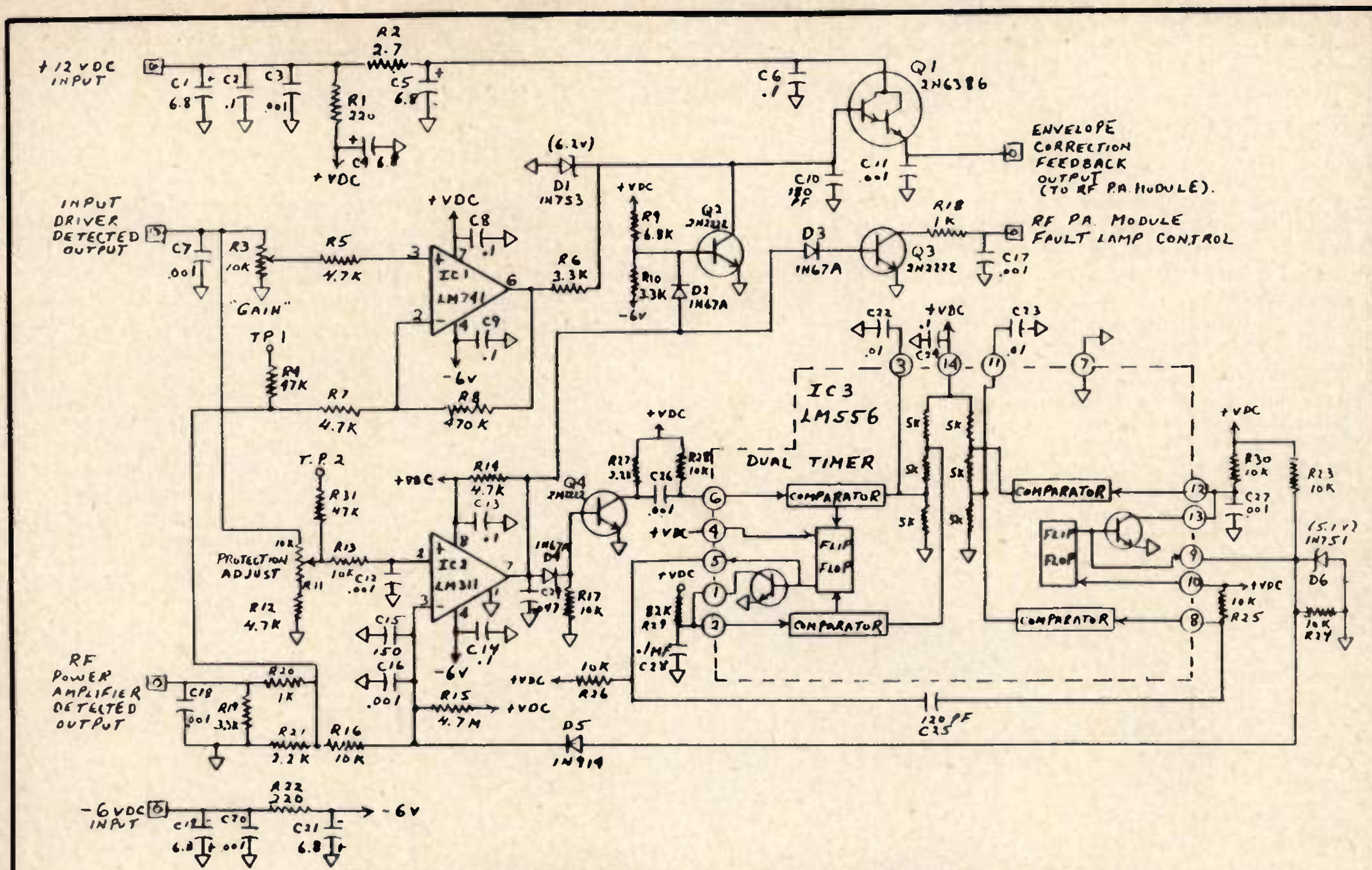

Figure 5. RF Power Amplifier Control Module Schematic.

ed power appears at the isolated port.

The 23 dB Input Directional Coupler delivers approximately 100 milliwatts PEP to the Input Envelope Detector after passing through a Band Limiting Filter. The 100 milliwatts drives the detector over the same dynamic range as the Output Envelope Detector and thereby minimizes distortion errors created by the detectors. The filter has steep skirts outside of the desired passband to prevent the feedback loop from responding to out-of-band signals (which may be emitted by the driving transmitter).

The Input and Output Directional Couplers are both shortened $\lambda/20$ printed circuit board couplers. They exhibit a 1 dB increase in output from 118 to 136 MHz. Since both have the same slope, the PA gain is constant over the band. The $\lambda/20$ coupler results in a length one-fifth of a quarter-wave coupler or a length of approximately 3 inches in the 118 to 136 MHz range.

The RF Power Amplifier module is shown in Figure 4 and is a balanced amplifier using 90° hybrids on the input and output to provide isolation between the two transistors and low input VSWR. The gain controlling element shown in Figure 1 is provided

(Continued on page 21.)

(Continued from page 16.)
by controlling the base current through the current driver Q1 in the Control Module. The gain is adjusted for 10 dB with 5 watts CW input by the gain adjust, R3, in the Control Module. The base current control has sufficient range to correct the non-linear gain characteristics of the PA to essentially duplicate the input envelope up to 200 watts PEP output. The ability to set the gain to 10 dB also greatly aids the combining efficiency when the eight modules are operated together. The output envelope of each of the eight modules is corrected by its own feedback loop referenced to the same input envelope. This technique provides very accurate power addition over the modulation envelope dynamic range and the output distortion is essentially the same as each individual module. The complete transmitter therefore provides the same output distortion as the driving transmitter.

The Output Directional Coupler and Envelope Detector are housed in the PA Module and connect the detected envelope to the error amplifier IC1 in the Control module. Here the difference between the input and output detected envelopes is amplified by IC1. Current driver Q1 then controls the base current to the two PA transistors in a manner which will correct the output envelope to match the input envelope.

ECFB Requires Protection

The control of the PA transistors gain by the base current is a very effective technique from all standpoints but requires protection from any failure mode which causes the feedback loop to lose control. This can be seen by examining the Control Module schematic in Figure 5. IC1 is the error amplifier of Figure 2, and under normal operation, the + and − inputs are virtually identical in magnitude because of the high gain feedback loop. The difference between these two inputs is amplified with a gain of 40 dB by IC1 and is used as the error signal to drive Q1, which in turn controls the base current of the PA transistors.

If the negative input to IC1 should fail for any reason with the positive input of IC1 still being driven, IC1's output will swing to its maximum positive value. Q1 will then turn the bases of the PA transistors fully on. Although the maximum current is limited by zener diode D1, the peak value required by the feedback circuit to provide adequate control is sufficiently high to cause destruction of the PA transistors when applied without feedback. This condition can be caused by relatively simple failures such as the PA input or output cables being disconnected which could happen under field operation. The purpose of the protection circuit is to eliminate this failure mode.

In Figure 5, both detector inputs are connected to comparator IC2. Under normal operation, R11 is adjusted so that the positive input is 3 dB less than the negative input and the output of IC2 is at its lowest potential. Q2, Q3 and Q4 are turned off. The inputs to IC2 are integrated with a cutoff of about 100 kHz permitting the circuit to respond in about 10 microseconds.

When a fault occurs and the negative input drops below the positive, the output of IC2 swings positive, turning on Q2, Q3 and Q4. Q3 turns on the front panel fault indicator diode and Q2 shorts the base of Q1 to ground, thereby disconnecting the feedback loop and preventing damage to the PA transistors. Without feedback, the RF output of the PA module is reduced to the point where it is impossible for the feedback loop to regain control when the originating fault is corrected without some reset method which temporarily disconnects the protection circuit. This is the purpose of Q4 and IC3.

When a fault occurs and Q4 turns on, it triggers a 10 millisecond delay pulse in 1/2 of IC3. At the end of the 10 millisecond delay, a 100 microsecond reset pulse is generated by the other half of IC3. The reset pulse drives the negative input of IC2 positive through diode D5, causing the protection circuit to be deactivated for 100 microseconds. If at the end of the 100 microsecond reset pulse the fault has been removed, the output of IC2 remains in its low state and the feedback loop returns to normal operation. If the fault has not been removed, the output of IC2 swings positive and the cycle is repeated.

The reset cycle continues until the fault is removed. The 100 microsecond reset pulse is short enough so that the peak current drawn by the PA transistors will not cause damage. The 100:1 duty cycle of the reset cycle prevents any damage due to average dissipation in the PA transistors.

The automatic reset circuit was developed because there are a number of transient conditions which may result in shut down of the feedback loop. Some of these are:

1. Sharply-keyed envelopes from the driving transmitter which causes the protection circuit to activate before the feedback loop gains control.
2. Temporary decreases in the power supply input.
3. Temporary high VSWR on the antenna.
4. Temporary overmodulation peaks from the driving transmitter.

Without the automatic reset circuit, these transient faults would have to be reset manually and could present a serious operational problem. Normal setting of the ECFB protection circuit requires a 3 dB decrease in the negative input of IC2 before activation. This greatly reduces the susceptibility to the transient activations above while still providing adequate protection.

ECFB Operates Without Quiescent Bias

It will be noted that no quiescent bias is applied to the PA transistors. This results in a threshold approximately 15-20 dB below the 5 watt input level, below which the feedback loop is not in control and the output is not a linear function of the input.

Since the Linear Amplifier is to be used in an AM application, the modulation will swing plus and minus around the CW output. If the downward modulation extends into the threshold region, it would only be after reaching about 90 percent downward modulation and the resulting distortion would be negligible. No problem during testing has been found because of this characteristic. The only problem which may occur is with driving transmitters with audio limiters which permit excessive negative overmodulation.

The sharply positive going wavefronts of the detected envelopes following the negative modulation peaks may cause activation of the protection circuit. In this case, a portion of the following syllable could be distorted because of the 10 millisecond time delay.

If true linear performance is desired for other applications, the addition of quiescent bias would have to be considered. This would require attention to the normally encountered bias thermal considerations not present in this scheme. Also the threshold reduces idle noise from appearing in the output when not being driven. □

Part 2 Next Installment: "System Description."

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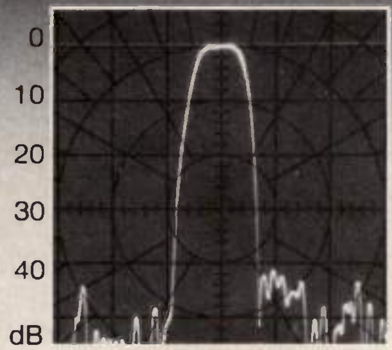

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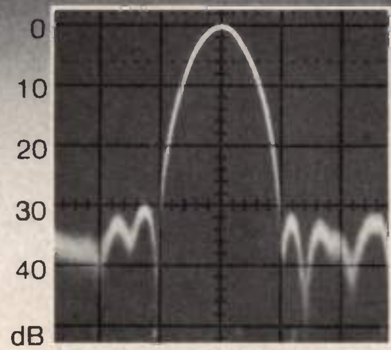

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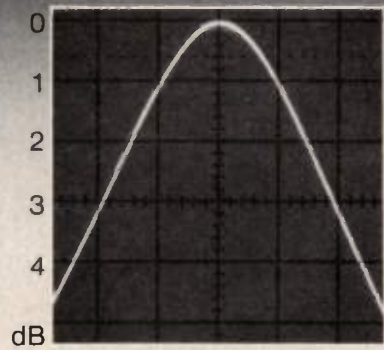

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The Interdigital Transducer

Jeffrey S. Schoenwald
Rockwell International/MRDC
Thousand Oaks, CA

The Equivalent Circuit Of a SAW Transducer

As I was saying in the first installment of this series,* White and Voltmer first demonstrated that a surface acoustic wave (SAW) could be excited on a piezoelectric substrate by means of a transducer consisting of two interdigitated comb-like electrode patterns formed by photolithographic techniques. We refer 'o this interdigital transducer as the IDT (Figure 1). In its original conception, each electrode finger is a quarter-wavelength long in the direction of propagation. Opposing electrode fingers have a quarter-wavelength gap between them and, of course, the period of repetition defines the wavelength. The period or wavelength, λ_0, defined by the dimension of the IDT in turn defines the frequency, f_0, at which a SAW is excited and travels down the substrate at the SAW velocity, V, namely

$$V = f_0 \lambda_0 \qquad (1)$$

Now several factors come into play that we must consider if the RF designer is to make use of this device. Viewed as an electrical component, we need a fair representation of the equivalent circuit of the IDT to understand its frequency behavior. First, regardless of the frequency of operation, the IDT looks like a capacitor. An IDT containing N finger pairs will have a static capacitance $C_0 = NC_{FP}$, where C_{FP} is the capacitance per finger pair. The bonding pads will certainly contribute additional capacitance, and we shall see that it is appropriate to include them with other sources of stray capacitance and not with the capacitance associated with the active portion of the transducer generating the SAW. So, to review, we must include a static capacitance, C_0, that is independent of frequency. Second, if the IDT radiates power at the frequency, f_0, the equivalent circuit must include a resistive component (R) that serves as a load that is frequency dependent and electrically significant only in the range of frequency near f_0. This is known as the *acoustic radiation resistance*, $R_a(\omega)$ Third, since the transducer has a finite number of finger pairs and, therefore, a finite dimension in the direction of propagation, this results in the IDT not having a singular frequency of SAW excitation, but rather a finite bandwidth. (Very shortly, we will derive a simple, theoretical model response of a transducer consisting of N pairs of identical electrodes. In the meantime, we may specify the series equivalent circuits for the IDT as shown in Figure 2a.)

The subscript s stands for series. For completeness, we may also show the parallel equivalent circuit in Figure 2b. The relationships between the components of two equivalent circuits are easily derived, and they are shown in the figure. If we can measure the parallel equivalent circuit elements (as for example, with an R-X meter), we may convert to series equivalent, or if we only have a

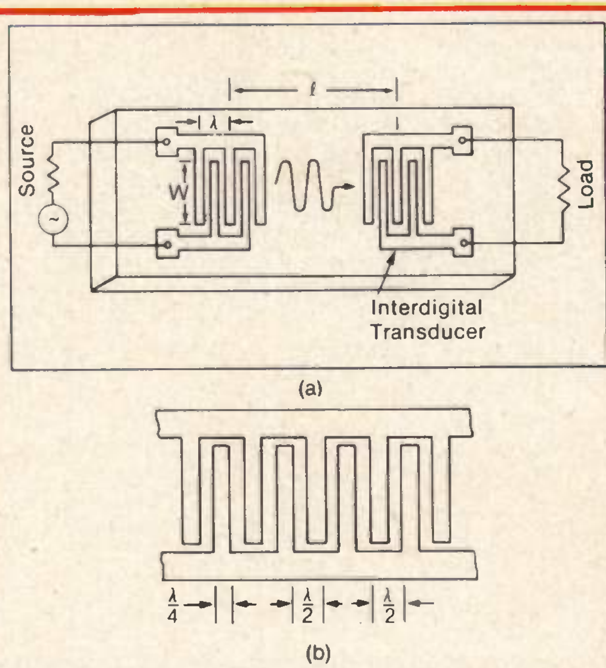
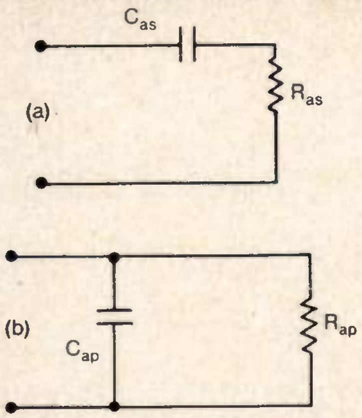


Figure 1. Schematic of an IDT delay line. In this example N = 2.5 for each IDT; in (b) N = 4.5, the aperture is W and each electrode and gap is λ/4 wide in the direction of propagation.

*March/April 1981 *r.f. design*.



$$R_{as} = \frac{R_{ap}}{1 + (\omega\, C_{ap}\, R_{ap})^2}$$

$$C_{as} = \frac{1 + (\omega\, C_{ap}\, R_{ap})^2}{C_{ap}\, \omega^2\, R_{ap}^2}$$

$$C_{ap} = \frac{C_{as}}{1 + (\omega\, C_{as}\, R_{as})^2}$$

$$R_{ap} = \frac{1 + (\omega\, C_{as}\, R_{as})^2}{R_{as}\, \omega^2\, C_{as}^2}$$

Figure 2. A simplified representation of the (a) series and (b) parallel equivalent circuit of the IDT near synchronous frequency f_0.

vector impedance meter, the series equivalent values are measured, and we can convert to parallel equivalent, if necessary. The choice usually depends on one's equivalent, if necessary. The choice usually depends on one's taste. I emphasize that the equivalent circuits in Figure 2 are approximate and relevant at or very close to the synchronous frequency of the transducer. Away from synchronism, R_a changes rapidly. Just how this change occurs will become clearer when the frequency behavior of the IDT is derived from mathematical arguments. Right now let us focus on the factors that determine the value of the acoustic radiation resistance. For this purpose, we will refer to the parallel equivalent circuit.

The admittance, Y, is given by

$$Y = \frac{1}{R_{ap}} + j\omega C_p = G + jB,\ \omega = 2\pi f. \qquad (2)$$

G, the real part of the admittance, is called the conductance. The more power the transducer can convert to surface acoustic wave power, the higher the conductance. What contributes to conductance? For one, the piezoelectric properties of the substrate result in an electromechanical coupling coefficient expressed as K^2. Without piezoelectricity, there is no K^2, no conductance and no SAW.

It is very simple: *G is proportional to* K^2 (an intrinsic property of the substrate, the crystallographic plane of the substrate surface, and the direction of SAW propagation). What else? Consider the geometry: an IDT has N finger pairs. With twice as many finger pairs, the SAW amplitude generated at synchronism is twice as large. *Then G will be proportional to* N^2. This is plausible and may be seen when we consider that a frequency dependent voltage applied across R_{ap} results in power dissipation.

$$P = V^2/R_{ap} = V^2\, G_a \qquad (3)$$

If the acoustic power goes as the square of the amplitude, and amplitude is proportional to N, the dissipated power is proportional to N^2, and hence so is G. The actual expression for G_0, *the conductance at synchronism* ($f = f_0$), is given by

$$G_0 = 8K^2 C_{FP} f_0 N^2. \qquad (4)$$

The factor, C_{FP}, is the static capacitance of one finger pair and takes into account the aperture of the device, which we label W. C_{FP} is proportional to W and the capacitance per unit length for one finger pair, which is determined by the dielectric constant of the substrate.

The Frequency Response Of an N Pair Transducer

Let us look at a transducer with N finger pairs of uniform aperture W in a mathematically abstract way. Let each electrode have an amplitude of unity and sign + or − depending on which pad it comes from. The position of each electrode (Figure 1), taking the first one as being located at X = 0, has a position

$$X_n = \frac{\lambda_0}{2} n \qquad n = 0, 1, 2, 3, 4, 5 \ldots, 2N-1 \qquad (5)$$

If we define the frequency dependent wave vector, k(f) of a SAW, as $k = 2\pi/\lambda$, we can define a function U

$$U = e^{jkx_n} = e^{j\pi n \lambda_0/\lambda} \qquad (6)$$

Note that $f_0\lambda_0 = V = f\lambda$ is the SAW velocity which is independent of frequency. Then we may substitute

$$U = e^{jk(f)x_n} = e^{j\pi n f/f_0} \qquad (7)$$

When $f = f_0$, i.e., synchronism,

$$U = e^{j\pi n} = +1 \quad n = 0, 2, 4 \ldots, 2N-2 \qquad (8)$$
$$= -1 \quad n = 1, 3, 5 \ldots, 2N-1$$

U can then represent the amplitude of each electrode, positioned at the center of each electrode and having amplitude ±1. We may think of this amplitude as the strength of exciting a SAW when a voltage is applied to the electrode or as a detection sensitivity when a SAW passed under it. The laws of reciprocity say that the transducer efficiency (for detection or excitation) is the same whether we go from electric to acoustic or acoustic to electric. Then, picture a continuous SAW passing under an IDT N wavelength long. Any particular crest in the wave train moves with the velocity, V, and can be found at a position $X = VT + X_0$, where X_0 is an arbitrarily defined position when T = 0. It is convenient to let $X_0 = 0$ at T = 0 when the crest of any particular wave passes under the first electrode. If τ is the time it takes for that crest to pass from one electrode to the next, we have

$$X_n = Vn\tau = \frac{\lambda_0}{2} n, \qquad (9)$$

since the spacing $V\tau$ is a half λ, the distance between adjacent electrode centers.

Let us say that we have some rectangular antenna in two dimensions (the substrate, naturally) with a constant aperture W (of no immediate importance) and extent in the direction of wave propagation equal to $N\lambda_0$ (see Figure 1). This physical strength of each detecting element, i.e., each electrode, within the array is constant, because the aperture is constant, and we arbitrarily define this strength as unity. We can renormalize the final results later to satisfy the electrical properties that we discussed before.

If we want to determine the frequency response of this array to SAW's, we must take the inverse fourier transform of the time domain (which is equivalent to the spacial domain) of the transducer array. Given that the time domain

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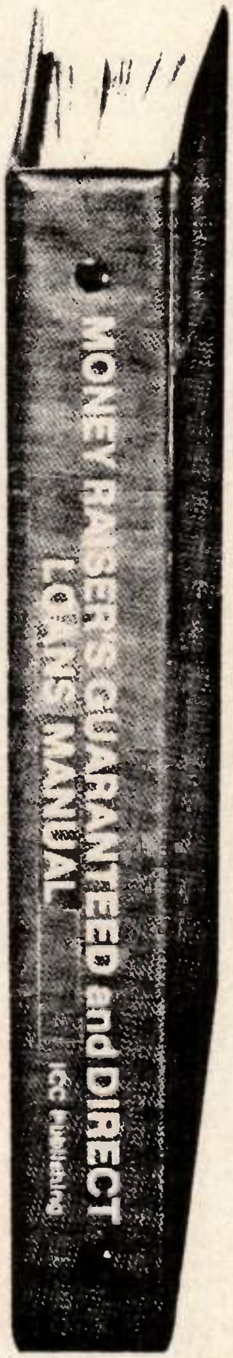
"...don't even try, it's just impossible — all those Business Loans Programs are strictly for the Chryslers, the Lockheeds, the big corporations ...not for the little guy or small companies ...etc."



Still there are those who declare:

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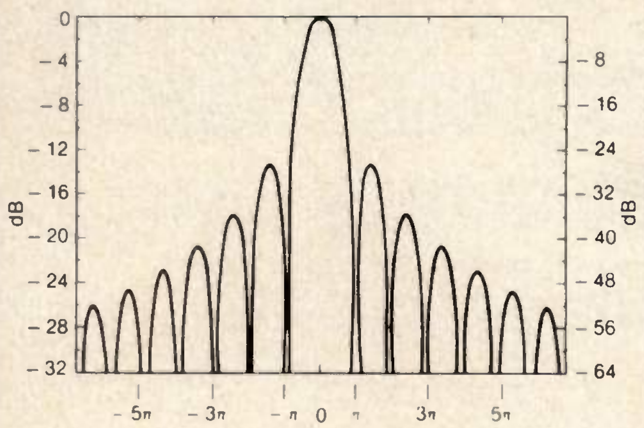


Figure 3. Power vs. frequency response of a single IDT (left scale) and a delay line consisting of two identical IDTs (right scale). Frequency is expressed in terms of the normalized quantity $X = N\pi(f - f_0)/f_0$.

extends from $t = 0$ to $t = T = N\lambda_0$, and the function value is + 1 or − 1 every half λ_0 and 0 elsewhere, we have

$$A(f) = \frac{1}{T}\int_{t=0}^{t=T} e^{i\omega_0 t} e^{-i\omega t}\,dt \tag{10}$$

$$= \frac{1}{T}\int e^{i(\omega_0 - \omega)t}\,dt$$

$$= \frac{1}{i(\omega_0 - \omega)T}\left[e^{i(\omega_0 - \omega)t}\right]_0^T$$

$$= \frac{1}{i(\omega_0 - \omega)T}\left[e^{i(\omega_0 - \omega)T} - 1\right]$$

$$= \frac{e^{i(\omega_0 - \omega)T}}{2}\;\frac{e^{i(\omega_0 - \omega)\frac{T}{2}} - e^{-i(\omega_0 - \omega)\frac{T}{2}}}{i(\omega_0 - \omega)T}$$

$$A(f) = e^{i(\omega_0 - \omega)\frac{T}{2}}\;\frac{\sin(\omega_0 - \omega)\frac{T}{2}}{(\omega_0 - \omega)\frac{T}{2}}$$

Recall that $T = N\lambda_0/V = N/f_0$. Then A (f) may be defined as

$$A(f) = e^{i(f_0 - f)N\pi/f_0}\;\frac{\sin \pi N\dfrac{(f_0 - f)}{f_0}}{\pi N\dfrac{(f_0 - f)}{f_0}} \tag{11}$$

A (f) is the amplitude response of the IDT and is complex, as the exponential factor indicates. The power that the IDT detects depends upon the square of the magnitude.

$$|A(f)|^2 = \frac{\sin^2 x}{x^2},\; x = \frac{N\pi(f_0 - f)}{f_0} \tag{12}$$

This is the celebrated sin x/x dependence that an unweighted (uniform electrode aperture) SAW IDT possesses.

Up to this point, we have assumed that the amplitude of the SAW incident on the IDT was constant amplitude, regardless of frequency. If that were so, then we would observe a power versus frequency response like the one shown in Figure 3. Whenever $x = \pm\pi$, $\pm 2\pi$, 3π, etc., there is null in the transducer response. Similarly, whenever x is nearly equal to $\pm 3\pi/2$, $\pm 5\pi/2$, etc., a local peak (sidelobe) occurs in the power. At $x = 0$ ($f = f_0$), the value of $|A(f_0)|$ is unity. Then $x = \pi/2$, we have

$$|A(f)|^2 = \left(\frac{\sin 3\pi/2}{3\pi/2}\right)^2 = \left(\frac{-1}{3\pi/2}\right)^2 = \frac{4}{9\pi^2} \tag{13}$$

Expressing the ratio of the power at this power of x to the power at center frequency (x = 0) in the familiar logarithmic form yields

$$10\log\frac{|A(f)|^2}{|A(f_0)|^2} = 10\log\left(\frac{4}{9\pi^2}\right) - 10\log 1$$

$$= -13.45\,dB \tag{14}$$

This is the magnitude of the power in the first sidelobe relative to the power at the center frequency of the transducer response.

Any filter or delay line, however, consists of two IDT's. The filter power response is the product of the power responses of the individual transducers:

$$P(f) \;\propto\; |A_1(f)|^2 \cdot |A_2(f)|^2 \tag{15}$$

If the transducers are identical, the first sidelobe in the frequency response is 27 dB below the main lobe (Figure 3). In real life, matters turn out to be very much like the simple model derived above. If we make a delay line with two transducers having different numbers of finger pairs, the response is a bit more complicated, but easy to model from the above analysis.

The IDT as a Frequency Dependent Element

In the previous section, we dealt with the IDT as an electrical element with an equivalent circuit. For simplicity, we restricted our interest to the synchronous frequency f_0 in order to focus on the main concept that a SAW IDT could be modeled (approximately) by a series or parallel equivalent circuit consisting of a capacitor and a resistor. Now that we understand from mathematical arguments how the acoustic power a transducer detects or radiates varies with frequency, we may include this frequency dependence in the model.

In the parallel equivalent, the conductance is G_0 at center frequency, but varies with frequency as

$$G(f) = G_0\frac{\sin x}{x},\; x = \frac{N\pi(f - f_0)}{f_0} \tag{16}$$

This conductance may also be referred to as an equivalent parallel resistance

$$R_p(f) = [G(f)]^{-1} = \left[(8K^2N^2C_{FP}f_0)\left(\frac{\sin x}{x}\right)\right]^{-1} \qquad (17)$$

Far from center frequency, R_p (f), the *acoustic radiation resistance* goes to infinity, and thus no power is radiated as a surface wave.

It turns out that bulk waves can be generated by the IDT as well, but a variety of techniques are available to suppress them, and they are not included in the foregoing model.

Keep in mind that the equivalent circuits in Figure 2 are still useful; we have only added frequency dependence to the equivalent resistance that represents the IDT's ability to dissipate power by acoustic transduction.

Suppose then that a voltage source with a characteristic impedance Z_0 (usually equal to 50Ω) drives an IDT, surface waves are excited, and another IDT detects the surface waves and converts the acoustic energy back to electrical, where it is delivered to a load impedance. For convenience, the load is also Z_0. What is the total insertion loss through the whole SAW device? Before we proceed with the details, I will outline the plan.

1. By reciprocity, the efficiency with which an IDT converts electrical power to acoustic is the same for the reverse — acoustic to electric; so the insertion efficiency need only be computed once for one IDT. In that case, I chose to analyze the case for acoustic excitation.

2. SAW IDT is a bidirectional device — acoustic waves emerge from both ends of the IDT. That means if there is only one receiving (output) IDT — to one side of the input IDT — half of the acoustic energy — 3 dB — is lost out the other side. It may seem difficult to accept at first, but by reciprocity another 3 dB will get lost at the output IDT, for a total of 6 dB. The 3 dB loss a. each transducer has to do with the fact that the SAW IDT is a 3 port device — one electrical and two acoustic — and we are following the acoustic signal into or out of only one acoustic port. When all is said and done, we add up the insertion loss for acoustic excitation, bidirectional losses, and output reconversion.

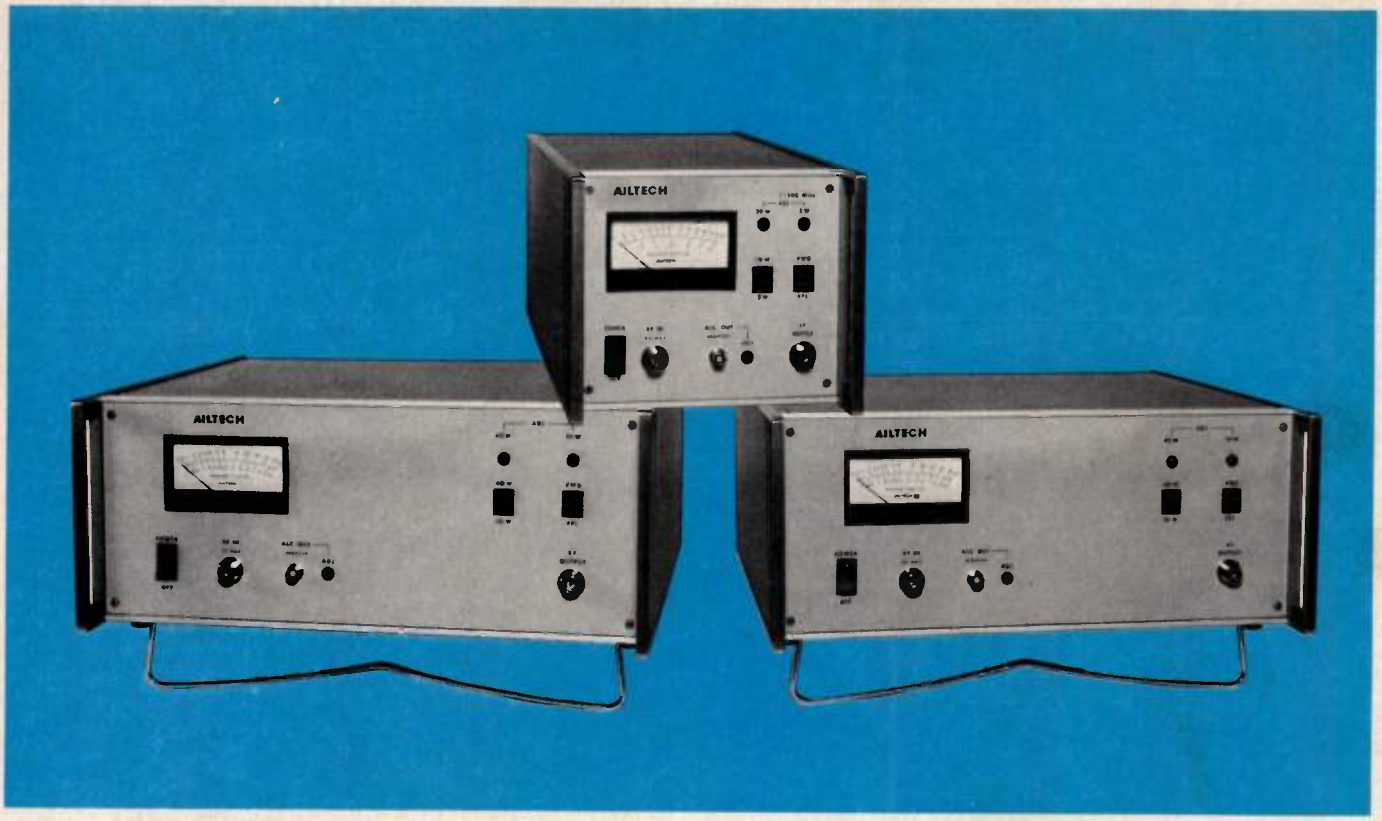
Referring to Figures 1 and 2 we will use the parallel equivalent circuit model for the IDT. The insertion loss is defined as the power dissipated by the IDT relative to that dissipated in the load conjugate to Z_0 — which would be Z_0^*, *the complex conjugate of* Z_0. Since we will assume Z_0 is real and equal to R_0, we see an immediate problem in that an IDT always has a static capacitance C_0. Based on the equivalent circuit, a SAW IDT can never be designed that is a perfect match to a pure real source or load. A capacitor doesn't dissipate energy. All the power that becomes acoustic is dissipated in the parallel resistance R_P which we have also referred to as the parallel equivalent acoustic radiation resistance. To determine how much power is radiated from R_P, we must determine the voltage across it and the current through it. Circuit analysis shows that the ratio of power dissipated in R_P relative to that dissipated in a matched load R_0 is

$$\frac{P(R_P)}{P(R_0)} = \frac{4\,R_S R_0}{R_0^2 + R_S R_P + 2\,R_0 R_s} \qquad (18)$$

where

$$R_S = \frac{R_P}{1 - (\omega C_P R_P)^2} \qquad (19)$$

is the same series equivalent resistance we found for the equivalance relations between the two formulations of the circuit (see Figure 2).



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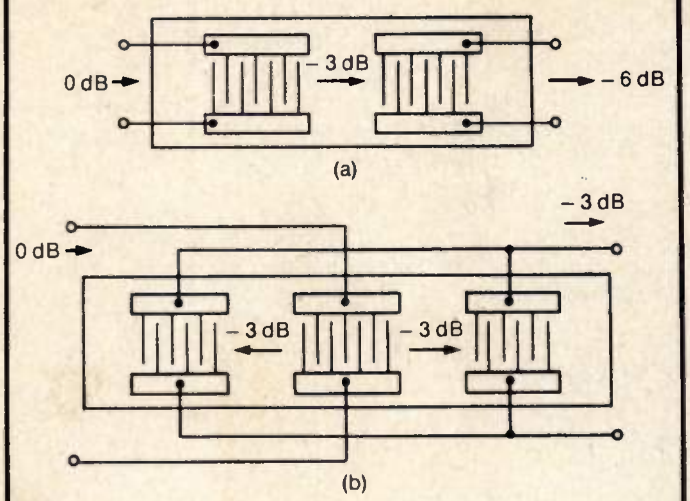

Figure 4. (a) Two IDT delay line. (b) Three IDT delay line.

Example

A measurement made on an R-X meter yields the following results for an IDT that is used in a delay line:

$$f_0 = 100 \text{ MHz}$$
$$R_p = 10^3\,\Omega$$
$$C_p = 5 \text{ pf} = 5 \times 10^{-12} \text{ Farads}$$

This results in

$$R_s = 92\,\Omega$$

Suppose the source and load connected to either IDT are the usual 50 Ω. The power that the 50 Ω source dissipates in R_P relative to a 50 Ω resistor is

$$\frac{P(R_P)}{P(R_0)} = 0.1774$$

In logarithmic terms, the insertion loss per IDT is

$$IL = -10 \log \frac{P(R_P)}{P(R_0)} = 7.51 \text{ dB}$$

Together, two transducers will contribute 15 dB of insertion loss, but half of the energy that makes it onto the substrate goes off in the wrong directions, which causes a 50 percent loss in signal — or another 3 dB. By reciprocity, another 3 dB is not absorbed by the receiving transducer and slips on down the substrate. This adds up to a grand total of 21 dB insertion loss.

If we wish to build a delay line with three equidistant-spaced IDT's (see Figure 4) and drive symmetrically — one input IDT and two output IDT's connected in parallel, or vice versa — we can reduce the total insertion loss by 3 dB to 18 dB, or so it would seem. Actually, by connecting the two outboard IDT's in parallel, R_P is reduced by a factor of 2, but C_P is doubled. This results in a R_s that is half of that for a single IDT. In that case, the power ratio for the two IDT's electrically in parallel is given by

$$\frac{P(R_P)}{P(50)} = 0.30565 \text{ or } IL = 5.5 \text{ dB}$$

The total IL would then be

$$7.5 + 6 + 5.5 = 19\ \text{dB}$$

We still include 6 dB loss for bidirectionability because half of the acoustic power at each of the *two* outer IDT's is still lost. This second method of computing insertion loss for a 3 IDT delay line is analytical and more correct than the hand waving argument given above about an additional IDT being used to capture 50 percent more energy, but the difference is only in error by 1 dB in this example.

Parasitics

Life is not so simple if we cannot ignore other factors that must be considered when specsmanship becomes important. For example, the IDT's are formed photolithographically in an evaporated film usually on the order of 1000 Å thickness of Al. This may lead to a static resistance, on the order of a few ohms in some cases, and must be properly considered as an element in the equivalent circuit. In addition, at high frequencies, lead inductance between the bonding pads and the package leads and the electrical ground can mask the true circuit parameters we are trying to measure at the port of a vector impedance meter, an R-X meter, or network analyzer. The capacitance to ground of the sum bars (pads) to which we make wire bonds was specifically excluded in the formulas for the equivalent circuit C_0, but will never-the-less be measured by any instrument. The best way to measure the pad capacitance is to photolithographically remove the fingers and measure the remaining capacitance — at as low a frequency as possible. This

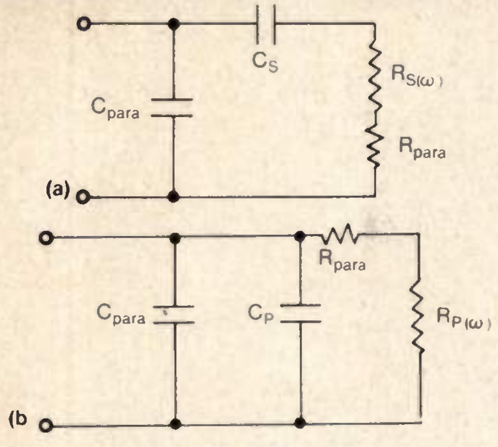

Figure 5.
Series **(a)** and parallel equivalent **(b)** circuit with parasitic pad capacitance and metalization sheet resistance included. The choice of placement of these lumped elements in the circuit is not unique. Other configurations could also be valid.

parasitic capacitance can then be subtracted from the measurement made on a complete IDT, since it is modeled as a parallel element. In Figure 5, we show how these parasitics might be incorporated in the parallel equivalent circuit. Parasitic electrode resistance is indicated in the series equivalent circuit as being in series with R_{ap}, but we could have placed it in parallel to C_0 and R_{ap}, and pad capacitances, in parallel with the rest of the circuit. (The reader can verify that C_0 is very nearly the same in the two versions — parallel or equivalent, but that R_{ap} and R_{as} are very different). One can see that what one obtains on a piece of impedance measurement equipment are not the relevant device parameters. Modeling of parasitics can become a full time job. In fact, describing the process of measuring parasitics can be a full time job, and I will leave a fuller discussion to a subsequent article.

Next Episode: Impedance Matching

Reduction of signal losses are often a critical requirement when the system dynamic range must be maximized against some limiting background noise floor, whether caused by the system, the device itself, or the package in which it comes. One very important technique in the implementation of SAW devices is the use of impedance matching (or transformation) networks between the device and the system in which it is placed. It is quite obvious, then, that the more we know about the equivalent circuit and the parasitic elements that cannot be ignored or eliminated, the more chance of success we have in optimizing the transfer efficiency of the SAW device. I will devote much of the next article to this matter. □

References

[1]H. Matthews, Ed, "Surface Wave Filters," Wiley-Interscience, New York, 1977.
[2]Proceedings, IEEE Ultrasonic Symposium, 1972-Present.

lowpass filters

Table of Precalculated Chebyshev Lowpass Filters with Inductive Input and Output.

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Introduction

Solid-state RF amplifiers usually require output filtering to attenuate their harmonic frequencies to acceptable levels[1,2]. This required filtering can be obtained by connecting a lowpass filter between the amplifier output and its load, and several articles have been published describing filters that are suitable for this purpose[3,4,5,6]. These filters consist of 7 elements (4 capacitors and 3 inductors) in a ladder configuration in which shunt capacitors alternate with series inductors. To simplify construction of these filters, the C input/output configuration was used (as opposed to the alternate L in/out configuration) and all these filter designs required only standard-value capacitors.

Unfortunately, there are situations where a C input filter (with its decreasing input impedance with increasing frequency) causes RF amplifier instability[7,8]. This amplifier instability sometimes can be eliminated by changing the cutoff frequency or the number of filter elements. However, in some situations these two solutions may be inconvenient or impractical. If that is the case, then the L input/output lowpass filter configuration should be considered.

To simplify the design, selection and construction of the L input/output filter, a table of precalculated lowpass designs, similar to those of the C input filter table, would be extremely useful. Although the three capacitors of the 7 element L input/output filter have standard values, the four inductors can have any odd value that the design might require. These odd inductor values present no problem because any required value can be realized by hand-winding iron-powdered toroidal cores using the procedures recently explained by DeMaw[9].

Computer-Calculated Values

A table of L input/output 7 element lowpass Chebyshev filter designs using only standard value capacitors has been computer-calculated using the same technique employed in the tabulation of the C input/output designs. A schematic diagram and typical attenuation response of these filters is shown in Figure 1. Table 1 lists 89 precalculated designs with their 3 dB frequencies covering the 1-10 MHz range. The 3 dB cutoff frequency increments are sufficiently small so that virtually any cutoff frequency can be realized. Designs for other frequency decades are easily scaled by shifting the decimal points in the tabulated data to the right or left. Although the designs are based on equal 50 ohm terminations *any value* of equal impedance termination can be accommodated using a simple scaling procedure *while still maintaining the advantage of standard-value capacitors*[10]. These inherent characteristics permit the table of precalculated designs to serve as a universal filter table for any cutoff frequency or impedance level. The L input/output table, in combination with the previously referenced C input/output table, provides an excellent solution to the design of the 7 element equally-terminated Chebyshev lowpass filter.

Explanation of Table 1

The Table 1 precalculated designs consist of 13 columns of filter data. The first and last columns list identification numbers of each design for convenient reference. Starting from the left of Table 1, the second to seventh columns list those frequencies (in MHz) associated with the attenuation levels (in dB) given in the column headings. Figure 1B depicts a typical attenuation response and shows the relationship between the attenuation levels in the column headings and the tabulated frequencies.

The A_P frequency (second column) denotes the end of the filter passband and the start of the stopband. Above this frequency, the filter input impedance departs from the nominal 50 ohm value, and the attenuation increases above the peak level of the passband ripple. The frequencies at the .01, 3, 10, 30 and 50 dB attenuation levels are listed so that the attenuation response of different designs can be conveniently compared.

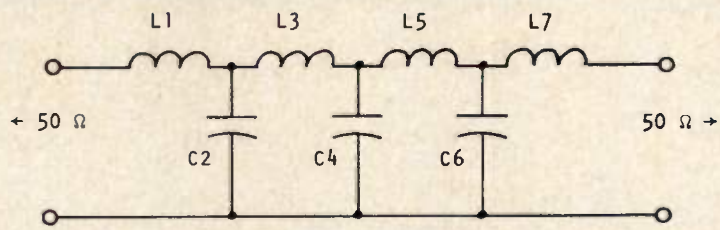
The .01 dB frequency corresponds to a VSWR of 1.10 and a reflection coefficient of 4.796 percent. This

VSWR is an arbitrarily selected value, and it is suggested as a suitable maximum level for RF filtering applications. Knowing this frequency, and assuming that a 1.10 VSWR is acceptable, it is possible to use the .01 dB frequency instead of the F_{Ap} frequency as the amplifier cutoff frequency. This will provide considerably greater second harmonic attenuation in those designs having a very low reflection coefficient. For those four designs having a reflection coefficient greater than 4.8 percent, the .01 dB frequency is not applicable because it is less than the F_{Ap} frequency, (indicated by a dashed line).

The *R.C. (percent) column* lists the reflection coefficient in percent. This is a parameter that is of particular interest to the RF engineer. Other filter parameters, such as the peak passband ripple amplitude (A_P), VSWR, ripple factor (ε) and stopband attenuation (A_s), can be calculated from the reflection coefficient. The equations relating these parameters are given in Appendix A. Because of the wide range of the R.C. values, they are tabulated in an exponential format.

The next four columns in Table 1 list the component values of the capacitors and inductors. The column headings of "C2,6," "C4," "L1,7" and "L3,5" correspond to similarly-labeled components in Figure 1A.

The ratios of C4/C2 are such that most designs have reflection coefficients of less than one percent. This limits the VSWR of these designs to less than 1.02 for all frequencies below the F_{Ap} cutoff frequency. Unfor-



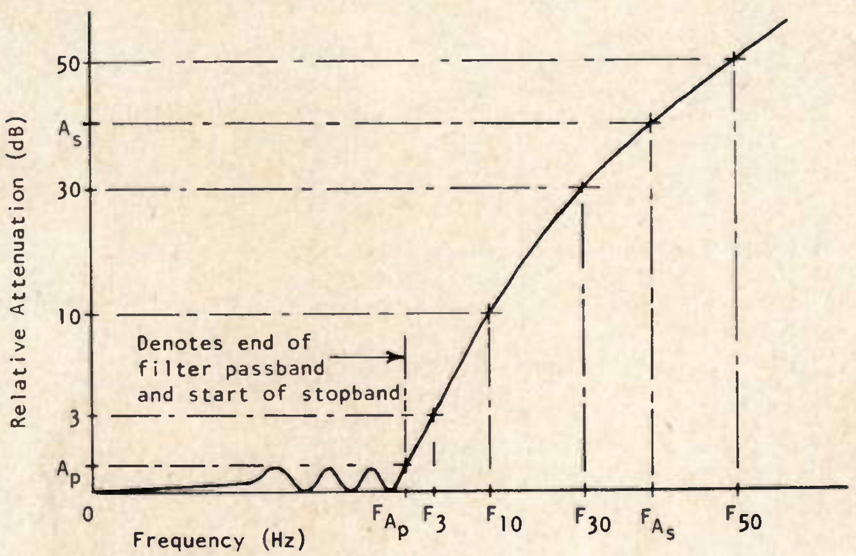
L1 = L7, L3 = L5, C2 = C6

RANGE OF C4/C2 RATIOS USED IN TABLE

1.604, R.C. = 0% (Butterworth)

1.125, R.C. = 10.7%

(A) Filter Schematic Diagram



(B) Typical Attenuation Response

Figure 1. 7-Element Chebyshev Lowpass Filter with Inductive Input and Output.

tunately, the selectivity of all designs becomes progressively poorer as the reflection coefficient becomes smaller. Consequently, when an amplifier second harmonic must be attenuated by more than 30 dB, those designs with reflection coefficients less than about 0.6 percent are not recommended for RF amplifier filtering because of their poor selectivity. (Nevertheless, perhaps a filter design with a very low reflection coefficient may be preferred because of convenient capacitor values, such as are found in designs #15, 31 or 60.) In either of these examples, a relatively low attenuation should be expected one octave above the F_{A_P} frequency. For example, design #60 has an F_{A_P} frequency of 1.836 MHz, and less than 1 dB of attenuation one octave above the F_{A_P} frequency. However, if a maximum VSWR of 1.10 is acceptable, then second harmonic attenuation can be improved by 20 dB if the desired cutoff frequency is changed from the F_{A_P} frequency to the .01 dB frequency. *A generally satisfactory procedure in selecting a lowpass filter to operate with an amplifier is to select a filter having both the highest acceptable reflection coefficient and an A_P or 0.01-dB cutoff frequency just above the amplifier operating frequency.*

Note that all capacitor values in Table 1 are standard except for designs #21 and 26.

Verification of Data In Table 1

Whenever computer-derived design data is published, such as shown in Table 1, an explanation of how the data can be verified should be provided. Since the computer calculates the data for each design in the same manner, the verification of only one or two designs should suffice to establish the validity of the entire table. The correctness of the attenuation vs. frequency data can be verified by plotting the 10, 30 and 50 dB frequencies of Design #81 (this is a Butterworth design) on semi-log graph paper. This plot has a 42 dB/octave slope, and the validity of the attenuation vs. frequency data is therefore confirmed. (The attenuation of any Butterworth filter has a slope of 6 dB/octave per each element.) Design #81 is theoretically not a Butterworth design because the C4/C2 ratio is not exactly equal to the ratio of a Butterworth design; however, because C2 and C6 are within 0.3 percent of the exact value, the approximate design is sufficiently close to a Butterworth to be included in the filter table. The other designs having higher values of reflection coefficient will have correspondingly higher slopes of attenuation.

Confirming the C and L component values is more involved, and two "dummy" designs (#21 and 26) are included in Table 1 to facilitate the calculations. This is the reason why C2 and C6 for these two designs are not standard values. Reflection coefficients of 5 percent and 1 percent were chosen[11,12] for these two designs. (Using the normalized component values from either of these two references for the 1 and 5 percent Chebyshev filters and the tabulated A_P frequency, it is possible for the reader to independently calculate the C and L values of the computer-calculated designs.)

Two verifications are provided — one for designs with reflection coefficients less than 1 percent, and the other for reflection coefficients greater than 1 percent. The computer program uses different equations to cover these two ranges. A demonstration of this verification procedure

(Continued on page 41.)

(Continued from page 36.)
follows for design #21 with an R.C. of 5 percent.

From the C0705 tables of Saal or Zverev, the normalized values of C2, C4, L1 and L3 are, respectively, 1.397F, 1.634F, 0.8068H and 1.757H. Using the A_P cutoff frequency (1.334 MHz) of design #21, the C_s and L_s scaling factors are calculated as follows:

$$C_s = 1/R\omega = 1/(50 \cdot 2\pi \cdot 1.334 \cdot 10^6) = 2386.1 \cdot 10^{-12}$$

$$L_s = R/\omega = 50/(2\pi \cdot 1.334 \cdot 10^6) = 5.965 \cdot 10^{-6}$$

The normalized C and L values are scaled to the desired cutoff frequency and impedance level by multiplying them by the appropriate scaling factors. The scaling equations and scaled component values are:

$$C2{,}6 = 1.397\,(2386.1)\,10^{-12}F = 3333.4 \text{ pF or } 3333 \text{ pF}$$

$$C4 = 1.634(2386.1)10^{-12}F = 3898.9 \text{ pF or } 3900 \text{ pF}$$

$$L1{,}7 = 0.8068(5.965)10^{-6}H = 4.81\ \mu H$$

$$L3{,}5 = 1.757(5.965)10^{-6}H = 10.48\ \mu H$$

(Note how these independently calculated values agree with the tabulated values.)

This concludes the data verification calculations.

How to Use Table 1 To Design for Any Termination Impedance

The tabulated filter data is based on equal terminations of 50 ohms since this impedance is most frequently encountered in RF design work.

However, the filter data can easily be scaled to other impedance levels while still maintaining standard capacitor values. The following example demonstrates the scaling procedure.

Scaling Example

Assume that an inductive input/output lowpass filter is desired to filter the 8 MHz output of a 75 ohm amplifier. Let $Z_x = 75$ ohms and the desired A_P cutoff frequency, $F_A^x = 8.0$ MHz.

1. Calculate the impedance scaling factor, $R = Z_x/50$.
2. Calculate the A_P cutoff frequency, F_A^{50}, of a 50 ohm filter from $F_A^{50} = RF_A^x$. If F_A^{50} exceeds ten, then scale the table into the next frequency decade.
3. Select the 50 ohm design from the table that most closely matches the calculated F_A^{50} value and has a high enough reflection coefficient to assure adequate selectivity at the 2nd harmonic of the A_P cutoff frequency. The tabulated values of C2,4 and 6 will be used in the new design, and the inductor values will be scaled.
4. Calculate the new inductor values by multiplying the tabulated values by the square of the impedance scaling factor, R.
5. Find the corresponding A_P, 0.01, 3, 10, 30 and 50 dB attenuation frequencies of the new filter by dividing the tabulated frequencies by the impedance scaling factor.

For $F_A^x = 8.0$ MHz and $Z_x = 75$ ohms, $R = 75/50 = 1.5$, $R^2 = 2.25$, and $F_A^{50} = 1.5$ $8.0) = 12.0$ MHz. Because F_A^{50} is greater than 10 MHz, the 50 ohm table is scaled to the 10-to-100 MHz decade by multiplying all frequencies by ten and by dividing all component values by ten.

Design #22 is selected because its reflection coefficient is high enough for good selectivity (4.16 percent) without being excessive, and the A_P cutoff frequency (13.28 MHz) is just above the F_A^{50} value of 12.0 MHz.

From design #22, C2,6 and C4 are 330 and 390 pF, respectively.

FLTR NO.	FREQ(MHZ) AT AP,.01,3,10,30 & 50-DB AP-DB	.01-DB	3-DB	10-DB	30-DB	50-DB	R.C. %	C2,6 (PF)	C4 (PF)	L1,7 (UH)	L3,5 (UH)	FLTR NO.
1	0.818	0.83	0.98	1.08	1.39	1.84	2.44E+00	5100	6200	6.45	15.72	1
2	0.721	0.79	1.01	1.13	1.50	2.02	4.62E-01	4700	6200	5.02	14.84	2
3	0.552	0.73	1.02	1.17	1.58	2.17	3.83E-02	4300	6200	4.14	14.40	3
4	0.165	0.67	1.03	1.20	1.67	2.32	4.60E-06	3900	6200	3.50	14.00	4
5	0.921	0.93	1.08	1.18	1.50	1.98	3.57E+00	4700	5600	6.33	14.58	5
6	0.819	0.88	1.11	1.25	1.64	2.21	6.14E-01	4300	5600	4.68	13.49	6
7	0.622	0.81	1.13	1.29	1.75	2.39	4.44E-02	3900	5600	3.77	13.02	7
8	0.337	0.76	1.13	1.32	1.83	2.53	3.77E-04	3600	5600	3.27	12.73	8
9	1.014	1.02	1.18	1.29	1.63	2.15	3.89E+00	4300	5100	5.89	13.37	9
10	0.892	0.97	1.22	1.37	1.81	2.43	5.61E-01	3900	5100	4.22	12.26	10
11	0.719	0.90	1.24	1.41	1.91	2.60	6.99E-02	3600	5100	3.52	11.91	11
12	0.407	0.84	1.24	1.45	2.00	2.77	7.56E-04	3300	5100	3.01	11.62	12
13	1.087	1.10	1.29	1.42	1.81	2.40	2.88E+00	3900	4700	5.06	12.04	13
14	0.971	1.05	1.33	1.49	1.96	2.64	5.82E-01	3600	4700	3.90	11.30	14
15	0.765	0.98	1.34	1.53	2.08	2.83	5.87E-02	3300	4700	3.21	10.96	15
16	0.353	0.90	1.35	1.57	2.19	3.03	1.46E-04	3000	4700	2.71	10.66	16
17	1.197	1.21	1.41	1.54	1.96	2.58	3.41E+00	3600	4300	4.81	11.15	17
18	1.065	1.15	1.45	1.62	2.14	2.88	6.07E-01	3300	4300	3.59	10.35	18
19	0.816	1.06	1.47	1.68	2.28	3.11	4.73E-02	3000	4300	2.90	10.01	19
20	0.213	0.97	1.48	1.73	2.42	3.35	2.08E-06	2700	4300	2.42	9.71	20
21	1.334	----	1.52	1.66	2.09	2.74	5.00E+00	3333	3900	4.81	10.48	21
22	1.328	1.33	1.54	1.68	2.12	2.80	4.16E+00	3300	3900	4.58	10.29	22
23	1.179	1.27	1.60	1.79	2.35	3.16	6.38E-01	3000	3900	3.28	9.40	23
24	0.870	1.16	1.62	1.85	2.52	3.45	3.58E-02	2700	3900	2.59	9.05	24
25	1.425	1.44	1.68	1.85	2.35	3.11	3.12E+00	3000	3600	3.95	9.27	25
26	1.326	1.40	1.72	1.92	2.51	3.35	9.99E-01	2829	3600	3.21	8.78	26
27	1.212	1.35	1.74	1.96	2.60	3.51	3.61E-01	2700	3600	2.84	8.58	27
28	0.765	1.22	1.76	2.03	2.79	3.84	6.57E-03	2400	3600	2.24	8.27	28
29	1.528	1.56	1.86	2.04	2.63	3.49	2.21E+00	2700	3300	3.36	8.32	29
30	1.223	1.43	1.91	2.16	2.89	3.93	1.67E-01	2400	3300	2.42	7.78	30
31	0.835	1.[illegible]	1.92	2.22	3.05	4.19	6.57E-03	2200	3300	2.05	7.58	31
32	1.634	1.69	2.06	2.28	2.96	3.95	1.43E+00	2400	3000	2.83	7.41	32
33	1.377	1.59	2.09	2.37	3.16	4.29	2.08E-01	2200	3000	2.24	7.09	33
34	0.918	1.46	2.11	2.44	3.35	4.61	6.57E-03	2000	3000	1.86	6.89	34
35	1.906	----	2.07	2.23	2.75	3.57	1.07E+01	2400	2700	4.31	8.19	35
36	1.859	1.90	2.27	2.51	3.22	4.28	2.04E+00	2200	2700	2.71	6.78	36
37	1.570	1.78	2.32	2.62	3.49	4.73	2.68E-01	2000	2700	2.07	6.40	37
38	1.020	1.62	2.35	2.71	3.72	5.12	6.57E-03	1800	2700	1.68	6.20	38
39	2.137	2.16	2.53	2.77	3.53	4.66	3.12E+00	2000	2400	2.63	6.18	39
40	1.818	2.02	2.61	2.94	3.90	5.26	3.61E-01	1800	2400	1.89	5.72	40
41	1.148	1.83	2.64	3.05	4.19	5.76	6.57E-03	1600	2400	1.49	5.51	41
42	0.224	1.72	2.65	3.10	4.34	6.03	4.82E-08	1500	2400	1.34	5.41	42
43	2.291	2.34	2.78	3.07	3.94	5.23	2.21E+00	1800	2200	2.24	5.54	43
44	1.835	2.15	2.86	3.24	4.34	5.89	1.67E-01	1600	2200	1.62	5.18	44
45	1.435	2.03	2.87	3.30	4.51	6.19	1.96E-02	1500	2200	1.42	5.09	45
46	2.452	2.54	3.09	3.42	4.44	5.92	1.43E+00	1600	2000	1.89	4.94	46
47	2.182	2.43	3.13	3.53	4.68	6.32	3.61E-01	1500	2000	1.58	4.77	47
48	1.097	2.14	3.17	3.68	5.10	7.04	1.15E-03	1300	2000	1.19	4.56	48

FLTR NO.	FREQ(MHZ) AT AP,.01,3,10,30 & 50-DB AP-DB	.01-DB	3-DB	10-DB	30-DB	50-DB
49	2.859	----	3.11	3.34	4.13	5.36
50	2.849	2.88	3.37	3.69	4.71	6.22
51	2.197	2.61	3.50	3.97	5.33	7.24
52	1.530	2.44	3.52	4.06	5.58	7.69
53	3.126	3.20	3.84	4.24	5.46	7.2[illegible]
54	2.728	3.03	3.92	4.41	5.84	7.9[illegible]
55	2.055	2.82	3.95	4.53	6.17	8.4[illegible]
56	0.336	2.58	3.98	4.65	6.51	9.04
57	3.475	----	3.90	4.22	5.29	6.91
58	3.269	3.39	4.12	4.56	5.92	7.89
59	2.753	3.18	4.19	4.74	6.33	8.58
60	1.836	2.92	4.22	4.87	6.70	9.22
61	3.985	4.00	4.61	5.03	6.37	8.39
62	3.538	3.81	4.80	5.37	7.06	9.49
63	2.734	3.52	4.85	5.55	7.52	10.27
64	0.902	3.21	4.89	5.71	7.97	11.06
65	4.274	4.32	5.05	5.54	7.06	9.33
66	3.726	4.08	5.21	5.85	7.73	10.42
67	2.660	3.74	5.27	6.05	8.26	11.33
68	0.448	3.45	5.30	6.20	8.68	12.05
69	4.633	4.70	5.53	6.08	7.78	10.29
70	3.912	4.39	5.70	6.42	8.54	11.54
71	2.869	4.07	5.75	6.60	9.02	12.38
72	5.053	5.15	6.12	6.73	8.64	11.47
73	4.364	4.85	6.26	7.05	9.35	12.63
74	3.112	4.46	6.32	7.27	9.94	13.64
75	5.581	5.67	6.70	7.37	9.44	12.51
76	4.756	5.32	6.89	7.76	10.30	13.93
77	3.455	4.91	6.95	7.98	10.91	14.97
78	6.229	6.31	7.41	8.14	10.40	13.76
79	5.418	5.95	7.63	8.57	11.34	15.29
80	3.881	5.47	7.71	8.86	12.09	16.59
81	N/A	5.03	7.76	9.08	12.71	17.67
82	6.791	6.89	8.12	8.92	11.41	15.11
83	5.760	6.45	8.36	9.42	12.51	16.91
84	4.149	5.95	8.43	9.69	13.25	18.19
85	1.037	5.53	8.48	9.92	13.86	19.25
86	7.463	7.59	8.97	9.87	12.65	16.76
87	6.418	7.14	9.21	10.37	13.75	18.58
88	4.956	6.66	9.29	10.64	14.48	19.83
89	1.899	6.15	9.35	10.91	15.22	21.10

Inductors L1,7 = R^2 (.458) = 1.031 μH, and L3,5 = R^2 (1.029) = 2.315 μH. The new frequencies are scaled by dividing the tabulated frequencies of design #22 by the impedance ratio, R = 1.5. Thus, the A_P, .01, 3, 10, 30 and 50 dB frequencies are 8.85, 8.87, 10.3, 11.2, 14.1 and 18.7 MHz, respectively. The amplifier operating frequency is 8.0 MHz, and the selected 8.85 MHz A_P cutoff frequency of scaled design #22 is about 10 percent above the desired A_P cutoff frequency of 8.0 MHz. The amplifier second harmonic (16 MHz) falls on the filter attenuation response curve about half-way between the scaled 30 and 50 dB frequencies of 14.1 and 18.7 MHz, such that the second harmonic attenuation is around 40 dB.

The exact value of VSWR seen by the amplifier due to the filter depends on the position of the amplifier frequency in the filter passband relative to the peaks (maximum attenuation) and troughs (minimum attenuation) of the passband ripple. That is, if the amplifier frequency coincides with one of the peaks of the filter passband ripple, the VSWR will be maximum, and equal to 1.087. On the other hand, if the amplifier frequency coincides with a trough of the ripple, the VSWR will be unity. Generally, it is sufficient to know that the maximum VSWR will not be exceeded anywhere in the filter passband. (See Equation (3a) in Appendix A for the calculation of VSWR.)

Part 2 continues with the construction and testing of filter design #31. ☐

References

[1]W. Mueller, "Linear Amplifier Design: Some General Considerations," *r.f. design*, March 1980.

[2]M.F. "Doug" DeMaw, *Practical RF Communications Data for Engineers and Technicians*, Howard W. Sams & Co., Inc., Indianapolis, Ind., 1978.

[3]E.E. Wetherhold, "7-Element 50-ohm Chebyshev Filters Using Standard-Value Capacitors," *r.f. design*, February 1980.

[4]E.E. Wetherhold, "Lowpass Chebyshev filters use standard-value capacitors," *Electronics*, Engineer's Notebook, pp. 160-161, June 19, 1980.

[5]E.E. Wetherhold, "Design 7-element lowpass filters using standard-value capacitors," *EDN*, Vol. 28, No. 27, Jan. 7, 1981.

R.C. %	C2,6 (PF)	C4 (PF)	L1,7 (UH)	L3,5 (UH)	FLTR NO.
1.07E+01	1600	1800	2.88	5.46	49
3.12E+00	1500	1800	1.97	4.64	50
1.38E-01	1300	1800	1.30	4.23	51
6.57E-03	1200	1800	1.12	4.13	52
1.93E+00	1300	1600	1.59	4.00	53
3.61E-01	1200	1600	1.26	3.81	54
2.75E-02	1100	1600	1.05	3.71	55
4.82E-08	1000	1600	.893	3.61	56
6.53E+00	1300	1500	2.00	4.17	57
1.43E+00	1200	1500	1.41	3.70	58
2.08E-01	1100	1500	1.12	3.55	59
6.57E-03	1000	1500	.931	3.45	60
4.16E+00	1100	1300	1.53	3.43	61
6.38E-01	1000	1300	1.09	3.13	62
5.30E-02	910	1300	.883	3.03	63
1.20E-05	820	1300	.736	2.94	64
3.12E+00	1000	1200	1.32	3.09	65
4.66E-01	910	1200	.972	2.87	66
2.15E-02	820	1200	.779	2.78	67
4.82E-08	750	1200	.670	2.70	68
2.72E+00	910	1100	1.17	2.81	69
3.13E-01	820	1100	.854	2.62	70
1.96E-02	750	1100	.711	2.54	71
2.30E+00	820	1000	1.03	2.53	72
3.61E-01	750	1000	.788	2.38	73
1.74E-02	680	1000	.643	2.31	74
2.54E+00	750	910	.954	2.31	75
3.31E-01	680	910	.711	2.17	76
1.90E-02	620	910	.587	2.10	77
2.85E+00	680	820	.881	2.10	78
4.36E-01	620	820	.659	1.96	79
2.09E-02	560	820	.532	1.90	80
ZERO	510	820	.456	1.85	81
2.68E+00	620	750	.796	1.91	82
3.25E-01	560	750	.585	1.78	83
1.74E-02	510	750	.482	1.73	84
6.54E-07	470	750	.421	1.69	85
2.50E+00	560	680	.711	1.73	86
3.61E-01	510	680	.536	1.62	87
3.37E-02	470	680	.451	1.58	88
2.39E-05	430	680	.387	1.54	89

Table 1.

[6]E.E. Wetherhold, "7-Element Chebyshev Filters for TEMPEST Testing," *Interference Technology Engineer's Master,* 1981, R & B Enterprises, POB 328, Plymouth Meeting, PA 19462.

[7]R.F. Frost, "Large-scale s parameters help analyze stability," *Electronic Design*, Vol. 28, No. 11, May 24, 1980.

[8]H.O. Granberg, "Good RF Construction Practices and Techniques," *r.f. design*, September/October 1980.

[9]M.F. DeMaw, "Magnetic Cores in RF Circuits, *r.f. design*, April 1980.

[10]E.E. Wetherhold, Letter to Editor, p. 19, *r.f. design*, June 1980.

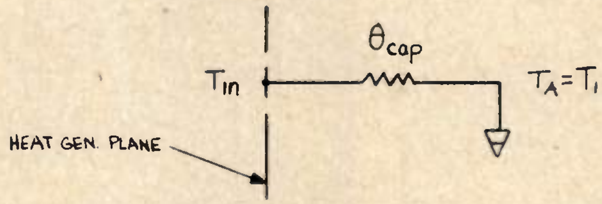
[11]R. Saal, *The Design of Filters Using the Catalog of Normalized Lowpass Filters*, Telefunken GmbH, Backnang, Western Germany, 1966.

[12]A. Zverev, *Handbook of Filter Synthesis*, John Wiley & Sons, N.Y., 1967.

Part 2

Ceramic and Porcelain

Power and Current Rating Determinations.

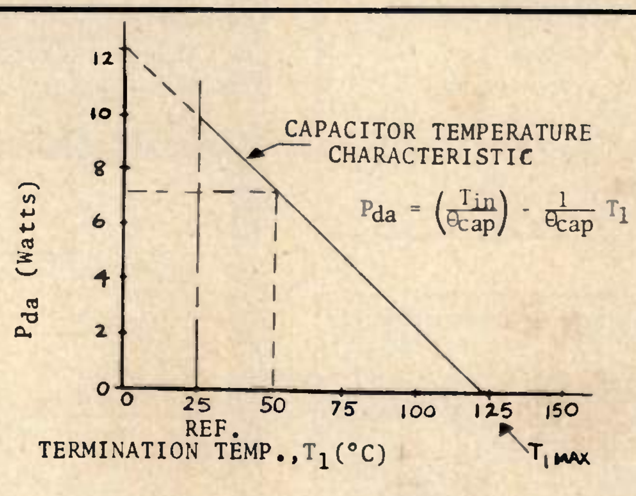


where, P_d = Power dissipated in capacitor (watts)

T_A = Ambient Temperature (°C)

∇ = INFINITE HEAT SINK (a mass at a temperature that cannot be changed, regardless of the amount of heat pumped into it)

Figure 6.



where, at $T_A = T_1 = 25°C$; $P_{dmax} = \left(\frac{1}{\theta_{cap}}\right) T_{in}$

and at $P_d = 0$; $T_{1max} = 125°C$

slope $= \frac{\Delta P_d}{\Delta T_1} = -\frac{1}{\theta_{cap}}$

where, P_{da} = Power dissipation allowed for an internal temperature of T_{1max}

Figure 7.

Multilayer Capacitors

F.M. Schabauer & R. Blumkin
American Technical Ceramics
Huntington, N.Y.

Power Rating

As previously stated, the allowable power dissipation can be determined by the knowledge of the thermal resistance Θ_{cap}, the equivalent series resistance ESR of the capacitor, the maximum allowable internal temperature, and the maximum temperature that the solder or epoxy on the termination can tolerate without destruction.

The simplified equivalent thermal circuit, *when the capacitor terminations are connected to an infinite heat sink*, is shown in Figure 6.

The thermal equation for the circuit in Figure 6 is given by:

$$\Theta_{cap}(P_d) = (T_{in} - T_1) \qquad (6)$$

and is plotted in Figure 7.

If the vertical scale name is changed from power dissipation P_d to power dissipation allowed P_{da}, this curve is really a maximum power rating curve for the capacitor, where the allowed internal temperature T_{in} is equal to $T_{1max} = 125°C$.

For example, if the heat sink and, therefore, the terminations are set to 50°C, then the internal temperature will be 125°C for a P_{da} of 7.2 watts. This is the particular condition shown by the dotted lines in Figure 7. Similarly, one can determine the power rating of the capacitor for any given heat sink temperature or termination temperature. It should be stressed that this equivalent circuit and curve is for the specific condition where the terminations are connected to an infinite heat sink. (Values of P_{da} for actual capacitors are plotted in power temperature rating curves (1 & 2) available from ATC).

The allowable power dissipation for the capacitors in Table 1 (see Part 1, May/June issue) with an infinite heat sink at 25°C connected to the termination is given in Table 2.

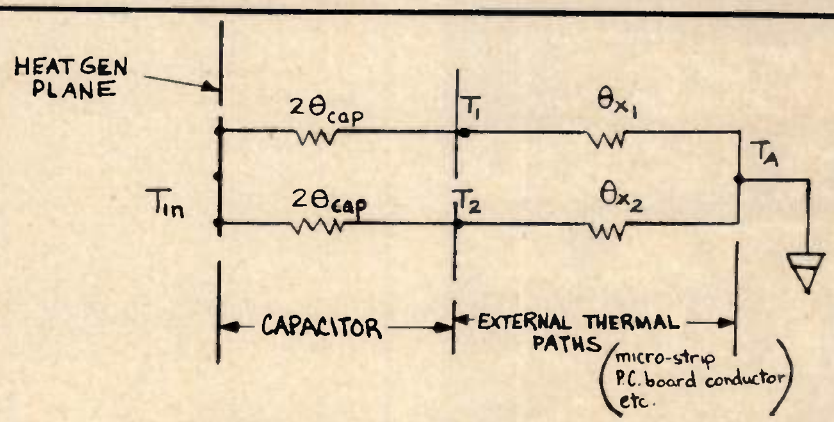
The thermal situation taking into account external thermal resistance is shown in Figure 8.

Assuming that $T_1 = T_2$, the thermal circuit becomes Figure 9.

The thermal circuit is described by:

$$P_d\Theta_{cap} + P_d\Theta_x = T_{in} - T_A \qquad (8)$$

Since T_{in} is a maximum of 125°C and both Θ_{cap} and Θ_x are known, the circuit designer can solve for the maximum allowable P_{da} either alge-

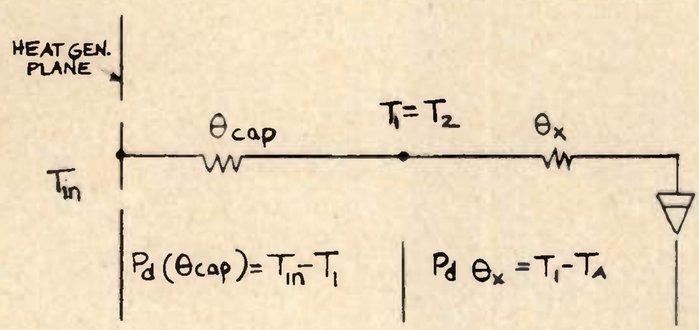


Θ_{x1} = External thermal resistance, from termination 1 to heat sink (°C/W)

Θ_{x2} = External thermal resistance, from termination 2 to heat sink (°C/W)

Figure 8.

Assuming that $T_1 = T_2$, the thermal circuit becomes FIGURE 10.



where,

$$\Theta_x = \frac{\Theta_{x1}\ \Theta_{x2}}{\Theta_{x1} + \Theta_{x2}} \qquad (7)$$

Figure 9.

SERIES	100A		100B		
Cap Value (pF)	1	100	1	100	1000
θ_{cap} (°C/W)	13.7	11.4	7.9	7.2	5.9
Max Power Diss. (watts) at 25°C	7.3	8.8	12.6	13.9	16.9

Table 2.

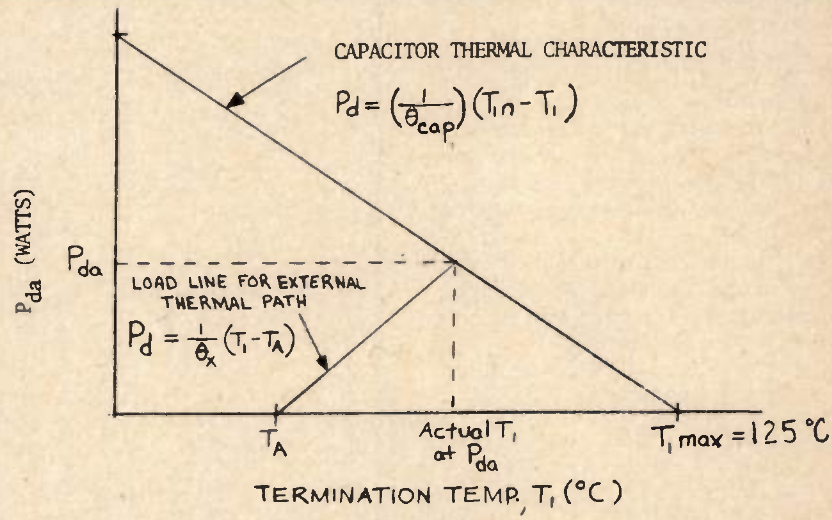


Figure 10.

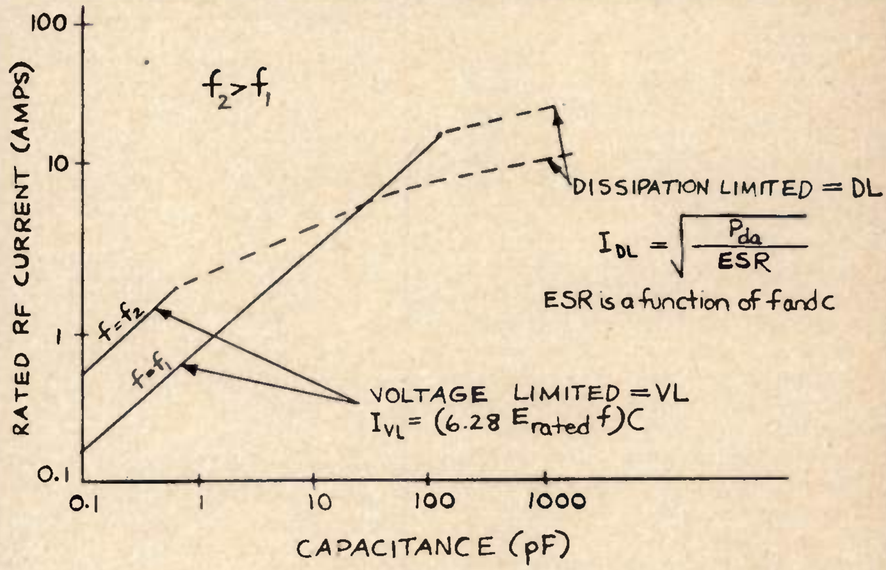


Figure 11.

braically or graphically. To solve graphically, use Figure 7 and then superimpose:

$$P_d = \frac{1}{\Theta_x}(T_1 - T_A) \qquad (9)$$

This is shown in Figure 10.

Starting at $T_1 = T_A$, plot a line whose slope is $1/\Theta_x$; the intersection of the two lines gives the allowed power dissipation and the actual termination temperature for this thermal circuit. The internal temperature (T_{in}) is 125°C.

Current Ratings

Knowing the allowed power dissipation (P_{da}) in the capacitor, for a given external thermal path, and knowing ESR at the frequency of interest, the dissipation limited current can then be calculated:

$$I_{DL} = \frac{P_{da}}{ESR} \qquad (10)$$

(ESR values can be obtained from ATC drawing 102-584*).

I_{DL} is valid as long as the maximum rated voltage of the capacitor is not exceeded. The voltage limited current due to the maximum rated voltage is calculated from Equation 11.

$$I_{VL} = \frac{E_{rated}}{X_C} = (E_{rated}\, 2\pi f)\, C \qquad (11)$$

A plot of maximum allowable current vs. capacitance from Equations 10 and 11 results in a family of curves as shown in Figure 11. From Figure 11, it is clear that when I_{VL} is smaller than I_{DL}, I_{VL} becomes the rated current. (Current-rating curves for the ATC 100A & 100B capacitors, curves 3, 4 & 5 are available from ATC).

Conclusion

Information and methods for arriving at RF current ratings of multilayer monolithic ceramic capacitors have been presented. It has been shown that the general shape of the current rating curves can be established. Expressions for the effect of various capacitor parameters (such as Equivalent Series Resistance, RF Voltage Rating and Thermal Resistance), on the current ratings have been developed. This data was developed theoretically and then verified experimentally. Examples of how to use this information to arrive at current ratings for specific thermal conditions have been included. □

Solid State Class A Amplifier

The ENI model A1000 Solid State Class A hybrid coupled power amplifier delivers over 1000 watts from 0.3 to 35 MHz. The unit has a gain of 60 dB with a variation of only ±1.5 dB over its entire bandwidth. It is designed to handle any amplitude modulated, frequency modulated or single-sideband signals. The amplifier is unconditionally stable and fully protected against both overdrive and overload, enabling it to drive non-linear loads without damage.

The A1000 has an input/output impedance of 50 ohms. The low distortion characteristics of the amplifier make it suitable for communication systems. At 800 watts all harmonics are more than 25 dB below the fundamental. It has a low noise figure of less than 12 dB. An RF voltmeter is provided on the front panel which also is calibrated to read watts into 50 ohms.

The 370-lb. unit, stands just over 4 feet tall measuring 52 1/2 by 16 3/4 by 21 1/2 inches. The A1000 is powered by its own highly regulated three-phase AC power supply from 208 to 230 VAC 50-60 Hz drawing about 22 amperes per phase. The power supply includes short circuit foldback features and extensive RFI/EMI filtering. The amplifier contains an integral forced air cooling system.

Contact: Electronic Navigation Industries Incorporated, 3000 Winton Road South, Rochester, NY 14623. INFO/CARD #140.

Siliconix Offers FET Design Kit

Siliconix is offering a new FET Designer's Kit which consists of:

(1) A copy of *Designing with Field-Effect Transistors* which covers theory and practical applications of FETs as well as a brief introduction to power MOSFETs. Published in February 1981 by McGraw-Hill and written by Siliconix, the book has a retail value of $24.50.

(2) A copy of the Siliconix FET Design Catalog, published July 1981, which includes data sheets and application notes on their entire FET product line.

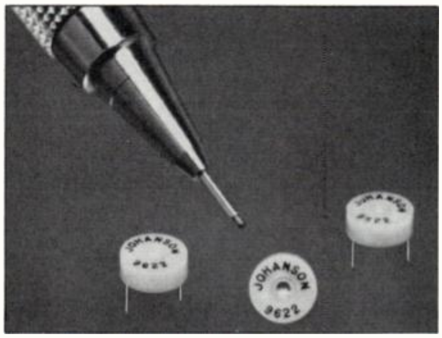
(3) Ten sample FETs: N-channel, P-channel, amps, switches, amp/

switch combinations, low leakage FETs, and high frequency FETs and design tips on how to use them.

The kit, a retail value of $42.50, is being offered to qualified engineers for $28.00 through the end of 1981. To order, send check and request to: Siliconix Incorporated, P.O. Box 4777, Santa Clara, CA 95054. INFO/CARD #139.

Ceramic Variable Capacitors

Seal-Trim high performance ceramic trimmer capacitors are now available in a smaller (.170" diameter) series.

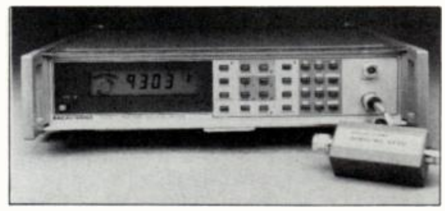


The moisture-proof housing eliminates intrusion of dirt, dust, solder flux, and cleaning agents. Available in capacity ranges from .5-2.5 pF through 6.0-25 pF with various Temperature Coefficents and Q values. Terminals for printed circuit mounting are spaced .140". Johanson square drive tuning insures captive, non-slip, tamper-proof, backlash free adjustment. Typical applications include crystal trimming, microprocessor clock oscillators, RF amplifiers, filter networks, and interstage coupling.

Contact: Johanson Manufacturing Corporation, 400 Rockaway Valley Road, Boonton, NJ 07005. INFO/CARD #138.

Microprocessor Based RF Level Meter

Model 9303 RF Level Meter introduced by Racal-Dana Instruments combines advanced RF technology with microprocessor LSI to achieve



major advances in the state-of-the-art of RF measurements. The unit functions as both a true RMS RF millivoltmeter and an RF power meter. Measurement features include an internal calibration source, automatic self-calibration, variable signal smoothing, two input channels, program memory, and an optional GPIB (IEEE-STD-488) interface with full programming capabilities.

Contact: Racal-Dana Instruments Inc., 18912 Von Karman Avenue, P.O. Box C-19541, Irvine, CA 92713. Circle INFO/CARD #137.

RF Wattmeters to IEEE-488 GPIB Interface

A new IEEE-488 Bus Interface Unit provides connection of the Bird directional RF Power Analyst™ Wattmeters to a General Purpose Instrumentation Bus (GPIB) for automatic measurement and recording of data. Digital Thruline® Wattmeters equipped with a bus-compatible output can now become a talker and a listener via GPIB central control.

Typical applications are unattended periodic measurement of one to nine RF signal parameters; hard-copy print-out of data for logging; scope, graphic terminal or curve plotter feed for dependent variables display such as attenuation or SWR vs. frequency, power vs. time, etc.; alarm or warning

functions of performance levels; remote RF Power Analyst™ interrogation and response readout; automatic testing of equipment or components; minimum/maximum level memory during a specified time period, etc.

The digital Thruline® RF Power Analyst™ Wattmeters operate from 1/2 to 2300 MHz at 100 milliwatts to 250 kilowatts depending on model and plug-in elements. The new Bus Interface Unit model 4380-488 opens up automatic testing to any industry utilizing radio frequency power within these wide limits of frequency and power.

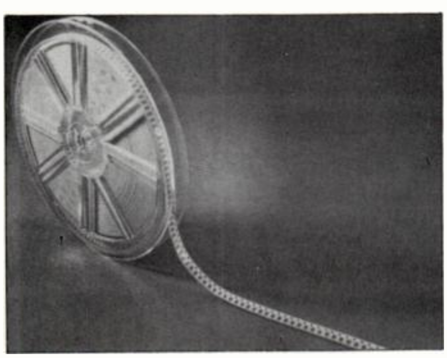
Contact: Bird Electronic Corporation, 30303 Aurora Road, Cleveland (Solon), OH 44139. INFO/CARD #136.

Chip Capacitors On Tape and Reel

A complete line of monolithic chip capacitors in 25 volt and 50 volt ratings, supplied in tape and reel packaging for high speed auto insertion, is now available from Murata Corporation of America. This new packaging configuration is compatible with a variety of automatic insertion equipment and allows very high speed component placement on PC boards.

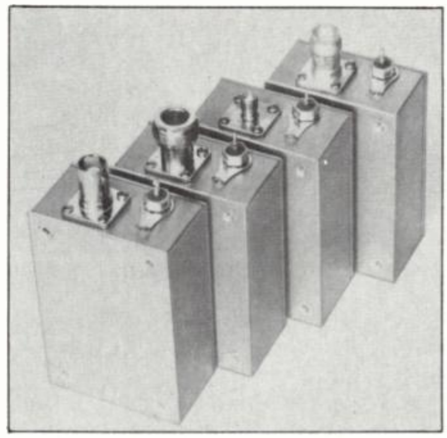
Chip capacitors offered in this line include units with NPO, X7R, Y5Z temperature characteristic covering the capacitance range from .5 pF to 330,000 pF. Chip dimensions are .079 x .049 x .049 and .126 x 0.63 x .049.

Contact: Murata Corporation of America, 1148 Franklin Road, S.E., Marietta, GA 30067. INFO/CARD #135.



RF Enclosures And Interchangeable Connectors

A new line of off-the-shelf RF enclosures with interchangeable SMA, BNC, TNC and N connectors is available from Compac, Deer Park, N.Y., manufacturers of RF shielded en-



closures and accessories.

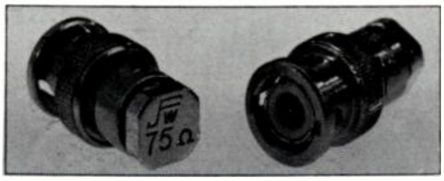
Compac's new product line enables manufacturers to readily switch connectors to meet changing customer requirements. Using a basic RF enclosure for his product the manufacturer can attach the specific Compac SMA, BNC, TNC or N connector, as needed. Enclosure users — laboratories, designers or OEM manufacturers — can purchase standard enclosures and additional quantities of the connectors to anticipate their requirements.

The interchangeable flange mounted "female" connectors have four uniform screw holes matching the screw holes in the enclosure. They are available for Compac's standard case series and custom series.

Contact: Compac, 279 Skidmore Road, Deer Park, NY 11729. Circle INFO/CARD #133.

Terminations And Feedthru Terminations

Frequency range DC-2000 MHz, 50 and 75 ohm available, VSWR 1.1:1 maximum. One-year warranty, delivery stock — 2 weeks. Price $10.00. Terminations (1-9 pc. qty.) $11.00 Feedthru (1-9 pc. qty.).



Contact: JFW Industries, Inc., P.O. Box 226, Beech Grove, IN 46107. INFO/CARD #134.

High Resolution Frequency Source

A 0.5 Hz to 50 MHz frequency source for pulse echo overlap ultrasonic velocity measurement systems and similar applications is available from Matec, Inc. of Warwick, Rhode Island.

The Matec Model 110 High Resolution Frequency Source produces square waves from below 0.5 Hz to above 50 MHz, and features a high stability oscillator tunable over a 12 to 50 MHz range. Output frequency, characterized by high spectral purity and low residual FM, adjusts to better than 6 places, even at low frequencies.

Suitable for phase-locked or frequency modulated operation, the Matec Model 110 High Resolution Frequency Source permits external frequencies to be divided by up to 4 x 10^7. Providing a 50 ohm output impedance, it is also ideal for use as a cw source for tuned RF gated amplifiers.

Contact: Matec, Inc., 60 Montebello Road, Warwick, RI 02886. Circle INFO/CARD #132.

New Wide Band Amplifier

Instruments for Industry, Inc. recently introduced its new self-contained, ultra-broadband laboratory amplifier, the IFI model 5300. It provides instantaneous bandwidth of 10 kHz to 320 MHz minimum; gain control of 40 dB (minimum); and unconditional stability.

Features which are standard on this new IFI solid-state amplifier include its electronic gain control with front panel adjustment and remote detector input for controlling gain by sampling power directly at load. For constant output, an automatic leveling system allows input from a remote sensing device, thereby enabling the amplifier to maintain a constant field level as the output is varied. Both the amplifier and its power supply have been designed to protect against overload with protection circuitry and automatic shut down. Using an efficient switching power supply, the amplifier operates as a standard pre-amplifier in kW amplifiers, or it can be used separately as a standard laboratory amplifier.

The IFI amplifier is rated for 15 watts continuous power output; and is applicable primarily for RFI/EMI susceptibility testing, as well as antenna testing; driving ultrasonic transducers; testing broadband communications; biological research testing;

Errata

In the "Need a Helical Filter," article in the May/June issue of *r.f. design* on page 46, the expression for Θ should read:

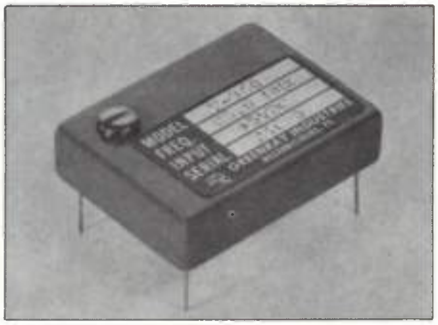
$$\Theta = \text{Arcsin}\sqrt{\frac{R_b R_{tap}}{2Z_0^2}}$$

and general laboratory applications.

Contact: IFI, 151 Toledo Street, Farmingdale, NY 11735. INFO/CARD #130.

Temperature Compensated Crystal Oscillator

Greenray Industries now offers a Temperature Compensated Crystal Oscillator holding a stability of ± 1 ppm over a 0 degrees to + 70 degrees C range utilizing a single + 5V DC input. This oscillator is available at any specific frequency in the range of 1.25 to 25 MHz. The output is TTL compatible to drive up to 5 standard TTL loads.



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